Recursive Least Squares Channel Estimation for Rapidly Time-Varying Scenarios in IEEE 802.11p

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Abstract*—Rapid time-varying channel estimation is one of the biggest challenges in IEEE 802.11p standard. It is a wireless vehicular communication standard which is used for outdoor applications. This paper proposes a novel decision-directed recursive least squares (RLS) time-domain channel estimation method that utilizes the guard interval of every orthogonal frequency division multiplexing (OFDM) symbol. Simulation results show considerably improved bit error rate (BER) performance with the proposed method that enables robust channel equalization in rapidly time-varying channel with high Doppler spread.

Keywords- IEEE 802.11p, Time-Varying Channel Estimation, OFDM, Vehicular-to-Vehicular Communication.

I. Introduction

IEEE 802.11p is a standard meant for dedicated short range communication (DSRC) in vehicular to vehicular (V2V) and vehicular to infrastructure (V2I) scenarios [1]. It is also known as wireless access in vehicular communication (WAVE) which supports many intelligent transport system (ITS) applications such as cooperative safety, smooth traffic flow, accident control, intersection collision avoidance, emergency warning, etc. [2]. Table 1 shows the physical parameter of IEEE 802.11p. It is similar to the IEEE 802.11a standard [1], with modifications in the carrier frequency and bandwidth.

Table I: Physical parameters of IEEE 802.11p

Parameter	Value
Bandwidth (MHz)	10
Modulation Scheme	BPSK, QPSK, 16QAM, 64QAM
Carrier Frequency (GHz)	5.9
FFT size	64
Total subcarriers	52
Pilot subcarriers	4
Data subcarriers	48
Subcarrier frequency spacing (MHz)	0.15625
GI duration (µs)	1.6 (16 samples)
FFT period (µs)	6.4 (64 samples)
OFDM Symbol duration (µs)	8 (80 samples)

Fig. 1 shows the OFDM frame structure. An OFDM frame starts with a short preamble (of 16μs) that is used for start of frame detection and, coarse time and frequency synchronisation [3]. It is followed by long preamble (of 16μs) that is used for fine time-synchronisation and channel estimation. Signal field of the frame contains information about modulation and coding. This is followed by variable number of OFDM symbols. In general, number of OFDM symbols can be chosen according to the coherence time of the channel.

Guard interval (GI) duration in the IEEE 802.11p standard is doubled (in duration) in comparison to the IEEE 802.11a standard, so as to support severe multipath delay spread of the vehicular channel. However, for channel tracking there is no amendment in this standard. This is to note that conventional channel estimation using long preamble is not sufficient for the rapid time-varying channel [4, 5]. Also, pilot density in the standard is not sufficient to track the fast time varying channel. Although in [6], pilot density has been increased with the insertion of pseudo pilots, it degrades the performance in terms of reduced data payload capacity.

To this end, methods have been proposed for channel estimation. In [7], a known Midamble is inserted between OFDM symbols for channel estimation. However, it deteriorates the performance by reducing the data payload capacity. In [8], guard interval (GI) of every OFDM symbol is utilized for channel tracking by replacing the cyclic prefix (CP) with the pseudo random (PR) sequence. Thus, a new time-domain least squares (TDLSE) channel estimation method is presented in [8]. Since CP helps to combat inter-carrier interference (ICI) in the received OFDM symbol, removal of CP in GI creates ICI. Thus, in [9], TDLSE method is modified using the ”overlap and add” method (OLA) with smoothing over 7 to 10 OFDM symbols. However, this method will introduce delays in real-time operations that may be unwarranted.

Thus, we notice that although it is encouraging to use TDLSE scheme for channel estimation in every OFDM symbol, it suffers with some limitations as: a) it does not mitigate ICI due to the replacement of CP; b) the smoothing method appears prohibitive in real-time applications because of delay; and c) it introduces additional computational time for every OFDM symbol. In order to overcome these limitations, we propose a novel time-domain recursive least squares (RLS) channel estimation method with decision directed approach in the GI of every OFDM symbol. The RLS algorithm requires computations of the order of $O(N^2)$ compared to LS that requires computations of the order of $O(N^3)$ [10]. Thus, it brings down the computational time complexity.

