The Channel Estimation based on FSC Method for IEEE 802.11p OFDM System

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***Abstract*—This paper investigates the channel estimation (CE) schemes to track rapid time varying channels in IEEE 802.11p systems. To overcome the problems of the conventional preamble based CE scheme, various constructed data pilot (CDP) based algorithms have been widely developed. However, their performance is not sufficient to apply in practice, especially in high mobility and high modulation order cases to error propagation issue. Therefore, we present a novel three-step channel estimation method which can outperform the conventional CDP based designs. For the first step, we develop the frequency-domain sinc-function convolution (FSC) type filter not only to cancel noise components but also to estimate the channel frequency responses (CFRs) of virtual subcarriers for long preamble parts. In the second step, we employ the modified time-domain reliability test and frequency-domain interpolation (TRFI) scheme considering virtual subcarriers' CFRs. Then, the final FSC step is to further attenuate the noise components from CFRs obtained in the second step and to generate virtual subcarriers' CFRs being utilized at the next symbol time. Finally, the simulation results verify the efficiency of our proposed CE method.**

***Index Terms*—IEEE 802.11p, CE, CDP, CFR, TRFI**

I. Introduction

Much research has been widely discussed on cooperative intelligent transportation systems (C-ITS) to actively respond to traffic conditions through real-time intercommunication with surrounding vehicles and infrastructure while vehicles are moving [1]. To support vehicle radio (i.e., vehicle-to-everything (V2X)) communications, the IEEE 802.11p standard was developed to define the medium access control (MAC) and physical layer (PHY) of a WLAN [2]. Notice that the IEEE 802.11p standard is based on a modification of the frequency bandwidth of the IEEE 802.11a standard from 20 MHz to 10 MHz [2]. It means that IEEE 802.11p has only four pilot subcarriers during one orthogonal frequency division multiplexing (OFDM) symbol period, and only four pilot subcarriers is insufficient to accurately estimate the channel variation occurring in the frequency domain. Therefore, it is necessary to accurately estimate the rapidly time-varying channel in order to stably provide the traffic information to the moving vehicle. Therefore, various channel estimation (CE) techniques for IEEE 802.11p/Wireless Access in Vehicular Environments (WAVE) systems have been widely researched [3] [4] [5]. To improve performance of construct data pilot (CDP) based scheme [1], the authors in [3] and [4] have proposed the time-domain reliability test & frequency-domain interpolation (TRFI) scheme and the modified TRFI scheme with virtual subcarrier aided channel estimation method [5]. But, the performance gain is not sufficient yet for high computational complexity.

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Note that the authors in [4] and [5] assumed in estimating the channel frequency response (CFR) of virtual subcarriers that the channel length of channel impulse response (CIR) is perfectly known, or is $L_{\text{cp}} - 1$, where L_{cp} is the cyclic prefix length. In this paper, based on the works in [4] and [5], we propose another virtual subcarrier aided channel estimation method utilizing the estimated channel length obtained by the algorithms in [6] and [7]. The proposed method has the structure of frequency-domain sinc-function convolution (FSC) type filter.

The remainder of this paper is organized as follows: Section II describes the IEEE 802.11p physical layer. The proposed channel estimation scheme is described at Section III. Section IV shows simulation results, and concluding remarks are given in Section V.

Notations: Throughout this paper, normal letters represent scalar values, and boldface letters denote vectors or matrices, respectively. Also, diag $(\cdot)$ means to convert an input diagonal matrix to an output column vector or an input column vector to an output diagonal matrix. Note that $\mathbf{A}./\mathbf{B}$ indicates the element by element division between two vectors. In addition, fft $(\cdot)$ and c-shift$\left[\cdot\right]$ indicate the N-point fast fourier transform (FFT) and the $N/2$-point circular shift for an input $N \times 1$ vector, respectively.

II. System Model

IEEE 802.11p has been standardized by modifying some specifications in the physical layer of the existing IEEE 802.11a [2]. The IEEE 802.11p standard utilizes a frequency of 5.9GHz (5.850 GHz $\sim$ 5.925 GHz) and a bandwidth of 10MHz, which is half of 802.11a bandwidth. The size of FFT and inverse FFT (IFFT) is $N = 64$. With one OFDM symbol duration equal to $T_s = 6.4\mu s$, the transmitter employs the convolutional encoder and the receiver adopts the Viterbi decoder [2] [5].

