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RESEARCH ARTICLE

ADAPTIVE TECHNIQUE FOR HIGH SPEED MOBILE ENVIRONMENT USING TFT-OFDM

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Abstract: Orthogonal frequency division multiplexing (OFDM) is widely recognized as the key technology for the next generation broadband wireless communication (BWC) systems. The high spectral efficiency, reliable performance over fast fading channels is becoming more and more important for OFDM-based Broadband Wireless Communication systems. The time domain synchronous OFDM (TDS-OFDM) has higher spectral efficiency than the standard cyclic prefix OFDM (CP-OFDM), but suffers from severe performance loss over high speed mobile channels since the required iterative interference cancellation between the training sequence (TS) and the OFDM data block, for the application of compressed sensing (CS) to the estimation of doubly selective channels within pulse-shaping multicarrier systems (which include orthogonal frequency- division multiplexing (OFDM) systems as a special case. Thus, the fundamentally distinct OFDM-based transmission scheme called time-frequency training OFDM (TFT-OFDM) is proposed, whereby every TFT-OFDM symbol has training information both in the time and frequency domains. The grouped pilots in TFT-OFDM occupy only about 3% of the signal bandwidth with high spectral efficiency as well as good performance over fast time-varying channels. The efficiency of the proposed approach is illustrated in a simulation environment that allows for a flexible, structured, and comparative performance evaluation.

Keywords: Orthogonal frequency division multiplexing (OFDM), time-frequency training (TFT), joint time frequency channel estimation, compressed sensing, interference cancellation, spectral efficiency, fast fading channels

I. INTRODUCTION

Due to the robustness to the frequency-selective multipath channel and the low complexity of the frequency domain equalizer, orthogonal frequency division multiplexing (OFDM) has been widely recognized as one of the key techniques for the next generation broadband wireless communication (BWC) systems [2]. One fundamental issue of OFDM is the block transmission scheme. Basically, there are three types of OFDM-based block transmission schemes:

Cyclic prefix OFDM (CP-OFDM) [3], zero padding OFDM (ZP-OFDM) [4], and Time Domain synchronous OFDM (TDS-OFDM) [5]. The broadly used CP-OFDM scheme utilizes the CP to eliminate the inter-block interference (IBI) as well as the inter-carrier-interference (ICI) [6]. For both the CP-OFDM and ZP-OFDM schemes, some dedicated frequency-domain pilots are required for synchronization and channel estimation, thus the spectral efficiency is reduced. To solve this problem, instead of the CP, the known training sequence (TS) such as the pseudorandom noise (PN) sequence, is used as the guard interval in the TDS-OFDM scheme [5]. Since the TS is known to the receiver, it can be also used for synchronization as well as channel estimation [7]. Consequently, the large amount of frequency-domain pilots used in CP-OFDM and ZP-OFDM could be saved. Thus, TDS-OFDM outperforms CP-OFDM and ZP-OFDM in spectral efficiency by about 10% [8]. However, the main drawback of TDS-OFDM is that, the time-domain TS and the OFDM data block will cause IBI to each other. Thus, the iterative interference cancellation algorithm has to be used for channel estimation and equalization [7], [8], i.e., the IBI from the OFDM data block to the TS must be eliminated before the TS-based time-domain channel estimation, while the IBI caused by the TS to the OFDM data has to be removed to achieve reliable channel equalization. On one hand, the interference cancellation before channel estimation needs the equalized OFDM data information to calculate the IBI caused by the OFDM block; while on the other hand, channel estimation is prerequisite to obtain the equalized OFDM block. One exciting solution to the interference problem of TDS-OFDM is the cyclic postfix OFDM scheme [10], [11], whereby the TS serving as the cyclic postfix is not independent of the OFDM block like that in TDS-OFDM, but is generated by the redundant frequency-domain comb-type pilots within the OFDM symbol. In this way, the IBI from the TS to the OFDM data block can be avoided. However, the cyclic postfix OFDM scheme does not solve the problem of the interference from the OFDM data block to the next TS, thus the iterative interference cancellation with poor performance over fast time-varying channels is still required for channel estimation and OFDM equalization [12]. In addition, the inserted redundant pilots have much higher average power than the normal OFDM data [13], thus the equivalent signal-to-noise ratio (SNR) at the receiver will be reduced if the identical transmitted signal power is permitted. Such SNR loss can be slightly alleviated by changing the positions of the redundant pilots or adding more pilots in the frequency domain [14], but the effect is not obvious. The most effective solution to the interference problem of TDS-OFDM is to duplicate the TS twice, resulting in the dual-PN OFDM (DPN-OFDM) scheme [15].