Moreover, we add decision-directed approach that utilizes the data sample estimates of previous OFDM symbol in time-domain convolution equation for current channel estimation. This mitigates the impact of ISI and ICI considerably. Thus, we propose a decision directed time-domain recursive least

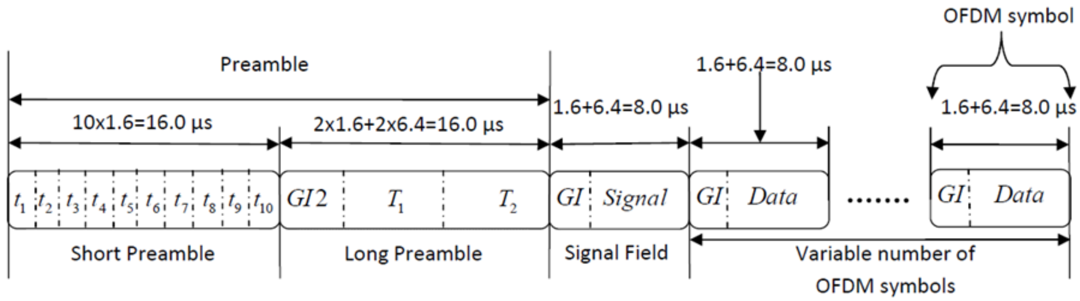

Fig. 1: OFDM Frame Structure

squares (RLS w/DD) channel estimation algorithm for rapidly fast fading channels.

Below are the contributions of this paper:

1) Proposed scheme guarantees better performance for time-varying channel estimation compared to the existing techniques due to the proposed decision-directed approach.
2) Proposed adaptive RLS algorithm is computationally less time-intensive compared to the existing LS based algorithms.
3) Since the proposed channel estimation in GI of every OFDM symbol is successful, we do not require channel estimation via long preamble and hence, effective data rate can be increased via the removal of long preamble. This may be useful in addressing higher wireless data rate challenge of 5G communication.

This paper is organised as follows. Section 2 describes the OFDM transceiver structure. Section 3 describes the proposed RLS w/DD algorithm. Simulation results are presented in Section 4. Finally, conclusions are presented in Section 5.

Notations:We use lower case bold letters for vectors, uppercase bold letters for matrices and lowercase italicized letter for scalar variables. $[.]^H$ *denotes the Hermitian conjugation of a vector or a matrix.* $[.]^T$*denotes the non-conjugate transpose of a vector.*

II. OFDM Transceiver Structure

The transceiver structure discussed in this section is identical to that proposed in [8] and [9]. Fig. 2 shows the OFDM transceiver structure for q^{th} OFDM symbol. First G (=16) samples of every OFDM symbol correspond to the guard interval (GI), while the last N (=64) samples denoted as $\{\hat{s}_{q,0}, \hat{s}_{q,1}, \cdots, \hat{s}_{q,N-1}\}$ correspond to the complex data samples to be transmitted in the frequency domain. This data is passed through the inverse Discrete Fourier Transform (IDFT) block to obtain time domain complex data samples as

$$x_{q,n} = IDFT_N\left(\hat{s}_{q,k}\right) \quad \text{for} \quad n,k = 0,1,\cdots,N-1. \quad (1)$$

where n and k are indices in time and frequency domain, respectively.We insert a pseudo-random (PR) sequence $\{p_0, p_1, \cdots, p_7\}$ of length $L_p = 8$ and zero padding (ZP) of length 8 in the guard interval of every OFDM symbol. Thus, the resultant data which is to be transmitted, in an OFDM symbol is:

$$x'_{q,n} = \begin{cases} p_n & n = 0,1,\cdots,7 \\ 0 & n = 8,9,\cdots,2L_p-1 \\ x_{q,n-2L_p} & n = 2L_p,\cdots,N+2L_p-1 \end{cases} \quad (2)$$