The IEEE 802.11p packet is composed of three parts: the preamble for time-frequency synchronization, the signal field for control information, and the data field for message signals [2]. The preamble located at the beginning of the packet has short training symbols for time synchronization and two long training symbols for the initial channel estimation. Other specific parameters in IEEE 802.11p can be found in [2].

IEEE 802.11p physical layer is based on OFDM. The guard interval (GI) is inserted as the cyclic prefix so as to reduce the inter symbol interference (ISI) due to multipath fading channels. The signal field composed of one OFDM symbol has information such as modulation order, code rate, and so on. On the other hand, the data field contains data to be transmitted and the number of OFDM symbols in data field can be variable according to data size. In the data field, each OFDM symbol contains $N (= 64)$ subcarriers including 4 pilot subcarriers with an index set of $\mathbb{S}_p = \{-21, -7, 7, 21\}$, 12 virtual (null) subcarriers with an index set of $\mathbb{S}_v = \{-32, \cdots, -27, 0, 27, \cdots, 31\}$, and 48 data subcarriers with an index set of $\mathbb{S}_d$. Then, the number of useful subcarriers is 52 with an index set of $\mathbb{S}_u = \mathbb{S}_d \cup \mathbb{S}_p = \{-26, -25, \cdots, -1, 1, \cdots, 25, 26\}$ [4] [5] [8]. From here, the diagonal matrices $\mathbf{p}_p$, $\mathbf{p}_v$, $\mathbf{p}_d$, and $\mathbf{p}_u (= \mathbf{p}_p + \mathbf{p}_d)$ are defined so that the non-zero diagonal elements of the matrix indicates the positions of pilot subcarriers, virtual subcarriers, data subcarriers, and useful subcarriers, respectively. For example, $\mathbf{p}_p$ can be presented as

$$\mathbf{p}_p(c, l) = \begin{cases} 1, & \text{if} \quad c = l \in \mathbb{S}_p \\ 0, & \text{else.} \end{cases} \tag{1}$$

By removing GI and applying FFT to the received signal, the frequency-domain received signal $N \times 1$ vector $\mathbf{Y}_i$ of the ith OFDM symbol can be expressed as

$$\mathbf{Y}_i = \text{diag}(\mathbf{X}_i)\mathbf{H}_i + \mathbf{W}_i \tag{2}$$

with $Y_i(k)|_{k\in\mathbb{S}_u} = H_i(k)X_i(k) + W_i(k)$, $i \in \{1, 2, \cdots, M\}$, $Y_i(k)|_{k\in\mathbb{S}_v} = 0$, and M is the number of OFDM symbols in data field (i.e., packet). $\mathbf{H}_i$, $\mathbf{X}_i$, and $\mathbf{W}_i$ denote the CFR $N \times 1$ vector, the data symbol $N \times 1$ vector, and the additive Gaussian noise (AWGN) $N \times 1$ vector in which the element $W_i(k)$ has zero mean and the variance of σ^2.

III. Proposed Channel Estimation Scheme for the IEEE 802.11p

A. Initial Step (Long preamble Part)

At first, we can obtain the initial CFR by the least square (LS) method for two long preambles, $\mathbf{Y}_1^{\text{LP}}$ and $\mathbf{Y}_2^{\text{LP}}$, as follow:

$$\hat{\mathbf{H}}_0 = \frac{1}{2}\left(\mathbf{Y}_1^{\text{LP}} + \mathbf{Y}_2^{\text{LP}}\right) ./ \left(\text{diag}(\mathbf{p}_v) + \mathbf{p}_u\mathbf{X}^{\text{LT}}\right) \tag{3}$$

where $Y_1^{\text{LP}}(k)\big|_{k\in\mathbb{S}_v} = Y_2^{\text{LP}}(k)\big|_{k\in\mathbb{S}_v} = 0$, $\hat{H}_0(k)\Big|_{k\in\mathbb{S}_v} = 0$, $\hat{H}_0(k)\Big|_{k\in\mathbb{S}_u} = \frac{1}{2}\left(Y_1^{\text{LP}}(k) + Y_2^{\text{LP}}(k)\right)/X^{\text{LT}}(k)$, and $X^{\text{LT}}(k)$ is the long training symbol of the kth subcarrier known to the receiver.