The second received PN sequence immune from the interference caused by the preceding OFDM data block can be directly used for channel estimation, and the interference cancellation before channel equalization can be replaced by the cyclic prefix reconstruction which is accomplished by the simple add-subtraction operation [15]. In this way, the iterative interference cancellation algorithm could be avoided, leading to the reduced complexity and improved performance over fast fading channels. However, the spectral efficiency of the DPN-OFDM solution is remarkably decreased by the doubled length of the TS.

II. TFT-OFDM SYSTEM MODEL

In this section, the basic concept of the proposed TFTOFDM system is generalized at first and then the TFT-OFDM system model is outlined.

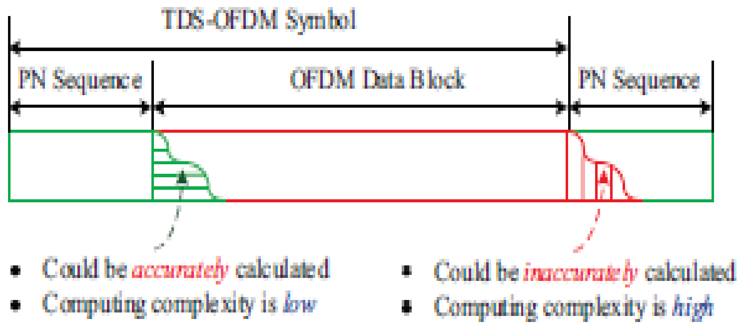

Fig. 2: Distinct features of the IBIs in TDS-OFDM [1]

1. Basic Concept of the TFT-OFDM System

As shown in Figure. 2, the IBI from the TS to the OFDM data block and the IBI caused by the OFDM block to the TS have distinct features in TDS-OFDM. The interference caused by the TS can be completely removed if the channel estimation is perfect, since the TS is known at the receiver. In addition, this IBI can be calculated with relatively low complexity since the TS length is not large. However, the interference caused by the OFDM data block has to be calculated with high complexity, since the OFDM block length is usually large. More importantly, such interference cannot be totally eliminated even when the channel estimation is ideal, because the OFDM data block is random and unknown, and perfect OFDM detection is difficult due to the noise, the ICI, the imperfect channel equalization, and so on, especially when the channel is varying fast. Therefore, the TS-based time-domain channel estimation in TDS-OFDM is not accurate over fast fading channels. Such estimation error would in turn result in the unreliable cancellation of the IBI caused by the TS, which would deteriorate the OFDM equalization performance in the next iteration. Consequently, the corresponding performance loss is unavoidable in TDS-OFDM.

Based on the observations that the IBI caused by the OFDM data block has to be removed for reliable channel estimation, and the complete cancellation of such IBI is difficult even when the channel estimation is perfect, TFTOFDM is derived in this paper to provide a fundamentally distinct solution. In the proposed TFT-OFDM scheme, unlike the conventional method where both the channel path delays and the channel path coefficients are estimated by using the “clean” received TS after IBI cancellation, we do not remove the IBI imposed on the TS, but directly use the “contaminated” TS without IBI cancellation to obtain the partial channel information: the path delays of the channel, while the rest part of the channel information: the path coefficients, are acquired by utilizing the small amount of grouped pilots in the frequency domain. In this way, the IBI caused by the OFDM data block needs not to be removed, leading to the breaking of the mutually conditional relationship between the channel estimation and channel equalization in TDS-OFDM. Consequently, the iterative interference cancellation algorithm with poor performance could be avoided. The only cost is the extra frequency-domain grouped pilots, which lead to the spectral efficiency loss compared with TDS-OFDM. However, such loss is negligible, since the pilots used to estimate the path coefficients only occupy about less than 3% of the total subcarriers in the proposed TFT-OFDM solution.

