A New Data Pilot-Aided Channel Estimation Scheme for Fast Time-Varying Channels in IEEE 802.11p Systems

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***Abstract*—The channel estimation in the IEEE 802.11p vehicle-to-everything (V2X) communication systems is generally a challenging task due to high mobility and insufficient number of pilots. Conventional data pilot-aided channel estimation schemes are difficult to apply in practice due to high complexity and the error propagation problem. In this correspondence paper, we propose a new data pilot-aided channel estimation scheme to overcome both issues. To this end, we develop a state feedback decision algorithm that enables us to extract reliable data pilots within a few received symbols. As a result, we can minimize both the error propagation effect and the computational complexity. We verify the complexity gain of the proposed scheme through the complexity analysis. Finally, we demonstrate the accuracy of the proposed channel estimation scheme in terms of the packet error rate performance.**

***Index Terms*—Channel estimation, WAVE, IEEE 802.11p, V2X, fast time-varying channels.**

I. Introduction

IEEE 802.11p has been developed as a dedicated physical layer (PHY) standard for vehicle-to-everything (V2X) communications in wireless access vehicular environments (WAVE) [1]. As the vehicle speed becomes faster, the V2X technologies are also required to have high reliability and low latency in fast moving environments. However, the current specification of 802.11p, which is based on convolutional coded orthogonal frequency division multiplexing (OFDM), hardly achieves both requirements due to structural constraint of the data packet. Specifically, the insufficient number of pilot subcarriers in each OFDM symbol makes it difficult for the receiver to estimate fast time-variant channels.

One possible way to tackle the problem is to insert more pilots into the packet [2]. However, such a modification may be unacceptable because of spectral efficiency loss, not to mention the cost for changing standards. In order to compensate for the insufficient number of pilots without modifying the packet structure, data-pilot aided successive (DPAS) channel estimation schemes, such as the spectral-temporal averaging (STA) method [3] and the constructed data pilot (CDP) algorithm [4], have been actively investigated in literature. The basic idea of the DPAS algorithm is to utilize the demapped data signals as virtual pilots for channel estimation, and the estimated channels are successively applied for data pilot construction (DPC) in the subsequent OFDM symbol. It is shown that the STA and CDP schemes improve the performance in the low and high signal-to-noise ratio (SNR) regimes, respectively. However, the performance gain is still marginal especially in the rapid time-variant (or fading) channels, because the detection errors during the DPC incur a serious error propagation effect.

To improve the reliability of the data pilots, the authors in [5] and [6] recently developed the iterative channel estimation and decoding (ICED) methods that exploit decoded data bits for DPC instead of the demapped signals. The ICED schemes exhibit improved channel estimation accuracy over the DPAS algorithms attributed to the decoding gain. However, they are practically less attractive, because the packet-wise iterative decoding gives rise to high complexity as well as a large processing delay at the receiver. In addition, their performance is not fully optimized, since the initial channel estimate as an input to the decoder is still dependent on the conventional DPAS algorithm.

In this correspondence, we propose a new channel estimation algorithm to further improve the performance of the ICED schemes with lower computational complexity. One important observation from the aforementioned DPAS algorithm is that the channel tracking performance in the fast time-varying channels is mainly determined by the reliability of the data pilots in the preceding OFDM symbol. Motivated by this observation, we develop a state feedback decision (SFD) algorithm that can extract reliable data pilots within a few received symbols without the aid of conventional decoders. The data pilots with improved reliability, in turn, deliver more accurate channel estimates to the subsequent OFDM symbol, thereby enhancing the channel estimation performance over a packet. Our method also offers great flexibility towards a receiver design because a useful tradeoff can be achieved between the reliability and the complexity by simply adjusting the number of information bits which the SFD algorithm processes at a time. Finally, we demonstrate the efficiency of the proposed scheme over the conventional DPAS and ICED methods in terms of the numerical packet error rate (PER) performance utilizing the actual V2X channel model.

Notation: Throughout the paper, normal letters represent scalar quantities and boldface letters indicate vectors. We use $[a(k)|k \in \mathbb{S}]$ to denote a row vector with $a(k)$ as entries in the order of $k \in \mathbb{S}$. The superscript $[\cdot]^T$ stands for the transpose operation.

