 Complexity Reduced Transmit Diversity Scheme for Time Domain Synchronous OFDM Systems

Zhaocheng WANG†a, Nonmember, Jintao WANG†, Member, and Linglong DAI†, Nonmember

SUMMARY This paper proposes a novel scheme to reduce the complexity of existing transmit diversity solutions to time domain synchronous OFDM (TDS-OFDM). The space shifted constant amplitude zero autocorrelation (CAZAC) sequence based preamble is proposed for channel estimation. Two flexible frame structures are proposed for adaptive system design as well as cyclicity reconstruction of the received inverse discrete Fourier transform (IDFT) block. With regard to channel estimation and cyclicity reconstruction, the complexity of the proposed scheme is only around 7.20% of that of the conventional solutions. Simulation results demonstrate that better bit error rate (BER) performance can be achieved over doubly selective channels.

key words: transmit diversity, time domain synchronous OFDM (TDS-OFDM), channel estimation, cyclicity reconstruction, low complexity

1. Introduction

In wireless communication systems, space diversity techniques are widely adopted to provide reliable transmission over fading channels [1]. Compared with receive diversity where every receiver has increased hardware complexity, transmit diversity attracts more attention in broadcasting systems where thousands of users are served by one or several transmit towers. In the recently issued European standard for the second generation digital video terrestrial broadcasting (DVB-T2) systems, transmit diversity is also recommended to provide high performance over wireless broadcasting channels [2].

Time domain synchronous OFDM (TDS-OFDM) is the key technology of the Chinese national digital television terrestrial broadcasting (DTTB) standard [3]. Classical space-time block coding (STBC) or space-time-frequency block coding (STFBC) can be directly used to realize transmit diversity for TDS-OFDM systems, but channel estimation is the essential issue for real applications. Inspired by the space-time (ST) coded training sequence (TS) [4], frequency-domain channel estimation based on ST coded pseudo-random-noise (PN) sequences was proposed [5], where the hypothesis of static channels along two consecutive signal frames was required. Similarly, space-frequency (SF) coded TS was used for channel estimation in [6], where constant channel frequency response (CFR) over two adjacent pilots was assumed. In both cases, channel estimation was achieved at the cost of high complexity. On the other hand, they both utilized the estimated channel impulse response (CIR) to remove the interferences caused by TS and then reconstruct the cyclicity of the received inverse discrete Fourier transform (IDFT) block [7], which resulted in poor reconstruction quality due to channel estimation errors as well as high signal processing complexity.

In this paper, an improved transmit diversity scheme with low complexity is proposed for TDS-OFDM systems, where channel estimation and cyclicity reconstruction can be efficiently realized. Specifically, the contributions of this paper are listed as below: 1) The preamble using space shifted constant amplitude zero autocorrelation (CAZAC) sequence is proposed for channel estimation; 2) Two flexible frame structures are proposed to facilitate channel estimation as well as cyclicity reconstruction of the received IDFT block, where no prior channel state information is required; 3) The proposed transmit diversity scheme achieves better bit error rate (BER) performance than conventional solutions over doubly selective channels.

The remainder of this paper is organized as follows. Section 2 proposes the transmit diversity scheme for TDS-OFDM systems. The corresponding receiver design is presented in Sect. 3, together with the spectral efficiency and computational complexity analysis. Section 4 investigates the performance of the proposed scheme. Finally, conclusions are drawn in Sect. 5.

2. TDS-OFDM Based Transmit Diversity

In this section, the TS and two flexible frame structures are presented, based on which the architecture of the TDS-OFDM transmit diversity scheme is outlined.

