Joint Time-Frequency Channel Estimation for Time Domain Synchronous OFDM Systems

Linglong Dai, Zhaocheng Wang, Jun Wang, and Zhixing Yang

***Abstract*—As the key technology of Chinese national digital television terrestrial broadcasting (DTTB) standard, time domain synchronous OFDM (TDS-OFDM) has high spectral efficiency at the cost of performance loss over fast fading channels. In this paper, a novel time-frequency signal structure is proposed to improve the system performance of TDS-OFDM over fast time-varying channels. Each TDS-OFDM symbol adopts the time-domain one-sample shifted training sequence (TS) and the frequency-domain grouped pilots as the time-frequency training information. The modulable orthogonal sequence (MOS) with perfect autocorrelation property is adopted as the TS, while each pilot group has only one non-zero central pilot surrounded by several zero pilots. The corresponding joint time-frequency channel estimation utilizes the time-domain TS for path delay estimation and the frequency-domain grouped pilots for path gain estimation, thus accurate tracking of the fast time-varying wireless channel could be achieved. Only about 1% of the total subcarriers will be occupied by the redundant grouped pilots, thus the loss in spectral efficiency is negligible. Simulation results demonstrate that the proposed scheme outperforms the conventional solutions over fast fading channels.**

***Index Terms*—Digital television terrestrial broadcasting (DTTB), fast fading channel, joint time-frequency channel estimation, time domain synchronous orthogonal frequency division multiplexing (TDS-OFDM), time-frequency training.**

I. Introduction

ORTHOGONAL frequency division multiplexing (OFDM) has received extensive attention for digital television terrestrial broadcasting (DTTB) systems due to its robustness to frequency-selective multipath channels. Basically, there are three types of block transmission schemes for OFDM systems: cyclic prefix OFDM (CP-OFDM) [1], zero padding OFDM (ZP-OFDM) [2], and time domain synchronous OFDM (TDS-OFDM) [3]. The widely used CP-OFDM scheme utilizes the CP as the guard interval to eliminate the inter-block-interference (IBI) as well as inter-carrier-interference (ICI) [4], [5]. ZP-OFDM replaces CP by zero padding to deal with the channel null problem [2]. In TDS-OFDM, the known sequence like the pseudorandom noise (PN) sequence is used as the guard interval as well as the training sequence (TS) for synchronization and channel estimation [3]. Consequently, TDS-OFDM could achieve higher spectral efficiency and faster synchronization [3]. As the key technology, TDS-OFDM has been successfully applied in Chinese national DTTB standard called digital television terrestrial multimedia broadcasting (DTMB) [3].

On the other hand, in TDS-OFDM, the IBI between the TS and the OFDM data block has to be cancelled by the iterative padding subtraction (IPS) method which involves the mutually conditional channel estimation and channel equalization [6]. As a result, TDS-OFDM suffers from performance loss even when the multipath channel is stationary. The unique word OFDM (UW-OFDM) scheme [7] could partly solve this problem by generating the time-domain TS via the redundant frequency-domain scattered pilots at the transmitter, but the equivalent signal-to-noise ratio (SNR) at the receiver is reduced due to the high average power required by those pilots [7]. The most attractive solution to the interference problem of TDS-OFDM is the dual PN padding (DPNP) scheme [8], whereby two identical PN sequences are used for simpler yet more reliable channel estimation. Due to its low complexity and reliable performance, the TDS-OFDM with DPNP is the most promising candidate of frame structure for Chinese next-generation DTTB standard [3], [9]. However, the DPNP-aided TDS-OFDM still assumes time-invariant channel within each TDS-OFDM symbol, thus the performance loss over fast fading channels is unavoidable.

In this paper, to improve the system performance of TDS-OFDM over fast time-varying channels, we propose a novel time-frequency signal structure and the corresponding joint time-frequency channel estimation method to accurately track the fast variation of the wireless fading channel. Specifically, the innovations and contributions of this paper are generalized in the following three aspects: 1) Unlike that the constant PN sequence is the training information only in the time domain for DPNP-aided TDS-OFDM, the novel time-frequency signal structure has training information in both the time and frequency domains, whereby the time-domain one-sample shifted TS with perfect autocorrelation and the frequency-domain grouped pilots with one non-zero element placed in the middle are used. 2) Based on the time-frequency signal structure, the proposed joint time-frequency channel estimation utilizes the time-domain TS for channel path delays estimation, while the path gains are estimated by the frequency-domain grouped pilots, thus accurate tracking of the fast fading channel could be achieved; 3) The proposed scheme outperforms the conventional solutions over fast fading channels at the cost of negligible loss in spectral efficiency, since it only requires that the grouped pilots occupy about 1% of the total subcarriers.