II. System Model

IEEE 802.11p is based on the bit-interleaved coded OFDM technology [7] with 64-point fast fourier transform (FFT) and 10 MHz sampling frequency in the $f_c = 5.9$ GHz carrier frequency band. With one OFDM symbol duration equal to $T_s =$6.4 μs, the transmitter employs the convolutional encoder with code-rate R and memory order $\delta = 6$, and the receivers adopt the Viterbi decoder.

The IEEE 802.11p packet structure is composed of three parts: the preamble for time-frequency synchronization, the signal field for control information, and the data field for message signals. The long training (LT) part in the preamble consists of two OFDM symbols for channel estimation, while the data field has N_{data} OFDM symbols, which are typically tens or hundreds of symbols. In the data field, each OFDM symbol contains 64 subcarriers with an index set $\mathbb{S} = \{-32, -31, \ldots, 31\}$ including 4 pilot subcarriers, 12 null subcarriers, and 48 data subcarriers, each of which belongs to index sets $\mathbb{S}_P = \{-21, -7, 7, 21\}$, $\mathbb{S}_N = \{-32, \ldots, -27, 0, 27, \ldots, 31\}$, and $\mathbb{S}_D = \mathbb{S} \cap (\mathbb{S}_P \cup \mathbb{S}_N)^c$, respectively. Other specific parameters in IEEE 802.11p can be found in [1].

By eliminating the cyclic prefix and applying the FFT to the received signals, the receiver observes $Y_i(k)$ at the k-th subcarrier of the i-th

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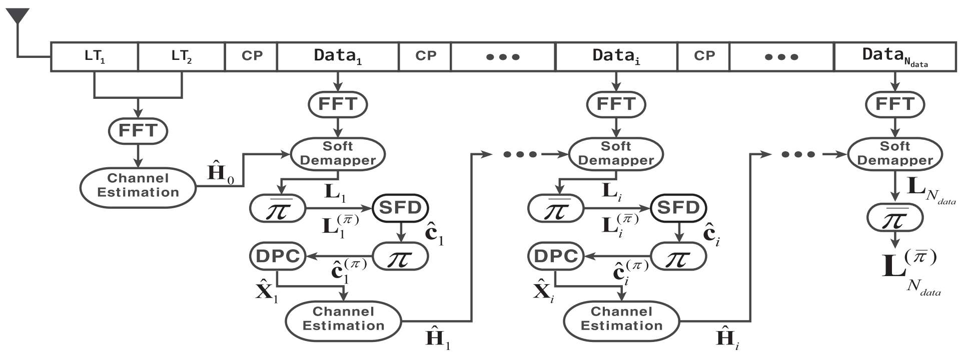

Fig. 1. Proposed DPAS channel estimation algorithm for WAVE 802.11p receivers.

OFDM symbol as

$$Y_i(k) = H_i(k)X_i(k) + W_i(k), \tag{1}$$

for $k \in \mathbb{S}_D$ and $i = 1, \ldots, N_{\text{data}}$, where $X_i(k)$, $H_i(k)$, and $W_i(k)$ denote the data symbol, the channel frequency response (CFR), and the additive Gaussian noise with zero mean and variance σ_W^2, respectively. Considering the M-ary modulation, each data symbol $X_i(k)$ is composed of $\log_2 M$ coded bits and each OFDM symbol in the data field has $J_c = 48 \log_2 M$ and $J_b = J_c R$ coded bits and information bits, respectively.

III. Proposed Scheme

In this section, we propose a new channel estimation scheme for IEEE 802.11p that outperforms conventional designs [3]–[6] with lower complexity. As shown in Figure 1, the proposed channel estimation method successively updates the CFRs based on the initial channel estimates in the preamble. In the figure, π and $\bar{\pi}$ denote the interleaving and deinterleaving operations, respectively. Unlike the conventional DPAS and ICED schemes, our method improves the data-pilot reliability in each OFDM symbol, thereby mitigating the error propagation effect without incurring any spectral efficiency loss. Details are described in the subsequent subsections.