2.1 Space Shifted CAZAC Training Sequence

PN sequence instead of cyclic prefix (CP) is used as the guard interval as well as the TS in the single-antenna TDS-OFDM scheme, where PN sequence has good but not ideal autocorrelation property. Two conventional TDS-OFDM transmit diversity schemes utilize ST/SF coded PN sequence as TS [5], [6]. To further improve the performance, CAZAC sequence with perfect autocorrelation property is recommended as the TS in this paper. According to [8], the CAZAC sequence \( c_i \) with length \( N_p \) for the \( i \)th transmit an-
tenna (denoted by Tx $i$) can be selected as

$$c_i(n) = \exp\left[j\pi r_i n^2/N_p\right], \quad 0 \leq n \leq N_p - 1,$$

where $r_i$ is relatively prime to $N_p$. The ideal autocorrelation of CAZAC sequence can be denoted by

$$R_t(k) = \sum_{n=0}^{N_p-1} c_i^*(n)c_i((n+k))N_p = \begin{cases} N_p, & k = 0, \\ 0, & k \neq 0, \end{cases}$$

where $(\cdot)^*$ means complex conjugate, $(\cdot)_{N_p}$ denotes the modulo-$N_p$ operation.

It is clear that CAZAC sequence has the lowest peak-to-average power ratio (PAPR) due to the constant amplitude in the time domain. Moreover, unlike the PN sequence whose frequency response is not flat, the CAZAC sequence also has constant amplitude in the frequency domain, which is optimal for channel estimation [9].

To distinguish the wireless channel associated with each transmit antenna, CAZAC sequences $c_i$ and $c_{i+1}$ for the adjacent $i$th and $(i+1)$th transmit antennas have a constant shift of $K$ symbols, which can be defined by

$$c_{i+1} = [c_i(K), c_i(K+1), \ldots, c_i(N_p-1), c_i(0), \ldots, c_i(K-1)].$$

Specifically, for transmit antennas located at different places in the space dimension, $c_{i+1}$ can be regarded as the space shifted version of $c_i$.

### 2.2 Frame Structure for Block Transmission

The frame structure of the conventional solutions [5],[6] is shown in Fig. 1(a). The TS is protected by its cyclic prefix (CP), which is similar to the dual PN-sequence padding scheme [10]. As shown in Figs. 1(b) and (c), two flexible frame structures are proposed for block transmission in TDS-OFDM transmit diversity scheme based on the space shifted CAZAC sequences. Without the loss of generality, the case of two transmit antennas is assumed, i.e., $i = 0, 1$.

The basic block transmission unit for the proposed scheme is the superframe. Each superframe is composed of one CAZAC based preamble and $M$ subframes. The $m$th subframe is composed of the IDFT block $x_m$ with length $N$ and the postfix with length of $K$, thus one subframe is equivalent to an OFDM symbol. The number of subframes $M$ could be adaptively adjusted according to the coherent time of the wireless channel, e.g., lower Doppler spread leads to larger $M$, and vice versa. Specifically, two types of frame structures are provided:

1. **Type 1: Identical postfix based frame structure (IPFS).**
   
   As shown in Fig. 1(b), the preamble is consisted of one CAZAC sequence $c_i$ and its postfix $c_{i,0}$, which is the second half of $c_i$. In the following $M$ subframes, every IDFT block is followed by the identical postfix $c_{i,0}$, which is exactly the same as the postfix in the preamble.

2. **Type 2: Alternate postfix based frame structure (APFS).**
   
   As shown in Fig. 1(c), the preamble is just the CAZAC sequence, which can be divided into two halves: the first half $c_{i,0}$ and the second half $c_{i,1}$. In the following $M$ subframes, each subframe alternatively selects $c_{i,0}$ or $c_{i,1}$ as its postfix. More specifically, the subframes with even indices choose $c_{i,0}$ as their postfix, while subframes with odd indices select $c_{i,1}$ as their postfix. The postfix of the last subframe should be $c_{i,1}$.

The comparison of these two types of frame structures will be discussed in detail in Sect. 3.5.

### 2.3 Transmit Diversity Scheme for TDS-OFDM

Based on the proposed TS and frame structures above, the
transmit diversity scheme for TDS-OFDM system is shown in Fig. 2. Without the loss of generality, we focus on the system with \( N_T = 2 \) transmit antennas (denoted by Tx 1 and Tx 2, respectively) and \( N_R = 1 \) receive antenna (denoted by Rx).