Symbols $x'_{q,n}$ are serially fed to a D/A converter and the resultant signal $x(t)$ is transmitted. The transmitted signal is passed through a time-varying multipath fading channel and is corrupted by white Gaussian noise. The A/D converter at the receiver converts this noisy analog information back into serial digital information which is denoted by:

$$r'_{q,n} = \{c_0, c_1, \cdots, c_{L_p+L-2}, \mathbf{0}_{1\times L_p-L+1}, r_{q,0}, r_{q,1}, \cdots, r_{q,N-1}\} \quad (3)$$

where L corresponds to the L-tap multipath fading channel and hence, leads to a maximum channel delay spread of $(L-1)$ x duration of each tap. We utilize the first $L_p + L - 1$ time domain samples of the GI for channel estimation. Details on the proposed channel estimation method are presented in section 3.

First $L_p + L - 1$ time domain samples of $r'_{q,n}$ can be represented as

$$c(n) = \begin{cases} \sum_{i=0}^{n} h_i p_{n-i} + \sum_{i=n+1}^{L-1} h_i x_{q-1,n+N-i} + w_{q,n}; \\ \qquad \text{for } n = 0,1,\cdots,L-2 \\ \sum_{i=0}^{L-1} h_i p_{n-i} + w_{q,n}; \\ \qquad \text{for } n = L-1, L, \cdots, L_p+L-2 \end{cases} \quad (4)$$

where $\{h_0, h_1, \cdots, h_{L-1}\}$ are the channel tap coefficients, $x_{q-1,N-1}$ is the $(N-1)^{th}$ sample of the $(q-1)^{th}$ OFDM symbol, and $w_{q,n}$ is the complex white Gaussian noise.

The last N time domain samples of $r'_{q,n}$ are fed to the Discrete Fourier Transform (DFT) block. The resultant output in the frequency domain is denoted as $\hat{r}_{q,k}$ where $k =$

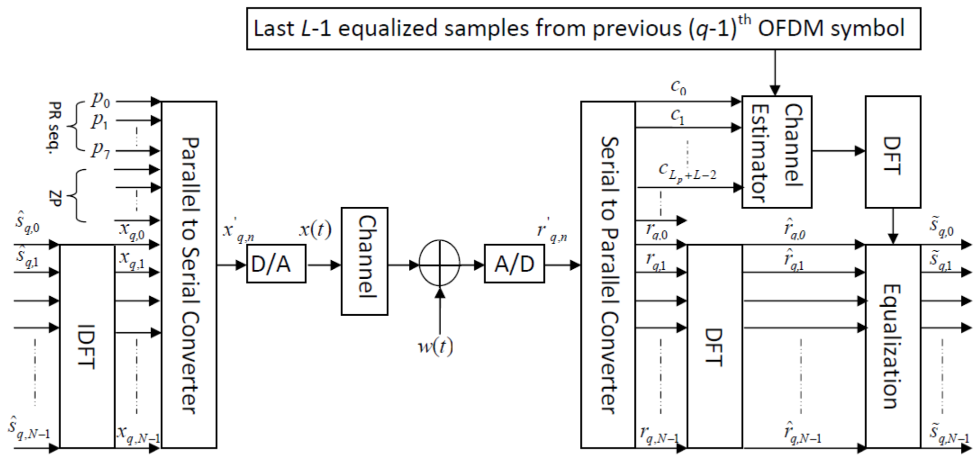

Fig. 2: OFDM Symbol Transceiver Structure

$0, 1, \cdots, N-1$. The estimated channel tap coefficients are denoted as $\tilde{h}_0, \tilde{h}_1, \cdots, \tilde{h}_{L-1}$ and the N-point DFT of the estimated channel tap coefficients are denoted as $\hat{h}_{q,k} = DFT_N\left(\tilde{h}_0, \tilde{h}_1, \cdots, \tilde{h}_{L-1}, 0, 0, \cdots, 0\right)$. Zero forcing channel equalization algorithms is used to estimate the transmitted frequency domain data samples:

$$\tilde{s}_{q,k} = \hat{r}_{q,k}/\hat{h}_{q,k} \quad \text{for} \quad k = 0, 1, \cdots, N-1. \tag{5}$$

