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Improved Maximum Access Delay Time, Noise Variance, and Power Delay Profile Estimations for OFDM Systems

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Abstract

In this paper, we propose improved maximum access delay time, noise variance, and power delay profile (PDP) estimation schemes for orthogonal frequency division multiplexing (OFDM) system in multipath fading channels. To this end, we adopt the approximate maximum likelihood (ML) estimation strategy. For the first step, the log-likelihood function (LLF) of the received OFDM symbols is derived by utilizing only the cyclic redundancy induced by cyclic prefix (CP) without additional information. Then, the set of the initial path powers is sub-optimally obtained to maximize the derived LLF. In the second step, we can select a subset of the initial path power set, i.e. the maximum access delay time, so as to maximize the modified LLF. Through numerical simulations, the benefit of the proposed method is verified by comparison with the existing methods in terms of normalized mean square error, erroneous detection, and good detection probabilities.

Keywords: maximum access delay time, PDP, OFDM, multipath fading channel, ML

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1. Introduction

Noise variance, signal to noise ratio (SNR), and maximum access delay time estimations are essential information to evaluate the channel quality. Only when accurate estimation of them is made, the optimal modulation order can be determined and then, the channel estimation (CE) performance and the error correction capability of channel coding schemes can be maximized [1][2]. The authors in [3] and [4] dealt with discrete Fourier transform (DFT)-based CE, respectively, related to SNR estimation and maximum access delay time for orthogonal frequency division multiplexing (OFDM) systems. The work in [5] addressed the noise power calculation using repeated nulled subcarriers. An adaptive minimum mean square error (MMSE)-CE was addressed in [6] related with maximum access delay time estimation.

For OFDM systems, several noise variance or SNR estimation schemes have been studied [1-11]. There are two kinds of estimation schemes: one is data-aided (DA) scheme requiring pilot symbols and the other is non-data aided (NDA) scheme using the cyclic redundancy induced by the cyclic prefix (CP).

The authors in [1] have presented a noise variance and power delay profile (PDP) estimator for OFDM systems using the CP, but there is a major disadvantage of using an arbitrarily chosen threshold α. There is a problem that the estimation performance is significantly affected depending on how this threshold is selected. In order to overcome this drawback, the authors in [2] presented a maximum access delay time estimator without ambiguity due to the subjective threshold selection, and then showed that noise variance and SNR estimation performance are improved in arbitrary multipath fading channels. An NDA-based noise variance estimation scheme has been proposed in [7].

DA-based works in [8] and [9] dealt with the delay spread estimation based on training symbols and the SNR estimation based on the preamble for OFDM systems, respectively. Unlike estimation techniques using the CP, estimators using specially designed training sequences or preambles not only consume radio resources additionally, but are also difficult to be applied to existing OFDM systems. In this estimation scheme, reference signals suitable for frequency selective fading were designed and utilized, and the noise variance was most accurately estimated. Although there has been continuous improvement in estimation performance in the studies disclosed in [2][6–9], these studies have not yet proposed an estimator considering the PDP. Estimation of the PDP is essential for the MMSE-CE of the OFDM system. If the PDP is first estimated and used, performance approaching the MMSE can be expected in estimating the noise variance. The conventional works in [6], [10], and [11] presented the PDP or MMSE-CE methods based on pilot symbols. In [12], the authors presented the CE scheme in IEEE 802.11p/WAVE system under the assumption that the maximum access delay time is known to the receiver.

Notice that the works in [6][8-11] cannot be used to estimate the PDP for the case of having insufficient pilot or training symbols such as IEEE 802.11p/WAVE system. Therefore, in this paper, we propose an improved PDP estimation method that can be applied to OFDM systems such as IEEE 802.11p where pilot symbols are not sufficient. Also, the proposed technique can be utilized to verify the performance of PDP based MMSE-CE schemes and to evaluate the practical performance of CE methods under ideal assumptions such as [12].

In this paper, based on the work in [1], we develop an improved maximum access delay time estimator in the form of a pilot symbol-free maximum likelihood (ML) using the parameters in [1] with no thresholds. For the first step, we derive the log-likelihood function (LLF) of the received OFDM symbols by using the periodic redundancy induced by the CP and then, the initial path powers are estimated to sub-optimally maximize the derived LLF. In the second step, a subset of the initially estimated path powers is selected to maximize the modified LLF. Notice that this step is to estimate the maximum excess delay time from which we can determine the estimated noise variance and the estimated PDP. Note that the correct detection (CD) of the maximum access delay time is important in order to fully utilize the correlation property of CP samples in estimation process but CD probability is not sufficient to explain the performance trend of estimators and its reason. In order to this, we present new performance metrics as the erroneous detection (ED) and the good detection (GD) probabilities of the estimated maximum access delay time.

The remainder of this paper is organized as follows: Section 2 describes discrete signal model for OFDM systems. Section 3 presents the proposed estimation method. Section 4 shows simulation results, and concluding remarks are given in Section 5.

2. Discrete Signal Model for OFDM systems

In OFDM systems, source data are grouped and mapped into a modulated symbol $X_m(k)$ with $E\left\{\left|X_m(k)\right|^2\right\}=1$, $k\in\{0,1,\cdots,N-1\}$, and $E\{\cdot\}$ denoting expectation. Then, by inverse discrete Fourier transform (IDFT) on N parallel subcarriers, the transmitted time-domain signal of the n th sample for the m th OFDM symbol can be expressed as

$$x_m(n)=\sqrt{\frac{E_s}{N}}\sum_{k=0}^{N-1}X_m(k)\exp(j2\pi kn/N) \tag{1}$$

where $n\in\{0,1,\cdots,N-1\}$ and E_s is the signal power [13-15].