From here, let us $\hat{L}$ define as the estimated channel length (i.e., maximum access delay time) in [6] or [7] for one packet having M OFDM symbols. The frequency domain filter gain can be obtained as the complex sinc-function type of

$$\mathbf{SINC}_f = \text{c-shift}\big[\text{fft}(\mathbf{rect}_t)/N\big] \tag{4}$$

with

$$\mathbf{rect}_t(l) = \begin{cases} 1, & \text{if} \quad 1 \le l \le \hat{L} \\ 0, & \text{else.} \end{cases} \tag{5}$$

The sum of filter coefficients at the given position can be expressed as

$$\mathbf{w}_{\text{sum}} = \text{c-shift}\big[\text{diag}\{\mathbf{p}_u\} \circledast \mathbf{SINC}_f\big] \tag{6}$$

where $\circledast$ denotes the circular convolution with the size of N.

The key ingredient of our method is to estimate the CFRs corresponding to the virtual subcarriers as well as the useful subcarriers. To this end, we initially estimate the CFR vector as

$$\hat{\mathbf{H}}_0^{\text{FSC}} = \text{c-shift}\big[\hat{\mathbf{H}}_0 \circledast \mathbf{SINC}_f\big]./\mathbf{w}_{\text{sum}}. \tag{7}$$

Notice that $\hat{\mathbf{H}}_0$ of (3) has zero at virtual subcarrier position. In (7), the circular convolution performs the operation of extrapolation to virtual subcarriers and $(\cdot)./\mathbf{w}_{\text{sum}}$ means the weight adjustment considering virtual subcarriers in terms of weighted sum.

B. Modified TRFI Step

1) Constructing Data Pilot: The received signal of the ith data OFDM symbol $\mathbf{Y}_i$ is equalized using the $(i-1)$th estimated channel value $\hat{\mathbf{H}}_{i-1}^{\text{FSC}}$ and then, by demapping, the constructed data pilot can be expressed as

$$\hat{\mathbf{X}}_i = \mathbf{p}_d D\left(\mathbf{Y}_i./\hat{\mathbf{H}}_{i-1}^{\text{FSC}}\right) + \mathbf{p}_p\mathbf{X}_i^P \tag{8}$$

where $D(\cdot)$ is a function that maps the equalized signal with regard to the corresponding modulation order [1]. Note in (8) that $\hat{X}_i(k)\Big|_{k\in\mathbb{S}_d} = D\left(Y_i(k)/\hat{H}_{i-1}^{\text{FSC}}(k)\right)$ and $\hat{X}_i(k)\Big|_{k\in\mathbb{S}_p} = X_i^P(k)$ which is a predefined frequency domain pilot symbol of the kth subcarrier in the ith OFDM symbol.

2) Instantaneous CFR by LS Method: The instantaneous channel coefficient $\hat{\mathbf{H}}_i$ can be expressed by equalizing $\mathbf{Y}_i$ with $\hat{\mathbf{X}}_i$ of (8) as

$$\hat{\mathbf{H}}_i = \mathbf{Y}_i./\left(\text{diag}(\mathbf{p}_v) + \mathbf{p}_u\hat{\mathbf{X}}_i\right) \tag{9}$$

with $\hat{Y}_i(k)\Big|_{k\in\mathbb{S}_v} = 0$, $\hat{H}_i(k)\Big|_{k\in\mathbb{S}_v} = 0$, and $\hat{H}_i(k)\Big|_{k\in\mathbb{S}_u} = Y_i(k)/\hat{X}_i(k)$.

3) Reliability Test: From the estimated instantaneous CFR $\hat{\mathbf{H}}_i$ of (9) and the $(i-1)$th OFDM symbol's CFR $\hat{\mathbf{H}}_{i-1}^{\mathrm{FSC}}$ of (7) , we can equalize the previous received signal $\mathbf{Y}_{i-1}$ and then, two symbols for the reliability test can be demapped as

$$\begin{aligned}\hat{\mathbf{X}}' &= \mathbf{p}_d D\left(\mathbf{Y}_{i-1}./\left(\mathrm{diag}\left(\mathbf{p}_v\right)+\hat{\mathbf{H}}_i\right)\right)\\ \hat{\mathbf{X}}'' &= \mathbf{p}_d D\left(\mathbf{Y}_{i-1}./\hat{\mathbf{H}}_{i-1}^{\mathrm{FSC}}\right)\end{aligned} \tag{10}$$