At the transmitter, the space-frequency block coding (SFBC) [11] is applied before IDFT. Then the space shifted CAZAC sequences are used to construct the superframe signals as shown in Figs. 1(b) or (c) for signal transmission via Tx 1 and Tx 2.

The multi-path CIR between the \( i \)th transmit antenna and the receive antenna during the \( m \)th block of a superframe, \( h_{i,m} \), is modeled as an \( L \)-order finite impulse response filter. The maximum delay spread \( L \) of the channel is assumed to be not larger than the postfix length \( K \) in IPFS or APFS, i.e., \( L \leq K \). The received IDFT block in the \( m \)th subframe is given by

\[
r_m = \sum_{i=0}^{1} r_{i,m} = \sum_{i=0}^{1} y_{i,m} + w_m = \sum_{i=0}^{1} x_{i,m} * h_{i,m} + w_m, \tag{4}
\]

where \( y_{i,m} \) denotes the response of the transmitted IDFT block \( x_{i,m} \) passing through the channel \( h_{i,m} \), \( * \) is the linear convolution operator, and \( w_m \) is the additive white Gaussian noise (AWGN) with zero mean and the variance of \( \sigma^2 \).

At the receiver, the received CAZAC sequences and the corresponding local CAZAC sequences are used for channel estimation. The proposed frame structure is utilized to reconstruct the cyclicity of the received IDFT block with low complexity. After that, the space-frequency decoding can finally restore the transmitted signal with diversity gain.

3. Receiver Design for Transmit Diversity

Channel estimation and cyclicity reconstruction algorithms at the receiver are addressed in this section. The performance of the proposed scheme is also analyzed in the sequel.

3.1 Channel Estimation

Both in the proposed IPFS and APFS, the postfix of the last subframe belonging to the previous superframe can naturally serve as the CP of the CAZAC sequence in the preamble. Assuming that channel is quasi-static during one superframe [12], the received CAZAC sequence \( \mathbf{d} \) takes the form

\[
\mathbf{d} = \sum_{i=0}^{1} \mathbf{d}_i + \mathbf{w} = \sum_{i=0}^{1} \mathbf{c}_i \otimes h_{i,m} + \mathbf{w}, \tag{5}
\]

where \( \otimes \) denotes the circular convolution, and \( \mathbf{w} \) is the noise term.

Circular convolution between one local CAZAC sequence \( \mathbf{c}_0 \) with the received sequence \( \mathbf{d} \) will generate the orthogonally separable CIR estimates \( \{ \hat{h}_{i,m} \}_{i=0}^{1} \) in the \( j \)th subframe as follows

\[
\hat{h}_j = \mathbf{c}_0 \otimes \mathbf{d} = \mathbf{c}_0 \otimes \left( \sum_{i=0}^{1} \mathbf{c}_i \otimes h_{i,m} + \mathbf{w} \right) = N_p (\mathbf{h}_0, m + \mathbf{h}_1, m \delta[n-K]) + \mathbf{w}', \tag{6}
\]

where \( \delta(n) \) denotes the Kronecker delta function, \( \mathbf{w}_m' = \mathbf{c}_0 \otimes \mathbf{w} \) is the noise term, and the perfect autocorrelation of the CAZAC sequence denoted by (2) is used, e.g., \( \mathbf{c}_0 \otimes \mathbf{c}_1 = N_p \delta(n-K) \). Intuitively, \( \mathbf{h}_{i,m} \) is shifted by \( K \) symbols in the time dimension. If \( L-1 \leq K \) and \( N_p \geq K+L-1 \), the CIR estimates \( \{ \hat{h}_{i,m} \}_{i=0}^{1} \) can be directly extracted from \( \hat{h}_j \) in (6).