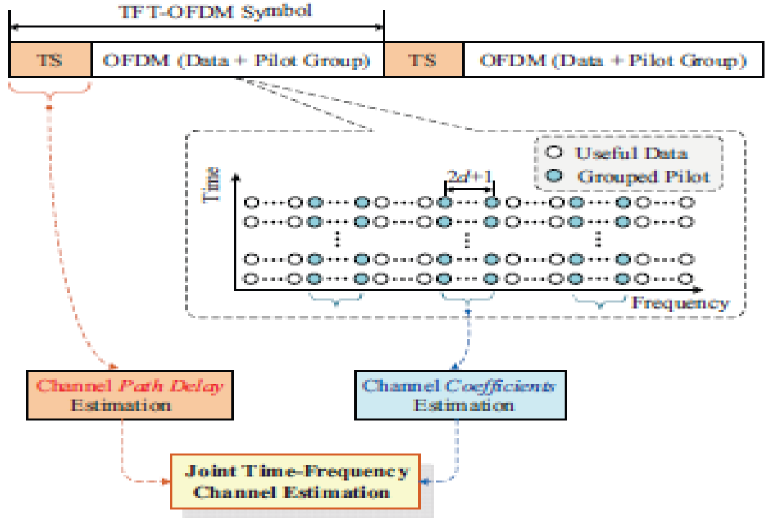

Fig. 3: Signal structure and the corresponding joint time-frequency channel estimation of the TFT-OFDM scheme [1]

In MIMO systems, for a certain receive antenna, the channel impulse response (CIR) $\mathbf{h_i^{(P)}}$ associated with the *p*th transmit antenna during the *i*th TFT-OFDM symbol can be denoted by

$$\mathbf{h_i^{(p)} = [h_{i,0}^{(p)}, h_{i,1}^{(p)},\ldots\ldots, h_{i,L-1}^{(p)}]^T} \qquad \mathbf{(1)}$$

where $h(p)\ i,l$ is the path gain of the lth path with the path delay $\tau(p)\ l$, L denotes the maximum channel spread, and $L = M$ is assumed to avoid the interference between two neighbouring OFDM data blocks, so we have $Np = NtM = NtL$.

2. TFT OFDM SYSTEM MODEL IN TIME DOMAIN

The received preamble $\mathbf{d}_0 = [d_{0,0}, d_{0,1}, \cdot\ \cdot\ \cdot, d_{0,N_{p-1}}]^T$ in the time domain at the receive antenna is immune from the inter-block-interference (IBI) due to the protection of the cyclic extension, so $\mathbf{d}_0$ can be expressed by

$$\mathbf{d_0} = \sum_{P=1}^{Nt} C^{(P)} * \mathbf{h_0} + \mathbf{V_0} = \sum_{P=1}^{Nt} C_0^{(P)} h_0^p + \mathbf{V_0} = \mathbf{C_0 h_0} + \mathbf{V_0} \qquad \mathbf{(2)}$$

where $\mathbf{c_0^{(p)}}$ is the $N_p \times L$ circular matrix with the first column being the preamble $\mathbf{c}(p)$, $C_0 = [C_0^{(1)}, C_0^{(2)}, \ldots, C_0^{(N_t)}]$ denotes the $Np \times NtL$ time-domain training matrix based on $\{C^{(p)}\}_{p=1}^{Nt}$, $h_0 = [(h_0^{(1)})^T, (h_0^{(2)})^T, \ldots, (h_0^{(Nt)})^T]^T$ presents the $NtL\times1$ equivalent "total" CIR for all Nt transmit antennas, and $V_0 = [v_{0,1}, v_{0,1}, \ldots, v_{0,Np-1}]^T$ stands for the channel's AWGN vector with each element having zero mean and the variance of $\sigma 2$.