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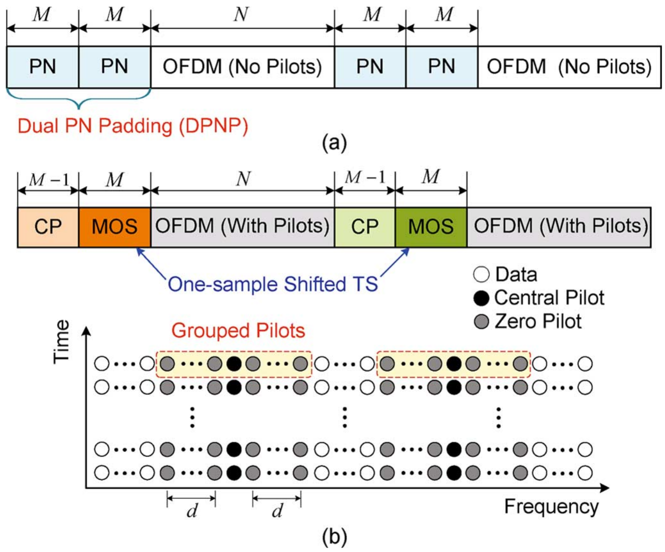

Fig. 1. Signal structure comparison: (a) TDS-OFDM with dual PN padding (DPNP) [8]; (b) proposed time-frequency signal structure for TDS-OFDM.

The remainder of this paper is organized as follows. The system model based on the proposed time-frequency signal structure is described in Section II. The corresponding joint time-frequency channel estimation is presented in Section III. Section IV addressed the performance analysis. Simulation results are shown in Section V. Finally, conclusions are drawn in Section VI.

Notation: The boldface letters are used to denote matrices and vectors. $\mathbf{F}_N$ is the $N \times N$ fast Fourier transform (FFT) matrix with the $(n+1, k+1)$th entry $\exp(-j2\pi nk/N)/\sqrt{N}$. $\mathbf{I}_N$ is the $N \times N$ identity matrix and $\mathbf{0}_{M\times N}$ is the $M \times N$ zero matrix. $\otimes$ means the circular correlation. The superscripts $(\cdot)^{\mathrm{T}}$, $(\cdot)^H$, $(\cdot)^{-1}$, and $(\cdot)^{\dagger}$ denote the transpose, conjugate transpose, matrix inversion, and Moore-Penrose matrix inversion operators, respectively. $\widehat{x}$ means the estimate of x, and $|x|$ denotes the absolute value of x. Finally, $\mathrm{diag}\{\mathbf{u}\}$ is a diagonal matrix with $\mathbf{u}$ at its main diagonal.

II. System Model

In this section, the proposed time-frequency signal structure for TDS-OFDM systems is introduced at first, and then the system model over fast fading channels is presented.

A. Time-Frequency Signal Structure

To improve the system performance over fast fading channels, a novel signal structure for TDS-OFDM as illustrated in Fig. 1 is proposed. For comparison, the conventional TDS-OFDM with DPNP is also presented. It is clear that the proposed signal structure differs from the conventional one in two aspects: the time-domain guard interval and frequency-domain subcarriers.

In the time domain, unlike DPNP where two identical PN sequences are used as the guard interval of the following OFDM data block, the ith TDS-OFDM symbol $\mathbf{s}_i = [s_{i,0}, s_{i,1}, \cdots, s_{i,P-1}]^T$ in the proposed signal structure is composed of the known guard interval $\mathbf{g}_i = [c_{i,1}, \cdots, c_{i,M-1}, c_{i,0}, c_{i,1}, \cdots, c_{i,M-1}]^T$ of length $2M-1$ and the OFDM data block $\mathbf{x}_i = [x_{i,0}, x_{i,1}, \cdots, x_{i,N-1}]^T$ of length N. The guard interval $\mathbf{g}_i$ is composed of the TS $\mathbf{c}_i = [c_{i,0}, c_{i,1}, \cdots, c_{i,M-1}]^T$ with the length of M and the corresponding cyclic prefix $[c_{i,1}, \cdots, c_{i,M-1}]^T$ with the length of $M-1$. The time-domain OFDM data block can be also presented by $\mathbf{x}_i = \mathbf{F}_N^H \mathbf{X}_i$, whereby $\mathbf{X}_i = [X_{i,0}, X_{i,1}, \cdots, X_{i,N-1}]^T$ denotes the FFT of $\mathbf{x}_i$. Note that $P = N + 2M - 1$ is the length of one TDS-OFDM symbol.

