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Training Sequence Design for Efficient Channel Estimation in MIMO-FBMC Systems

SU HU[1], ZILONG LIU[2], YONG LIANG GUAN[2], (Member, IEEE), CHUANXUE JIN[1], YIXUAN HUANG[1], AND JEN-MING WU[3]

[1]National Key Laboratory on Communications, University of Electronic Science and Technology of China, 611731 Chengdu, China
[2]School of Electrical and Electronic Engineering, Nanyang Technological University, 639798 Singapore
[3]Department of Electrical Engineering, Institute of Communications Engineering, National Tsing Hua University, 30013 Hsinchu, Taiwan

Corresponding author: S. Hu (husu@uestc.edu.cn)

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ABSTRACT This paper is focused on training sequence design for efficient channel estimation in multiple-input multiple-output filterbank multicarrier (MIMO-FBMC) communications using offset quadrature amplitude modulation (OQAM). MIMO-FBMC is a promising technique to achieve high spectrum efficiency as well as strong robustness against dispersive channels due to its feature of time-frequency localization. A salient drawback of FBMC/OQAM signals is that only real-field orthogonality can be kept, leading to the intrinsic imaginary interference being a barrier for high-performance channel estimations. Also, conventional channel estimations in the MIMO-FBMC systems mostly suffer from high training overhead especially for large number of transmit antennas. Motivated by these problems, in this paper, we propose a new class of training sequences, which are formed by concatenation of two identical zero-correlation zone sequences whose auto-correlation and cross correlation are zero within a time-shift window around the in-phase position. Since only real-valued symbols can be transmitted in MIMO-FBMC systems, we propose ''complex training sequence decomposition (CTSD)'' to facilitate the reconstruction of the complex-field orthogonality of MIMO-FBMC signals. Our simulations validate that the proposed CTSD is an efficient channel estimation approach for practical preamble-based MIMO-FBMC systems.

INDEX TERMS Filterbank multicarrier (FBMC), multiple-input multiple-output (MIMO), channel estimation, preamble, zero-correlation zone (ZCZ) sequences.

I. INTRODUCTION

A. BACKGROUND

Orthogonal frequency division multiplexing (OFDM) is an efficient multicarrier modulation scheme which is resilient to the effect of multipath fading channels. Although all the OFDM subcarriers are modulated by waveforms that are limited in the time-domain, in practice, there is unavoidable power leakage in the frequency-domain and because of this, the guard band has to be placed so as to minimize adjacent channel interference to other coexisting wireless systems [1]. Furthermore, OFDM's robustness against multipath propagation relies on the insertion of cyclic prefix (CP) which is a loss of spectrum efficiency.[1]

[1]In practice, this may go up to a typical value of 25%.

As an alternative modulation scheme to CP-OFDM systems, this paper focuses on the filterbank multicarrier (FBMC) systems employing offset quadrature amplitude modulation (OQAM) without inserting CP, called FBMC/OQAM [2], [3]. For ease of presentation, from now on, we use FBMC to denote FBMC/OQAM. The subcarriers in an FBMC system are modulated with staggered OQAM symbols, i.e., real-valued symbols at twice the symbol rate of CP-OFDM systems [4]. This allows FBMC pulse shaping filters to be well localized in time- and frequency- domains for increased robustness against carrier frequency offset and Doppler spread, as well as better spectral containment in bandwidth sensitive applications [5]. However, FBMC signals suffer from the intrinsic imaginary interference [6] [in the form of inter-carrier interference (ICI) and inter-symbol interference (ISI)] as the subcarrier functions in FBMC systems are orthogonal in the real field only, i.e., the real-field orthogonality. Because of this, FBMC signal processing tasks (such as channel estimation) are in general more complicated than that in CP-OFDM systems [7].

B. LITERATURE OVERVIEW

Numerous channel estimation methods for preamble-based FBMC systems have been proposed in the literature [8]–[18]. Based on the assumption of low frequency selectivity, interference approximation method (IAM) which is capable of computing an approximation of the interference from neighboring symbols, has been proposed in [9]–[12]. A second approach to cope with the intrinsic imaginary interference is called interference cancellation method (ICM), which was presented in [13] by taking advantage of the odd symmetry property owned by the ambiguity function of pulse shaping filters. By observing ICI mainly from the nearest subcarriers, another ICM was developed by constructing preambles with only odd- (or even-) indexed subcarriers [14]. [15] introduces challenges of channel estimation in FBMC systems and reviews several representative preamble- and pilot-aided approaches as well as some results on "sparse" preamble designs. A linear minimum mean square error channel estimation method and its mean square error (MSE) analysis has been presented in [16]. In addition, a time-domain approach (as opposed to the frequency-domain approach) for preamble-based channel estimation in FBMC systems has been reported in [17] and [18].

Combining the advantages of FBMC, multiple-input multiple-output filterbank multicarrier (MIMO-FBMC) transmission emerges as a promising technique to achieve high spectrum efficiency as well as strong robustness against dispersive channels. An excellent work by A. Pérez-Neira *et al.* [19] which provides an overview of recent advances on a series of signal processing problems in MIMO-FBMC systems. It is noted that significant research attention on MIMO-FBMC systems has been focused on how to remove/suppress the effect of multi-antenna interference for high-performance channel estimations in the presence of intrinsic imaginary interference [20]–[23]. Reference [21] suggested two subcarriers assignment schemes to avoid multi-stream interference by allocating the subcarriers among different base stations during the training phase. Following their preamble structure, every receiver can only sense a part of the channel frequency response and therefore frequency-domain interpolation is required in their scheme in order to recover the full channel frequency response. In [22], the IAM-C channel estimation method has been extended to MIMO cases by simply selecting a corresponding Hadamard transform matrix. Reference [23] proposed a class of FBMC preambles, each of which is comprised of a single nonzero pilot FBMC symbol plus the null guard(s), for highly frequency selective MIMO channels. These preambles are optimized to achieve minimum channel estimation MSE. Recently, a low-overhead channel estimation method has been proposed in [24] by employing intrinsic interference pre-cancellation at the transmitter side.

C. MOTIVATIONS AND CONTRIBUTIONS

In general, channel estimation in MIMO systems is more challenging than that in single transmit antenna scenarios as more channel fading coefficients need to be estimated. It has been shown in [25]–[28] that zero-correlation zone (ZCZ) sequences are an excellent training sequence family in MIMO-OFDM systems and are capable of achieving optimal channel estimation performance under the condition that all the received signals are quasi-synchronous within the ZCZ. Here, a ZCZ refers to a zero correlation window (centered around the in-phase timing position). In the literature, ZCZ sequences were first proposed as spreading sequences to achieve interference-free quasi-synchronous code-division multiple-access communications [29]–[31] and can be efficiently constructed from complementary codes which have zero correlation sum properties [33]–[35].

In contrast to MIMO-OFDM systems, the application of ZCZ sequences to MIMO-FBMC systems is not straightforward. For a MIMO-FBMC system, all transmit symbols are real-valued and with the real-field orthogonality only, conflicting with the feature of ZCZ sequences which is defined in complex-field. In fact, this becomes a barrier of MIMO-FBMC systems for directly using any complex-field based orthogonal sequences for channel estimation. According to the transmultiplexer response of pulse shaping filter, the intrinsic imaginary interference including ISI and ICI is mainly from the nearest training- and data- symbols. By inserting enough guard symbols, both IAM and ICM can be easily extended to MIMO-FBMC systems. However, when the number of transmit antennas becomes large, huge amount of time-frequency resources is required for the training purpose, resulting in *highly inefficient* channel estimation schemes in MIMO-FBMC systems.