The expectation of the channel estimation error is

\[
\text{Mean} = \frac{1}{N_p} E(\hat{h}_{i,m} - h_{i,m}) = \frac{1}{N_p} E(\mathbf{c}_0 \otimes \mathbf{w}) = 0, \tag{7}
\]

where \( E(\cdot) \) is the expectation operator. The mean square error (MSE) of the channel estimation error is

\[
\text{MSE} = \frac{1}{N_p} E\left( (\hat{h}_{i,m} - h_{i,m})^H (\hat{h}_{i,m} - h_{i,m}) \right) = \frac{1}{N_p} E\left( \mathbf{c}_0 \otimes \mathbf{w}_m \right)^2 = \frac{\sigma^2}{\lambda}, \tag{8}
\]

where \( \cdot^H \) denotes the Hermitian transpose, \( \lambda \) is the transmitted signal power, and \( \lambda = \frac{P_s}{\sigma^2} \) is the averaged signal-to-noise ratio (SNR).

To reduce the complexity, the time-domain circular convolution in (6) can also be implemented by

\[
\hat{h}_j = \text{IFFT} \left( \frac{\text{FFT} (\mathbf{d} \otimes \mathbf{c}_0)}{\text{FFT} (\mathbf{c}_0)} \right), \tag{9}
\]
where IFFT denotes inverse fast Fourier transform. FFT($\mathbf{c}_0$) can be pre-stored at the receiver, thus only one FFT and one IFFT operations are required.

When the channel is slowly time-varying and can be considered to be static over the duration of one superframe, the preamble based CIR estimate obtained from (6) can directly be used for decoding the $M$ subframes following the preamble. In other words, $\{\hat{\mathbf{h}}_{i,m}(m)\}_{m=0}^{M-1}$ extracted from $\hat{\mathbf{h}}^{(j)}$ are all identical. However, if the channel is varying fast, this mechanism would lead to inaccurate channel estimation and consequently poor BER performance. The following low-complexity linear interpolation can be used to reduce the channel estimation error

$$\hat{\mathbf{h}}_{i,m}(m) = \hat{\mathbf{h}}^{(j)} + \frac{2m+1}{2M}(\hat{\mathbf{h}}^{(j+1)}(m) - \hat{\mathbf{h}}^{(j)}), \quad 0 \leq m \leq M-1,$$

(10)

where $\hat{\mathbf{h}}_{i,m}(m)$ is the interpolated CIR for the $m$th subframe of the $j$th superframe. Then $\{\hat{\mathbf{h}}_{i,m}(m)\}_{m=0}^{M-1}$ can be extracted from $\hat{\mathbf{h}}_{i,m}(m)$ instead of $\hat{\mathbf{h}}^{(j)}$. Note that in broadcasting systems where the data streaming is transmitted successively, the preamble-based channel estimates $\hat{\mathbf{h}}^{(j)}$ and $\hat{\mathbf{h}}^{(j+1)}$ in two adjacent superframes can be obtained before the interpolation. More sophisticated signal processing techniques, such as Kalman filter or higher order interpolation [13], can be also used for more accurate channel tracking at the cost of increased complexity.

3.2 Cyclicity Reconstruction

For both the ST and SF coded TS in TDS-OFDM transmit diversity schemes [5], [6], the CIR estimates were used for interference removal and consequently for the cyclicity reconstruction of the received IDFT block [7]. However, the computational complexity is high, especially when the transmit antenna number $N_T$ becomes large.

To solve this problem, the proposed frame structures can be used for cyclicity reconstruction with low complexity. We take IPFS as an example in Fig. 3, since the similar procedure is applicable to APFS.

In Fig. 3(a), the received $m$th subframe from the $i$th transmit antenna is shown by $\mathbf{r}_{i,m} = \{r_{i,m}(n)\}_{n=0}^{N-1}$, whereby $\{r_{i,m}(n)\}_{n=0}^{N-1}$ is the received IDFT block and $\{r_{i,m}(n + N)\}_{n=0}^{K-1}$ is the received postfix of this subframe. The received preamble is presented as $\mathbf{r}_{i,0} = \{r_{i,0}(n)\}_{n=0}^{N_P-1}$, whereby $\{r_{i,0}(n + N_P)\}_{n=0}^{K-1}$ is the received postfix of the preamble. The following one step add-subtraction operation of three received blocks will accomplish the joint cyclicity reconstruction of the received IDFT block

$$y_{i,m}(n) = \begin{cases} \sum_{i=0}^{m} [r_{i,m}(n) + r_{i,m}(n+N) - r_{i,0}(n+N_P)], & 0 \leq n \leq K-1, \\ \sum_{i=0}^{m} r_{i,m}(n), & K \leq n \leq N-1. \end{cases}$$