The TS $\mathbf{c}_{i+1}$ for the $(i+1)$th TDS-OFDM symbol is generated by one-sample cyclically shifting the TS $\mathbf{c}_i$ for the ith TDS-OFDM symbol to the right as below

$$\mathbf{c}_{i+1} = \begin{bmatrix} \mathbf{0}_{1\times(M-1)} & 1 \\ \mathbf{I}_{M-1} & \mathbf{0}_{(M-1)\times 1} \end{bmatrix} \mathbf{c}_i. \tag{1}$$

Therefore, non-constant TS will be used in the proposed scheme, since it has been proved in [10] that the constant TS is not optimal for channel tracking. On the other hand, we can derive from (1) that the guard interval $\mathbf{g}_{i+1} = [c_{i,0}, c_{i,1}, \cdots, c_{i,M-2}, c_{i,M-1}, c_{i,0}, c_{i,1}, \cdots, c_{i,M-2}]^T$, which indicates that the first M samples of $\mathbf{g}_{i+1}$ is identical with the last M samples of $\mathbf{g}_i$. That is to say, the one-sample shifted TS above could also preserve the cyclic property of the transmitted data stream when constant PN sequence is used for every TDS-OFDM symbol in [8]. Thus, the low-complexity equalization facilitated by the constant TS in DPNP could still be achieved both for multi-carrier and single-carrier transmission [11]. In addition, the cyclic property is also useful for accurate timing/frequency synchronization [12].

The PN sequence with non-ideal autocorrelation property in conventional TDS-OFDM systems is not optimal for channel estimation [13], [14]. Instead, the modulable orthogonal sequence (MOS) [15] will be adopted as the TS

$$c_{i,n} = b(n_1)\exp\left(\frac{2\pi}{\sqrt{M}} m n_0 n_1\right), \quad 0 \le n \le M-1, \tag{2}$$

where $0 \le n_0 \le \sqrt{M}-1$, $0 \le n_1 \le \sqrt{M}-1$, $n = n_0\sqrt{M} + n_1$, m is relatively prime to $\sqrt{M}$, and $|b(n_1)| = 1$. In this paper, $m = 1$ and $b(n_1) = 1$ are used for simplicity. The MOS has perfect autocorrelation property denoted by

$$\mathbf{c}_i \otimes \mathbf{c}_i = M \begin{bmatrix} 1 & \mathbf{0}_{1\times(M-1)} \end{bmatrix}^T. \tag{3}$$

The MOS also achieves the theoretical lower bound of the cross-correlation $1/\sqrt{M}$ [15]. Moreover, the MOS has the lowest peak-to-average power ratio (PAPR) of 0 dB due to its constant envelope in the time domain.

In the frequency domain, as illustrated in Fig. 1, unlike the conventional TDS-OFDM scheme where all active subcarriers are used to carry data and no pilot is adopted [6], [8], [9], the proposed signal structure uses N_d data subcarriers and N_{group} grouped pilots randomly scattered within the signal bandwidth. Each pilot group occupies $2d+1$ subcarriers, whereby only one non-zero central pilot is located in the middle, and there are $2d$ zero pilots around the central pilot. The $2d$ zero pilots are used to alleviate the potential ICI imposed on the central pilot [16]. The index set of the central pilots can be denoted by $\mathbf{P} = \{p_0, p_1, \cdots, p_{N_{group}-1}\}$. So the grouped pilots will occupy $N_p = N_{group}(2d+1)$ subcarriers, and $N = N_d + N_p$. We will show later in Section IV-A that N_p is very small compared with N.

B. System Model Over Fast Fading Channels

After cyclic prefix reconstruction (the add-subtract method used in [8] is still effective for cyclic prefix reconstruction due to the cyclic property contributed by the one-sample shifted TSs denoted by (1), or the hybrid domain cyclic prefix reconstruction method in [17] can be also used), the received time-domain OFDM block $\mathbf{y}_i = [y_{i,0}, y_{i,1}, \cdots, y_{i,N-1}]^T$ has the elements

$$y_{i,n} = \sum_{l=0}^{L-1} h_{i,n,l} x_{i,(n-n_l)_N} + w_{i,n}, \tag{4}$$

where $w_{i,n}$ is the additive white Gaussian noise (AWGN) with zero mean and the variance of σ^2, $h_{i,n,l}$ denotes the path gain of the lth non-zero path with the delay of n_l at the time instant of n within the ith OFDM block ($n_0 = 0$ is assumed in this paper without loss of generality), L is the number of resolvable paths. To avoid IBI, the maximum channel length n_{L-1} is assumed to be smaller than the TS length M, i.e., $n_{L-1} < M$. Note that the number of resolvable paths L is usually much smaller than the maximum channel length n_L due to the nature of wireless channels, i.e., $L \ll n_{L-1}$ [17], [18].