The main contribution of this paper is a novel preamble design approach based on *complex training sequences decomposition* (CTSD) to facilitate the reconstruction of complex-field orthogonality in MIMO-FBMC systems. We propose to cascade two identical ZCZ sequences in the time domain forming preambles with non-zero values in odd or even subcarriers only (in the frequency domain). By doing so, ICI can be mostly self-cancelled as it is mainly from the neighboring subcarriers. Nevertheless, these newly designed training sequences, which are complex-valued in the frequency domain, cannot be directly applied to MIMO-FBMC systems because only real-valued transmitted symbols are allowed. This motivates us to propose the CTSD method by decomposing each complex-valued number into two real-valued numbers (i.e., real and imaginary parts) and sending them over two FBMC symbols (which are separated by some zero guard symbols), while maintaining the ZCZ properties of preambles. In this way, training sequences for different transmit antennas can be sent out simultaneously at the cost of two non-zero FBMC sym-

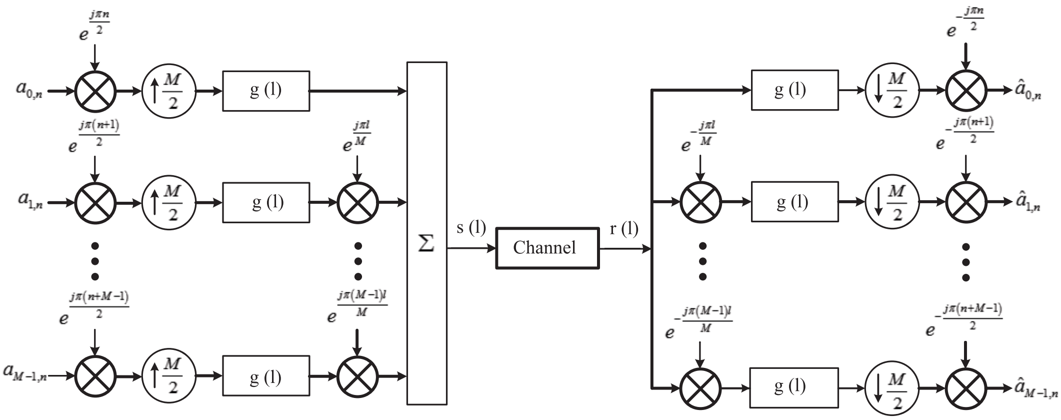

FIGURE 1. Block diagram of an equivalent FBMC baseband model.

bols only (one for real part, the other for imaginary part). Compared to existing preamble-based channel estimation methods, the proposed CTSD method is coded-division multiplexing based channel estimation and thus much less training overhead.

This paper is organized as follows. Section II presents the system model of FBMC systems and introduces existing preamble-based channel estimation methods. In Section III, a design of training symbols with the capability of interference self-cancellation is presented, followed by detailed description of the CTSD based channel estimation scheme. Numerical simulations are presented in Section IV, followed by the summary of this paper in Section V.

Notations: $\mathbf{X}^{\mathrm{T}}$ and $\mathbf{X}^{\mathrm{H}}$ denote the transpose and the Hermitian transpose of matrix $\mathbf{X}$, respectively; $\mathcal{I}_N$ denotes the identity matrix of order N; $\Re\{z\}$ and $\Im\{z\}$ denote the real and the imaginary parts of complex number z, respectively; $\lfloor p/N \rfloor$ denotes the floor function, i.e., the largest integer smaller than p/N.

II. SYSTEM MODEL AND PROBLEM FORMULATION

A. FBMC SYSTEM MODEL

As shown in Fig. 1, an equivalent FBMC baseband model with M subcarriers is considered, in which the subcarrier spacing is $1/T$, with T being the complex symbol interval. The transmitted symbol $a_{m,n}$ is real-valued with frequency index m and time index n, and $T/2$ is the interval of real-valued symbols. $a_{m,n}$ and $a_{m,2n+1}$ are obtained by taking the real and imaginary parts of complex-valued symbol from quadrature amplitude modulation (QAM) constellation.[2] $g(l)$ is the employed symmetrical real-valued prototype filter impulse response with total energy of one and with length of L_g. When the critical sampling period is defined by T/M, the equivalent discrete-time FBMC signal is expressed as [6]

$$s(l) = \sum_{m=0}^{M-1} \sum_{n\in\mathbb{Z}} a_{m,n} \underbrace{e^{j\pi(m+n)/2} e^{j2\pi ml/M} g\left(l - n\frac{M}{2}\right)}_{g_{m,n}(l)}, \quad (1)$$

where $j = \sqrt{-1}$, $g_{m,n}(l)$ represents the synthesis basis which is obtained by a time-frequency translated version of $g(l)$, and the transmultiplexer response is defined as

$$\zeta_{m,n}^{p,q} = \sum_{l=-\infty}^{\infty} g_{m,n}(l) g_{p,q}^{*}(l). \quad (2)$$

It is noted that all the values of $\zeta_{m,n}^{p,q}$ are purely imaginary, except at $\{m = p, n = q\}$ [6]. This shows that FBMC systems only satisfy the real-field orthogonality, meaning that perfect reconstruction of real-valued symbol $a_{m,n}$ is obtained if and only if the following orthogonal condition holds,

$$\Re\left\{ \sum_{l=-\infty}^{\infty} g_{m,n}(l) g_{p,q}^{*}(l) \right\} = \delta_{m,p}\delta_{n,q}, \quad (3)$$

where $\delta_{m,p}$ denotes the Kronecker delta function which is equal to 1 if and only if $m = p$. On the other hand, there will be *intrinsic imaginary interference* in FBMC systems for any $(m, n) \neq (p, q)$, even passing through an AWGN channel.

B. PROBLEM FORMULATION

Let $[h(0), h(1), \cdots, h(L_h - 1)]^{\mathrm{T}}$ be the discrete impulse response of a multipath fading channel, where L_h denotes the maximum channel delay. According to (1), the baseband received signal therefore can be written as

$$r(l) = \sum_{\tau=0}^{L_h-1} h(\tau) s(l-\tau) + \eta(l), \quad (4)$$

where $\eta(l)$ denotes the complex additive white Gaussian noise with zero mean and variance of σ^2. The demodulation of received signal at the (m, n)th time-frequency point provides

[2]It is noticed that pseudo-pilots that are either purely real or imaginary at all the subcarriers are also allowed for the purpose of preamble-based channel estimation [7].

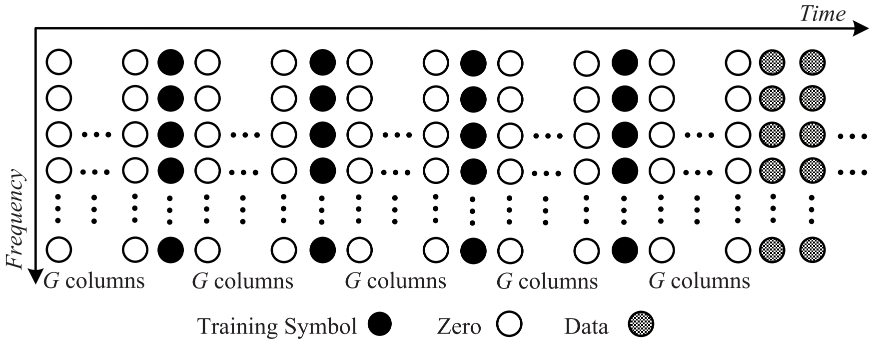

FIGURE 2. Conventional preambles for (4 × 4) MIMO-FBMC systems by inserting G zero symbols.

a complex-valued symbol as follows.

$$y_{m,n} = \sum_{l=-\infty}^{\infty} r(l) g\left(l - \frac{nM}{2}\right) e^{-j2\pi ml/M} j^{-m-n}. \tag{5}$$

We assume the symbol interval is much longer than the maximum channel delay spread, i.e., $L_h \ll L_g$. Therefore, the channel may be viewed as frequency flat at each subcarrier over the prototype filter duration, approximately [10]. Specifically, the prototype filter function has good time-frequency localization with relatively low variation over any time interval $[l, l + L_h]$, i.e., $g(l) \approx g(l+\tau)$, for $\tau \in \{0, 1, \cdots, L_h - 1\}$. Hence, $y_{m,n}$ in (5) can be simplified to [18]

$$y_{m,n} = H_{m,n} a_{m,n} + \underbrace{\sum_{(m,n)\neq(p,q)} H_{p,q} a_{p,q} \zeta_{m,n}^{p,q}}_{I_{m,n}} + \eta_{m,n}, \tag{6}$$

where $H_{m,n}$ denotes the channel frequency response at the (m, n)th time-frequency point, and $I_{m,n}$ and $\eta_{m,n}$ stand for the intrinsic imaginary interference and the noise term, respectively. Different from that in OFDM systems, the noise term in an FBMC system is non-white due to the prototype filtering at each subcarrier channel.