The guard interval can be inserted as the CP that replicates the end of the IDFT output samples. N_g is the number of guard interval samples assumed to be larger than the delay spread of the channel. The signal is transmitted over a multipath fading channel and its low-pass channel impulse response is expressed as

$$h(t;\tau)=\sum_{l=0}^{L-1}h_l(t)\delta(\tau-\tau_l) \tag{2}$$

where t, τ, $\delta(\cdot)$, L, and τ_l are the time, the delay, a Dirac delta function, the number of multipaths, and the propagation delay of the l th path, respectively [13-15]. The correlation relationship between the paths can be expressed by the wide-sense stationary uncorrelated scattering (WSSUS) model [14][16].

The received signal after removing CP is given by

$$y_m(n)=\sum_{l=0}^{L-1}h_{l,m}(n)x_m\left((n-d_l)_N\right)+w_m(n) \tag{3}$$

where $(\cdot)_N$ represents a cyclic shift in the base of N, $w_m(n)\sim\mathcal{N}\left(0,\sigma^2\right)$ is an Additive White Gaussian Noise (AWGN), $h_{l,m}(n)=h_l(t)\big|_{t=\left[m\left(N_g+N\right)+n\right]T_S}$ is the l th path channel gain of the n th sample for the m th OFDM symbol, and $d_l=\lfloor\tau_l/T_s\rfloor$ is the delay normalized by the sampling time T_s [14][15]. For simplicity, we round d_l to an integer without considering leakage. However, the correlation approach in this paper may also be extended to fractional d_l [1].

When we assume perfect synchronization with $d_0 = 0$, and that the channel is time-invariant within two consecutive OFDM symbols, indexes m and (n) in $h_{l,m}(n)$ of (3) can be omitted as $h_{l,m}(n) \to h_l$. For $-N_g \le n < 0$, the received signal samples of (3) can be expressed as

$$\begin{aligned} y_m(n) &= \sum_{l=0}^{L-1} h_l\, x_{m-1}(N+n-d_l)U(d_l-n) \\ &= \sum_{l=0}^{L-1} h_l\, x_m(n-d_l)U(n-d_l) + w_m(n) \end{aligned} \tag{4}$$

where $h_l \sim \mathcal{N}\left(0,\sigma_l^2\right)$, $\sigma_h^2 = \sum_{l=0}^{L-1}\sigma_l^2$, and $U(\cdot)$ is the unit step function [1]. Let us define the maximum number of paths including paths with zero channel gain as $L_{\max} = \max\{d_l\}+1$ and then, the maximum access delay time, normalized by T_s, is $d_{\max} = \max\{d_l\} = L_{\max}-1$.

Note that the samples in the CP and their copies are pairwise correlated as

$$E\left\{y_m(-k)\,y_m^*(N-k)\right\} = \begin{cases} \sigma_h^2, & 0 < k \le N_g - d_{L-1} \\ \sum_{l=0}^{L-1}\sigma_l^2 U\left(N_g-k-d_l\right), & N_g - d_{L-1} < k \le N_g - d_0 \\ 0, & N_g - d_0 < k \le N_g \end{cases} \tag{5}$$

where $k = 1,\cdots,N_g$ and the expectation in (5) is taken with regard to both $\{h_l\}$ and $\{x_m(n)\}$ [1][17].

For large L, $y_m(n)\big|_{n=0}^{N-1-N_g}$ can be assumed as complex Gaussian by the central limited theorem, and the probability density function (PDF) is written as

$$f\left(y_m(n)\right) = \frac{1}{\pi\left(\sigma_h^2+\sigma^2\right)}\exp\left(-\frac{\left|y_m(n)\right|^2}{\sigma_h^2+\sigma^2}\right). \tag{6}$$

From the correlation property of (5), samples $y_m(-k)$ and $y_m(N-k)$ can be jointly Gaussian with the PDF of

$$f\left(y_m(-k), y_m(N-k)\right) = \frac{\exp\left(-\dfrac{\left|y_m(-k)\right|^2 + \left|y_m(N-k)\right|^2 - 2\rho_k \Re\left\{y_m(-k)\,y_m^*(N-k)\right\}}{\sigma_h^2+\sigma^2}\right)}{\pi^2\left(\sigma_h^2+\sigma^2\right)\left(1-\rho_k^2\right)} \tag{7}$$

where

$$\rho_k = \frac{\left|E\left\{y_m(-k)\,y_m^*(N-k)\right\}\right|}{\sqrt{E\left\{\left|y_m(-k)\right|^2\right\}E\left\{\left|y_m(N-k)\right|^2\right\}}} = \frac{\sum_{l=0}^{L-1}\sigma_l^2 U\left(N_g-k-d_l\right)}{\sigma_h^2+\sigma^2}. \tag{8}$$

Note that $0 < \rho_k < 1$ and $\rho_k \ge \rho_{k+1}$ (i.e., ρ_k is a non-increasing in proportion to k).

3. Proposed Estimation Method

Under the perfect synchronization at reception and a time-invariant channel over two OFDM symbol times, N_g noise variance estimators can be written as

$$\hat{\sigma}_u^2 = J(u),\ 1 \le u \le N_g$$
$$J(u) = \frac{1}{2M\left(N_g - (u-1)\right)} \sum_{m=1}^{M} \sum_{k=u}^{N_g} \left| y_m(N-k) - y_m(-k) \right|^2 \tag{9}$$

where M is the number of OFDM symbols in the observation window. Under given environment, **Fig. 1** shows normalized mean square errors (NMSEs) for N_g noise variance estimators of (9) and the estimator with the smallest NMSE is found at $u = L_{\max}$ [2].