After FFT at the receiver, the frequency-domain OFDM block $\mathbf{Y}_i = [Y_{i,0}, Y_{i,1}, \cdots, Y_{i,N-1}]^T$ is [5]

$$\begin{aligned} Y_{i,k} &= \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} y_{i,n} e^{-j\frac{2\pi}{N}nk} \\ &= \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} \left(\sum_{l=0}^{L-1} h_{i,n,l} x_{i,(n-n_l)} + w_{i,n} \right) e^{-j\frac{2\pi}{N}nk} \\ &= X_{i,k} H_{i,k,k} + \underbrace{\sum_{q=0, q\neq k}^{N-1} X_{i,q} H_{i,k,q}}_{\text{ICI}} + W_{i,k}, \end{aligned} \tag{5}$$

where $W_{i,k} = (1/\sqrt{N}) \sum_{n=0}^{N-1} w_{i,n} e^{-j(2\pi/N)nk}$ is the noise term, and

$$H_{i,k,q} = \frac{1}{N} \sum_{l=0}^{L-1} \left(\sum_{n=0}^{N-1} h_{i,n,l} e^{-j\frac{2\pi}{N}n(k-q)} \right) e^{-j\frac{2\pi}{N}qn_l}. \tag{6}$$

The conventional TDS-OFDM systems [6], [8], [9] all assume the time-invariant channel within each TDS-OFDM symbol, i.e., $h_{i,0,l} = h_{i,0,l} = \cdots = h_{i,N-1,l} = h_{i,l}$ $(0 \le l \le L-1)$, where $h_{i,l}$ presents the averaged path gain of the lth path, i.e., $h_{i,l} = (1/N) \sum_{n=0}^{N-1} h_{i,n,l}$. Then, the ICI term $H_{i,k,q}$ $(q \neq k)$ becomes zero, i.e., $H_{i,k,q} = 0$, and the signal model (5) is consequently simplified as

$$Y_{i,k} = X_{i,k} H_{i,k} + W_{i,k}. \tag{7}$$

where $H_{i,k} = H_{i,k,k}$.

However, the assumption of time-invariant channel within each TDS-OFDM symbol would inevitably lead to the performance loss over fast fading channels due to the obvious deviation of the simple system model (7) compared with the actual one (5). This problem can be solved in the following section to improve the system performance of TDS-OFDM.

III. Joint Time-Frequency Channel Estimation

Unlike the conventional TDS-OFDM schemes where both the path delays and the path gains are estimated by the time-domain TS, in this section we propose the joint time-frequency channel estimation of the fast fading channels, namely, the path delay estimation via time-domain TS and the path gain estimation via frequency-domain grouped pilots.

A. Path Delay Estimation Using Time-Domain TS

Due to the protection of the CP in the guard interval, the received TS $\mathbf{d}_i = [d_{i,0}, d_{i,1}, \cdots, d_{i,M-1}]^T$ is immune to the IBI caused by the preceding OFDM data block, so $\mathbf{d}_i$ is the circular convolution between the transmitted TS and the channel plus the noise

$$\mathbf{d}_i = \mathbf{c}_i^{(s)} \otimes \mathbf{h}_i + \mathbf{v}_i, \tag{8}$$

where $\mathbf{c}_i^{(s)} = [c_{i,0}, c_{i,M-1}, c_{i,M-2}, \cdots, c_{i,1}]^T$ is the rearranged vector based on $\mathbf{c}_i$ according to the relationship between the circular correlation and circular convolution, $\mathbf{v}_i = [v_{i,0}, v_{i,1}, \cdots, v_{i,M-1}]^T$ denotes the channel's AWGN vector whose every element has zero mean and the variance of σ^2, $\mathbf{h}_i = [h_{i,0}, \underbrace{0, \cdots, 0}_{n_1 - n_0 - 1}, h_{i,1}, 0, \cdots, 0, h_{i,L-1}, \underbrace{0, \cdots, 0}_{M - n_{L-1} - 1}]^T$ denotes the $M \times 1$ zero padded channel impulse response (CIR) during the TS. Note that the time-invariant channel model is used here for simplicity, since only the channel delays are to be estimated in this stage.

The rough channel estimate $\widehat{\mathbf{h}}_i$ can be obtained by the circular correlation between the local sequence $\mathbf{c}_i^{(s)}$ and the received TS $\mathbf{d}_i$ as

$$\begin{aligned} \widehat{\mathbf{h}}_i &= \frac{1}{M} \mathbf{c}_i^{(s)} \otimes \mathbf{d}_i = \frac{1}{M} \mathbf{c}_i^{(s)} \otimes \left(\mathbf{c}_i^{(s)} \otimes \mathbf{h}_i + \mathbf{v}_i \right) \\ &= \mathbf{h}_i + \frac{1}{M} \mathbf{c}_i^{(s)} \otimes \mathbf{v}_i, \end{aligned} \tag{9}$$

where the perfect autocorrelation property of the TS given by (3) has been utilized. Then, the time delays of the L most significant taps of $\widehat{\mathbf{h}}_i$ are saved in the path delay set $\mathbf{D}$ as

$$\mathbf{D} = \left\{ n_l : \left| \widehat{h}_{i,n_l} \right|^2 \ge T_{th} \right\}_{l=0}^{L-1}, \tag{10}$$

where T_{th} is the pre-defined power threshold.