Based on the transmultiplexer response of pulse shaping filter, the intrinsic imaginary interference is mainly from the nearest training- and data- symbols. Numerous preamble structures and their associated estimation methods have been proposed in [9]–[13]. A precondition for these methods is that G zero symbols have to be inserted to suppress ISI and ICI.

The above FBMC transmission model in SISO scenario can be easily extended to MIMO-FBMC systems. Consider an $N_T \times N_R$ MIMO-FBMC system. The received signal in each receive antenna $k = 1, 2, \cdots, N_R$ can be expressed as

$$y_{m,n}^{k} = \sum_{i=1}^{N_T} \left(H_{m,n}^{k,i} a_{m,n}^{i} + \sum_{(m,n)\neq(p,q)} H_{p,q}^{k,i} a_{p,q}^{i} \zeta_{m,n}^{p,q} \right) + \eta_{m,n}^{k}, \tag{7}$$

where $H_{m,n}^{k,i}$ denotes the channel frequency response from the ith transmit antenna to the kth receive antenna, and $\eta_{m,n}^{k}$ denotes the corresponding noise term in the kth receive antenna.

Let us recall the channel estimation methods for MIMO-OFDM systems, in which orthogonal training sequences (such as ZCZ sequences) are simultaneously applied to all transmit antennas, i.e., less time-frequency resources are consumed [28]. However, this approach requires the complex-field orthogonality (other than the real-field orthogonality) and thus cannot be directly applied to MIMO-FBMC systems. Alternatively, conventional preamble structures for SISO-FBMC systems, such as IAM [12] and ICM [14] schemes, may be extended to MIMO-FBMC by transmitting training symbols in turn over different transmit antennas. As shown in Fig. 2, for a MIMO-FBMC system with N_T transmit antennas, both IAM and ICM require $(G + 1)N_T + G$ FBMC symbols, where G zero symbols are inserted in between every two nearest (non-zero) training symbols. Clearly, there would be huge waste in time-frequency resources when the number of transmitting antennas becomes large.

III. COMPLEX TRAINING SEQUENCES DECOMPOSITION CHANNEL ESTIMATION

Motivated by the aforementioned inefficient channel estimation approaches, we propose CTSD channel estimation method which enables: (1) zero values for all odd-indexed subcarriers for ICI suppression by concatenation of two identical ZCZ sequences in the time-domain and (2) separate transmissions of the real- and imaginary parts of the frequency-domain training sequences to facilitate the reconstruction of the complex-field orthogonality of the concatenated ZCZ sequences. In what follows, we present the CTSD preamble structure as well as the corresponding channel estimation scheme.

A. PROPOSED PREAMBLE STRUCTURE

Let $\mathcal{A} = \{\mathbf{a}_1, \mathbf{a}_2, \cdots, \mathbf{a}_{N_T}\}$ be a set of N_T sequences, each of length L, i.e.,

$$a_i = \left[a_{i,0}, a_{i,1}, \cdots, a_{i,L-1}\right]^{\mathrm{T}}, \quad 1 \leq i \leq N_T. \tag{8}$$

Let us now construct a new sequence set $\mathcal{C} = \{\mathbf{c}_1, \mathbf{c}_2, \cdots, \mathbf{c}_{N_T}\}$, each formed by cascading two identical

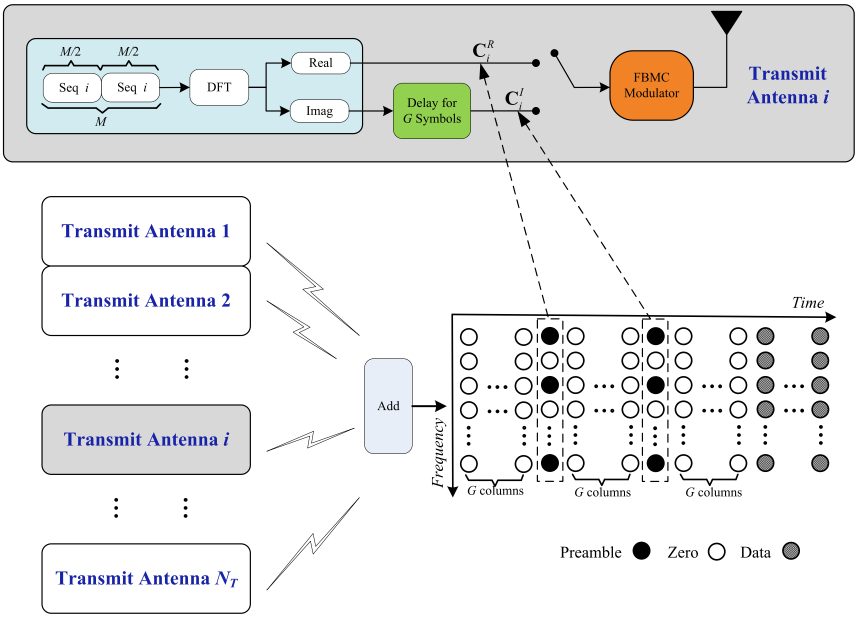

FIGURE 3. Proposed CTSD preamble structure in a MIMO-FBMC system with N_T transmit antennas.

sequences from $\mathcal{A}$, i.e.,

$$\begin{aligned}\mathbf{c}_i &= \left[c_{i,0}, c_{i,1}, \cdots, c_{i,M-1}\right]^{\mathrm{T}} \\ &= \left[a_{i,0}, \cdots, a_{i,L-1}, a_{i,0}, \cdots, a_{i,L-1}\right]^{\mathrm{T}}, \end{aligned} \tag{9}$$

where $M = 2L$ equals the total number of FBMC subcarriers. Applying discrete Fourier transform (DFT) to $\mathbf{c}_i$ $(1 \le i \le N_T)$, the corresponding frequency-domain training sequences $\{\mathbf{C}_1, \mathbf{C}_2, \cdots, \mathbf{C}_{N_T}\}$ can be shown as

$$\begin{aligned}\mathbf{C}_i &= \left[C_{i,0}, C_{i,1}, \cdots, C_{i,M-1}\right]^{\mathrm{T}} \\ C_{i,m} &= \sum_{n=0}^{M-1} c_{i,n} e^{-j2\pi mn/M}, \quad 0 \le m \le M-1. \end{aligned} \tag{10}$$

Substituting (9) into (10), it is easy to obtain

$$C_{i,m} = \begin{cases} 2\sum_{n=0}^{L-1} a_{i,n} e^{-j2\pi mn/L}, & m \text{ is even}; \\ 0, & m \text{ is odd}. \end{cases} \tag{11}$$

Our idea here is to respectively apply these N_T sequences as frequency domain training sequences to the N_T transmit antennas. Without loss of generality, assume $\mathbf{C}_i$ will be sent out over transmit antenna i. However, each element in $\mathbf{C}_i$ may be complex-valued, the transmission of which is not supported in MIMO-FBMC systems. For this, we introduce a novel CTSD preamble design by decomposing each complex-valued symbol into two real-valued symbols that are to be transmitted as FBMC symbols. For the ith complex-valued training sequence $\mathbf{C}_i$, the corresponding real-valued pair of $\left(\mathbf{C}_i^R, \mathbf{C}_i^I\right)$ is obtained as follows,

$$\begin{aligned}\mathbf{C}_i^R &= \Re\{\mathbf{C}_i\} = \left[C_{i,0}^R, C_{i,1}^R, \cdots, C_{i,M-1}^R\right]^{\mathrm{T}}, \\ \mathbf{C}_i^I &= \Im\{\mathbf{C}_i\} = \left[C_{i,0}^I, C_{i,1}^I, \cdots, C_{i,M-1}^I\right]^{\mathrm{T}}, \end{aligned} \tag{12}$$

where $C_{i,m}^R$ and $C_{i,m}^I$ $(0 \le m \le M-1)$ denote the real- and imaginary- parts of $C_{i,m}$, respectively. To illustrate the transmission of these N_T training sequences, the CTSD preamble structure is plotted in Fig. 3. We have the following notes: (1) $\mathbf{C}_i^R$ (or $\mathbf{C}_i^I$) for $1 \le i \le N_T$ occupy the same FBMC symbol, meaning that they will be simultaneously sent out; (2) The addition of these FBMC training sequences (from different transmit antennas) takes place over the air; (3) The role of the switch is to split the real- and imaginary- parts of the preambles such that they can be transmitted over two FBMC symbols (separated by G zero symbols).