with $X'(k)|_{k\in\mathbb{S}_d} = D\left(Y_{i-1}(k)/\hat{H}_i(k)\right)$, $X''(k)|_{k\in\mathbb{S}_d} = D\left(Y_{i-1}(k)/\hat{H}_{i-1}^{\mathrm{FSC}}(k)\right)$, and $X'(k)|_{k\in\mathbb{S}_v} = X''(k)|_{k\in\mathbb{S}_v} = 0$ [1]. If $\hat{H}_i(k)$ is correctly estimated, $\hat{H}_i(k)$ and $\hat{H}_{i-1}^{\mathrm{FSC}}(k)$ can be similar because of the highly correlated characteristics for two adjacent subcarriers in the time domain. Using this relationship, the unreliable channel value in $\hat{\mathbf{H}}_i$ of (9), which will be estimated by interpolation in the next step, can be initialized as

$$\hat{H}_i(k)\Big|_{k\in\mathbb{S}_d} = 0 \ \text{ if } \ X'(k) \neq X''(k). \tag{11}$$

4) Frequency Interpolation: Once the reliability test step is completed, we perform the frequency domain 64-point interpolation to estimate the CFRs in the remaining subcarriers of (11). Unlike the conventional TRFI scheme (i.e., 52-point interpolation in [3]), we leverage the virtual CFRs estimated in the previous OFDM symbol for $k \in \mathbb{S}_v$ to improve the interpolation accuracy. We can obtain the estimated CFRs in the current ith OFDM symbol as [5]

$$\bar{\mathbf{H}}_i = \mathrm{interpo.}\left\{\mathbf{p}_v\hat{\mathbf{H}}_{i-1}^{\mathrm{FSC}} + \mathbf{p}_u\hat{\mathbf{H}}_i\right\}. \tag{12}$$

Notice that when some CFR information is determined to be unreliable near the edge of data subcarriers, the CE performance can be improved by performing interpolation considering both directions (i.e., the estimated CFR values to be retaining the property of circular symmetry) in the proposed method, unlike the existing scheme that interpolates in one direction [4].

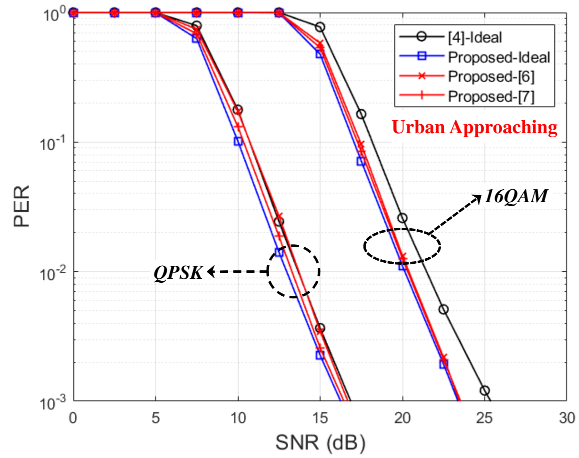
C. FSC (Noise Attenuation) Step

In this step, we attenuate the noise components from the noisy CFRs in (12). To do this, we can perform the circular convolution on $\bar{\mathbf{H}}_i$ and $\mathbf{SINC}_f$ of (4) to obtain

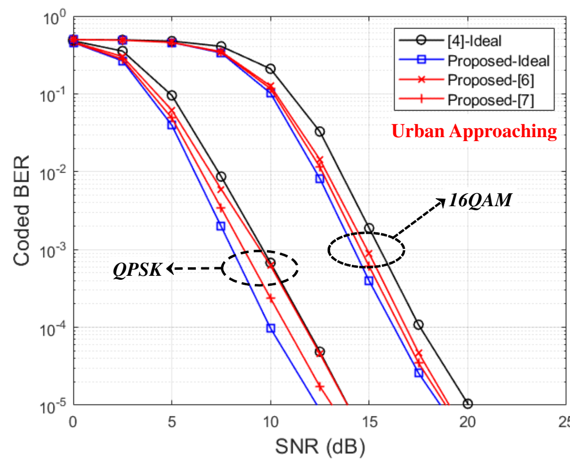
$$\hat{\mathbf{H}}_i^{\mathrm{FSC}} = \text{c-shift}\left[\bar{\mathbf{H}}_i \circledast \mathbf{SINC}_f\right]. \tag{13}$$