After the path delays $\{n_l\}_{l=0}^{L-1}$ have been obtained in (10), the remained work for the complete channel estimation is to estimate the path gains $h_{i,n,l}$ in the next subsection.

B. Path Gain Estimation Using Frequency-Domain Grouped Pilots

According to [18], the path gain $h_{i,n,l}$ in (6) of fast time-varying channels can be modeled by the Q-order Taylor series expansion as

$$h_{i,n,l} = \sum_{v=0}^{Q} b_{i,n,v} \gamma_{i,l,v} + \xi_{i,n,l} = \mathbf{b}_n \boldsymbol{\gamma}_{i,l} + \xi_{i,n,l}, \tag{11}$$

where $\mathbf{b}_n = [b_{i,n,0}, b_{i,n,1}, \cdots, b_{i,n,Q}]_{1 \times (Q+1)}$ denotes the basis function, $\boldsymbol{\gamma}_{i,l} = [\gamma_{i,l,0}, \gamma_{i,l,1}, \cdots, \gamma_{i,l,Q}]^T_{(Q+1) \times 1}$ with the entry

$\gamma_{i,l,v}$ being the vth polynomial coefficient, and $\xi_{i,n,l}$ presents the approximation error. The approximation order Q depends on the maximum Doppler spread f_d of the channel. For fast fading channels, $Q = 1$ ensures good approximation performance if the normalized Doppler spread $f_d T \leq 0.1$, where T is the OFDM block duration [18]. Take the typical value $T = 500\ \mu$s specified in [3] as an example, $f_d \leq 200$ Hz meets such criterion. In addition, $b_{i,n,v} = n^v$ provides sufficient approximation accuracy of the path gain of the channels subject to the Jake's power spectrum [18].

Since ICI is dominantly caused by the neighboring subcarriers [4], [5], it can be assumed that the ICI coefficient $H_{i,k,q} = 0$ if $|k-q| > d$ (that's the reason why the $2d$ zero pilots are configured within each pilot group as illustrated in Fig. 1). Thus, by substituting (11) and (6) into (5), $Y_{i,k}$ becomes

$$\begin{aligned} Y_{i,k} &= \sum_{q=k-d}^{k+d} H_{i,k,q} X_{i,q} + W_{i,k} \\ &= \sum_{q=k-d}^{k+d} \left[\frac{1}{N} \sum_{l=0}^{L-1} \left(\sum_{n=0}^{N-1} h_{i,n,l} e^{-j\frac{2\pi}{N} n(k-q)} \right) e^{-j\frac{2\pi}{N} q n_l} \right] \\ &\quad \times X_{i,q} + W_{i,k} \\ &= \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} \sum_{v=0}^{Q} \gamma_{i,l,v} b_{i,n,v} \mu_{i,n,l,k} + \varepsilon_{i,k}, \end{aligned} \tag{12}$$

where $\varepsilon_{i,k} = \sum_{n=0}^{N-1} \sum_{l=0}^{L-1} \xi_{i,n,l} \mu_{i,n,l,k} + W_{i,k}$, and

$$\mu_{i,n,l,k} = \frac{1}{N} \sum_{q=k-d}^{k+d} e^{-j\frac{2\pi}{N} n(k-q)} e^{-j\frac{2\pi}{N} q n_l} X_{i,q}. \tag{13}$$

By denoting

$$\begin{aligned} \boldsymbol{\mu}_{i,k} &= [\mu_{i,0,0,k}, \cdots, \mu_{i,0,L-1,k}, \mu_{i,1,0,k}, \cdots, \\ &\qquad \mu_{i,N-1,L-1,k}]_{1 \times LN}, \\ \mathbf{b}_i &= \left[\mathbf{b}_{i,0}^T, \mathbf{b}_{i,1}^T, \cdots, \mathbf{b}_{i,N-1}^T \right]_{LN \times (Q+1)L}^T, \\ \mathbf{b}_{i,n} &= [\mathrm{diag}\{\mathbf{b}_n, \mathbf{b}_n, \cdots, \mathbf{b}_n\}]_{L \times (Q+1)L}, \\ \boldsymbol{\gamma}_i &= \left[\boldsymbol{\gamma}_{i,0}^T, \boldsymbol{\gamma}_{i,1}^T, \cdots, \boldsymbol{\gamma}_{i,L-1}^T \right]_{(Q+1)L \times 1}^T, \end{aligned} \tag{14}$$

(12) can be also expressed in a more compact matrix form as

$$Y_{i,k} = \boldsymbol{\mu}_{i,k} \mathbf{b}_i \boldsymbol{\gamma}_i + \varepsilon_{i,k}. \tag{15}$$