III. Proposed RLS based Decision-Directed Channel Estimation Method

A robust channel estimation algorithm should be able to track rapid time varying channel. In this paper, we utilize the pseudo-random sequence inserted with ZP (PRwZP) in the GI for channel estimation in every OFDM symbol.

The first $L_p + L - 1$ received time domain samples of $r'_{q,n}$ after A/D conversion denoted in (4) can be written as

$$\mathbf{c} = \mathbf{M}\mathbf{h} + \mathbf{w} \tag{6}$$

where $\mathbf{c} = \left[c_0, c_1, \cdots, c_{L_p+L-2}\right]^T$,

$$\mathbf{M} = \begin{bmatrix} p_0 & x_{q-1,N-1} & \cdots & x_{q-1,N-L+2} & x_{q-1,N-L+1} \\ p_1 & p_0 & \ddots & \vdots & \vdots \\ p_2 & p_1 & \ddots & p_0 & x_{q-1,N-1} \\ \vdots & \vdots & \ddots & \vdots & p_0 \\ \vdots & \vdots & \ddots & p_7 & \vdots \\ 0 & 0 & \cdots & 0 & p_7 \end{bmatrix}_{(L_p+L-1)\times L},$$

$\mathbf{w}$ is the vector of complex baseband additive white Gaussian noise (AWGN) that is assumed to be uncorrelated with the channel $\mathbf{h}$.

This is to note that the first $L-1$ rows of matrix $\mathbf{M}$ contain the last $L-1$ data samples of the previous OFDM symbol. Since these data samples of the previous OFDM symbol have already been estimated, the use of the estimated samples in the channel estimation during the current OFDM symbol makes the scheme decision-directed. This proposed solution of channel estimation is labeled as RLS w/DD time-domain channel estimation (RLS w/DD). The proposed channel tracking scheme is illustrated with block diagram in Fig. 3.

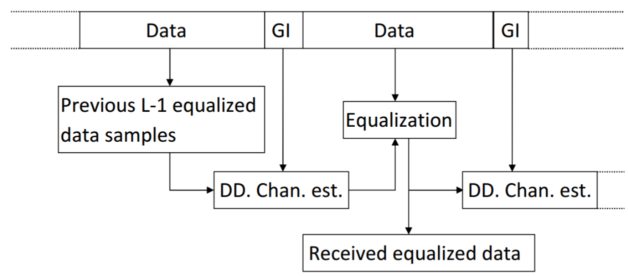

Fig. 3: Block Diagram of the Proposed Channel Estimation Method

With the availability of the PR training sequence and the estimates of $\{\hat{x}_{q-1,N-1}, \hat{x}_{q-1,N-2}, \cdots, \hat{x}_{q-1,N-L+1}\}$ from the previous OFDM symbol, the matrix $\mathbf{M}$ is estimated as $\tilde{\mathbf{M}}$ at the receiver end during the GI of the current OFDM symbol. Recursive least squares solution of (6) minimizes the total squared error, $E(j)$ at any time instant j given as

$$E(j) = \sum_{i=0}^{j} \left|c(i) - \mathbf{M}(i,:)\mathbf{h}(i)\right|^2 \tag{7}$$

and provides the solution at time index j as below [11]:

$$\tilde{\mathbf{h}}(j) = \hat{\mathbf{R}}^{-1}(j)\hat{\mathbf{d}}(j) \quad \text{for} \quad j = 0, 1, \cdots, L_p + L - 2. \tag{8}$$

where $\tilde{\mathbf{h}}(j)$ is the estimated channel tap coefficient vector $[\tilde{h}_0, \tilde{h}_1, \cdots, \tilde{h}_{L-1}]^T$ in the current OFDM symbol and, $\hat{\mathbf{R}}(j)$