Let us compare (13) to (7). $\hat{\mathbf{H}}_0$ in (7) has zero virtual subcarrier component and the virtual subcarrier component of $\hat{\mathbf{H}}_0^{\mathrm{FSC}}$ is estimated by extrapolation of (7). But, $\bar{\mathbf{H}}_i$ in (13) has the virtual subcarrier component $\mathbf{p}_v\hat{\mathbf{H}}_{i-1}^{\mathrm{FSC}}$ determined in the previous OFDM symbol time as shown in (12) and then, the virtual subcarrier component of $\hat{\mathbf{H}}_i^{\mathrm{FSC}}$ is newly obtained and then, can be used at the next time CE process.

For the next $(i+1)$th data field, we can go to the step of (8). The process is repeated until we arrive at the Mth OFDM symbol.



(a) PER vs. SNR(dB)



(b) BER vs. SNR(dB)

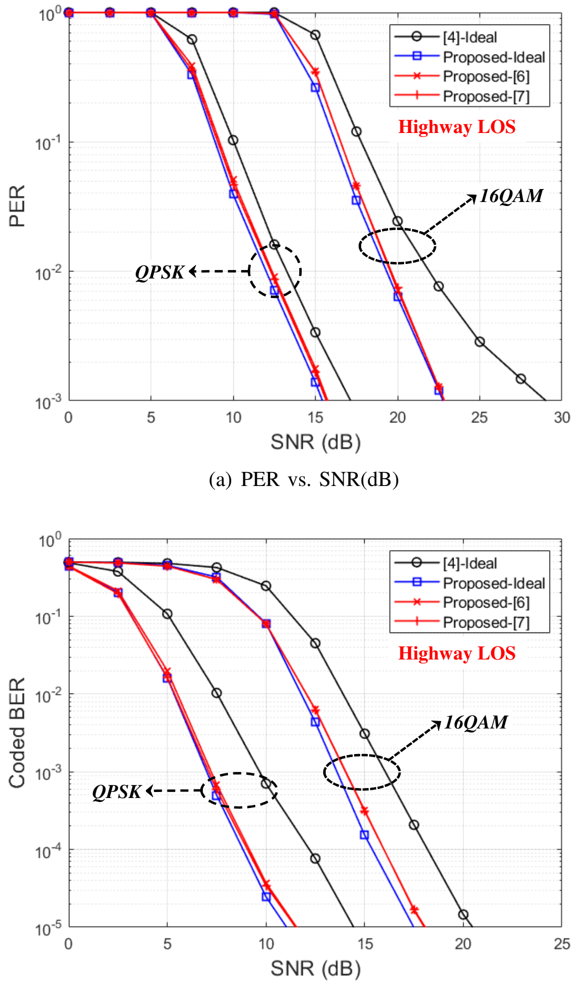
Fig. 1. Error Performance Comparison at 'Urban Approaching with 119km/h' (QPSK, 16QAM, CR=1/2, $M = 100$).

IV. Simulation Results

In this section, we present simulation results to verify the error performance of the proposed scheme based on the IEEE 802.11p standard with $N = 64$, $L_{\mathrm{cp}} = 16$, and $T_s = 0.1\mu s$ [2] [8]. We assume that one packet consists of $M\ (=100)$ OFDM symbols, quaternary phase shift keying (QPSK) and 16-QAM with coding rate of 1/2. For all cases, it is averaged over 10^5 packet transmissions. Among five scenarios of 'CohdaWireless V2V channel model' in [9], we consider 'Urban Approaching LOS with 119km/h', 'Street Crossing NLOS with 126km/h', 'Highway LOS with 252km/h', and 'Highway NLOS with 252km/h' channel environments. The other parameters such as the delay time, the relative power, and the Doppler spectrum for each channel tap are listed in [9].

From here, 'X$-$Ideal', 'X$-$[6]', and 'X$-$[7]' represent the estimated channel length $\hat{L}$ to be ideal channel length and to be obtained by the algorithms in [6] and [7], respectively.