If the subcarrier index k belongs to the central pilots set $\mathbf{P}$, i.e., $k \in \mathbf{P} = \{p_0, p_1, \cdots, p_{N_{group}-1}\}$, $\mu_{i,n,l,k}$ in (13) can be calculated according to the known frequency-domain grouped pilots $\{X_{i,q}\}_{q=k-d}^{k+d}$ as well as the path delay set $\mathbf{D} = \{n_l\}_{l=0}^{L-1}$ obtained in (10), and then $\boldsymbol{\mu}_i = [\boldsymbol{\mu}_{i,p_0}, \boldsymbol{\mu}_{i,p_1}, \cdots, \boldsymbol{\mu}_{i,p_{N_{group}-1}}]_{N_{group} \times LN}^T$ is consequently known. Therefore, the received central pilots $\mathbf{Y}_p = [Y_{i,p_0}, Y_{i,p_1}, \cdots, Y_{i,p_{N_{group}-1}}]_{N_{group} \times 1}^T$ can be expressed by

$$\mathbf{Y}_p = \boldsymbol{\mu}_i \mathbf{b}_i \boldsymbol{\gamma}_i + \boldsymbol{\varepsilon}_i, \tag{16}$$

where $\boldsymbol{\varepsilon}_i = [\varepsilon_{i,p_0}, \varepsilon_{i,p_1}, \cdots, \varepsilon_{i,p_{N_{group}-1}}]_{N_{group} \times 1}^T$.

Since there are $(Q+1)L$ unknown parameters in $\boldsymbol{\gamma}_i$ and N_{group} observations in $\mathbf{Y}_p$, $N_{group} \geq (Q+1)L$ is required to guarantee the matrix $\boldsymbol{\mu}_i \mathbf{b}_i$ to have the full column rank. Then, the least square (LS) estimation of $\boldsymbol{\gamma}_i$ can be obtained by

$$\widehat{\boldsymbol{\gamma}}_i = (\boldsymbol{\mu}_i \mathbf{b}_i)^{\dagger} \mathbf{Y}_p = \left[(\boldsymbol{\mu}_i \mathbf{b}_i)^H \boldsymbol{\mu}_i \mathbf{b}_i \right]^{-1} (\boldsymbol{\mu}_i \mathbf{b}_i)^H \mathbf{Y}_p. \tag{17}$$

Consequently, $h_{i,n,l}$ in (11) is known, and the channel frequency response $H_{i,k,q}$ in (6) can be calculated based on the path gains $h_{i,n,l}$ in (11) and the path delays n_l in (10).

IV. Performance Evaluation

This section compares the spectral efficiency and computational complexity of the proposed scheme with those of the conventional DPNP scheme for TDS-OFDM systems.

A. Spectral Efficiency

The Karhunen-Loeve theorem [1] requires that the number of pilots should be not smaller than the channel length n_L for frequency-domain channel estimation based on pilots in CP-OFDM systems, while the proposed joint time-frequency channel estimation only requires $N_p = (Q+1)(2d+1)L$ subcarriers for path gains estimation after the path delays have been obtained in (10). Compared with the conventional DPNP scheme for TDS-OFDM [8], the redundant grouped pilots in the proposed time-frequency signal structure leads to the loss in spectral efficiency by

$$\mathrm{E}_{\text{loss}} = \frac{N}{N+2M} - \frac{N-(Q+1)(2d+1)L}{N+2M-1}. \tag{18}$$

For typical DTTB systems, large FFT size, e.g., $N = 4096$ is usually used [3]. Since all channel models used for digital television system evaluation [19] have less than six resolvable paths, we adopt $L = 6$ without loss of generality. In addition, through simulation study we find that $Q = 1$, $d = 1$ could already provide satisfying performance over fading channels. In this case, the grouped pilots in the proposed scheme only occupy $N_p = 36$ subcarriers, which is only 0.88% of the total subcarriers. Therefore, the spectral efficiency loss is negligible. For example, the spectral efficiency loss is 0.69% and 0.76% when $M = N/8$ and $M = N/16$, respectively.