Let us define $\mathbf{y} = \left[y_1(-N_g), y_1(-N_g+1), \cdots, y_M(N-1) \right]$, $\mathbf{p} = \left[\sigma_0^2, \cdots, \sigma_{L-1}^2 \right]$, and $\mathbf{d} = \left[d_0, \cdots, d_{L-1} \right]$. From the fact that M OFDM symbols are mutually independent, (6), and (7), the log-likelihood function of $\mathbf{y}$ can be presented, conditioned on σ^2, $\mathbf{p}$, and $\mathbf{d}$, as

$$\begin{aligned} \Lambda\left(\mathbf{y} \middle| \sigma^2, \mathbf{p}, \mathbf{d}\right) &= \sum_{m=1}^{M} \log\left(\prod_{k=1}^{N_g} f\left(y_m(-k), y_m(N-k)\right) \prod_{k=0}^{N-1-N_g} f\left(y_m(k)\right) \right) \\ &= -M \left(\sum_{k=1}^{N_g} \left[\frac{a_k - 2\rho_k b_k}{c\left(1-\rho_k^2\right)} + \log\left(\pi^2 c\left(1-\rho_k^2\right)\right) \right] + \sum_{k=0}^{N-1-N_g} \frac{g_k}{c} + \log(\pi c) \right) \end{aligned} \tag{10}$$

where

$$\begin{aligned} a_k &= \frac{1}{M} \sum_{m=1}^{M} \left(\left| y_m(-k) \right|^2 + \left| y_m(N-k) \right|^2 \right) \\ b_k &= \frac{1}{M} \sum_{m=1}^{M} \Re\left\{ y_m(-k) y_m^*(N-k) \right\}^2 \\ g_k &= \frac{1}{M} \sum_{m=1}^{M} \left| y_m(k) \right|^2 \\ c &= \sigma_h^2 + \sigma^2. \end{aligned} \tag{11}$$

The authors in [1] showed a suboptimal way for the joint parameters' estimation of (10). As the time average estimation for $\sigma_h^2 + \sigma^2$, we can estimate c in (10) as

$$\hat{c} = \frac{1}{N - N_g} \sum_{k=0}^{N-1-N_g} g_k = \frac{1}{\left(N - N_g\right) M} \sum_{k=0}^{N-1-N_g} \sum_{m=1}^{M} \left| y_m(k) \right|^2. \tag{12}$$

By substituting $\hat{c}$ into the first summation in (10) and maximizing ρ_k individually, ρ_k can be estimated as the real root of the equation

$$\hat{c}\rho_k^3 - b_k \rho_k^2 + \left(a_k - \hat{c}\right)\rho_k - b_k = 0. \tag{13}$$

By letting $\left\{\hat{\rho}_k\right\}_{k=1}^{N_g}$ be the real roots of N_g cubic equations of (13), the temporary estimated path power can be written as

$$\hat{p}_0 = \hat{\sigma}_0^2 = \hat{\rho}_{N_g} / \hat{c}$$
$$\hat{p}_k\Big|_{k=1}^{N_g-1} = \hat{\sigma}_k^2 = \begin{cases} \left(\hat{\rho}_{N_g-k} - \hat{\rho}_{N_g-k+1}\right)/\hat{c} & \text{if} \quad \hat{\rho}_{N_g-k} > \hat{\rho}_{N_g-k+1} \\ 0 & \text{else.} \end{cases} \tag{14}$$

Table 1. $\left\{\rho_u(k)\right\}_{k=1}^{N_g}$ Calculation

Algorithm	**Comments**
for $u = 1 : N_g$	
$\mathbf{p}_u = \mathbf{0}_{1\times N_g}$	$1\times N_g$ zero vector
$p_u(k)\Big\|_{k=0}^{u-1} = \hat{p}_k = \hat{\sigma}_k^2$	u paths power of (14)
$\sigma^2 = J(u)$	u th noise power of (9)
$\mathbf{p}_u = \mathbf{p}_u \times \left(\hat{c} - \sigma^2\right) / \sum_{k=0}^{u-1} \hat{p}_k$	Power normalization
$\rho_u(k)\Big\|_{k=1}^{N_g} = \sum_{l=0}^{N_g-k} p_u(l) / \hat{c}$	u th ρ calculation
end for	

Table 2. Channel profile due to scenario in [19]

Ch. Type	**Item**	**Tap0**	**Tap1**	**Tap2**	**Tap3**	**Units**
Street Crossing NLOS ($126\ km/h$, $L_{\max} = 7$)	Power	0	-3	-5	-10	dB
	Delay (τ_l)	0	267	400	533	ns
	d_l	$d_0 = 0$	$d_1 \in \{2,3\}$	$d_2 = 4$	$d_3 \in \{5,6\}$	$\times T_s$
Highway NLOS ($252\ km/h$, $L_{\max} = 8$)	Power	0	-2	-5	-7	dB
	Delay (τ_l)	0	200	433	700	ns
	d_l	$d_0 = 0$	$d_1 = 2$	$d_2 \in \{4,5\}$	$d_3 = 7$	$\times T_s$

In [1], the authors suggested the $\hat{L}_{\max}$ estimation scheme based on a threshold value. If $\hat{p}_k\left(=\hat{\sigma}_k^2\right) > \alpha\hat{c}$, it is identified as a path having the estimated path power of $\hat{p}_k$ and the estimated delay time of k.

