Spectrally Efficient Time-Frequency Training OFDM for MIMO Systems

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***Abstract*—The large number of pilots commonly used in OFDM MIMO systems reduces the spectral efficiency in practice. This paper proposes the time-frequency training OFDM (TFT-OFDM) transmission scheme for MIMO systems to solve this problem. The transmission frame is composed of one preamble and the following TFT-OFDM symbols, where each TFT-OFDM symbol without cyclic prefix adopts the time-domain training sequence (TS) and the frequency-domain orthogonal grouped pilots as the time-frequency training information. At the receiver, the time-frequency joint channel estimation directly exploits the "contaminated" time-domain TS to estimate the path delays only, while the path gains are acquired by the frequency-domain grouped pilots. The Cramér-Rao lower bound (CRLB) of the proposed estimator is also derived. Compared with standard OFDM MIMO systems in typical applications, the proposed scheme has about 17% higher spectral efficiency, and has better performance over doubly selective fading channels as indicated by the simulation results.**

I. Introduction

Orthogonal frequency division multiplexing (OFDM) and multiple-input multiple-output (MIMO) are widely recognized as two fundamental physical layer technologies for wireless communications [1]. In MIMO systems, the standard OFDM transmission scheme is called cyclic prefix OFDM (CP-OFDM), where a CP is used as the guard interval to eliminate inter-block-interference (IBI) as well as inter-carrier-interference (ICI) [2], [3]. Normally, frequency-domain orthogonal pilots are adopted to convert the complex MIMO channel into much simpler single-input single-output (SISO) channels [4]. However, the increased number of pilots with respect to the number of transmit antennas obviously reduces the spectral efficiency of MIMO systems, especially when a large number of transmit antennas is used. It is common to reduce the pilot overhead by limiting the pilot density within a certain degree, but the channel estimation accuracy and consequently the system performance will deteriorate accordingly [5].

As an alternative to standard CP-OFDM, time domain synchronous OFDM (TDS-OFDM) increases the spectral efficiency by utilizing the pseudorandom noise (PN) sequence as both the guard interval and the time-domain training sequence (TS) for channel estimation [6], [7]. However, in TDS-OFDM SISO systems, the required iterative padding subtraction (IPS) [8] of the IBI between the TS and the OFDM data block suffers from high complexity as well as performance degradation over fading channels. This issue becomes more severe when TDS-OFDM is extended from SISO to MIMO systems due to more complex interferences caused by multiple antennas [9]. Some TDS-OFDM MIMO schemes propose the space-time or space-frequency coded TS for channel discrimination, and the CP of those TS is used to avoid severe interferences [10]. However, the increased TS length and the re-adoption of CP dismiss the merit of high spectral efficiency of TDS-OFDM.

To solve those problems, we propose a time-frequency training OFDM (TFT-OFDM) scheme with the following contributions: 1) At the transmitter, unlike CP-OFDM or TDS-OFDM, the proposed TFT-OFDM MIMO scheme jointly adopts the one-sample shifted time-domain TS and the frequency-domain orthogonal grouped pilots as the time-frequency training information for every TFT-OFDM symbol; 2) At the receiver, the corresponding time-frequency joint channel estimation scheme directly exploited the "contaminated" time-domain TS without interference cancellation to merely acquire the information about path delays of the channel, while path gains are estimated by the frequency-domain grouped pilots only; 3) Simulation results indicate that the proposed channel estimator has higher accuracy than conventional schemes, and approaches the theoretical Cramér-Rao lower bound (CRLB) derived in this paper. As a result, the proposed scheme with reliable performance requires much fewer pilots than that in common OFDM MIMO systems, so the spectral efficiency can be increased by about 17% in typical applications.

The rest of this paper is organized as follows. The system model of the proposed TFT-OFDM MIMO scheme is presented in Section II. The corresponding receiver design is discussed in Section III. Section IV studies the spectral efficiency of the proposed scheme. Simulation results are shown in Section V, and conclusions are finally drawn in Section VI.

Notation: We use upper and lower boldface letters to denote matrices and column vectors, respectively. $\mathbf{F}_N$ is the $N \times N$ normalized discrete Fourier transform (DFT) 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 $(\cdot)^*$, $(\cdot)^{\mathrm{T}}$, $(\cdot)^H$, $(\cdot)^{-1}$, and $(\cdot)^\dagger$ denote the complex conjugate, transpose, conjugate transpose, matrix inversion, and Moore-Penrose matrix inversion, respectively. $\mathrm{Tr}\{\cdot\}$ and $\mathrm{E}\{\cdot\}$ stand respectively for trace and expectation operators. $\widehat{x}$ means the estimate of x. Finally,

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diag$\{\mathbf{u}\}$ is a diagonal matrix with $\mathbf{u}$ at its main diagonal.