and $\hat{\mathbf{d}}(j)$ are the correlation matrix and cross-correlation vector, respectively. These are given as:

$$\hat{\mathbf{R}}(j) = \sum_{i=0}^{j} \lambda^{j-i} (\tilde{\mathbf{M}}(i,:))^H \tilde{\mathbf{M}}(i,:) \tag{9}$$

and

$$\hat{\mathbf{d}}(j) = \sum_{i=0}^{j} \lambda^{j-i} (\tilde{\mathbf{M}}(i,:))^H c(i) \tag{10}$$

where λ is the forgetting factor with the range $0 < \lambda \leq 1$. In the RLS algorithm, $\hat{\mathbf{R}}(j)$ and $\hat{\mathbf{d}}(j)$ are updated iteratively as

$$\hat{\mathbf{R}}(j) = \lambda \hat{\mathbf{R}}(j-1) + (\tilde{\mathbf{M}}(j,:))^H \tilde{\mathbf{M}}(j,:) \tag{11}$$

$$\hat{\mathbf{d}}(j) = \lambda \hat{\mathbf{d}}(j-1) + (\tilde{\mathbf{M}}(j,:))^H c(j) \tag{12}$$

Since matrix inversion is time-intensive, calculation of $\hat{\mathbf{R}}^{-1}(j)$ in (8) adds more time-complexity to the RLS solution. Thus, $\hat{\mathbf{R}}^{-1}(j)$ required in equation (8) is also updated iteratively as below:

$$\hat{\mathbf{R}}^{-1}(j) = \lambda \hat{\mathbf{R}}^{-1}(j-1) - \alpha(j)\mathbf{g}(j)\mathbf{g}^H(j) \tag{13}$$

where

$$\mathbf{g}(j) = \lambda^{-1} \hat{\mathbf{R}}^{-1}(j-1)(\tilde{\mathbf{M}}(j,:))^H \tag{14}$$

and

$$\alpha(j) = 1 + \tilde{\mathbf{M}}(j,:)\mathbf{g}(j) \tag{15}$$

On using the above equations in (8), RLS update equation for channel tap coefficients is given as [11]:

$$\tilde{\mathbf{h}}(j) = \tilde{\mathbf{h}}(j-1) + \frac{\mathbf{g}(j)e(j)}{\alpha(j)} \tag{16}$$

where

$$e(j) = c(j) - \tilde{\mathbf{M}}(j,:)\tilde{\mathbf{h}}(j-1) \tag{17}$$

In RLS channel tap coefficients are updated recursively upon receiving the new training sample. RLS is used when the data samples are received sequentially and the estimated coefficients are required to be updated with the arrival of new measurements. Utilizing this method for channel estimation in time varying environment requires no a-priori knowledge of channel statistics. For the simulation, initial estimate of λ and $\hat{\mathbf{R}}^{-1}(0) = \frac{\mathbf{I}_{L \times L}}{\epsilon}$ are chosen empirically with λ=0.995 and $\epsilon = 10^{-3}$ [12].

IV. Simulation Results

In this section, we present simulation results to validate the working of the proposed channel estimation algorithm. Simulation results are carried out via the transmission of 100 OFDM frames over 500 channel realizations and 200 noise realizations. The number of OFDM symbols per frame is chosen to be 10 and 64, respectively. Data is modulated via quadrature phase-shift keying (QPSK). Results are generated with channel coherence time of 120μs (equal to 15 OFDM symbol duration) and 552μs (equal to 69 OFDM symbol duration). We simulated data communication via IEEE 802.11p standard for two wireless channel models (as shown in Table-II) of DSRC [12]. We used tapped delay line model for generating channel taps with the desired power spectrum profile as shown in Table-II. We generated each tap after 100ns delay. Maximum channel delay spread is 700ns for the 8-tap channel model and 400ns for the 5-tap channel model, respectively.