From $\left\{\hat{p}_k\right\}_{k=0}^{N_g-1}$ in (14), we propose $L_{\max}$ estimation method as a way to find $\hat{L}_{\max}$ that maximizes the log likelihood function of (10) regardless of a threshold level as shown in **Table 1**. By ignoring constant term in (10), we can represent (10), from $\left\{\rho_u(k)\right\}_{k=1}^{N_g}$ in **Table 1**, a_k, b_k, and $\hat{c}$, as

$$\Lambda_p\left(\mathbf{y}, \left\{\rho_u(k)\right\}_{k=1}^{N_g} \Big| L_{\max} = u\right) = -M\sum_{k=1}^{N_g}\left[\frac{a_k - 2\rho_u(k)b_k}{\hat{c}\left(1-\rho_u^2(k)\right)} + \log\left(1-\rho_u^2(k)\right)\right] \tag{15}$$

and then, $L_{\max}$ can be estimated as

$$\hat{L}_{\max} = \arg \max_{u} \left[\Lambda_p \left(\mathbf{y}, \left\{ \rho_u(k) \right\}_{k=1}^{N_g} \middle| L_{\max} = u \right) \right]. \tag{16}$$

From (16), the estimated noise variance, the estimated maximum access delay time, and the estimated PDP can be obtained as $\hat{\sigma}^2 = J\left(\hat{L}_{\max}\right)$, $\hat{d}_{\max} = \left(\hat{L}_{\max} - 1\right)$, and $\mathbf{p}_{\hat{L}_{\max}} = \mathbf{p}_u\big|_{u=\hat{L}_{\max}}$, respectively.

4. Simulation Results

From here, we demonstrate the efficiency of the proposed estimation schemes through simulations based on the IEEE 802.11p standard with $N = 64$, $N_g = 16$, and $T_s = 0.1\mu s$ [18]. We assume that one packet consists of 100 OFDM symbols and quaternary phase shift keying (QPSK) with coding rate of 1/2. For all cases, it is averaged over 5×10^5 packet transmissions with SNR $= E_s\sigma_h^2/\sigma^2$. In [19], five scenarios of 'CohdaWireless V2V channel model' were presented as 'Rural LOS', 'Urban Approaching LOS', 'Street Crossing NLOS', 'Highway LOS', and 'Highway NLOS'. Among the five channel environments, 'Street Crossing NLOS' and 'Highway LOS' have a relatively long delay spread and are more frequency selective channels without line of sight (LOS) component. In order to verify the performance of the proposed scheme at severe channel environment, we have employed 'Street Crossing NLOS with $126km/h$' and 'Highway NLOS with $252km/h$' of which the channel profiles are presented in **Table 2**. The other parameters such as the Doppler frequency for each channel tap are listed in [19]. In our simulation, we employ the fractional d_l by considering $T_s = 0.1\mu s$ in Table 1 so that 'Street Crossing NLOS' has $d_1 \in \{2,3\}$ and $d_3 \in \{5,6\}$ and 'Highway NLOS' has $d_2 \in \{4,5\}$ as shown in **Table 2**. For the fractional case, the given path power is divided into two according to the relative distance of two adjacent sampling time locations [20].

When we define the NMSE of $J(u)$ in (9) as $E\left[\left|J(u)-\sigma^2\right|^2\right]/\sigma^4$, **Fig. 1(a)** and **Fig. 1(b)** show it for 'Street Crossing NLOS' and 'Highway NLOS', respectively. From **Fig. 1**, we can find that there is a different trend depending on the region to which u belongs. For $1 \le u < L_{\max}$, the NMSE of $J(u)$ increases with respect to SNR because of the residual interference (i.e, inter-symbol interference(ISI)). Notice that $u = L_{\max}$ gives the smallest NMSE. For $L_{\max} < u \le 16$, the NMSE of $J(u)$ is slightly increased compared to $J(L_{\max})$ but it is maintained according to the SNR. Moreover, even for $J(L_{\max})$, it can be seen that the NMSE slightly increases at high SNR, which is due to the time-varying effect of the channel.

For $\hat{L}_{\max}$, we define ED, CD, and GD probabilities, respectively, as follows:

$$\begin{aligned} \text{Pr. of ED} &= \Pr\left\{\hat{L}_{\max} < L_{\max}\right\} \\ \text{Pr. of CD} &= \Pr\left\{\hat{L}_{\max} = L_{\max}\right\} \\ \text{Pr. of GD} &= \Pr\left\{L_{\max} \le \hat{L}_{\max} < L_{\max} + \left(N_g - L_{\max}\right)/2\right\} \end{aligned}$$