II. TFT-OFDM MIMO System Model

As illustrated by the time-frequency frame structure in Fig. 1, the TFT-OFDM signal is transmitted frame by frame, and each frame is composed of one preamble with its cyclic extension and the subsequent U TFT-OFDM symbols (subframes). We assume N_t transmit antennas and N_r receive antennas in MIMO systems.

In the time domain, the ith TFT-OFDM symbol ($1 \leq i \leq U$) for the pth transmit antenna ($1 \leq p \leq N_t$) is composed of the length-N OFDM symbol $\mathbf{x}_i^{(p)} = [x_{i,0}^{(p)}, x_{i,1}^{(p)}, \cdots, x_{i,N-1}^{(p)}]^T$ (whose DFT is $\widetilde{\mathbf{x}}_i^{(p)} = \mathbf{F}_N \mathbf{x}_i^{(p)}$) and the followed length-M TS $\mathbf{z}_i^{(p)} = [z_{i,0}^{(p)}, z_{i,1}^{(p)}, \cdots, z_{i,M-1}^{(p)}]^T$. Since it has been proved in [11] that constant TS is not optimal for channel tracking, the TS $\mathbf{z}_i^{(p)}$ will be generated by cyclically shifting the basic TS $\mathbf{z}^{(p)} = [z_0^{(p)}, z_1^{(p)}, \cdots, z_{M-1}^{(p)}]^T$ by i samples to the left as

$$\mathbf{z}_i^{(p)} = \begin{bmatrix} \mathbf{0}_{(M-i)\times i} & \mathbf{I}_{M-i} \\ \mathbf{I}_i & \mathbf{0}_{i\times(M-i)} \end{bmatrix} \mathbf{z}^{(p)}, \tag{1}$$

where $\mathbf{z}^{(p)}$ is the Zadoff-Chu sequence (also known as generalized Chirp-like (GCL) sequence) [12] defined by

$$z_n^{(p)} = \exp\left(j\frac{M-1}{M}\pi n^2 r_p\right), 0 \leq n \leq M-1, \tag{2}$$

where r_p is relatively prime to M. The Zadoff-Chu sequences have ideal autocorrelation and the optimal crosscorrelation equals to the theoretical Sarwate bound [13].

In the frequency domain, TFT-OFDM for MIMO systems adopts G orthogonal pilot groups, where each pilot group has only one non-zero central pilot with the amplitude of a surrounded by $2d$ zero pilots (in Fig. 1 (c), $d = 1$ is used as an example). Based on [14], $d = 1$ zero pilot will be sufficient to alleviate the ICI imposed on the central pilot in fast fading channels. Due to the existence of zero pilots, the central pilot naturally allows power boosting without sacrificing the transmit power, namely, $a = \sqrt{2d+1}$ if $|\widetilde{x}_{i,k}^{(p)}| = 1$. In addition, the subcarrier index set of the central pilots can be denoted by $\mathbf{g}^{(p)} = [g_0 + O_p, g_1 + O_p, \cdots, g_{G-1} + O_p]^T$ with $O_p = (d+1)(p-1)$ being the subcarrier offset for the pth transmit antenna, which indicates that the grouped pilots for two adjacent transmit antennas can overlap d zero pilots.

The length-N_p basic preamble $\mathbf{c} = \mathbf{F}_{N_p}^H \widetilde{\mathbf{c}}$ is also a Zadoff-Chu sequence defined by (2), where N_p instead of M and $r_p = 1$ are used. Based on $\mathbf{c}$, the preamble for the pth transmit antenna $\mathbf{c}^{(p)} = [c_0^{(p)}, c_1^{(p)}, \cdots, c_{N_p-1}^{(p)}]^T$ is then generated by

$$\mathbf{c}^{(p)} = \mathbf{F}_{N_p} \widetilde{\mathbf{c}}^{(p)} = \mathbf{F}_{N_p}^H \text{diag}\{\widetilde{\mathbf{c}}\} \mathbf{w}^{(p)}, \tag{3}$$

where $\mathbf{w}^{(p)} = [0, e^{-j\frac{2\pi}{N_t}(p-1)}, \cdots, e^{-j\frac{2\pi}{N_t}(N_p-1)(p-1)}]^T$, $\widetilde{\mathbf{c}}^{(p)} = \text{diag}\{\widetilde{\mathbf{c}}\} \mathbf{w}^{(p)}$ denotes the DFT of $\mathbf{c}^{(p)}$, and $N_p = N_t M$ is assumed. Then, the last $M-1$ samples of $\mathbf{c}^{(p)}$ is used to form the corresponding cyclic extension as shown in Fig. 1.