B. Computational Complexity

The complexity of the proposed scheme is evaluated by the number of complex multiplications. Eq. (9) needs one M-point FFT and one M-point inverse FFT (IFFT) to implement the M-point circular correlation (one M-point FFT/IFFT requires $(M/2)\log_2 M$ complex multiplications). Eq. (17) requires $2N_{group}(Q+1)^2L^2 + (Q+1)^3L^3$ multiplications to compute the Moore-Penrose inverse matrix of $\boldsymbol{\mu}_i \mathbf{b}_i$ and other $N_{group}(Q+1)L$ multiplications for matrix multiplication. Eq. (6) needs $2L(2d+1)N$ multiplications to calculate the ICI coefficients since ICI is mainly introduced by the adjacent $2d$ subcarriers [5], [16]. Thus, considering the parameters mentioned above, the proposed scheme requires 152,784 more multiplications (which are equivalent to about three 8192-point FFTs used in [6]) than the conventional DPNP-aided TDS-OFDM scheme [8]. However, with the fast

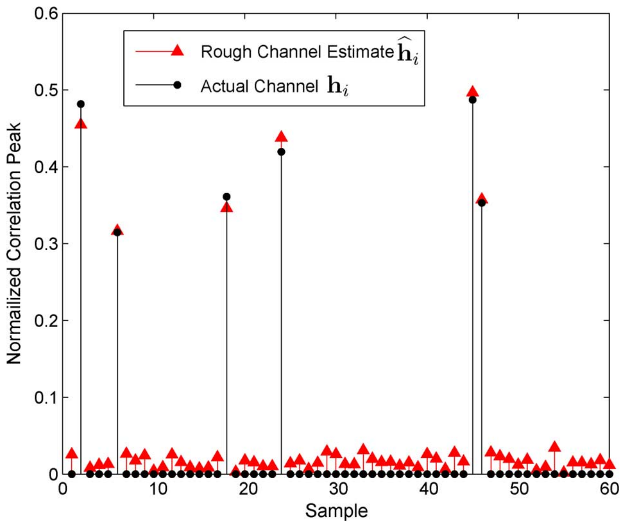

Fig. 2. TS-based path delay estimation over the Brazil D channel with the SNR of 10 dB.

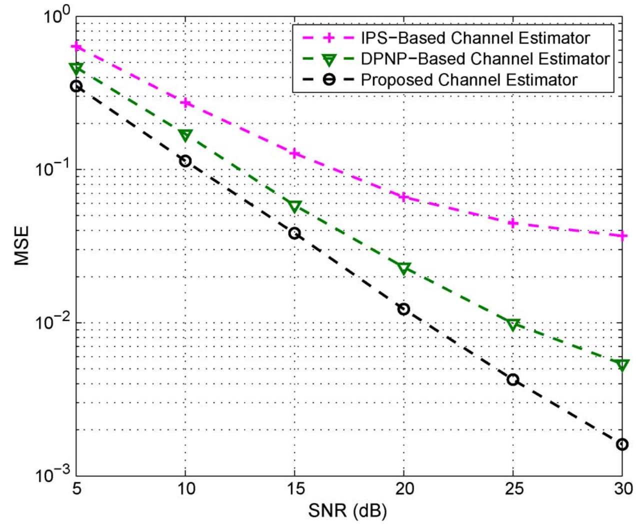

Fig. 3. MSE performance comparison over the Brazil D channel with the receiver velocity of 140 km/h.

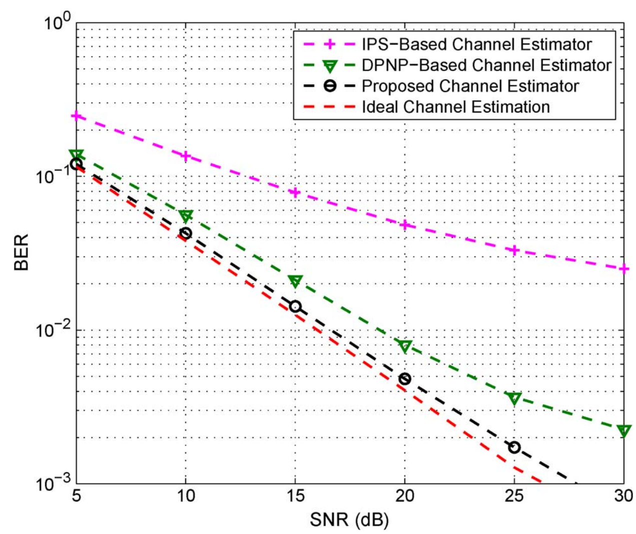

Fig. 4. BER performance comparison over the Brazil D channel with the receiver velocity of 140 km/h.

increase of the digital processor capability, such complexity burden is affordable and worthy of the obvious performance improvement as shown in the next section.

V. Simulation Results

The performance of the proposed time-frequency signal structure and the corresponding joint time-frequency channel estimation for TDS-OFDM systems is investigated over fast fading channels. The signal bandwidth is 7.56 MHz located at the central radio frequency of 770 MHz. The modulation scheme is QPSK. Other system parameters are consistent with those specified in Section IV-A, e.g., $N = 4096$, $M = 256$, $L = 6$, $Q = 1$, $d = 1$. The zero-forcing equalization with ICI removal [18] is employed for data detection. The Brazil D [19] Rayleigh fading channel with the maximum Doppler spread f_d of 100 Hz corresponding to the relative receiver velocity of 140 km/h is considered.