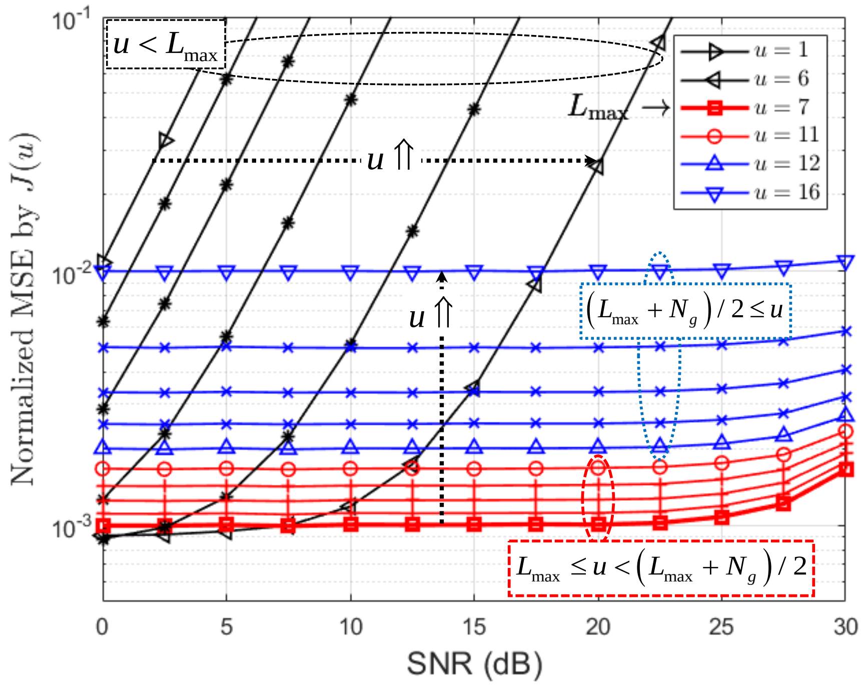
Fig. 2 and **Fig. 3** show Pr. of CD, Pr. of ED, and Pr. of GD, respectively, for 'Street Crossing NLOS' and 'Highway NLOS' with regard to different methods and α. When the NMSE of $\hat{\sigma}^2$ is defined as $E\left[\left|\hat{\sigma}^2-\sigma^2\right|^2\right]/\sigma^4$, **Fig. 4(a)** and **Fig. 4(b)** show the NMSE of $\hat{\sigma}^2$ for 'Street Crossing NLOS' and 'Highway NLOS', respectively, in order to compare different methods.

At first, let us consider 'Street Crossing NLOS' of **Fig. 2** and **Fig. 4(a).** In the high SNR region at **Fig. 2(a)**, 'Ref. [1]' with $\alpha = 0.01$ and $\alpha = 0.005$ can have a higher CD probability rather than other methods but a non-zero ED probability, which gives a low GD probability. Note that a non-zero ED probability means the case of $\hat{L}_{\max} < L_{\max}$ at **Fig. 1(a)** and hence the NMSE of 'Ref. [1]' is observed to greatly increase at **Fig. 4(a)** in the high SNR region. In the low SNR region at **Fig. 2**, 'Ref. [1]' has a low ED probabilities but low GD and CD probabilities, which means the occurrence of the case $\left(L_{\max}+N_g\right)/2 \le \hat{L}_{\max}$ in **Fig. 1(a)**. It causes that 'Ref. [1]' gives higher NMSE rather than other methods at **Fig. 4(a)**. The proposed scheme has a lower CD probability in the high SNR region but a higher CD (or GD) probability in the low SNR region. Furthermore, it gives a higher GD probability for all SNR region as shown in **Fig. 2(c)**.

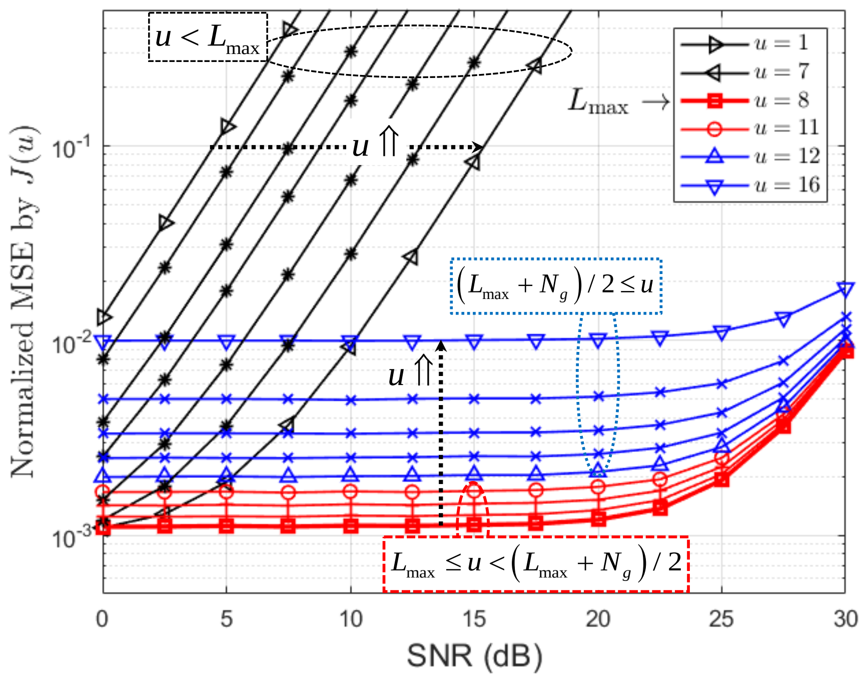
Second, let us describe the performance for 'Highway NLOS' of **Fig. 3** and **Fig. 4(b).** Similar to **Fig. 2**, it can be seen from **Fig. 3(a)** that the proposed scheme has a higher CD probability in the low SNR region and a lower CD probability in the high SNR region rather than 'Ref. [1]'. From **Fig. 3(b)**, the proposed method outperforms 'Ref. [2]' in terms of ED probability at a low SNR. From **Fig. 3(c)**, the proposed method outperforms 'Ref. [1]' and 'Ref. [2]' in terms of GD probability at a low SNR.

As shown in **Fig. 4**, both 'Ref. [2]' and the proposed method show the stable NMSE performance in all SNR ranges regardless of the channel environment. Notice from **Fig. 2(c)** and **Fig. 3(c)** that the proposed method gives more stable GD performance rather than both 'Ref. [1]' and 'Ref. [2]' in all SNR ranges without a threshold level.