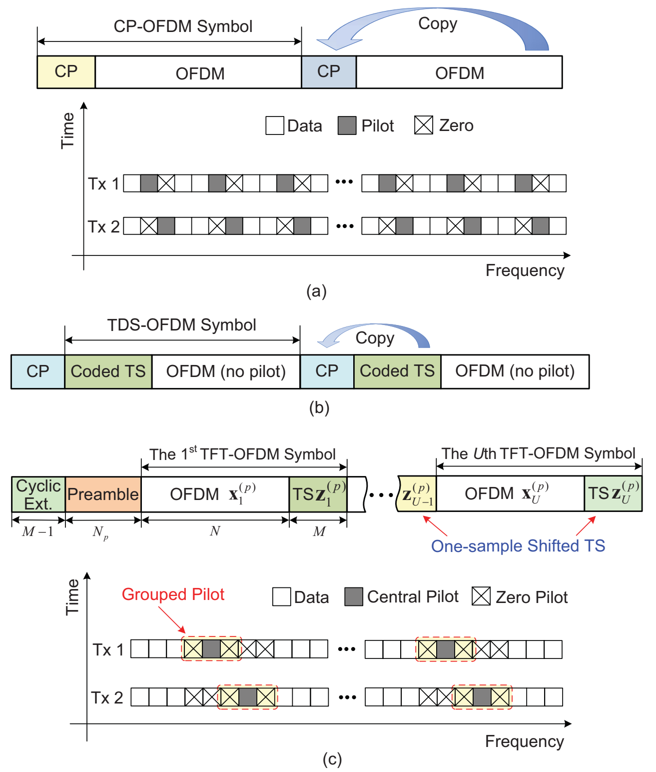

Fig. 1. Time-frequency frame structure comparison of the MIMO systems based on: a) Conventional CP-OFDM [1]; b) Conventional TDS-OFDM [10]; c) The proposed TFT-OFDM.

III. TFT-OFDM MIMO Receiver Design

A. Preamble-Based Initial Channel Estimation

The channel impulse response (CIR) $\mathbf{h}_i^{(p)}$ between the pth transmit antenna and a certain receive antenna (we explicitly drop the receive antenna index since every receive antenna adopts the same processing) during the ith TFT-OFDM symbol can be denoted by

$$\mathbf{h}_i^{(p)} = [h_{i,0}^{(p)}, h_{i,1}^{(p)}, \cdots, h_{i,L-1}^{(p)}]^T, \tag{4}$$

where $h_{i,l}^{(p)}$ is the path gain of the lth path with the path delay of τ_l, L denotes the channel memory length, and $L \leq M$ is assumed in this paper, so we have $N_p = N_t M = N_t L$. Note that $\mathbf{h}_0^{(p)}$ is used to denote the CIR during the preamble. Then, due to the protection of the cyclic extension, the received time-domain preamble $\mathbf{d}_0 = [d_{0,0}, d_{0,1}, \cdots, d_{0,N_p-1}]^T$ can be expressed by

$$\mathbf{d}_0 = \sum_{p=1}^{N_t} \mathbf{c}^{(p)} \otimes \mathbf{h}_0^{(p)} + \mathbf{v}_0 = \sum_{p=1}^{N_t} \mathbf{c}_0^{(p)} \mathbf{h}_0^{(p)} + \mathbf{v}_0 = \mathbf{C}_0 \mathbf{h}_0 + \mathbf{v}_0, \tag{5}$$

where $\mathbf{c}_0^{(p)}$ is the $N_p \times L$ circulant matrix with the first column being the preamble $\mathbf{c}^{(p)}$, $\mathbf{C}_0 = \left[\mathbf{c}_0^{(1)}, \mathbf{c}_0^{(2)}, \cdots, \mathbf{c}_0^{(N_t)}\right]$ denotes the $N_p \times N_t L$ time-domain training matrix based on $\{\mathbf{c}^{(p)}\}_{p=1}^{N_t}$, $\mathbf{h}_0 = \left[\left(\mathbf{h}_0^{(1)}\right)^T, \left(\mathbf{h}_0^{(2)}\right)^T, \cdots, \left(\mathbf{h}_0^{(N_t)}\right)^T\right]^T$ presents the