3. TFT OFDM System Model in Frequency Domain

The frequency-domain signal model is also valid when Np-point DFT instead of N-point DFT is used to produce the received preamble $D_0 = [D_{0,0}, D_{0,1}, ., D_{0,Np-1}]^T$ in the frequency domain, i.e.,

$$\mathbf{D}_0 = \mathbf{C}_0 H_0 + \mathbf{V}_0. \qquad \textbf{(3)}$$

where $\mathbf{V}_0 = \mathbf{F}_{\mathbf{Np}}\mathbf{V}_\mathbf{0}$ denotes AWGN, $\mathbf{D}_\mathbf{0} = \mathbf{F}_{Np}\, \mathbf{D}_\mathbf{0}$ presents the Np-point DFT of the time-domain received preamble $\mathbf{d}_0$, C_0 = [diag{$C^{(1)}$, diag{$C^{(2)}$,…, diag{$C^{(N_t)}$}] denotes the Np × NtNp frequency-domain training matrix based on {**C**(p)}Nt p=1, and the CFR $\mathbf{H}_\mathbf{0}$ during the preamble can be related to the corresponding CIR $\mathbf{h}_0$ by using (11) as $\mathbf{H}_0 = \mathbf{F}_\mathbf{0}\mathbf{H}_\mathbf{0}$, where

$$F_0 = \begin{bmatrix} F_{Np,L} & \cdots & 0_{Np\times L} \\ \vdots & \ddots & \vdots \\ 0_{Np\times L} & \cdots & F_{Np,L} \end{bmatrix}$$

Since there are NtNp unknown parameters in $\mathbf{H}_0$ and only Np observations in $\mathbf{D}_0$, eq. (3) is an underdetermined problem without unique solution.

However, this problem can be solved by using the relationship between the CFR $\mathbf{H}_0$ and the CIR $\mathbf{h}_0$ as below

$$\mathbf{D_0 = C_0F_0h_0 + V_0 = A_0h_0 + V_0} \qquad \textbf{(4)}$$

where $\mathbf{A}_0 = \mathbf{C_0F_0}$, and the number of unknown parameters is reduced from NtNp to NtL in (20). If Np ≥ NtL, the channel estimation can be achieved by

$$\hat{h} = \boldsymbol{A_0^{\dagger}}\mathbf{D_0} = ((\boldsymbol{C_0F_0})^{\boldsymbol{H}}\boldsymbol{C_0F_0})^{-1}(\boldsymbol{C_0F_0})^{\boldsymbol{H}}\boldsymbol{D_0} \qquad \textbf{(5)}$$

Then, the CFR can be obtained by

$$\hat{h} = \tilde{_{H0}} = \mathbf{F_0}\,\hat{h}0 = \mathbf{F_0}A_0^{\dagger}\mathbf{D_0} \qquad \textbf{(6)}$$

4. JOINT TIME-FREQUENCY CHANNEL ESTIMATION

The time-domain channel estimator is based on time domain signals $\mathbf{d}_0$ and $\mathbf{C_0}$, while the frequency-domain channel estimator depends on the frequency-domain signals $\mathbf{D}_0$ and $\mathbf{C_0}$. Since $\mathbf{D}_0$ ($\mathbf{C}_0$) can be obtained once $\mathbf{d}_0$ ($\mathbf{c}_0$) is known, and vice versa, the time- and frequency-domain channel estimators and can be directly unified by the extracted DFT matrix **F**0. Using the well-known shift property of DFT, it can be derived that $F_{Np}C_0 = C_0F_0 = A_0$. Using this equation, the unification of the optimal design criteria and for the time- and frequency-domain channel estimators, respectively, can be revealed by

$$\mathbf{C_0}^{\mathbf{H}}\mathbf{C_0} = \left(F_{Np}^H A_0\right)^{\mathbf{H}} F_{Np}^H A_0 = A_0^H\left(F_{Np}F_{Np}^H\right)A_0 = A_0^H \qquad \textbf{(7)}$$

4.1. TS-Based Path Delay Estimation

The received TS will be contaminated by the IBI from the previous OFDM data block after multi-path propagation, i.e., the received TS $\mathbf{d}_i = [d_{i,0}\ d_{i,1}\ ....\ D_{i,M-1}]^T$ in the ith TFT-OFDM symbol containing the contributions from the preceding (i − 1)th unknown OFDM should be

$$\mathbf{d_i} = B_{i,ISIC_i} + B_{i-1,IBIX_{i-1}N-M:N-1+V_i} \qquad .$$

To achieve reliable channel estimation in TDS-OFDM, IBI cancellation and cyclic prefix reconstruction of the received TS are required to fully utilize the good autocorrelation property of the TS.