In case of V2V communication, it has been observed that maximum channel taps can be 8-10 [13-16]. Although channel can have longer delay spread, there is sufficient power in only first few taps. In general, results have been shown in the literature with only 3-7 tap length channels [13-16]. We have shown simulations over 5 tap (common for V2V) and 8-tap channel. Thus, we can easily fix the length of PR sequence to be 8 (=Lp). This provides us matrix $\mathbf{M}$ in (8) of maximum size $(Lp + L - 1 = 16)$ x L with maximum channel length to be 9.

In order to assess the un-coded performance of the proposed channel estimation scheme, no error correction codes are used in the simulation. Perfect timing synchronisation is assumed at the receiver end. In this section, we compare the performance of PRwZP TDLSE and its variants [9] and the proposed PRwZP RLS w/DD scheme with reference to Bit Error Rate (BER) versus energy per bit to noise power spectral density (E_b/N_0). In order to assess the performance of the decision-directed scheme, we have also shown results of PRwZP RLS (i.e., without decision-directed) scheme.

Table II: Parameters of vehicular channel models

Tap	Time (ns)	**Channel-1** Suburban street (120 km/h)	**Channel-2** Expressway (140 km/h)
1	0	0.0 dB, Rician, K = 3.3 dB	0.0 dB, Rician, K =−5.3 dB
2	100	−9.3 dB, Rayleigh	−9.3 dB, Rayleigh
3	200	−14.0 dB, Rayleigh	−20.3 dB, Rayleigh
4	300	−18.0 dB, Rayleigh	−21.3 dB, Rayleigh
5	400	−19.4 dB, Rayleigh	−28.8 dB, Rayleigh
6	500	−24.9 dB, Rayleigh	0
7	600	−27.5 dB, Rayleigh	0
8	700	−29.8 dB, Rayleigh	0

K = ratio of the specular to diffuse component power of the received signal

Fig-4 and Fig-5 show the simulation results for channel model-1 (8-tap) under the channel coherence times of 120μs and 552μs, respectively. Similarly, Fig.-6 and Fig.-7 show the simulation results for channel model-2 (5-tap) under the channel coherence times of 120μs and 552μs, respectively.

From these figures, the following observations are in order:

1) The proposed RLS and RLS w/DD schemes work better than the existing schemes that utilize LS for channel estimation.
2) Number of channel taps affects the performance of the proposed algorithm. For the 5-tap channel model, BER obtained is much closer to that of perfect channel state information (CSI). Thus, maximum excess delay spread

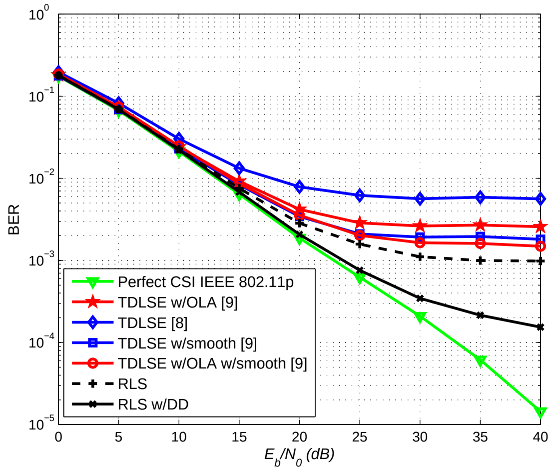

Fig. 4: BER vs. E_b/N_0, coherence time 120μs, 10 OFDM symbol per frame, 8-tap channel model (Channel-1), 120 km/h