$N_tL \times 1$ equivalent "total" CIR for all N_t transmit antennas, and $\mathbf{v}_0 = [v_{0,1}, v_{0,1}, \cdots, v_{0,N_p-1}]^T$ denotes the channel's AWGN. Thus, the time-domain channel estimate $\widehat{\mathbf{h}}_0$ during the preamble can be obtained by

$$\widehat{\mathbf{h}}_0 = \mathbf{C}_0^{\dagger}\mathbf{d}_0 = \left(\mathbf{C}_0^H\mathbf{C}_0\right)^{-1}\mathbf{c}_0^H\mathbf{d}_0. \tag{6}$$

Due to the perfect autocorrelation of Zadoff-Chu sequences, the preamble satisfies $\mathbf{c}_0^H\mathbf{c}_0 = N_tL\mathbf{I}_{N_tL}$, so the preamble-based initial channel estimator (6) can be simplified as

$$\widehat{\mathbf{h}}_0 = \frac{1}{N_tL}\mathbf{c}_0^H\mathbf{d}_0 = \frac{1}{N_tL}\mathbf{c} \otimes \mathbf{d}_0. \tag{7}$$

B. Cyclictiy Reconstruction of the OFDM Symbol

Due to the absence of CP, the cyclicity of the received OFDM symbol in TFT-OFDM will be destroyed in multipath fading channels. The hybrid-domain cyclictiy reconstruction method [15] for SISO systems can be directly extended for MIMO systems by using N_t times of the overlap and add (OLA) algorithm [16]. Note that the required channel information can be either simply approximated by preamble-based initial channel estimation or predicted by the Kalman filter exploiting the temporal correlation nature of the channel and the obtained channel estimates [17].

C. Time-Frequency Joint Channel Estimation

The preamble-based channel estimate (7) may be outdated over fast fading channels. Thus, the data detection in the TFT-OFDM transmission scheme will rely on the time-frequency joint channel estimation algorithm as described in this subsection, whereby the time-frequency training information within every TFT-OFDM symbol is fully exploited by two sequential steps: the TS-based path delay estimation at first and then the pilot-based path gain estimation.

Being different from TDS-OFDM where iterative interference cancellation is exploited to achieve the complete CIR estimation based on the TS [8], we directly use the "contaminated" TS $\mathbf{d}_i = [d_{i,0}, d_{i,1}, \cdots, d_{i,M-1}]^T$ without interference cancellation to just estimate the path delays of the channel as

$$\widehat{\mathbf{h}}_i^{(p)} = \frac{1}{M}\mathbf{z}_i^{(p)} \otimes \mathbf{d}_i = \mathbf{h}_i^{(p)} + \mathbf{n}_i^{(p)} + \mathbf{v}_i^{(p)}, \tag{8}$$

where $\mathbf{v}_i^{(p)} = \frac{1}{M}\mathbf{z}_i^{(p)} \otimes \mathbf{v}_i$, and $\mathbf{n}_i^{(p)}$ denotes the interferences caused by the non-zero crosscorrelation among different Zadoff-Chu sequences as well as the IBIs from previous OFDM symbols. Note that the ideal autocorrelation property of the Zadoff-Chu sequence has been utilized in (8).

Then, the path gains in $\widehat{\mathbf{h}}_i^{(p)}$ are discarded due to the interferences mentioned above, and the path delays of the Q most significant taps of $\widehat{\mathbf{h}}_i^{(p)}$ are stored in the path delay set

$$\mathbf{\Gamma}_i^{(p)} = \{\tau_l : \left|\widehat{h}_{i,l}^{(p)}\right|^2 \geq T_{th}\}_{l=0}^{L-1}, \quad 1 \leq p \leq N_t, \tag{9}$$

where T_{th} is pre-defined power threshold [18], and Q is usually much smaller than the channel length L, i.e., $Q \ll L$. For example, the ITU Vehicular B channel [19] with the maximum delay spread of 20 μs, which is equivalent to $L = 200$ symbols at the symbol rate of 10 MHz, has only $Q = 6$ resolvable paths. That is to say, the number of unknown parameters in the CIR $\mathbf{h}_i^{(p)}$ is substantially reduced from L to Q ($Q \ll L$) after the path delays $\mathbf{\Gamma}_i^{(p)}$ have been obtained.