4.2. Pilot-Based Path Coefficients Estimation

For fast time-varying channels which vary even within each OFDM symbol, the nonzero path coefficients $h_{i,n,l}$ with the path delay ɼ = $\{n_l\}_{l=0}^{s-1}$ can be modeled by the Q-order Taylor .

The proposed joint time-frequency channel estimation scheme also differs from the compressive sensing (CS) based channel estimation technique implemented either in the time domain or in the frequency domain in the following aspects:

4.2.1. Our proposal is a linear over determined problem whose solution can be uniquely determined by solving the linear equation array, while the CS-based technique is a nonlinear underdetermined problem whose non-unique solution can be acquired by several convex optimization strategies

4.2.2. Our proposal achieves the final CIR estimate by two sequential steps, i.e., the path delay estimation at first, and then the path coefficients estimation, while the CS-based technique jointly estimate the path delays and path gains in one step.

5. Channel Equalization

Soft-decision aided MMSE channel equalization with iterative ICI cancellation exploiting the extrinsic information from the soft-in soft-out (SISO) channel decoder is the widely used scheme with reliable equalization performance [23]. Here, we adopt the $O(N)$-complexity equalizer.

It should be pointed out that, although the iterative ICI removal based channel equalization above is used in TFTOFDM to remove the ICI over the fast fading channels, it is essentially different from the iterative IBI cancellation method in TDS-OFDM where both the channel estimation and channel equalization are involved and coupled together. If the channel can be assumed quasi-stationary during each TFTOFDM symbol, the iterative ICI removal is not necessary any more since no significant ICI will be introduced. However, even when the channel is static, TDS-OFDM still requires the iterative IBI cancellation, whereby channel estimation and channel equalization are iteratively involved to remove the IBIs as completely as possible.

III. PERFORMANCE ANALYSIS OF TFT-OFDM

The system performances of the proposed TFT-OFDM scheme, including the spectral efficiency, pilot power and the corresponding SNR loss, the equivalent SNR loss due to the cyclic prefix reconstruction, the receiver performance over time-varying channels, and the receiver complexity, are analysed in this section

TABLE: I

Spectral Efficiency Comparison [1]

TS LENGTH	CP-OFDM	TDS-OFDM	DPN-OFDM	TFT-OFDM
K=N/4	**60.00 %**	**80.00 %**	**66.67 %**	**77.66 %**

K=N/8	**77.78 %**	**88.89 %**	**80.00 %**	**86.28 %**
K=N/16	**88.23 %**	**94.12 %**	**88.89 %**	**91.36 %**

A. Spectral Efficiency

One major merit of OFDM is its high spectral efficiency due to the orthogonality between the subcarriers although they are overlapped. The spectral efficiency is defined as the net bit rate over a certain signal bandwidth, i.e., the ideal OFDM system without guard interval or pilots has the spectral efficiency [25]

$$E_{\text{ideal}} = \frac{N\alpha/T}{N/T} = \alpha \text{ (bit/s/Hz)}$$

where 2α denotes the constellation points of the modulation scheme, e.g., $\alpha = 4$ for 16QAM, $N\alpha/T$ stands for the net bit rate, and N/T is the signal bandwidth. However, both the time-domain guard interval and the frequency-domain pilots would reduce the actual spectral efficiency of the practical OFDM systems [25]. So the spectral efficiency of TFT-OFDM is

$$E_{\text{real}} = E_{\text{ideal}} \frac{N-N_P}{M+N} \text{ (bit/s/Hz)}$$

When the same modulation scheme is taken into account, we define the *normalized spectral efficiency* in the form of percentage as below:

$$E_0 = \frac{E_{real}}{E_{ideal}} = \frac{N-(Q+1)(2d+1)S}{M+N} \times 100\%$$

The key advantage of TDS-OFDM over CP-OFDM is the increased spectral efficiency since no pilot is used in TDSOFDM, but the IBIs between the TS and the OFDM data block deteriorate the system performance over fast fading channels. The DPN-OFDM solution can solve the performance loss problem, but the doubled TS length obviously reduces the spectral efficiency. Regarding to the TFT-OFDM scheme proposed in this paper, the TS has the same length as the guard interval in TDS-OFDM and CP-OFDM systems, while only N_p = 120 frequency-domain grouped pilots can be configured with some design margin. For digital broadcasting systems like DVB-T2 [19], typically N = 4096 (4K mode) is used (In Chinese national digital television standard [9], N = 3780 is adopted), which means that the grouped pilots only occupy less than 3% of the signal bandwidth. Table I shows the spectral efficiency comparison between CP-OFDM, TDS-OFDM, DPN-OFDM and the proposed TFTOFDM schemes.

TABLE: II
The SNR loss due to pilot power boosting [1]

Guard Interval Length	CP-OFDM	TFT-OFDM
K=N/4	**0.77 dB**	**0.098 dB**
K=N/8	**0.40 dB**	**0.098 dB**

K=N/16	**0.21 dB**	**0.098 dB**
K=N/32	**0.10 dB**	**0.98**

B. Pilot Power and the Corresponding SNR Loss

In standard CP-OFDM systems, the power boosting technique [19] is commonly used to increase the average power of the pilots to achieve more reliable channel estimation, which leads to the equivalent SNR loss at the receiver

$$\text{SNR}_{\text{loss}} = 10\log_{10}\left(\frac{N_P E_p + (N - N_P)E_D}{N E_D}\right)$$

where E*p* and E*d* denote the average power of the pilot and data, respectively. Such SNR loss is not negligible, especially when the pilot number is large in CP-OFDM. However, the proposed TFT-OFDM scheme only requires a small amount of grouped pilots, so the SNR loss will be small. Table II compares the SNR loss in CP-OFDM and TFT-OFDM, where the case that E_p is 2.5 dB higher than E_d as specified by DVBT2 [19] is taken as an example. It can be found that the SNR loss in CP-OFDM is relative to the guard interval length, and 0.40 dB SNR loss will be introduced when $M = N/8$, while the negligible SNR loss in TFT-OFDM is 0.098 dB, which is independent of the guard interval length. Furthermore, if the same SNR loss is permitted in TFTOFDM as that in CP-OFDM, the pilot power in TFT-OFDM could be much higher than that in CP-OFDM. As shown in Table III, if the SNR loss of 0.40 dB is affordable when $M = N/8$, the boosted pilot power in TFT-OFDM could be 3.86 dB higher than that in CP-OFDM, which is beneficial for more accurate channel estimation. Due to the lower pilot occupation ratio in TFT-OFDM than CP-OFDM, the pilot power boosting technique is more efficient for more reliable channel estimation, e.g., the boosted pilot power from 2.5 dB to 3.0 dB in CP-OFDM is equivalent to the boosted pilot power from 6.36 dB to 7.20 dB in TFT-OFDM.

C. SNR Loss Due to Cyclic Prefix Reconstruction

The cyclic prefix reconstruction with the OLA method [4] would result in the noise enhancement effect caused by removing the "tail" of the OFDM data block to its head. Thus, similar to ZP-OFDM, TFT-OFDM also suffers from the SNR loss

$$\text{SNR}_{\text{TFT, loss}} = 10\log_{10}\left(\frac{M+N}{N}\right)$$

When $M = N/8$, the SNR loss is 0.51 dB, and when $M = N/16$, the SNR loss is 0.26 dB. Thus, the SNR loss in TFT-OFDM will be smaller than that in TDS-OFDM.