A. Initial Channel Estimation Using Long Training Symbols

The receiver first obtains the initial channel estimates for the k-th subcarrier $H_0(k)$ from the two long training symbols, which are expressed by

$$Y_0^{(l)}(k) = H_0^{(l)}(k)X_0(k) + W_0^{(l)}(k),$$

for $l = 1, 2$ and $k \in \mathbb{S}_D$ where $Y_0^{(l)}(k)$, $X_0(k)$, and $W_0^{(l)}(k)$ represent the received signals in the long training period, the known training symbols, and the Gaussian noise with zero-mean and variance σ_W^2, respectively. Then, we estimate the initial CFRs by using the least square estimation strategy as

$$\hat{H}_0(k) = \frac{Y_0^{(1)}(k) + Y_0^{(2)}(k)}{2X_0(k)}, \ \forall k. \tag{2}$$

B. Channel Estimation in Data Field

Now, we estimate the CFRs $H_i(k)$ in the data field. As the original pilots in each OFDM symbol are confined to a few subcarriers in a fixed position, the channel estimation needs to be accomplished with the aid of the data pilots. As illustrated in Figure 1, the proposed channel estimation in the i-th OFDM symbol occurs in three steps: 1) the soft-demapping step that extracts the coded-bit log-likelihood ratio (LLR) values based on the channel estimates in the previous OFDM symbol, 2) the SFD step which makes a decision on the most probable coded bit sequence among multiple local codeword candidates, and 3) the DPC step that constructs the data pilots from the chosen codeword.

1) Soft-demapping Step: Defining $c_{m,i}$ as the m-th interleaved coded bit in the i-th OFDM symbol, we can compute the coded-bit LLR values as

$$L_{m,i} = \log\left(\frac{P(c_{m,i} = 0|\mathbf{Y}_i, \hat{\mathbf{H}}_{i-1})}{P(c_{m,i} = 1|\mathbf{Y}_i, \hat{\mathbf{H}}_{i-1})}\right) \text{ for } m = 1, \ldots, J_c \tag{3}$$

for the given observations $\mathbf{Y}_i = [Y_i(k)|k \in \mathbb{S}_D]$ and the channel estimates in the preceding OFDM symbol $\hat{\mathbf{H}}_{i-1} = [\hat{H}_{i-1}(k)|k \in \mathbb{S}_D]$. Note that if $i = 1$, we use the initial channel estimates $\hat{\mathbf{H}}_0$ constructed by (2). Let us denote $\mathbf{L}_i \triangleq [L_{1,i}, \ldots, L_{J_c,i}]$ as an LLR vector whose entries correspond to the interleaved coded bits. Finally, the deinterleaver delivers its deinterleaved version $\mathbf{L}_i^{(\bar{\pi})} \triangleq [L_{\bar{\pi}(1),i}, \ldots, L_{\bar{\pi}(J_c),i}]$ to the next step.

2) SFD Step: In this step, we recover more reliable coded bit sequences in the i-th OFDM symbol by correcting the erroneous coded bits. To this end, we first need to construct a look-up table that shows the relationship between the input and output of the encoder for a given initial encoder state. Then, we choose the best one among multiple output codewords according to the maximum a posteriori (MAP) decision rule by utilizing the LLR values in $\mathbf{L}_i^{(\bar{\pi})}$, and then save the current encoder state. We design the SFD algorithm which can process $1 \leq P \leq J_b$ information bits at a time. Thus, the algorithm needs to be performed $U = J_b/P$ times per OFDM symbol.[1]

Specifically, in the u-th SFD step with an encoder state $0 \leq \lambda_u \leq 2^\delta - 1$ for $u \in \{0, \ldots, U-1\}$, we can construct a look-up table corresponding to P information bits, which represents the relationship between all possible input bit sequences $\mathbb{B}_{u,i} = \{\mathbf{b}_{u,i}^{(1)}, \ldots, \mathbf{b}_{u,i}^{(2^P)}\}$ and their corresponding output coded bit sequences $\mathbb{C}_{u,i} = \{\mathbf{c}_{u,i}^{(1)}, \ldots, \mathbf{c}_{u,i}^{(2^P)}\}$ where $\mathbf{b}_{u,i}^{(j)} = [b_{Pu+1,i}^{(j)}, \ldots, b_{Pu+P,i}^{(j)}]^T$ and $\mathbf{c}_{u,i}^{(j)} = [c_{Qu+1,i}^{(j)}, \ldots, c_{Qu+Q,i}^{(j)}]^T$ indicate the vectors containing binary elements with $Q \triangleq P/R$. Table I shows a look-up table example of $\lambda_u = 0$ for a system with $P = 2$ and $R = 1/2$. Here, λ_{u+1} stands for the current encoder state when the u-th encoding is finished.