(11)

If the wireless channel is quasi-static during one superframe, the “tails” caused by multi-path propagation shown by the shadows with the same form in Figs. 3(a) and (b) can be assumed to be identical [14]. Therefore, (11) leads to

$$y'_{i,m}(n) = \sum_{i=0}^{1} y'_{i,m}(n), \quad 0 \leq n \leq N-1,$$

(12)

where

$$y'_{i,m}(n) = \begin{cases} y_{i,m}(n + N) + y_{i,m}(n), & 0 \leq n < K - 1, \\ y_{i,m}(n), & K \leq n \leq N - 1. \end{cases}$$

(13)

The procedure to generate $y'_{i,m} = \{y'_{i,m}(n)\}_{n=0}^{N-1}$ in (13) is illustrated in Fig. 3(b), whereby $i = 1$ is taken as an example. The consequent process to obtain $y'_m = \{y'_{i,m}(n)\}_{n=0}^{N-1}$ in (12) is illustrated in Fig. 3(c). Clearly, $\{y'_m\}_{m=0}^{K-1}$ is the cyclicity reconstructed IDFT block for Tx $i$, thus $y'_m$ also holds the cyclicity property. Therefore, joint cyclicity reconstruction of the received IDFT block is achieved via the simple add-subtraction operation (11) with substantially low complexity, whereby no CIR information is required.

Since the last postfix of the previous $(i - 1)$th superframe naturally serves as the cyclic prefix of the preamble of the current $i$th superframe in the APFS in the broadcasting data streaming, the add-subtraction operation for cyclicity reconstruction in IPFS proposed above can be similarly applied to APFS. The only difference is that, the block used for subtraction in (11) is not the received postfix of the preamble.
in IPFS, but the first or second half of the received CAZAC sequence. More specifically, if $e_{i,0}$ is the postfix of one subframe, the first half of the received CAZAC sequence should be used for subtraction in (11), while the second half of the received CAZAC sequence will be used instead if $e_{i,1}$ is the postfix of the subframe.

Note that the joint cyclicity reconstruction method (11) can be directly used for more transmit antennas scenarios. What is more, this can be achieved within only one step, no matter how many transmit antennas are supported, thus the complexity is low.

After channel estimation and cyclicity reconstruction have been obtained, $y_m'$ and $[\mathbf{h}_m]_{i=0}^1$ are fed into the space-frequency decoding block to restore the transmitted data with diversity gain.

### 3.3 Spectral Efficiency

The TS as well as the CP/postfix would decrease the spectral efficiency of TDS-OFDM systems [14]. Table 1 compares the spectral efficiency of the conventional solutions [5], [6] with the proposed IPFS and APFS schemes.

<table>
<thead>
<tr>
<th>ST/SF coded TS [5],[6]</th>
<th>Proposed IPFS</th>
<th>Proposed APFS</th>
</tr>
</thead>
<tbody>
<tr>
<td>$N/(N + 2K)$</td>
<td>$N/(N + K + 3K/M)$</td>
<td>$N/(N + K + 2K/M)$</td>
</tr>
</tbody>
</table>

The ST/SF coded TS solutions have the same efficiency $N/(N + 2K)$ due to their identical frame structure. The proposed APFS has the spectral efficiency of $N/(N + K + 2K/M)$, which is higher than that of the conventional solutions when the subframe number $M > 2$. The spectral efficiency $N/(N + K + 3K/M)$ of the proposed IPFS is higher than that of the traditional solutions if $M > 3$, but slightly lower than that of the APFS. However, the loss is small, since $N$ is usually much larger than $K$. For example, when $N=3780$, $K=256$ and $M=2$, the spectral efficiency is 85.52% for IPFS, while 88.07% for APFS. When $M$ is increased to 3, the spectral efficiency is 88.07% for IPFS, which is 1.79% lower than the efficiency of 89.86% for APFS. When the number of subframe becomes large under slowly time-varying channels, IPFS and APFS have very close spectral efficiencies, both of which are higher than that of the conventional solutions.