D. Receiver Performance over Time-Varying Channels

Compared with TDS-OFDM, TFT-OFDM could improve the receiver performance at the cost of marginally reduced spectral efficiency due to the small amount of the grouped pilots. When equalizing the OFDM data block between two adjacent TSs over fast fading channels, only the linear interpolation or other more complicated interpolation methods can be used to track the channel variation [26]. However, the proposed joint time-frequency channel estimation can accurately track the fast time-varying channel during the OFDM block by using the scattered pilots in the frequency domain. TDS-OFDM requires that the IBIs between the TS and OFDM block should be removed completely, leading to the mutually conditional relationship between channel estimation and channel equalization, and performance loss is unavoidable over fast fading channels. However, in TFT-OFDM, the joint time-frequency channel estimation is achieved by using the received "contaminated" TS without IBI cancellation and the frequency domain grouped pilots, so the channel estimation performance is independent of the channel equalization quality.

IV. SIMULATION RESULTS AND DISCUSSION

Simulations were carried out to investigate the performance of the proposed TFT-OFDM transmission scheme. The signal bandwidth was 7.56 MHz at the central radio frequency of 770 MHz, and the subcarrier spacing was 2 kHz. The modulation scheme 64QAM was adopted. Other system parameters were consistent with those specified as $N = 3780$, $M = 420$, $N_{group} = 40$, $Q = 1$, $d = 1$, $S = 20$, $J_0 = 3$. Based on the fact that now days almost all OFDM based systems use channel coding for reliable performance, we adopted the powerful low-density parity-check (LDPC) code with the block length of 64, 8000 bits and code rate of 2/3 as specified by the standard [19]. The maximum Doppler spread of 20 Hz and 100 Hz were considered, which corresponded to the relative receiver velocity of 28 km/h and 140 km/h @ 770 MHz, respectively. In the simulations, we assumed *M* equally spaced combo type pilots were used in CP-OFDM, since it has been proved that such scheme could achieve the best channel estimation performance under static channels. The classical iterative algorithm in [7] was used for TDS-OFDM. DPN-OFDM adopted the receiver algorithm proposed in [15].

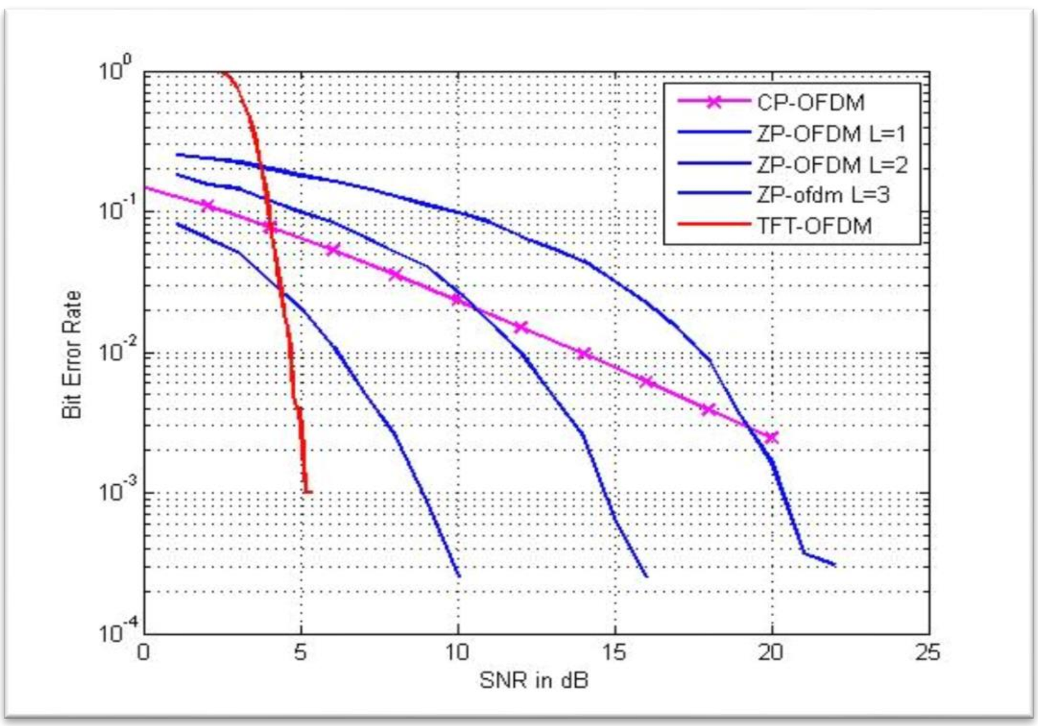

Fig.4. BER performance comparison between the proposed TFT-OFDM scheme and the traditional schemes