Once the look-up table is determined, we can find the most probable coded bit sequence among 2^P candidates in $\mathbb{C}_{u,i}$ such that the posterior probability is maximized as

$$\hat{\mathbf{c}}_{u,i} = \underset{\mathbf{c}_{u,i}^{(j)} \in \mathbb{C}_{u,i}}{\arg\max}\, P(\mathbf{c}_{u,i}^{(j)}|\mathbf{Y}_i, \hat{\mathbf{H}}_{i-1})$$

[1]Here, we assume that P is chosen such that U is an integer number.

TABLE I
A Look-up Table Example at the u-th SFD Step With $P = 2, R = 1/2$, and $\lambda_u = 0$

	$c^{(j)}_{4u+1,i}$	$c^{(j)}_{4u+2,i}$	$c^{(j)}_{4u+3,i}$	$c^{(j)}_{4u+4,i}$	λ_{u+1}
$\mathbf{b}^{(1)}_{u,i} = [0\ 0]^T$	0	0	0	0	0
$\mathbf{b}^{(2)}_{u,i} = [0\ 1]^T$	0	0	1	1	16
$\mathbf{b}^{(3)}_{u,i} = [1\ 0]^T$	1	1	1	0	32
$\mathbf{b}^{(4)}_{u,i} = [1\ 1]^T$	1	1	0	1	48

$$= \arg\max_{\mathbf{c}^{(j)}_{u,i} \in \mathbb{C}_{u,i}} \prod_{q=1}^{Q} P(c_{\bar{\pi}(Qu+q),i} = c^{(j)}_{Qu+q,i} | \mathbf{Y}_i, \hat{\mathbf{H}}_{i-1})$$

$$= \arg\max_{\mathbf{c}^{(j)}_{u,i} \in \mathbb{C}_{u,i}} \sum_{q=1}^{Q} (2c^{(j)}_{Qu+q,i} - 1) L_{\bar{\pi}(Qu+q),i}, \quad (4)$$

where $\{c_{\bar{\pi}(m),i} | \forall m\}$ denotes the deinterleaved coded bit sequence at the receiver. The second equality holds true because the elements in $\{c_{\bar{\pi}(m),i} | \forall m\}$ are independent of each other due to the deinterleaving operation at the receiver and the last equality follows from the well-known equivalence between the posterior probability and the LLR value as [8]

$$P(c_{\bar{\pi}(Qu+q),i} = d | \mathbf{Y}_i, \hat{\mathbf{H}}_{i-1}) = \frac{\exp(\frac{(2d-1)}{2} L_{\bar{\pi}(Qu+q),i})}{\exp(-\frac{L_{\bar{\pi}(Qu+q),i}}{2}) + \exp(\frac{L_{\bar{\pi}(Qu+q),i}}{2})} \quad \text{for } d \in \{0, 1\}.$$

After the codeword selection is finished, we provide the corresponding encoder state λ_{u+1} to the next SFD step for the $(u+1)$-th look-up table construction. The algorithm is repeated until u reaches $U-1$.[2] This approach enables us to obtain reliable data pilots within each OFDM symbol, and thus can transfer more accurate CFRs to the subsequent OFDM symbol. We note that as P grows, the decision accuracy of (4) becomes higher. Therefore, we can utilize more reliable data pilots at the expense of increased complexity.

3) DPC Step: In this step, we construct the data-pilots based on the selected codeword $\hat{\mathbf{c}}_i = [\hat{\mathbf{c}}_{0,i}, \ldots, \hat{\mathbf{c}}_{U-1,i}]$ in the previous step for estimating the CFRs in $\mathbf{H}_i$. First, we obtain $\hat{\mathbf{c}}^{(\pi)}_i$ that is an interleaved version of $\hat{\mathbf{c}}_i$. Then, we compute the data pilots $\hat{X}_i(k)$ for $k \in \mathbb{S}_D$, and estimate the CFRs in the i-th OFDM symbol as

$$\hat{H}_i(k) = \frac{Y_i(k)}{\hat{X}_i(k)}, \text{ for } k \in \mathbb{S}_D. \quad (5)$$

Finally, the result is passed to the soft-demapping step in the $(i+1)$-th OFDM symbol. We repeat the algorithm until the end of a packet. The entire process of the proposed channel estimation scheme is summarized in Algorithm 1.