The received central pilots $Y_{i,k}^{(p)}$ corresponding to the central pilots for the pth transmit antenna should be

$$Y_{i,k}^{(p)} = \sum_{p=1}^{N_t} X_{i,k}^{(p)}H_{i,k}^{(p)} + W_{i,k} = aH_{i,k}^{(p)} + W_{i,k}, \quad k \in \mathbf{g}^{(p)}, \tag{10}$$

where $X_{i,k}^{(u)} = 0$ ($u \neq p, k \in \mathbf{g}^{(p)}$) has been utilized due to the orthogonality of the grouped pilots among different transmit antennas. Eq. (10) can be rewritten in a more compact notation as

$$\mathbf{Y}_i^{(p)} = a\mathbf{F}_N^{(p)}\mathbf{h}_{i,\Gamma}^{(p)} + \mathbf{W}_i^{(p)}, \tag{11}$$

where $\mathbf{Y}_i^{(p)} = [Y_{i,g_0+O_p}^{(p)}, Y_{i,g_1+O_p}^{(p)}, \cdots, Y_{i,g_{G-1}+O_p}^{(p)}]_{G\times 1}^T$, $\mathbf{h}_{i,\Gamma}^{(p)} = [h_{i,\tau_0}^{(p)}, h_{i,\tau_1}^{(p)}, \cdots, h_{i,\tau_{Q-1}}^{(p)}]_{Q\times 1}^T$ denotes the "selected" CIR out of $\mathbf{h}_i^{(p)}$ according to the path delays $\mathbf{\Gamma}_i^{(p)}$ obtained in (9), $\mathbf{F}_N^{(p)}$ presents the $G \times Q$ "extracted" matrix generated by selecting the $\mathbf{g}^{(p)}$ rows and $\mathbf{\Gamma}_i^{(p)}$ columns out of the DFT matrix $\mathbf{F}_N$, and $\mathbf{W}_i^{(p)} = [W_{i,g_0+O_p}^{(p)}, W_{i,g_1+O_p}^{(p)}, \cdots, W_{i,g_{G-1}+O_p}^{(p)}]_{G\times 1}^T$ presents the AWGN vector.

It is clear from (11) that only Q unknown path gains in $\mathbf{h}_{i,\Gamma}^{(p)}$ have to be estimated by the G observations in $\mathbf{Y}_i^{(p)}$. If $G > Q$, the Q (not L , and usually $Q \ll L$) path gains of $\mathbf{h}_i^{(p)}$ can be estimated by the received central pilots $\mathbf{Y}_i^{(p)}$ as

$$\widehat{\mathbf{h}}_{i,\Gamma}^{(p)} = \frac{1}{a}\left(\mathbf{F}_N^{(p)}\right)^{\dagger}\mathbf{Y}_i^{(p)} = \frac{1}{a}\left[\left(\mathbf{F}_N^{(p)}\right)^H\mathbf{F}_N^{(p)}\right]^{-1}\left(\mathbf{F}_N^{(p)}\right)^H\mathbf{Y}_i^{(p)}. \tag{12}$$

After that, the obtained OFDM symbol $\mathbf{Y}_i$ and the complete CIR estimate $\widehat{\mathbf{h}}_i$ (based on $\{\mathbf{\Gamma}_i^{(p)}\}_{p=1}^{N_t}$ and $\{\widehat{\mathbf{h}}_{i,\Gamma}^{(p)}\}_{p=1}^{N_t}$) are fed into the standard MIMO detector to develop the soft-likelihood estimates of the signal, which are then used by the channel decoder to ultimately recover the transmitted signal.

It can be derived that the CRLB of the proposed time-frequency joint channel estimator is

$$\mathrm{CRLB} = \mathrm{E}\left\{\left\|\widehat{\mathbf{h}}_{i,\Gamma}^{(p)} - \mathbf{h}_{i,\Gamma}^{(p)}\right\|^2\right\} = \frac{Q\sigma^2}{G}, \tag{13}$$

which is lower than the noise level σ^2 (the best performance that can be achieved by conventional TDS-OFDM MIMO systems if the interferences are completely removed [9]) due to we usually have $G > Q$.