The cyclic postfix OFDM used the PN sequence as the unique word [13], and the channel estimation method in [12] was used. Figure 4 compares the coded bit error rate (BER) performance of the conventional CP-OFDM, TDS-OFDM, and cyclic postfix OFDM schemes with the proposed TFT-OFDM solution over the AWGN channel. The ideal channel estimation is assumed for all those systems. We can find that TFT-OFDM and TDS-OFDM have very close BER performance, and they have the SNR gain of 0.18 dB compared with CP-OFDM. The reason is that, the equivalent SNR at the receiver is reduced by the large amount of pilot with boosted power in CP-OFDM.

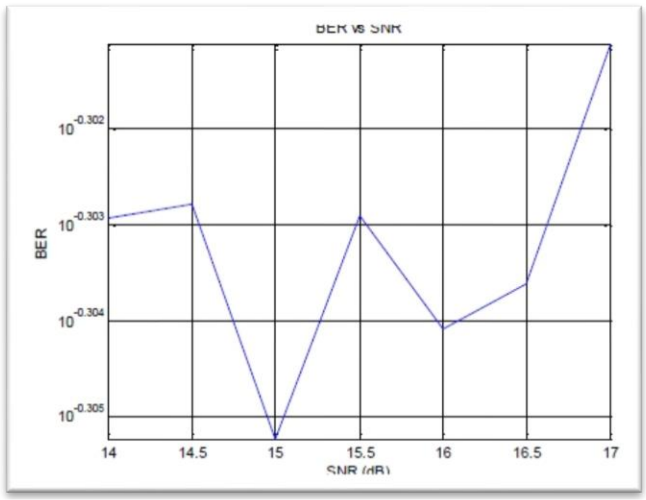

Fig.5 BER performance proposed new TFT-OFDM scheme.

The performance of CP-OFDM is between that of TDS-OFDM and DPN-OFDM, while the proposed TFT-OFDM scheme has superior BER performance to those three conventional OFDM transmission schemes. For example, when the BER equals to 10^{-4}, TFT-OFDM outperforms DPN-OFDM, CP-OFDM and TDS-OFDM by the SNR gain of 0.95 dB, 1.15 dB and 2.40 dB respectively. Compared with DPN-OFDM, CP-OFDM and TDS-OFDM, the SNR gain achieved by TFT-OFDM is increased to be about 1.15 dB, 2.25 dB and 4.40 dB, respectively. Compared with CP-OFDM and DPN-OFDM, TFT-OFDM achieves the performance improvement because the proposed joint channel estimation can accurately track the channel variation, and ICI is removed before the frequency domain equalization.

V. CONCLUSION

Thus, a novel OFDM-based transmission scheme called TFT-OFDM, whereby the training information exists in both time and frequency domains. The corresponding joint time-frequency channel estimation utilizes the time domain TS without interference cancellation to estimate the channel path delays, while the channel path coefficients are acquired by using the pilot groups scattered within the OFDM symbol. In this way, the conventional iterative interference cancellation algorithm with high complexity and poor performance is avoided and the variation of the fast fading channels within every TFT-OFDM MIMO symbol can be well tracked. The grouped pilots in TFT-OFDM MIMO occupy only about 3% of the signal bandwidth. Therefore, high spectral efficiency as well as good performance is achieved over fast time-varying channels. Simulation results indicate that the proposed scheme enjoys the BER performance close to the theoretical ergodic capacity. The proposed TFT-OFDM MIMO scheme can be also directly applied in multiple access systems in both the uplink and downlink, and the principle of joint time-frequency processing behind TFT-OFDM can be adapted for other OFDM MIMO systems (including large- and small scale systems) to achieve higher spectral efficiency as well as more reliable performance over severe fading channels.

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