### 3.4 Computational Complexity

Regarding to channel estimation and cyclicity reconstruction, Table 2 compares the computational complexity between the proposed scheme and the conventional solutions based on ST/SF coded TS [5],[6]. All the operations in Table 2 are based on $N_p$-symbol block.

The detail of Table 2 can be explained as follows. For recovering the transmitted signal from one received IDFT block, the SF coded method [6] needs two times of FFT operations to obtain the frequency-domain received TSs surrounding the corresponding IDFT block, and two times of divisions are used to divide the pre-stored TSs in the frequency domain to derive the corresponding CFRs around the IDFT block. Then $2N_T$ times of IFFT operations are used to convert the antenna-specific CFRs into CIRs. Next, $2N_T$ times of convolutions between the estimated CIRs and the corresponding local time-domain TSs are required to calculate the interferences between the IDFT block and its adjacent TSs, and $2N_T$ times of subtraction are used for interference removal. Finally, one add operation is used for cyclicity reconstruction using the overlap-add (OLA) method [10]. Similarly, the complexity of the ST coded scheme [5] can be calculated as shown in Table 2.

On the other hand, the proposed scheme only needs one time-domain circular convolution to acquire the CIR estimates (which can be implemented by FFT/IFFT shown in (9) for more efficient computation), and cyclicity reconstruction is accomplished by the one-step add-subtraction operation (11) without any prior CIR information.

Regarding to the computational complexity in terms of multiplications, [5] needs $(1 + 6.5N_pN_T)\log_2N_p + N_p + 10N_pN_T$ multiplications, [6] needs $(1 + 7N_pN_T)\log_2N_p + 2N_p + 10N_pN_T$ multiplications, while the proposed scheme only requires $N_p\log_2N_p + N_p$ multiplications. Take the typical values $N_p = 256$, $N_T = 2$ for example, the complexity of the proposed scheme is only 6.71% of the SF coded TS solution [6], and 7.20% of the ST coded PN solution [5].

### 3.5 Discussion about Proposed Frame Structures

IPFS and APFS have identical channel estimation and quite similar cyclicity reconstruction algorithms. They also have the identical computational complexity, but IPFS has slightly lower spectral efficiency than that of APFS. However, IPFS is preferred for TDS-OFDM transmit diversity scheme due to its following merits:

1. **Flexible adaptation to channel variation.** It is required that the subframe number $M$ should be an integer times of the transmit antenna number $N_T$ for APFS, i.e., $M = uN_T$, where $u$ is a positive integer. However, for APFS, $M$ could be an arbitrary number, leading to more flexible frame structure design. Thus IPFS is more adaptive to the time-varying channels due to the arbitrary selection of $M$. Note that APFS can be also designed to support any subframe number larger than $N_T$ by shifting the CAZAC sequences in adjacent superframes. For example, when $N_T = 3$, if $M = 4$ subframes are to be
supported, it is required that the CAZAC sequences for adjacent two superframes should have the left shift of \(N_p/3\) symbols in the time dimension, and \(2N_p/3\) symbols should be shifted to support \(M = 5\) subframes. The left shift value is calculated according to the principle of guaranteeing the postfix of the last subframe is just the last part of the preamble. However, this mechanism would lead to the variable preamble in the superframe, thus more complex control must be required.

2. **Superior subframe structure.** The identical postfixes in IPFS can be regarded as “unique words” [15]. Actually, one postfix acts as CP of the following subframe, and this property can be used for low complexity frequency domain equalization for both single-carrier [15] and multi-carrier transmissions [16]. In APFS, such property does not exist. Meanwhile, many timing and carrier synchronization algorithms for OFDM systems require cyclic property in the frame structure [17]. The identical postfixes in IPFS naturally provide such property, while APFS can not.