For an $N_T \times N_R$ MIMO-FBMC system, there are $N_T N_R$ independent channels to be estimated, where every channel is modelled as a finite impulse response filter with L_h taps. For ease of analysis, we assume that $N_T = N_R$ and these channel taps remain constant over the transmission of each block and vary independently from one block to another (i.e., quasi-static). In this paper, denote by

$$\mathbf{h}_{k,i} = \left[h_{k,i}(0), h_{k,i}(1), \cdots, h_{k,i}(L_h-1)\right], \tag{13}$$

the channel impulse response vector from the ith $(i = 1, 2, \cdots, N_T)$ transmit antenna to the kth $(k = 1, 2, \cdots, N_R)$ receive antenna. Similar to SISO-FBMC systems, the demodulation of the (m, n)th symbol at the kth receive antenna can

TABLE 1. Training overhead comparison for channel estimation in MIMO-FBMC systems.

	MIMO-FBMC ($G=1$)		MIMO-FBMC ($G=2$)		MIMO-FBMC ($G=3$)	
	IAM & ICM	CTSD (proposed)	IAM & ICM	CTSD	IAM & ICM	CTSD
$N_T=2$	5	5	8	8	11	11
$N_T=4$	9	5	14	8	19	11
$N_T=8$	17	5	26	8	35	11

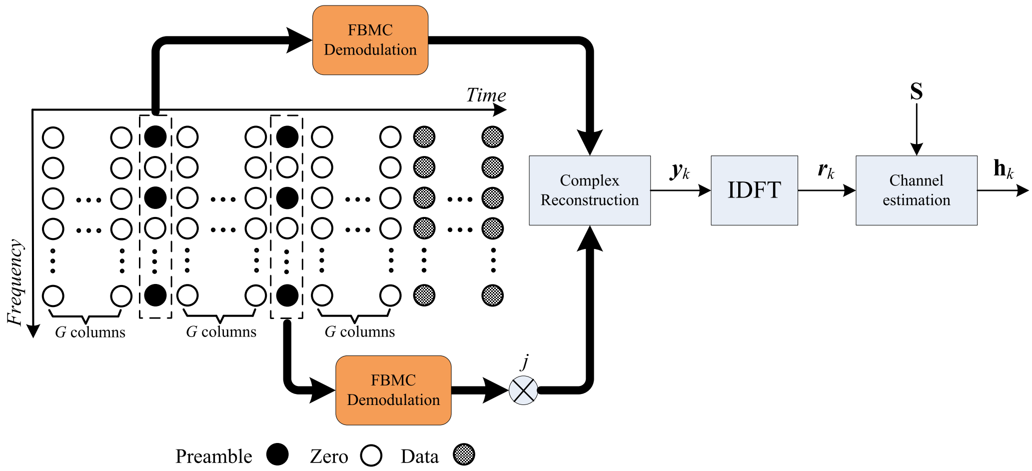

FIGURE 4. Channel estimation of the kth receive antenna of a MIMO-FBMC system.

be written as [18]

$$y^k_{m,n} = \sum_{i=1}^{N_T}\sum_{\tau=0}^{L_h-1}\sum_{l=-\infty}^{\infty}\sum_{p=0}^{M-1}\sum_{q\in\mathbb{Z}} a^i_{p,q} g\left(l-\tau-v\frac{qM}{2}\right) g\left(l-\frac{nM}{2}\right) \times e^{j2\pi((p-m)l-p\tau)/M} j^{p+q-m-n} h_{k,i}(\tau) + \eta^k_{m,n}.$$

Since the prototype filter function has the property of time-frequency localization and relatively low variation over certain time interval (cf. Section II), i.e., $g(l) \approx g(l+\tau)$, for $\tau \in \{0, 1, \cdots, L_h-1\}$, (14) can be approximated as

$$y^k_{m,n} = \sum_{i=1}^{N_T}\sum_{\tau=0}^{L_h-1}\sum_{p=0}^{M-1}\sum_{q\in\mathbb{Z}} a^i_{p,q}\zeta^{m,n}_{p,q} e^{-j2\pi p\tau/M} h_{k,i}(\tau) + \eta^k_{m,n}. \tag{14}$$

In practice, each preamble is interfered by adjacent training- and/or data- symbols. Recalling from the transmultiplexer response in Table 1, both ISI and ICI received at the (m, n)th time-frequency point are mainly from the nearest neighboring points specified in the set below.

$$\{(m\pm 1, n\pm 1), (m, n\pm 1), (m\pm 1, n)\}.$$

Remark 1: The proposed CTSD method is capable of significant ISI- and ICI- suppression. To explain, let us take the (m, n)th time-frequency point for example. ISI arising from the $(m, n\pm 1)$th neighboring points is negligible because they are G zero symbols away to the (m, n)th point, as shown in Fig. 3. Secondly, since all the odd-indexed subcarriers are set to zero [c.f. (11)], ICI to the (m, n)th grid is also substantially suppressed.

B. CHANNEL ESTIMATOR DESIGN

Note that $\mathbf{C}^R_i$ and $\mathbf{C}^I_i$ are the real- and imaginary- parts of complex-valued training symbol $\mathbf{C}_i$, respectively. According to (14), the received signal associated to $\mathbf{C}^R_i$ $(1 \le i \le N_T)$ can be expressed as

$$y^R_{k,m} = \sum_{i=1}^{N_T}\sum_{\tau=0}^{L_h-1} C^R_{i,m} e^{-j2\pi m\tau/M} h_{k,i}(\tau) + \eta^R_{k,m}, \tag{15}$$

where $\eta^R_{k,m}$ denotes the noise term, the subscripts k, m denote the kth receive antenna and the mth subcarrier, respectively. Similarly, the received signal associated to $\mathbf{C}^I_i$ $(1 \le i \le N_T)$ can be expressed as

$$y^I_{k,m} = \sum_{i=1}^{N_T}\sum_{\tau=0}^{L_h-1} C^I_{i,m} e^{-j2\pi m\tau/M} h_{k,i}(\tau) + \eta^I_{k,m}, \tag{16}$$

with $\eta^I_{k,m}$ being the noise term.

As shown in Fig. 4, the complex-valued symbol $\mathbf{y}_k = \left[y_{k,0}, y_{k,1}, \cdots, y_{k,M-1}\right]^{\mathrm{T}}$ can now be reconstructed as follows,

$$\begin{aligned} y_{k,m} &= y^R_{k,m} + jy^I_{k,m} \\ &= \sum_{i=1}^{N_T}\sum_{\tau=0}^{L_h-1} \underbrace{\left(C^R_{i,m} + jC^I_{i,m}\right)}_{=C_{i,m}} e^{-j2\pi m\tau/M} h_{k,i}(\tau) \\ &\quad + \eta^R_{k,m} + j\eta^I_{k,m}. \end{aligned} \tag{17}$$

Let $\eta_{k,m} = \eta^R_{k,m} + j\eta^I_{k,m}$. Then (17) can be rewritten as

$$y_{k,m} = \sum_{i=1}^{N_T}\sum_{\tau=0}^{L_h-1} C_{i,m}e^{-j2\pi m\tau/M}h_{k,i}(\tau) + \eta_{k,m}. \quad (18)$$