When the NMSE is defined as $E\left[\left|\hat{\sigma}_k^2-\sigma_k^2\right|^2\right]/\sigma_k^4\Big|_{k\in\{6,7\}}$, **Fig. 5(a)** and **Fig. 5(b)** present the NMSEs of the 7th path power $\left(\hat{\sigma}_7^2\right)$ for 'Street Crossing NLOS' and the 6th path power $\left(\hat{\sigma}_6^2\right)$ for 'Highway NLOS', respectively. It is confirmed from **Fig. 5** that the proposed scheme outperforms 'Ref. [1]' regardless of both α and SNR.



(a) Street Crossing NLOS



(b) Highway NLOS

Fig. 1. NMSE of $J(u)$ ($u \in \{1, 2, \cdots, N_g\}$, $M = 100$).

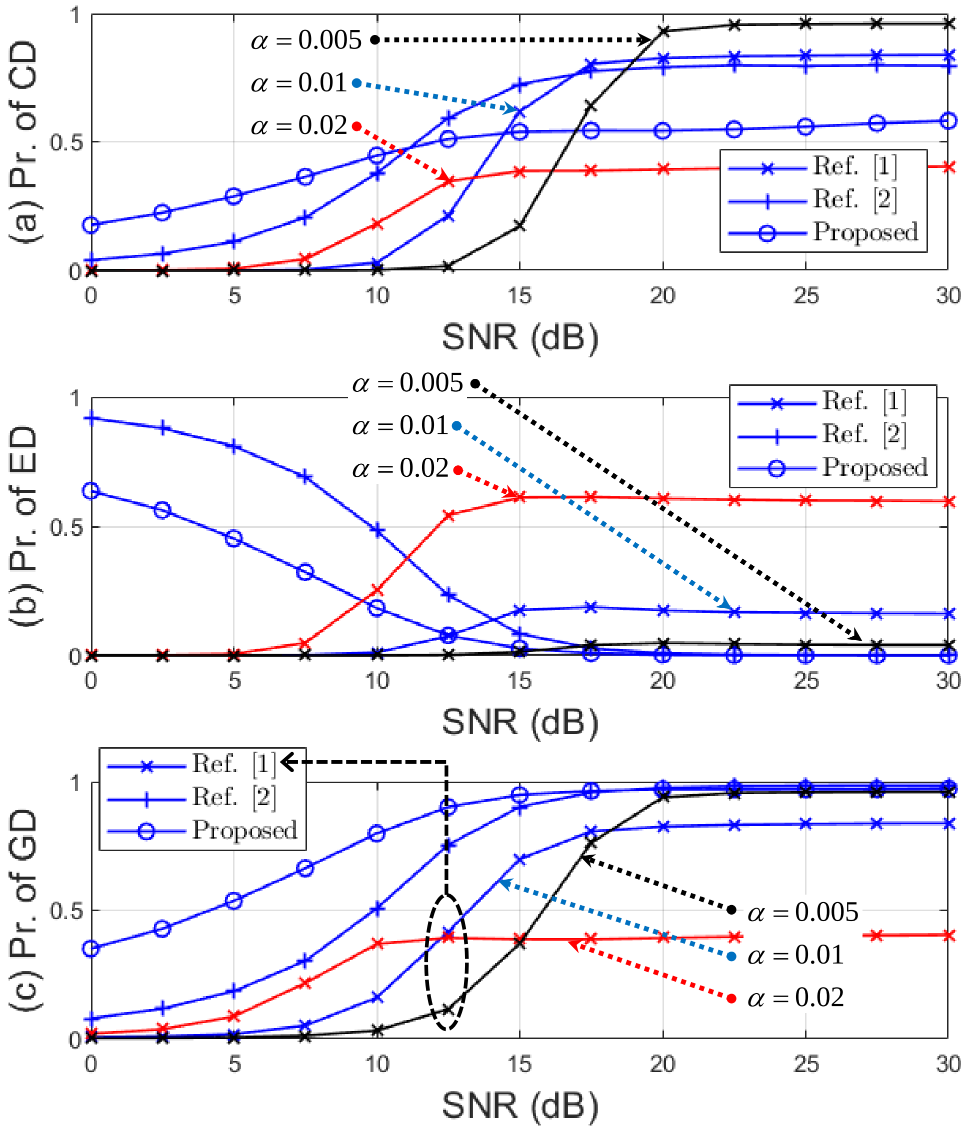


Fig. 2. Probabilities of CD, ED, and GD for $\hat{L}_{\max}$ with respect to α (Street Crossing NLOS, $M = 100$, $\alpha \in \{0.005, 0.01, 0.02\}$).