3. **Easy extension to more transmit antennas.** The preambles both in IPFS and APFS can be prolonged to support more transmit antennas to achieve higher diversity gain. However, even when the preamble in every superframe can be changed in APFS, the minimum supportable subframe number is \(N_r\), i.e., \(M = 4\) is the smallest number required by APFS in the four transmit antennas scenario. In such cases, APFS may be not able to work if the channel is varying fast. On the other hand, \(M\) could be selected as small as \(M = 1\) for SPSF to combat fast time-varying channels.

Due to these merits discussed above, IPFS is more suitable for TDS-OFDM transmit diversity. In addition, IPFS can also be used for single-carrier block transmission (whereby postfix is the unique word), which means a uniform frame structure is provided in this paper for both single-carrier and multi-carrier transmissions.

### 4. Simulation Results

Simulations were carried out to investigate the performance of the proposed transmit diversity scheme for TDS-OFDM systems. The major system parameters are configured as below: 1) The signal bandwidth is 8 MHz at the central frequency of 770 MHz in the television ultra high frequency (UHF) band; 2) \(N = 3780, K = 256, M = 2; 3) QPSK modulation is selected; 4) The symbol rate is 7.56 Msps, and the sub-carrier spacing is 2 kHz; 5) The DTTB test channel model Brazil E [18] with the maximum delay of 2 μs and the ITU defined channel model Vehicular B [19] with the maximum delay of 20 μs were used in the simulation. For Rayleigh fading channels, we use the maximum Doppler spread of 20 Hz and 100 Hz with the corresponding receiver velocity of 28 km/h (slow time-varying) and 140 km/h (fast time-varying) @770 MHz, respectively.

The proposed scheme is compared with other three types of systems: the single-antenna TDS-OFDM system [10], the ST coded PN sequence [5] and the SF coded TS [6] based TDS-OFDM transmit diversity systems, in terms of BER performance as a function of SNR.

Figure 4 compares the BER performance between the proposed IPFS and APFS over Brazil E channel. In both cases of 28 km/h and 140 km/h receiver velocity, IPFS and APFS have very close performance due to the identical channel estimation algorithm and similar cyclicity reconstruction algorithms. Therefore, considering the superior merits of IPFS to APFS as mentioned in Sect. 3.5, it is recommended to adopt IPFS in the actual implementation. Only the proposed IPFS scheme (denoted by “Proposed” for simplicity) is simulated in the following study.

Figures 5 and 6 show the BER performance over the weakly frequency-selective Brazil E channel with the receiver velocity of 28 km/h and 140 km/h, respectively. Compared with the single-antenna scheme [10], obvious BER improvement can be achieved by transmit diversity. Figure 5 indicates that both the ST and SF coded TS based solutions [5, 6] work well under the weakly frequency-selective and slow fading channels, while the proposed scheme has the best performance. Figure 6 shows that the performances of all schemes degrade in fast fading environment, especially for the ST coded TS based solution due to its hypothesis has been severely deviated. It also shows that the proposed scheme still has the best performance.

Figures 7 and 8 present the BER performance over the deeply frequency-selective Vehicular B channel with the receiver velocity of 28 km/h and 140 km/h, respectively. The SF coded TS based transmit diversity solution [6] works even worse than the single-antenna system without transmit diversity [10] over the deeply frequency-selective channels like Vehicular B. That is because the CFR of the Vehicular B channel over adjacent subcarriers may have large differences, thus the assumption of [6] does not hold any more.
Similar observations as the case of Brazil E channel can be found, that the proposed scheme has superior performance to the conventional solutions.