IV. Spectral Efficiency Comparison

Compared with the ideal OFDM systems without any overhead, the spectral efficiency of the proposed TFT-OFDM MIMO scheme in percentage notation [9], [10] is

$$\mathrm{E}_0 = \frac{U(N-K)}{U(N+M) + N_p + M - 1}, \tag{14}$$

where $K = G\left(N_t(d+1)+d\right)$ is the number of used pilots.

For typical wireless digital television systems, large DFT size, e.g., $N = 4096$ is usually adopted [20]. Since all channel models defined by ITU [19] for cellular network investigation and all channel models used for digital television system evaluation [21] have no more than six resolvable paths, we can assume $G = Q = 6$ without loss of generality. However, in practical applications, the path number may be large, so we configure $G = 20$ for system design with some margin. Thus, in the case of two transmit antennas, i.e., $N_t = 2$, the number of used pilots in TFT-OFDM is $K = 100$, which is only 2.44% of the total subcarrier number $N = 4096$. On the contrary, the Karhunen-Loeve theorem [14] requires $N_t L$ frequency-domain pilots to estimate the length-L channels for N_t transmit antennas in CP-OFDM MIMO systems[1], which indicates that $N_t L = 1024$ pilots occupying 25% of the total subcarriers are required in CP-OFDM MIMO systems when the typical guard interval length $M = N/8$ is applied.

TABLE I
SPECTRAL EFFICIENCY IN MIMO SYSTEMS.

Guard Interval Length	CP-OFDM	TDS-OFDM	TFT-OFDM
$M = N/8$	66.67%	72.73%	83.92%
$M = N/16$	82.35%	84.21%	90.23%
$M = N/32$	90.91%	91.43%	93.75%

With the parameters above, Table I compares the spectral efficiency of the MIMO systems based on CP-OFDM, TDS-OFDM[2], and TFT-OFDM schemes with different guard interval lengths. In TFT-OFDM, the subframe number U can be configured large due to the good channel tracking capability of the joint time-frequency channel estimation in every TFT-OFDM symbol, and $U = 10$ is assumed in this paper. It is clear that the proposed TFT-OFDM scheme outperforms CP-OFDM and TDS-OFDM in spectral efficiency in all cases. In the typical application scenario of $M = N/8$, the spectral efficiency of TFT-OFDM is 17.32% and 11.19% higher than that of CP-OFDM and TDS-OFDM, respectively.

V. SIMULATION RESULTS

This section investigates the performance of the proposed TFT-OFDM scheme for MIMO systems with the low density parity check (LDPC) channel coding. The signal bandwidth is 7.56 MHz with the central radio frequency of 770 MHz, and other system parameters are consistent with those specified in Section IV-B. Similar to the typical configuration of the standard [20], the classical Alamouti space-time block coding scheme [22] is used at the transmitter, and the corresponding optimum maximum-likelihood detection is used at the receiver.

[1]In some applications like long term evolution (LTE) [5], fewer pilots can be used due to interpolation can be applied to estimate the CFRs at data subcarriers at the cost of performance degradation, especially when N_t is large.

[2]The length of the coded TS in TDS-OFDM MIMO is twice that of the CP to ensure the equivalent channel estimation performance as that in SISO scenarios [10].

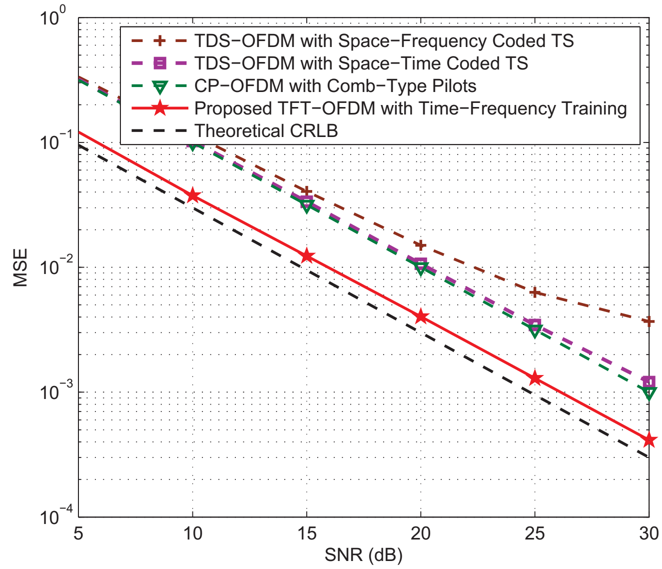

Fig. 2. MSE performance comparison between the proposed time-frequency joint channel estimation method for TFT-OFDM with those for CP-OFDM and TDS-OFDM.