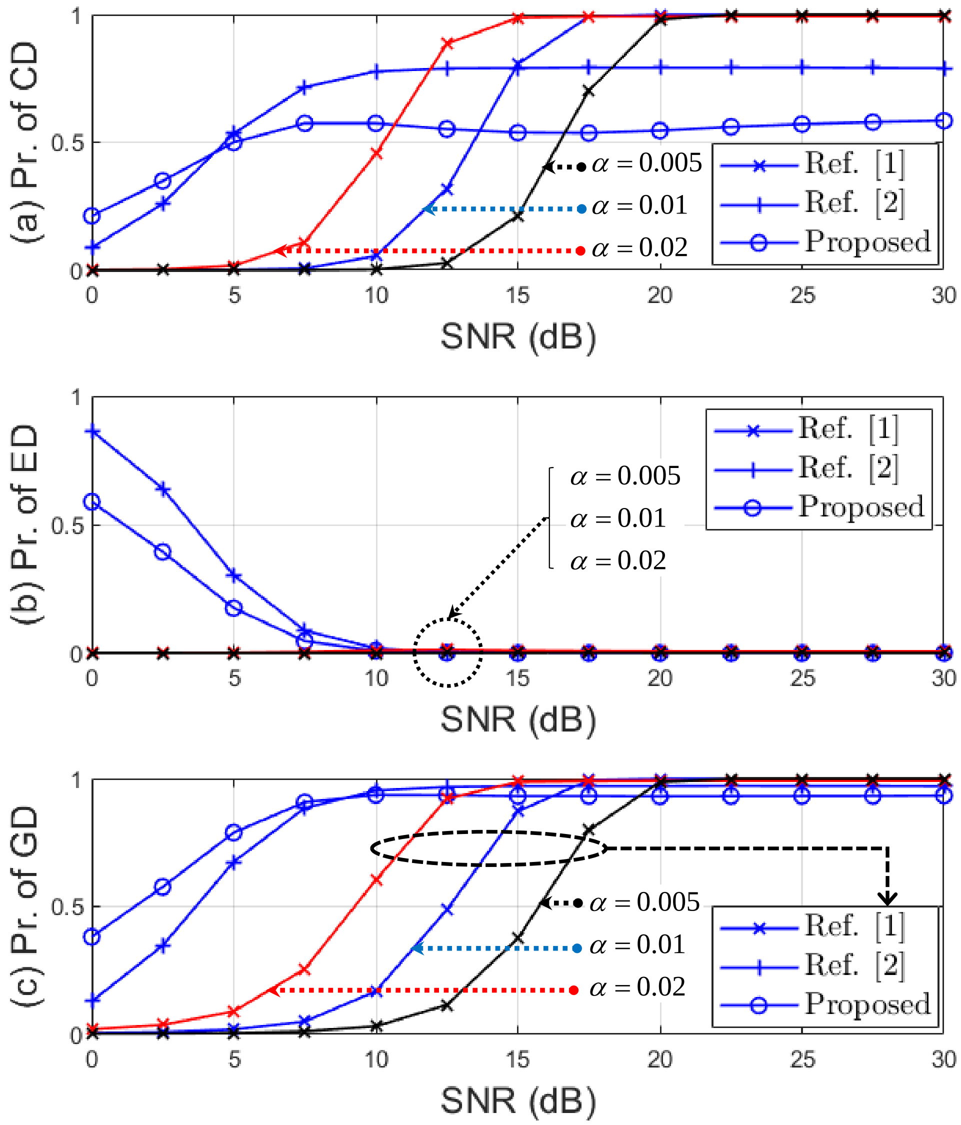


Fig. 3. Probabilities of CD, ED, and GD for $\hat{L}_{\max}$ with respect to α (Highway NLOS, $M = 100$, $\alpha \in \{0.005, 0.01, 0.02\}$).