Applying inverse discrete Fourier transform (IDFT) on $\mathbf{r}_k$, we obtain $\mathbf{r}_k = \left[r_{k,0}, r_{k,1}, \cdots, r_{k,M-1}\right]^{\mathrm{T}}$,

$$r_{k,n} = \sum_{i=1}^{N_T}\sum_{\tau=0}^{L_h-1} c_{i,(n-\tau)_M}h_{k,i}(\tau) + \omega_{k,n}, \quad (19)$$

where $(\cdot)_M$ denotes the modulo M operation, and

$$\omega_{k,n} = \sum_{m=0}^{M-1} \underbrace{\left(\eta^R_{k,m} + \eta^I_{k,m}\right)}_{=\eta_{k,m}} e^{j2\pi mn/M}. \quad (20)$$

Similar to [23], the above formula can be written in matrix form as

$$\mathbf{r}_k = \mathbf{S}\mathbf{h}_k + \boldsymbol{\omega}_k, \quad (21)$$

where

$$\begin{aligned}\mathbf{S} &= \left[\mathbf{S}_1 \ \mathbf{S}_2 \ \cdots \ \mathbf{S}_{N_T}\right],\\ \mathbf{h}_k &= \left[\mathbf{h}_{k,1} \ \mathbf{h}_{k,2} \ \cdots \ \mathbf{h}_{k,N_T}\right]^{\mathrm{T}},\\ \boldsymbol{\omega}_k &= \left[\omega_{k,0}, \omega_{k,1}, \cdots, \omega_{k,M-1}\right]^{\mathrm{T}}. \end{aligned} \quad (22)$$

Here, $\mathbf{S}_i$, $i = 1, 2, \cdots, N_T$, are matrices of order $M \times L_h$,

$$S_i = \begin{bmatrix} c_{i,0} & c_{i,M-1} & \cdots & c_{i,M-L_h+2} & c_{i,M-L_h+1} \\ c_{i,1} & c_{i,0} & \cdots & c_{i,M-L_h+3} & c_{i,M-L_h+2} \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ c_{i,M-2} & c_{i,M-3} & \cdots & c_{i,M-L_h} & c_{i,M-L_h-1} \\ c_{i,M-1} & c_{i,M-2} & \cdots & c_{i,M-L_h+1} & c_{i,M-L_h} \end{bmatrix}. \quad (23)$$

Since in general the rank of $\mathbf{S}$ is equal to the number of columns, the linear least square channel estimator for the kth receive antenna is shown below [37]

$$\tilde{\mathbf{h}}_k = \left[\tilde{\mathbf{h}}_{k,1} \ \tilde{\mathbf{h}}_{k,2} \ \cdots \ \tilde{\mathbf{h}}_{k,N_T}\right]^{\mathrm{T}} = \left(\mathbf{S}^{\mathrm{H}}\mathbf{S}\right)^{-1}\mathbf{S}^{\mathrm{H}}\mathbf{r}_k. \quad (24)$$

In order to derive the correlation properties of the training sequences $\mathbf{c}_1, \mathbf{c}_2, \cdots, \mathbf{c}_K$ (cf. Section III.A), let us consider FBMC noise which may be (approximately) independent and identically distributed. In this case, the channel estimation MSE can be defined as

$$\mathrm{MSE} = \mathbb{E}\left[(\mathbf{h}_k - \tilde{\mathbf{h}}_k)^{\mathrm{H}}(\mathbf{h}_k - \tilde{\mathbf{h}}_k)\right] = 2\sigma^2\mathrm{Tr}\left(\left(\mathbf{S}^{\mathrm{H}}\mathbf{S}\right)^{-1}\right), \quad (25)$$

where $\mathbb{E}(\cdot)$ denotes the expectation of a random variable and $\mathrm{Tr}(\cdot)$ denotes the matrix trace operation. Note that the minimum MSE is achieved if and only if

$$\mathbf{S}^{\mathrm{H}}\mathbf{S} = \begin{bmatrix} \mathbf{S}_1^{\mathrm{H}}\mathbf{S}_1 & \mathbf{S}_2^{\mathrm{H}}\mathbf{S}_1 & \cdots & \mathbf{S}_{N_T}^{\mathrm{H}}\mathbf{S}_1 \\ \mathbf{S}_1^{\mathrm{H}}\mathbf{S}_2 & \mathbf{S}_2^{\mathrm{H}}\mathbf{S}_2 & \cdots & \mathbf{S}_{N_T}^{\mathrm{H}}\mathbf{S}_2 \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{S}_1^{\mathrm{H}}\mathbf{S}_{N_T} & \mathbf{S}_2^{\mathrm{H}}\mathbf{S}_{N_T} & \cdots & \mathbf{S}_{N_T}^{\mathrm{H}}\mathbf{S}_{N_T} \end{bmatrix} = M\mathcal{I}_{N_T L_h}. \quad (26)$$

In the next subsection, we shall specifically design training sequences which allows (26) to be met. As we will see later, these training sequences are key to achieve very low channel estimation MSE (compared to certain MSE bound), given the white-noise approximation applied in (25).[3]

C. TRAINING SEQUENCE DESIGN

It has been shown in (26) that the proposed CTSD channel estimation approach requires the following two conditions.

$$\begin{aligned} &\text{Condition 1: } \mathbf{S}_i^{\mathrm{H}}\mathbf{S}_i = M\mathcal{I}_{L_h}, \quad \text{for all } i \in \{1, 2, \cdots, N_T\}.\\ &\text{Condition 2: } \mathbf{S}_i^{\mathrm{H}}\mathbf{S}_j = \mathbf{0}_{L_h}, \quad \text{for all } i \neq j. \end{aligned} \quad (27)$$

To relate both conditions with training sequences, recall the "seed" sequence set $\mathcal{A}$ in Section III-A and define the periodic cross-correlation function (PCCF) between $\mathbf{a}_i$ and $\mathbf{a}_j$ as

$$R_{\mathbf{a}_i,\mathbf{a}_j}(\tau) = \sum_{n=0}^{L-1} a_{i,n}\left(a_{j,n+\tau}\right)^*, \quad (28)$$

where the addition $n + \tau$ is performed modulo L. If $i = j$, $R_{\mathbf{a}_i,\mathbf{a}_j}(\tau)$ reduces to the periodic auto-correlation function (PACF) of $\mathbf{a}_i$ and will be written as $R_{\mathbf{a}_i}(\tau)$ for simplicity. Note that $R_{\mathbf{a}_i,\mathbf{a}_j}(\tau)$ is equal to the inner product between $\mathbf{a}_i$ and the τ-shifted version of $\mathbf{a}_j$ (which is obtained by cyclically shifting $\mathbf{a}_j$ to the left for τ positions). Keeping this in mind and recalling (23), Conditions 1 and 2 in (27) are equivalent to

$$R_{\mathbf{c}_i}(\tau) = \begin{cases} M, & \text{for } \tau = 0; \\ 0, & \text{for any } 0 < \tau < L_h, \end{cases} \quad (29)$$

and

$$R_{\mathbf{c}_i,\mathbf{c}_j}(\tau) = 0, \quad \text{for any } i \neq j, \ 0 \le \tau < L_h. \quad (30)$$

respectively. Based on (29) and (30), we have the following remark.

Remark 2: To carry out the CTSD based channel estimation, it is required that the training sequence set $\mathcal{C} = \{\mathbf{c}_1, \mathbf{c}_2, \cdots, \mathbf{c}_{N_T}\}$ should have zero PACF and PCCF for all time-shifts less than L_h. In the literature, a sequence set satisfying (29) and (30) is said to have ZCZ width of L_h.

Construction: Note that each sequence in $\mathcal{C}$ is formed by cascading two identical sequences from $\mathcal{A}$, i.e., $\mathbf{c}_i = \left[a_{i,0}, \cdots, a_{i,L-1}, a_{i,0}, \cdots, a_{i,L-1}\right]^{\mathrm{T}}$, where $i \in \{1, 2, \cdots, N_T\}$. It is easy to show that

$$R_{\mathbf{c}_i,\mathbf{c}_j}(\tau) = \begin{cases} 2R_{\mathbf{a}_i,\mathbf{a}_j}(\tau), & \text{for } 0 \le \tau \le L-1; \\ 2R_{\mathbf{a}_i,\mathbf{a}_j}(\tau - L), & \text{for } L \le \tau < 2L = M. \end{cases} \quad (31)$$

[3] In fact, the noise in an FBMC system is colored due to the prototype filtering at the receiver and this is different from that of OFDM systems [8].