IV. Complexity Analysis

In what follows, we briefly address the complexity gain of the proposed channel estimation scheme compared to the conventional ICED methods [5], [6]. As the soft-demapping and DPC steps are employed both in our method and the ICED, we only compare the complexity of the SFD algorithm with that of the iterative decoding in the ICED. For the rest of the paper, we assume $R = 1/2$. To compute the decoding complexity of the ICED, we adopt the analytical method in [9], in which the required number of basic operations such as addition (ADD), multiplication (MUL), maximum (MAX), and look-up table (LKUP) construction have been calculated according to the memory order δ. Then, it turns out that the ICED methods based on Viterbi decoder [5] and turbo decoder [6] require $I_{\text{iter}}(10 \cdot 2^{\delta} + 3)$ and $I_{\text{iter}}(48 \cdot 2^{\delta} - 13)$ number of equivalent additions (EAs) per information bit[3], respectively, where I_{iter} represents the number of decoding iterations. Note that $I_{\text{iter}} = 0$ means no iteration, i.e., no improvement from the initial channel estimates of the ICED schemes. The complexity for the initial channel estimation is ignored for simplicity.

[2]In the $(U-1)$-th SFD step, we deliver the current encoder state λ_U to the initial encoder state λ_0 in the next OFDM symbol.

Algorithm 1: The Proposed Channel Estimation Scheme.

Obtain initial channel estimates $\{\hat{H}_0(k) | k \in \mathbb{S}_D\}$ as in (2). Set the initial encoder state as $\lambda_0 = 0$.
for $i = 1 : N_{\text{data}}$
  Obtain $\boldsymbol{L}^{(\bar{\pi})}_i$ as in Section III-B1 based on $\hat{\mathbf{H}}_{i-1}$.
  for $u = 0 : U-1$
    Construct a look-up table for a given encoder state λ_u.
    Select $\hat{\mathbf{c}}_{u,i}$ according to the rule in (4).
    Save the current encoder state λ_{u+1} associated with the selected codeword $\hat{\mathbf{c}}_{u,i}$.
  end
  Construct the data pilots and estimate CFRs $\hat{\mathbf{H}}_i$.
  Set $\lambda_0 := \lambda_U$.
end
Restore the message bits based on the estimated CFRs $\{\hat{\mathbf{H}}_i | i = 1, \ldots, N_{\text{data}}\}$.

TABLE II
Required Number of EAs

ICED with Viterbi	$I_{\text{iter}}(10 \cdot 2^{\delta} + 3)$
ICED with Turbo	$I_{\text{iter}}(48 \cdot 2^{\delta} - 13)$
SFD	$\frac{1}{P}[(7P-1)2^P + 5]$

According to [9], the complexity of the proposed SFD scheme requires $(Q-1)2^P/P$ ADD, $Q2^P/P$ MUL, $1/P$ MAX, and $2^P + 1/P$ LKUP operations per information bit. We compare the proposed scheme with the ICED methods in terms of the required number of EAs per information bit as shown in Table II. We see from the result that as P reduces or δ grows, the complexity gain of the proposed method will be more pronounced. For example, the proposed SFD algorithms for $P = 2$ and 6 achieve 95% and 31% reduction in terms of EA over the *ICED with Viterbi* when $I_{\text{iter}} = 1$ and $\delta = 6$. Note that $\delta = 6$ is a standard value proposed in IEEE 802.11p. It is worth emphasizing that the performance of the proposed scheme with only $P = 2$ significantly outperforms the ICED schemes with much reduced computational complexity as will be shown in the subsequent section.

V. Simulation Results

In this section, we demonstrate the efficiency of the proposed channel estimation scheme through numerical simulation results. For our simulations, we have employed the $CohdaWireless$ V2X channel model in none line-of-sight (NLOS) environments. The specific channel parameters such as the tap delay line and the Doppler frequency for each channel tap are listed in [10]. We adopt 4-QAM with $R = 1/2$. As for the ICED schemes, the initial channel estimates are built upon the CDP method. For all cases, the PER performance is averaged over 10^5 packet transmissions.

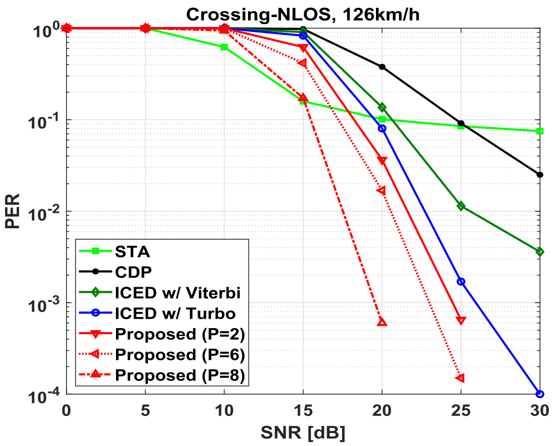

Fig. 2. PER performance comparison of various channel estimation schemes with relative velocity 126 km/h.