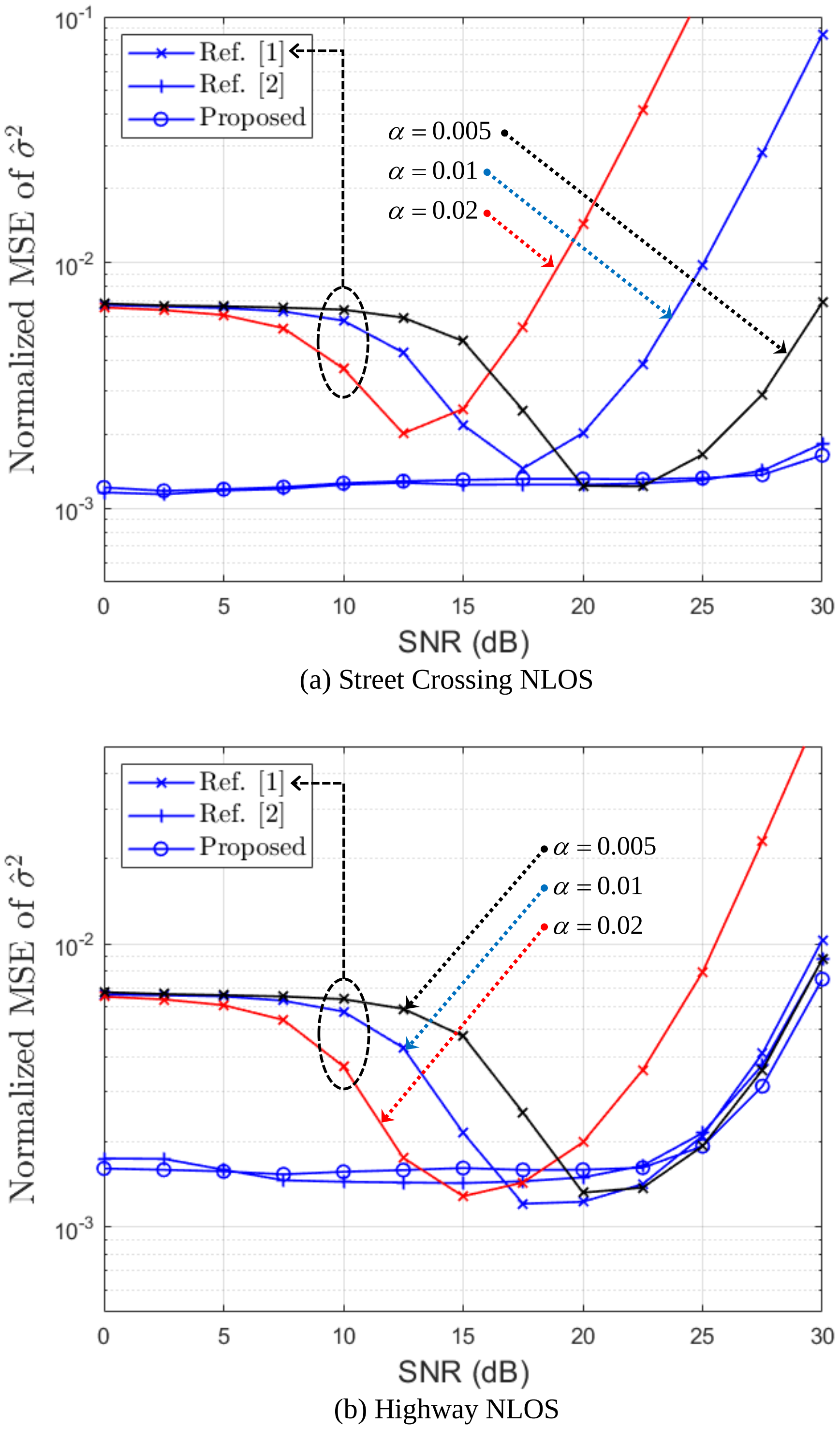
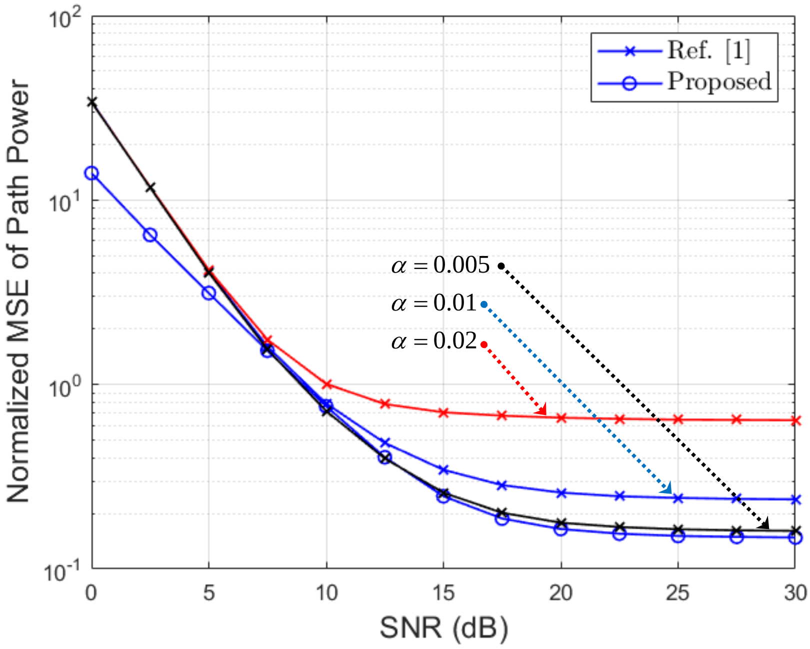
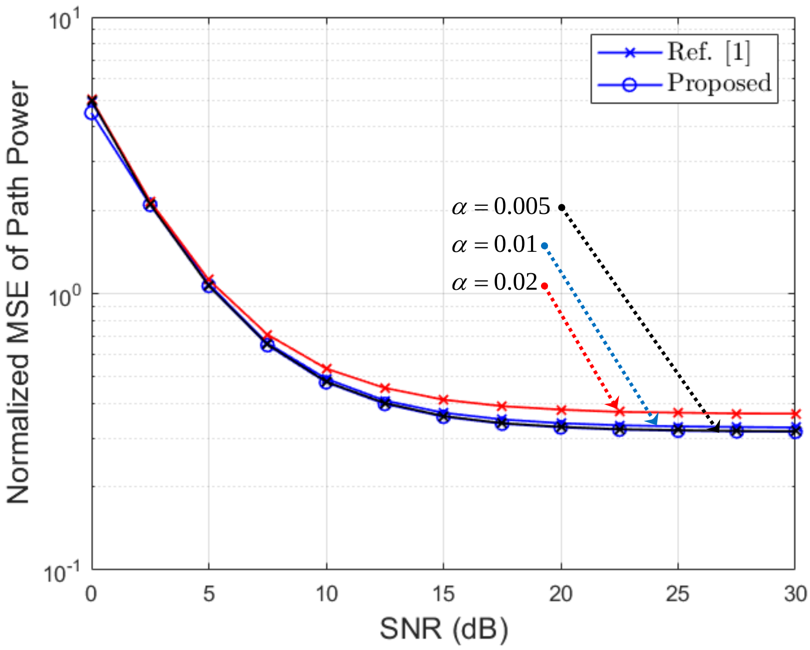


Fig. 4. NMSE of $\hat{\sigma}^2$ with respect to α ($M = 100$, $\alpha \in \{0.005, 0.01, 0.02\}$).



(a) Street Crossing NLOS $\left(\hat{\sigma}_7^2 = \mathbf{p}_{\hat{L}_{\max}}(7)\right)$



(b) Highway NLOS $\left(\hat{\sigma}_6^2 = \mathbf{p}_{\hat{L}_{\max}}(6)\right)$

Fig. 5. NMSE of path power with respect to α ($M = 100$, $\alpha \in \{0.005, 0.01, 0.02\}$).

5. Conclusions

In this contribution, we suggest improved estimation schemes of maximum access delay time, noise variance, and PDP utilizing only the correlation property of CP in each OFDM block. Through the observation results, it is confirmed that the proposed estimation scheme is showing a stable performance regardless of both SNR and α so as to outperform the methods described in [1] and [2]. Therefore, we can say that the robustness of the proposed method is confirmed. The results in this paper can be used for applications such as MMSE channel estimation for IEEE 802.11p/Wireless Access in Vehicular Environments (WAVE) system having insufficient number of pilots within OFDM symbols [18].

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