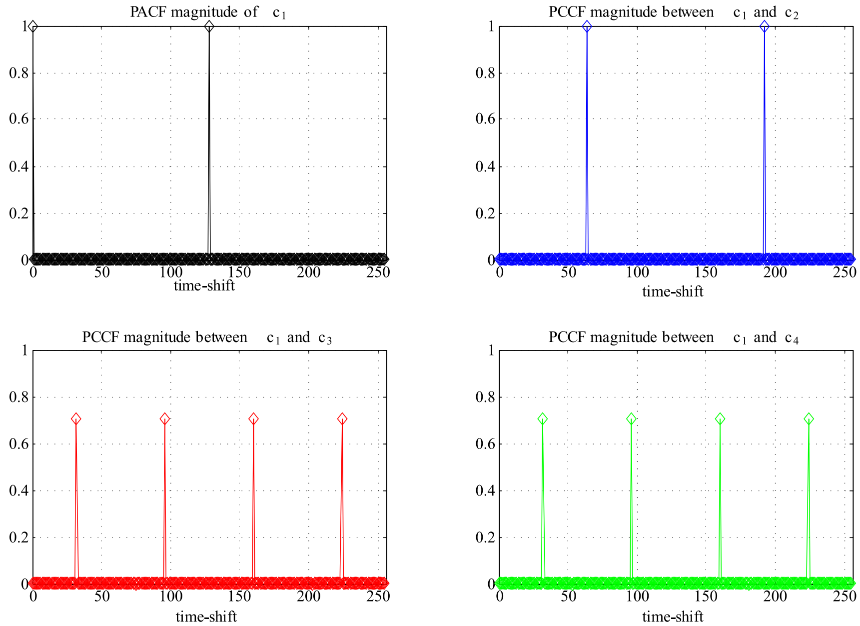

FIGURE 5. **Correlation plots of training sequences $\mathbf{c}_1, \mathbf{c}_2, \mathbf{c}_3, \mathbf{c}_4$ which are constructed from Popovic-Mauritz ZCZ sequences.**

Hence, $\mathcal{C}$ will have ZCZ width of L_h provided that $\mathcal{A}$ also has ZCZ width of L_h, i.e.,

$$R_{\mathbf{a}_i,\mathbf{a}_j}(\tau) = 0, \quad \text{for } i = j, 1 \le \tau < L_h \text{ and } i \neq j, \ 0 \le \tau < L_h. \tag{32}$$

In this paper, we employ the Popovic-Mauritz ZCZ sequences [38] as the "seed" sequence set $\mathcal{A}$ to generate the training sequence set $\mathcal{C}$. The Popovic-Mauritz ZCZ family is optimal in the sense that the set size meets its upper bound [30] with equality, i.e., $N_T \le L/L_h$. In other words, for a MIMO-FBMC system with M subcarriers, at most $N_T = \left\lfloor \frac{M}{2L_h} \right\rfloor$ ZCZ training sequences (which will be simultaneously over N_T transmit antennas) can be generated. In what follows, we provide a step-by-step example to specifically illustrate the training sequence generation.

Example 1: Let $N_T = 4, L = 128, M = 2L = 256$.

Step 1: Generate a Popovic-Mauritz sequence $\mathbf{z} = [z_0, z_1, \cdots, z_{L-1}]^{\mathrm{T}}$ with zero PACFs as follows.

$$z_k = \exp\left(\frac{-\sqrt{-1}2\pi k^2}{2L}\right). \tag{33}$$

Step 2: Generate the $N_T \times N_T$ Hadamard matrix $\mathbb{H}$ over ± 1. Let $\mathbb{H}_i$ be the ith column of $\mathbb{H}$. Based on $\mathbb{H}$, generate a sequence set $\mathcal{V} = \{\mathbf{v}_1, \mathbf{v}_2, \cdots, \mathbf{v}_{N_T}\}$ by

$$\mathbf{v}_i = \mathbf{1}_{L/N_T} \otimes \mathbb{H}_i, \quad 1 \le i \le N_T, \tag{34}$$

where $\mathbf{1}_{L/N_T}$ denotes the all-one column vector of length L/N_T and $\otimes$ denotes the Kronecker product.

Step 3: Generate sequence set $\mathcal{A} = \left\{\mathbf{a}_1, \mathbf{a}_2, \cdots, \mathbf{a}_{N_T}\right\}$ as follows.

$$\mathbf{a}_i = \mathbf{v}_i \circ \mathbf{z}, \quad 1 \le i \le N_T, \tag{35}$$

where $\circ$ denotes the Hadamard product. It is shown in [38] that $\mathcal{A}$ has ZCZ width of L/N_T.

Step 4: Generate training sequence set $\mathcal{C}$ from $\mathcal{A}$ through the concatenation operation in (9).

In Fig. 5, we have plotted the PACF of $\mathbf{c}_1$, the PCCFs between $\mathbf{c}_1$ and $\mathbf{c}_2, \mathbf{c}_3, \mathbf{c}_4$, respectively. It is seen that each sequence has zero PACF for any non-zero time-shift (see the top-left sub-figure) and the PCCFs between any distinct sequence pair are zero over a zone centered around the in-phase position with width for at least 32 (see the two bottom sub-figures). This indicates that the ZCZ width equals to $128/4 = 32$, implying the training sequences $\mathbf{c}_1, \mathbf{c}_2, \mathbf{c}_3, \mathbf{c}_4$ can be used for channel estimation in a $4 \times N_R$ MIMO-FBMC system with maximal channel delay up to 32. One can also verify the correlation properties shown in (31) and (32) based on Fig. 5. It is noted that the positions of peaks are inherently brought in due to the application of $N_T \times N_T$ Hadamard matrix that splits the entire correlation zone (of length $2L$) into blocks with identical width of $2L/N_T$.

IV. DISCUSSIONS AND NUMERICAL SIMULATIONS

A. COMPARISON OF TRAINING SEQUENCE OVERHEAD

Recalling conventional training structures shown in Fig. 2, a MIMO-FBMC system with N_T transmit antennas requires $(G+1)N_T + G$ training symbols, where G zero symbols are inserted at the two ends of training block, and in between every pair of adjacent (non-zero) training symbols. Clearly, huge amount of time-frequency resources is needed for channel estimation with large N_T. However, the proposed CTSD preamble only requires $2(G+1)+G$ symbols, the number of which does not depend on the number

TABLE 2. Transmultiplexer response of Bellanger's filter [36].

	$n-3$	$n-2$	$n-1$	n	$n+1$	$n+2$	$n+3$
$m-2$	$0.0006j$	-0.0001	0	0	0	0.0001	$0.0006j$
$m-1$	$0.0429j$	$-0.125j$	$0.2058j$	$-0.2393j$	$0.2058j$	$-0.125j$	$-0.0429j$
m	$-0.0668j$	$0.0002j$	$-0.5644j$	1	$0.5644j$	$0.0002j$	$-0.0668j$
$m+1$	$-0.0429j$	$-0.125j$	$0.2058j$	$0.2393j$	$0.2058j$	$-0.125j$	$0.0429j$
$m+2$	$0.0006j$	$0.0001j$	0	0	0	$-0.0001j$	$0.0006j$

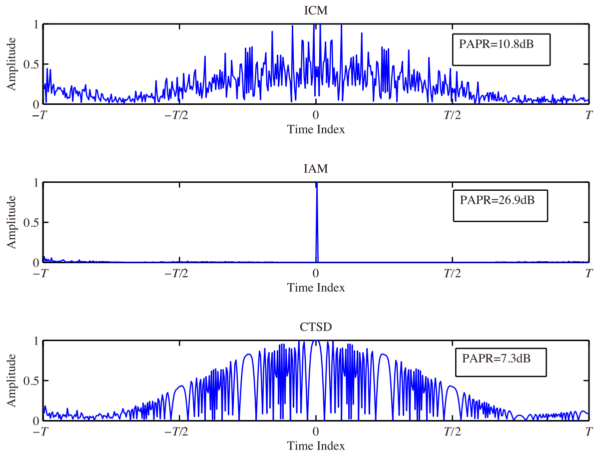

FIGURE 6. PAPR comparison between different channel estimation schemes for MIMO-FBMC systems.