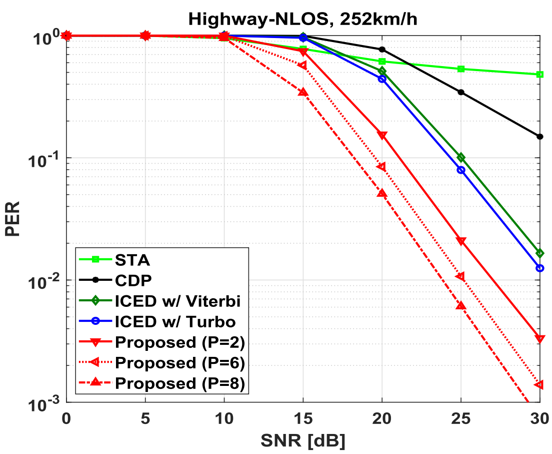

Fig. 3. PER performance comparison of various channel estimation schemes with relative velocity 252 km/h.

Figures 2 and 3 exhibit the PER performance of various channel estimation schemes according to the SNR defined by $1/\sigma_w^2$ with the relative velocity $v = 126$ and 252 km/h in the crossing and highway NLOS environments, respectively. In these scenarios, the conventional DPAS algorithms such as STA and CDP show the worst performance due to the error propagation effect. Although the ICED schemes provide improved performance, the gain is still limited owing to poor accuracy of the initial channel estimates that depend on CDP. We confirm from the figures that the proposed scheme outperforms all the aforementioned designs regardless of moving speed. Also, we observe that as P grows, the proposed scheme attains better performance at the cost of increased complexity. Interestingly, the proposed scheme with only $P = 2$ achieves at least 2 and 5 dB gains over the ICED methods with much lower computational complexity in the case of $v = 126$ and 252 km/h, respectively.

The performance gain of the proposed scheme will be more pronounced as the packet length becomes longer. This is because in this case, the error propagation issue becomes more intensive. To see this, in Figure 4, we present the PER performance of various channel estimation schemes in terms of the packet length in the 252 km/h highway-NLOS environment. As expected, the conventional DPAS and ICED designs become vulnerable to the mobility as the packet length increases, while the proposed scheme shows a robustness to the packet length change attributed to the error propagation mitigation effect.

[3] 1 MUL, 1 MAX, and 1 LKUP operations are counted as 1, 2, and 3 EAs, respectively [9].

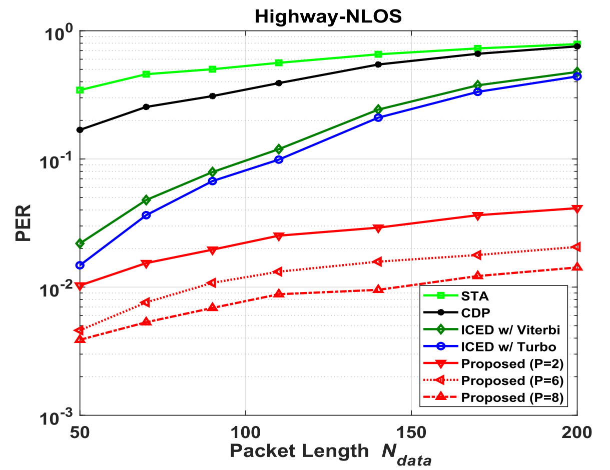

Fig. 4. PER performance comparison according to the packet length with SNR = 25 dB.

VI. Conclusion

In this correspondence, we have proposed a new channel estimation scheme for tracking fast time-varying channels in the IEEE 802.11p systems. To resolve the error propagation issue of the conventional DPAS algorithms, we have introduced an SFD algorithm that enhances the channel estimation accuracy further by extracting more reliable data pilots in each OFDM symbol without the aid of conventional decoders. Also, we have analyzed computational complexity and verified the complexity gain of the proposed scheme over the conventional ICED methods. Finally, we have demonstrated the efficiency of the proposed scheme through simulation results. Our method is practically attractive, since the reliability level of the data and the receiver complexity can be adjusted depending on the situation. It is also of interest to mathematically characterize such a trade-off and find the optimal value of P as future works.

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