Fig. 2 illustrates the TS-based channel delay estimation as addressed in Section III-A over the Brazil D channel with the SNR of 10 dB. The actual channel is also included for comparison. It is clear that although the rough channel estimate $\widehat{\mathbf{h}}_i$ deviates a little from the actual channel $\mathbf{h}_i$, especially for the channel gains, the path delays of the actual channel $\mathbf{h}_i$ could be perfectly preserved by $\widehat{\mathbf{h}}_i$ due to the ideal autocorrelation property of the MOS. The path delay information would facilitate the pilot-based path gain estimation to finally acquire the complete CIR estimate over fast fading channels.

Fig. 3 compares the mean square error (MSE) performance of the conventional IPS-based channel estimator [6], the DPNP-based channel estimator [8], and the proposed joint time-frequency channel estimator over the Brazil D channel with the receiver velocity of 140 km/h. It is clear that the proposed method obviously outperforms the conventional schemes in terms of MSE over fast fading channels. The superior performance of [8] to [6] is mainly contributed by the avoidance of the IPS algorithm having poor performance. However, [8] still assumes time-invariant channel within every TDS-OFDM symbol, which is not valid for fast fading channels. On the contrary, the proposed signal structure having training information both in the time and frequency domains enables the joint time-frequency channel estimation method, which could accurately track the variation of fast fading channels due to the hybrid processing in both the time and frequency domains [20].

Fig. 4 shows the bit error rate (BER) performance comparison of those three schemes above over the Brazil D channel with the mobile speed of 140 km/h. The BER performance with ideal channel estimation is also included as the benchmark for performance evaluation. We could observe that the obvious SNR loss will be imposed on the conventional TDS-OFDM scheme compared with [8] and our proposal due to the impact of residual interference over fast time-varying channels, and the proposed solution has the best performance in such scenario. For example, when the BER is 10^{-2}, compared with the DPNP-based channel estimator [8], the SNR gain achieved by the proposed scheme is about 2.3 dB. Furthermore, it also shows that the proposed scheme has the BER performance

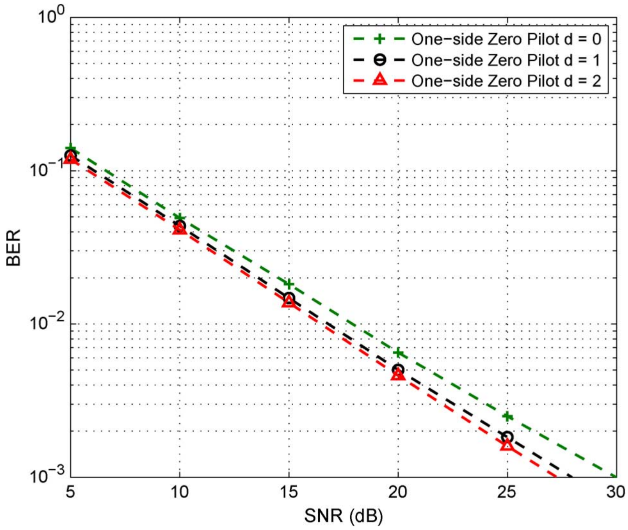

Fig. 5. Impact of the zero pilot number on the BER performance.

close to that of the ideal channel estimation case. The improved BER performance is mainly contributed by the obviously enhanced channel tracking capability as illustrated by Fig. 3. Additionally, the MOS with ideal autocorrelation property, as well as the signal frame structure with non-constant TS also contribute to the overall system performance improvement.

Fig. 5 presents the impact of the one-side zero pilot number d on the final BER performance. As mentioned in Section II-A, the $2d$ zero pilots surrounding the central pilot within each pilot group are dedicated to alleviate the potential ICI imposed on the central pilot. When the channel is varying slowly, the ICI due to Doppler spread is small, and $d = 0$ can be used. For fast time-varying channels, the ICI would increase with respect to the maximum Doppler spread f_d, and obvious performance gain could be achieved by increasing d from 0 to 1, as shown in Fig. 5. However, when d is further increased to 2, the SNR gain becomes negligible, which validates the statements that the ICI is dominantly introduced by the neighboring subcarriers and $d = 1$ could provide sufficient accuracy even when the channel is varying fast.

VI. Conclusions

In this paper, we have investigated a novel time-frequency signal structure and the corresponding joint time-frequency channel estimation method to improve the performance of TDS-OFDM systems over fast fading channels. The proposed scheme utilizes the one-sample shifted TS with perfect autocorrelation in the time domain to estimate the path delays, while the path gains are estimated by the frequency-domain grouped pilots with only single active pilot placed in the middle. It has been demonstrated that the proposed method could accurately track the channel variation, and outperforms the conventional solutions over fast time-varying channels. It is also shown that the proposed scheme has the close BER performance to that of the ideal channel estimation case.

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