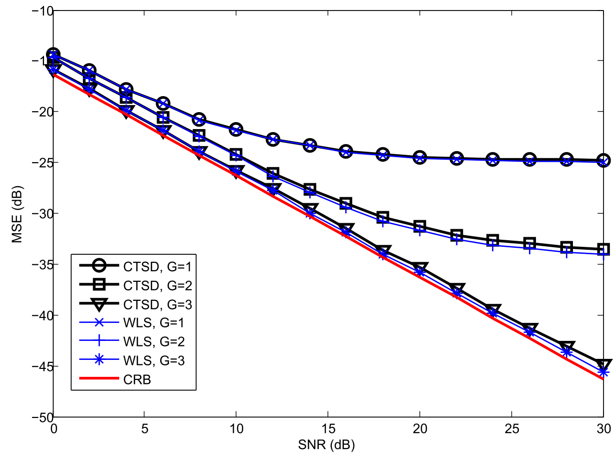

FIGURE 7. MSE performance of different channel estimators $L_h = 6$ and $G \in \{1, 2, 3\}$.

of transmit antennas. Because of this, our proposed CTSD channel estimation scheme leads to lower training overhead in MIMO-FBMC systems. A comparison of training sequence overhead is shown in Table 1 for different values of N_T and G.

B. COMPARISON OF PEAK-TO-AVERAGE POWER RATIO (PAPR)

A well-known drawback of multicarrier transmitters is that they may suffer from high peak-to-average power ratio (PAPR) problem. To compare the PAPR of the proposed CTSD scheme with that of ICM and IAM, the instantaneous FBMC signal envelopes of these three schemes over $[-T, T]$ are shown in Fig. 6. A common setting of these schemes is that the number of subcarriers $M = 256$ and each training symbol is protected by two zero symbols (one at the front, the other at the back), i.e., $G = 1$. The ICM method in Fig. 6 adopts a frequency-domain training sequence in the form of $[C_0, 0, C_1, 0, \cdots, C_{M/2-1}, 0]^{\mathrm{T}}$, where $C_0, C_1, \cdots, C_{M/2-1}$ are randomly generated binary numbers over $\{-1, 1\}$ [14]. The IAM-C method with frequency-domain training sequence of $[1, -j, -1, j, \cdots, 1, -j, -1, j]^{\mathrm{T}}$ is used as it has the best performance among the three IAM types. $\mathbf{C}_1^R$, the frequency-domain sequence associated to $\mathbf{c}_1$ in *Example 1*, has been chosen to discuss the PAPR property of the proposed CTSD scheme. One can see that the proposed CTSD scheme achieves the lowest PAPR of 7.3dB, smaller than the PAPR of ICM method (i.e., 10.8dB) and significantly smaller than that of IAM-C method (i.e., 26.9dB).

C. NUMERICAL SIMULATIONS FOR CHANNEL ESTIMATION MSE

In this section, we evaluate the performance of the proposed CTSD channel estimation method with regard to the channel estimation MSE. The transmitted real-valued data symbols are obtained by extracting the real- and imaginary- parts of QPSK modulated complex-valued symbols. A MIMO-FBMC system with $M = 256$ subcarriers and different number of transmit antennas $N_T = \{2, 4, 8\}$ are considered. Bellanger's pulse shaping filter with overlapping factor of 4 is employed, where the transmultiplexer response is shown in Table 2 [36]. Moreover, multipath channels with L_h sample-spaced fading coefficients (uniform power delay profile) are adopted, where the channel tap coefficients are assumed to be independent complex Gaussian random variable with uniformly distributed phase and Rayleigh distributed envelope. The transmitted power from each transmit antenna is assumed to be identical. Meanwhile, a frequency-domain per-subcarrier single-tap zero-forcing equalizer which has length of the entire FBMC block has been used.

Fig. 7 shows the MSE performances of the proposed CTSD method for 4×4 MIMO-FBMC systems, in which $G \in \{1, 2, 3\}$ and $L_h = 6$ are set. When $G = 3$, the CTSD method achieves MSE performance close to the performance CRB and comparable to that of the Weighted Least Square (WLS) method (although slightly worse) [18]. However, the WLS method suffers from very high complexity for the calculation of matrix inverses. In contrast, our proposed CTSD method has advantage of low complexity [cf. (24)] because: (1) matrix inverse is avoided as $\mathbf{S}^{\mathrm{H}}\mathbf{S}$ is an identity matrix

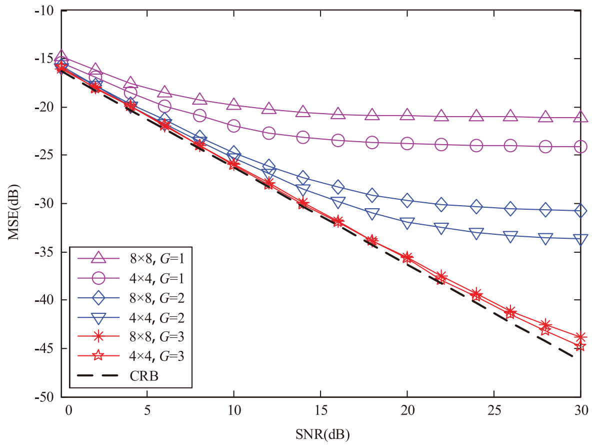

FIGURE 8. MSE performance of the proposed CTSD method for MIMO-FBMC systems with $L_h = 6$ and $G \in \{1, 2, 3\}$.

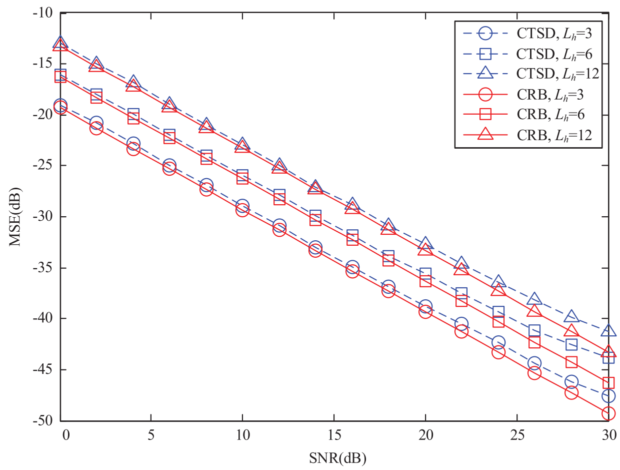

FIGURE 9. MSE performance of the proposed CTSD method for MIMO-FBMC systems (4x4) with $G = 3$ and $L_h \in \{3, 6, 12\}$.

multiplied by M, and (2) (inverse) fast Fourier transform (IFFT/FFT) can be used at both transmitter and receiver [cf. Figures 3 and 4]. For $G = 1$, the CTSD method suffers from the highest amount of residual ISI (compared with $G \in \{2, 3\}$) and therefore achieves the worst MSE performance. Since the residual ISI is a dominating interference in this case, the MSE performance of the proposed CTSD method almost matches with that of the WLS one.