We adopt the modulation scheme 64 QAM with the code rate 0.6 for LDPC, which is usually used to deliver high-definition television (HDTV) services in wireless broadcasting systems. The Brazil D Rayleigh fading channel [21] with the maximum delay spread of 5.93 μs is considered in the simulation.

Fig. 2 compares the mean square error (MSE) performance of the proposed time-frequency joint channel estimation method with its counterparts in conventional CP-OFDM and TDS-OFDM systems over the static Brazil D channel. The theoretical CRLB derived in (13) is also included as the benchmark for MSE comparison. It is clear that TFT-OFDM outperforms CP-OFDM and TDS-OFDM by more than 4 dB when the channel estimation MSE is considered. For example, when the MSE is 10^{-2}, CP-OFDM requires the signal-to-noise ratio (SNR) of 20 dB, TDS-OFDM with space-time and space-frequency coded TS requires the SNR of 20.3 dB and 22.3 dB, respectively, while TFT-OFDM only needs the SNR of 16 dB. In addition, the proposed channel estimation performs closely to the theoretical CRLB with a small SNR gap, since the "extracted" DFT matrix $\mathbf{F}_N^{(p)}$ has imperfect but approximate orthogonal columns.

Fig. 3 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. We can observe that TDS-OFDM MIMO scheme with the space-frequency coded TS could not work over such deeply frequency-selective channel. Compared with CP-OFDM and TDS-OFDM with the space-time coded TS, the performance gain achieved by the proposed TFT-OFDM scheme is 0.75 dB and 1.60 dB at the BER of 10^{-4} over the doubly-selective fading channel, respectively. In addition, compared with the case of ideal channel state

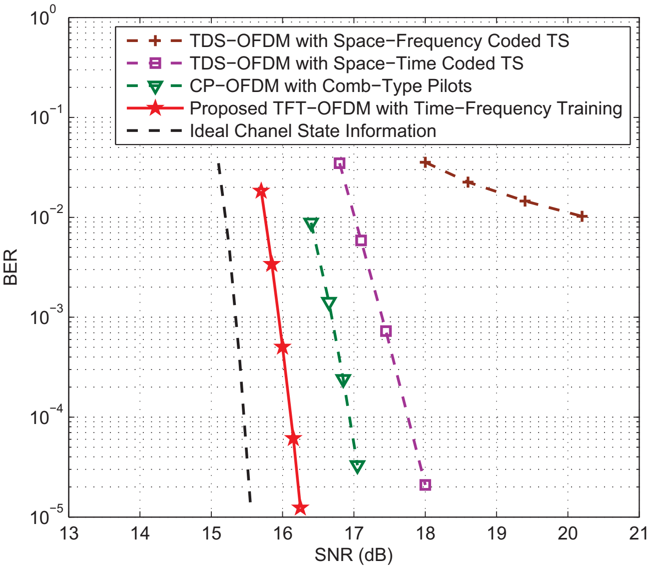

Fig. 3. BER performance comparison over the Brazil D channel with the receiver velocity of 140 km/h.

information, only 0.6 dB SNR loss will be imposed on TFT-OFDM. This is contributed by the fact that the CIR can be accurately tracked in every TFT-OFDM symbol, and the central pilot guarded by zero pilots is robust to fast time variation of the channel.

VI. Conclusions

In this paper, we propose a spectrally efficient TFT-OFDM transmission scheme for MIMO systems. Each TFT-OFDM symbol has training information in both time and frequency domains, and the frequency-domain grouped pilots occupy much fewer subcarriers than that in standard OFDM MIMO systems. This is achieved by the joint time-frequency channel estimation scheme, whereby the path delays are firstly acquired by the time-domain received TS without interference cancellation, then there remains much fewer channel parameters to be estimated by the substantially reduced number of frequency-domain pilots. In addition, the proposed TFT-OFDM scheme can be also directly applied in multiple access systems in both the uplink and downlink, and the design principle of joint time-frequency processing behind TFT-OFDM can be adapted for other OFDM MIMO systems to achieve higher spectral efficiency and more reliable performance as well.

Acknowledgments

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