Compared with the SF coded TS based solution [6] whereby the CIR is estimated at every IDFT block, the better performance of the proposed scheme is due to the fact that only half of the subcarriers in TS is effectively used for frequency-domain CFR estimation in [6]. That is to say, only the CFR over 128 out of the total 3780 subcarriers are estimated, which might be insufficient for channel estimation, especially when the channel is deeply frequency-selective. On the other hand, the assumption of identical CFR over adjacent two subcarriers of the TS (more specifically, $2 \times 3780/256 \approx 30$ actual subcarriers of the IDFT block) further introduces large errors for channel estimation. However, the proposed time-domain channel estimation directly acquires the CIRs, without the assumption of [6].

The ST coded PN based solution [5] assumes the constant CIRs over two adjacent signal frames, which is equivalent to the hypothesis of quasi-static channel during one superframe in the proposed scheme with the subframe number $M = 2$. However, the channel estimation of the proposed scheme is better than that of [5]. The conventional solution [5] uses two received TSs in two consecutive signal frames (see Fig. 1(a) in Sect. 2) corrupted by quite different channels, but assumes these channels to be identical. Therefore, the estimated channel result is quite different with the actual channels during both frames. This would lead to severe performance degradation for both channel estimation and cyclicity reconstruction. However, for the proposed scheme, the preamble-based channel estimation can achieve relatively more accurate estimates of the channels over the
preambles in the superframes, and channel variation can be partly compensated by the interpolation in (10), thus more reliable channel estimation can be achieved.

In addition, the performance of the proposed cyclicity reconstruction is only affected by the channel variation, while the conventional way in [5, 6] is influenced by both the channel fluctuation and channel estimation errors.

5. Conclusions

A complexity reduced transmit diversity scheme is proposed for TDS-OFDM systems in this paper. The space shifted CAZAC sequence is proposed for channel estimation. Two flexible frame structures (IPFS and APFS) are proposed to increase the spectral efficiency and reconstruct the cyclicity of the received IDFT block without prior CIR information. IPFS is preferred due to its robustness to channel variation. Compared with the existing solutions, the proposed scheme achieves much lower complexity as well as better BER performance over doubly selective fading channels. The proposed scheme could be directly extended to support more antennas for both single-carrier and multi-carrier transmissions in the uplink or downlink. The ongoing research is investigating the preamble-based channel estimates as the initial weight vectors for the semi-blind equalization algorithm [20] to further improve the performance of the transmit diversity scheme.

References


Zhaocheng Wang received his B.S.E., M.S.E. and Ph.D. degrees in 1991, 1993 and 1996 respectively, from the Department of Electronic Engineering, Tsinghua University. He was a Post Doctoral Fellow with Nanyang Technological University (NTU) in Singapore from 1996 to 1997. After that, he was with OKI Technic Center (Singapore) Pte. Ltd. from 1997 to 1999, firstly as a research engineer and then as a senior engineer from August 1998. From 1999 to 2009, he worked at SONY Deutschland GmbH/SONY International Europe in Germany, firstly as a senior engineer and then as a principal engineer from October 2002. Since April 2009, he has been with the Department of Electronic Engineering at Tsinghua University as a Full Professor. His general research interests include wireless communications, digital video broadcasting and signal processing, with emphasis on OFDM, single carrier with frequency domain equalization (SC-FDE) and MIMO techniques. Within these areas, he filed 35 US/EU patent applications, 18 of them have been granted. He has published tens of journal and conference papers.
**Jintao Wang** was born in Hebei, P.R. China. He received his B.Eng. and Ph.D. degrees in Electronic Engineering both from Tsinghua University, Beijing, China in 2001 and 2006, respectively. Since 2009, he has been an associate professor of Tsinghua’s DTV Technology R&D center. He is the standard committee member for the Chinese national digital terrestrial television broadcasting standard. His current research interest is in the area of the broadband wireless transmission, especially the channel estimation and space-time coding techniques.

**Linglong Dai** is a Ph.D. candidate at the Department of Electronic Engineering as well as the Tsinghua National Laboratory of Information Science and Technology (TNList), Tsinghua University, Beijing, China. His research interests lie in the field of synchronization, channel estimation for wireless communication system, space-time coding and diversity techniques, multiple access techniques, as well as wireless positioning.