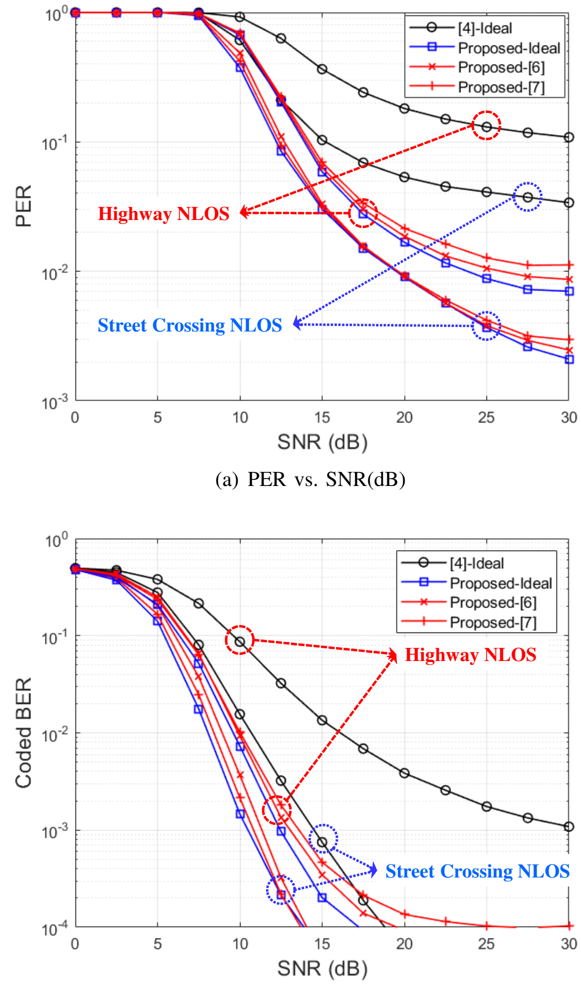
Fig. 1(a) and Fig. 1(b) show PER and BER performance



(a) PER vs. SNR(dB)

(b) BER vs. SNR(dB)

Fig. 2. Error Performance Comparison at 'Highway LOS with 252km/h' (QPSK, 16QAM, CR=1/2, $M = 100$).



(a) PER vs. SNR(dB)

(b) BER vs. SNR(dB)

Fig. 3. Error Performance Comparison at 'Highway NLOS with 252km/h' & 'Street Crossing NLOS with 126km/h' (QPSK, 16QAM, CR=1/2, $M = 100$).

comparison with respect to CE schemes under 'Urban Approaching with 119km/h' for QPSK and 16-QAM, respectively. Fig. 2(a) and Fig. 2(b) show PER and BER performance comparison with respect to CE methods under 'Highway LOS with 252km/h' for QPSK and 16-QAM, respectively. For QPSK, Fig. 3(a) and Fig. 3(b) show PER and BER performance comparison with respect to CE schemes under both 'Street Crossing NLOS with 126km/h' and 'Highway NLOS with 252km/h', respectively. From three figures, it can be confirmed that the proposed method using the channel length $\hat{L}$ estimated by [6] and [7] exhibits better error rate performance than '[4]$-$Ideal' which denotes the CE scheme of [4] with the perfectly estimated channel length. In addition, the proposed method using the estimated $\hat{L}$ presents performance similar to that when using the perfect channel length. Furthermore, from Fig. 3, it is verified that the proposed methods can achieve PER $= 10^{-2}$ in a reasonable SNR area even at severe channel environment of 'Street Crossing NLOS' and 'Highway NLOS'.

V. Conclusions

In this paper, we have studied a novel channel estimation scheme that outperforms the conventional design of [4] with the ideal channel length value over all practical SNR range. To this end, we developed three-step channel estimation process. In the first step, we proposed the noise attenuating and virtual subcarrier estimating scheme for long preamble parts, which has the complex sinc-function type frequency domain filter structure. Then, in the second step, we adopt the modified TRFI scheme with 64-point interpolation using the CFRs of virtual subcarriers obtained in the previous OFDM symbol to attenuate the noise components from the instantaneously estimated noisy CFRs. The final is the FSC step to further remove the noise components from CFRs obtained after the modified TRFI. Finally, the simulation results verify the efficiency of our proposed method.

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