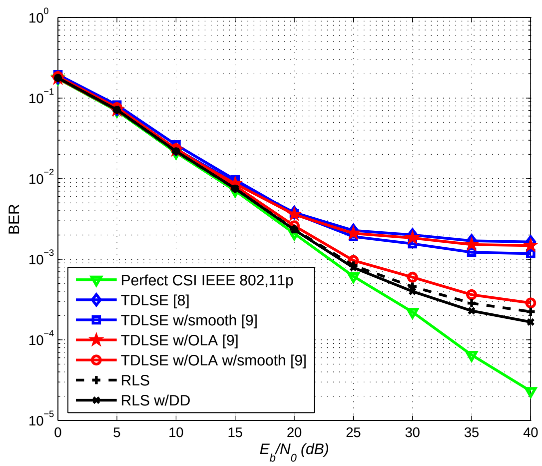

Fig. 5: BER vs. E_b/N_0, coherence time 552μs, 64 OFDM symbol per frame, 8-tap channel model (Channel-1), 120 km/h

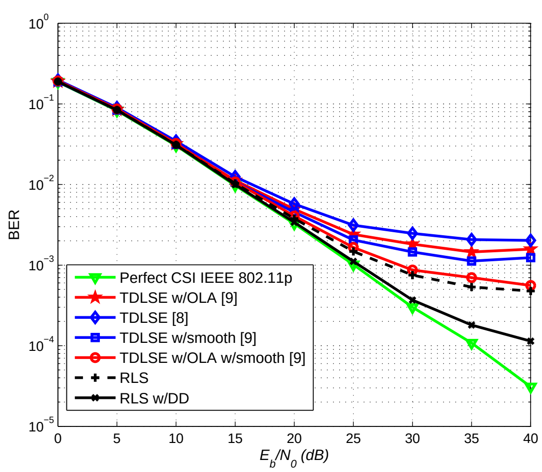

Fig. 6: BER vs. E_b/N_0, coherence time 120μs, 10 OFDM symbol per frame, 5-tap channel model (Channel-2), 140 km/h

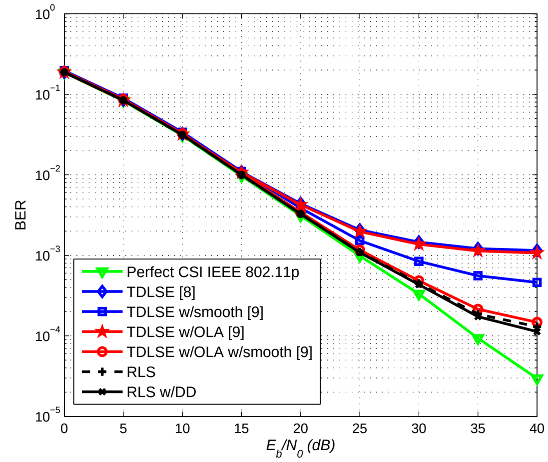

Fig. 7: BER vs. E_b/N_0, coherence time 552μs, 64 OFDM symbol per frame, 5-tap channel model (Channel-2), 140 km/h

of the channel impacts the performance of the proposed system.

3) In the case of PRwZP, orthogonality among subcarriers is lost. Although the existing PRwZP TDLSE with OLA is able to overcome this problem to some extent, the performance of the proposed RLS w/DD is even better. This shows that the proposed decision-directed scheme is able to mitigate the effect of ISI and ICI considerably compared to the existing OLA scheme.
4) Smoothing based techniques will fail when the channel varies within the smoothing window. Also, large smoothing windows will introduce longer delays that may be undesirable in real-time operations. From the simulation results, it is evident that the proposed algorithm works better than the smoothing based TDLSE method.
5) Simulation results show that the proposed algorithm provides better performance (very close to perfect channel state information (CSI)) compared to the existing GI based channel estimation methods.

V. CONCLUSION

In this paper, a novel time domain decision directed channel estimation based on RLS is proposed for rapid time varying channel estimation in IEEE 802.11p. The computational complexity is significantly reduced by recursively updating the channel on the arrival of new sample and with no matrix inversion. Due to the inclusion of decision-directed scheme in the proposed method, the channel estimation method is able to mitigate the effect of ISI and ICI considerably with no overhead of midamble insertion. Moreover, the proposed method does not require channel estimation via preamble and hence, effective data rate can be increased via removal of preamble in the future standards. This can address higher wireless data rate challenge for 5G communication.

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