Fig. 8 shows the MSE performance of the proposed CTSD channel estimation method with $G \in \{1, 2, 3\}$. Our proposed CTSD method exhibits MSE floor at high signal-to-noise ratio (SNR) region for $G \in \{1, 2\}$. This is mainly caused by the residual ISI from neighboring training- or data- symbols as shown in Fig. 3. As G increases, the MSE floor drops quickly owing to the substantial ISI reduction. Because of this, our proposed CTSD method can obtain almost the same MSE performance for $N_T \in \{4, 8\}$. This shows that the inter-antenna interference of MIMO-FBMC systems can be significantly suppressed by our proposed training sequences in Section III. Fig. 9 shows the MSE performance of our

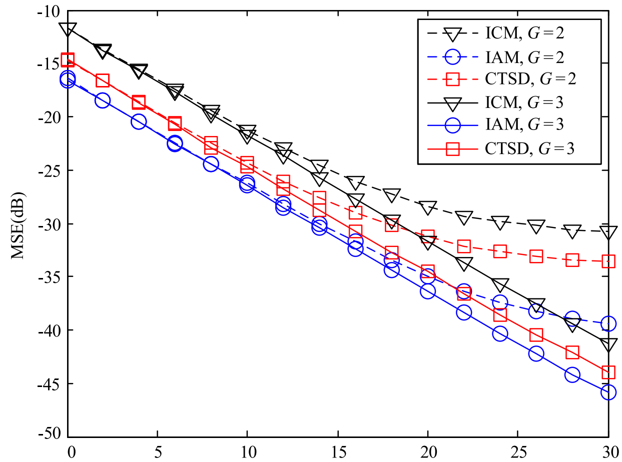

FIGURE 10. MSE performance of different channel estimation methods for MIMO-FBMC systems ($L_h = 6$).

proposed CTSD method for $L_h \in \{3, 6, 12\}$. One can see that, both the MSE lower bound and the actual MSE increase for larger L_h. Meanwhile, there is an MSE performance gap between the achievable results and the MSE lower bound. This is because the noise vector is correlated and a least square channel estimator may not be optimal [18]. Also, the correlated noise leads to increased inter-antenna interference for larger number of transmit antennas, in particular, when the number of guard symbols (G) is small. This explains the worse MSE performance for 8×8 MIMO-FBMC systems compared to that of the 4×4 ones.

Fig. 10 compares our proposed CTSD channel estimation with IAM and ICM schemes. Here, the MIMO-IAM-C preamble (which is designed based on Walsh-Hadamard matrix) is considered following the method in [22]. One can see that (1) there exists an MSE floor for each channel estimation scheme with $G = 2$ owing to the residual ISI from neighboring symbols; (2) Our proposed CTSD method achieves channel estimation MSE smaller than ICM and comparable to IAM, but less time-frequency resources.

Fig 11 shows uncoded bit error rate (BER) comparison between preamble-based MIMO-OFDM and MIMO-FBMC systems. In this simulation, we assume QPSK modulation and 4 transmit antennas. We also assume the CP duration takes 1/4 of the useful data part in MIMO-OFDM system. It is shown that MIMO-FBMC system achieves the better BER performance as no energy is wasted in CP part in low SNR region. One can see that the BER of MIMO-FBMC system with CTSD preamble structure has 1.7dB performance gain over that of MIMO-OFDM system. However in high SNR region (i.e., SNR>30dB), MIMO-OFDM systems outperform MIMO-FBMC systems. This is because MIMO-FBMC systems cannot completely remove ISI from next data symbols, whereas the CP in MIMO-OFDM systems can remove the effect of ISI. Moreover, the CTSD-based MIMO-FBMC system achieves the better BER performance than ICM-based systems, and is comparable to that of the IAM-based systems.

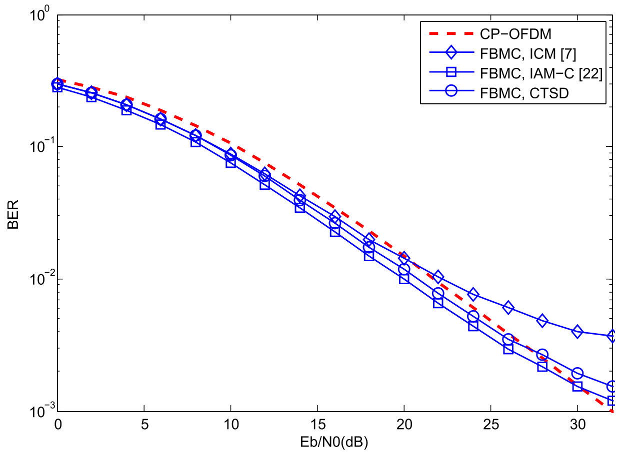

FIGURE 11. BER performance comparison between uncoded MIMO-OFDM and MIMO-FBMC systems ($L_h = 12$).

V. CONCLUSIONS

In this paper, we have proposed a new class of training sequences for efficient channel estimation in MIMO-FBMC systems. Each training sequence is formed by concatenating two identical zero-correlation zone (ZCZ) sequences and because of this, (1) all the odd-indexed subcarriers are set to zero, leading to substantial ICI suppression; (2) the ZCZ property these training sequences inherited from the "seed" set allows them to be simultaneously sent out from all transmit antennas, i.e., coded-division multiplexing based channel estimation. In order to attain the complex-field orthogonality of training sequences, we have proposed complex training sequence decomposition (CTSD) which splits the real- and imaginary- parts of every training sequence and then transmit them separately over two FBMC symbols only. This saves huge amount of precious time-frequency resources as the resultant training overhead does not depend the number of transmit antennas. Simulation results have validated that our proposed CTSD method is an efficient channel estimation scheme for practical MIMO-FBMC applications.

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SU HU received the M.S. and Ph.D. degrees from the National Key Laboratory on Communications, University of Electronic Science and Technology of China (UESTC) in 2007 and 2010, respectively. From 2011 to 2012, he was a Research Fellow with the School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore. He is currently an Associate Professor with UESTC. His research interests include sequence design with good correlation properties and physical layer design for wireless communication systems, such as filterbank multicarrier systems and cognitive radio networks.

ZILONG LIU received the B.S. degree from the School of Electronics and Information Engineering, Huazhong University of Science and Technology, in 2004, the M.S. degree from the Department of Electronic Engineering, Tsinghua University, China, in 2007, and the Ph.D. degree, with a focus on perfect- and quasi- complementary sequences, under the supervision of Prof. Y. L. Guan. Since 2008, he has been with the School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore, as a Research Associate. He was a Visiting Ph.D. student with the University of Melbourne from 2012 to 2013 [with Prof. U. Parampalli and Prof. S. Boztas (RMIT University)], and with The Hong Kong University of Science and Technology in 2013 (with Prof. W. H. Mow). He is generally interested in coding and modulation for (asynchronous/robust) communication systems, with emphasis on sequence design, interference-free multiuser communications (CDMA and OFDM), and physical layer receiver design.

YONG LIANG GUAN (M'94) received the B.E. degree (Hons.) from the National University of Singapore and the Ph.D. degree from Imperial College London, U.K. He is currently an Associate Professor and the Head of Communication Engineering Division with the School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore. His research interests broadly include modulation, coding, and signal processing for communication systems and information security systems.

CHUANXUE JIN received the M.S. degree from the National Key Laboratory on Communications, University of Electronic Science and Technology of China (UESTC) in 2006. He is currently a Lecture with UESTC. His research interests include hardware design for wireless communication systems, such as filterbank multicarrier systems and cognitive radio networks.

YIXUAN HUANG received the bachelor's degree from the University of Electronic Science and Technology of China, China, in 2015, where he is currently pursuing the master's degree with the National Key Laboratory of Science and Technology on Communication. His main fields of research interest include wireless communication (OFDM and CE-OFDM), and detecting and communication integrative system.

JEN-MING WU received the B.S. degree from National Taiwan University, Taipei, Taiwan, the M.S. degree from the Polytechnic Institute of New York University, and the Ph.D. degree from the University of Southern California, all in electrical engineering. From 1998 to 2003, he was with Sun Microsystems Inc., Sunnyvale, CA, USA, as member of Technical Staff. Since 2003, he has been a Faculty Member with the Department of Electrical Engineering, Institute of Communications Engineering, National Tsing Hua University, Taiwan, where he currently holds an associate professor position. He has involved on various fields of Electrical Engineering, including signal processing for communications, wireless communication transceiver IC design, high speed I/O interface IC design, and microprocessor architecture. He has authored over 80 technical papers in IEEE journals and conferences. He has led the design and received the Golden Medal of the Golden Silicon Award in 2006 and 2008, and the National CIC Outstanding Chip Design Award in 2010 and 2011. He has served on the Technical Program Committee of the IEEE ICC, the Globecom, the VTC and many other IEEE conferences. His research interests focus on MIMO signal processing, wireless interference mitigation, software defined radio, and high speed interface technologies.