Real-Valued Orthogonal Sequences for Ultra-low Overhead Channel Estimation in MIMO-FBMC Systems

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Abstract. Multiple-input multiple-output filterbank multicarrier communication (MIMO-FBMC) is a promising technique to achieve very tight spectrum confinement (thus, higher spectral efficiency) as well as strong robustness against dispersive channels. In this paper, we present a novel training design for MIMO-FBMC system which enables efficient estimate of frequency-selective channels (associated to multiple transmit antennas) with only one non-zero FBMC symbol. Our key idea is to design real-valued orthogonal training sequences (in the frequency domain) which displaying zero-correlation zone properties in the time-domain. Compared to our earlier proposed training scheme requiring at least two non-zero FBMC symbols (separated by several zero guard symbols), the proposed scheme features ultra-low training overhead yet achieves channel estimation performance comparable to our earlier proposed complex training sequence decomposition (CTSD). Our simulations validate that the proposed method is an efficient channel estimation approach for practical preamble-based MIMO-FBMC systems.

Keywords: Filterbank multicarrier · MIMO-FBMC
Channel estimation · Preamble · Real-valued orthogonality
Zero-correlation zone sequences

1 Introduction

Future wireless communications must be highly spectral efficient. For multi-carrier communications, we require waveform pulses with ultra-tight spectrum confinement as well as efficient channel estimation scheme having ultra-low training overhead. As an enhanced modulation scheme to orthogonal frequency-division multiplexing (OFDM), filterbank multicarrier (FBMC) systems employing offset quadrature amplitude modulation (OQAM), called OFDM/OQAM

L. Meng and Y. Zhang (Eds.): MLICOM 2018, LNICST 251, pp. 162–171, 2018.
https://doi.org/10.1007/978-3-030-00557-3_17

or FBMC/OQAM, achieves very high spectrum efficiency and have attracted increased research attention in recent years [1].

Aiming to eliminate the intrinsic imaginary interference incurred from adjacent symbols, a number of training schemes and associated estimation methods for preamble-based FBMC systems have been proposed [2–6]. These methods, however, generally suffer from heavy training overhead, especially when the number of transmit antennas becomes large. For improved channel estimation performance, there are three obstacles to be attacked: (1), inter-carrier interference (ICI) and inter-symbol interference (ISI); (2), correlation properties of orthogonal sequences could be damaged by the intrinsic imaginary interference; (3), the "real-field orthogonality" (which is pertinent to FBMC) generally prevents the use of any channel estimation methods which require the "complex-field orthogonality".

Recently, a training method called complex training sequence decomposition(CTSD) [7], which requires two non-zero FBMC symbols for channel estimation in MIMO-FBMC has been proposed. CTSD is specifically designed to facilitate the reconstruction of the complex-field orthogonality of MIMO-FBMC signals by superimposing different training sequences over the air (based on code-division multiplexing, CDM). This paper aims to improve our earlier proposed CTSD by designing real-valued orthogonal sequences for novel training scheme with one FBMC symbol only.

The paper is organized as follows: Sect. 2 presents the system model of FBMC systems and introduces the constraint conditions for channel estimation in MIMO-FBMC systems. In Sect. 3, we present a design of training symbols having real-valued orthogonality and capability of interference self-cancellation, followed by detailed description of corresponding training scheme. Numerical simulations are presented in Sect. 4. In the end, this paper is summarized in Sect. 5.

2 System Model and Constraint Conditions

2.1 System Model

In this paper, we consider an equivalent FBMC baseband model with M subcarriers, where the subcarrier spacing is $1/T$ with T being complex symbol interval. The equivalent discrete-time FBMC signal is expressed as [8]

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

where $j = \sqrt{-1}$, $a_{m,n}$ is the real-valued offset QAM symbol transmitted over the mth subcarrier and the nth time-slot, and $T/2$ is the interval of real-valued symbols. Meanwhile, $g\,(l)$ is the employed symmetrical real-valued prototype filter impulse response with length of $L_g = KM$ and K is the overlapping factor.

$g_{m,n}(l)$ represents the synthesis basis which is obtained by the time-frequency translated version of $g(l)$.

Let $\mathbf{h} = [h(0), h(1), \cdots, h(L_h - 1)]^T$ 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

$$r(l) = \sum_{\tau=0}^{L_h-1} h(\tau) s(l-\tau) + \eta(l), \tag{2}$$

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 lattice provides a complex symbol given as

$$y_{m,n} = \sum_{l=-\infty}^{\infty} r(l) g\left(l - \frac{nM}{2}\right) e^{-j2\pi ml/M} j^{m+n}. \tag{3}$$

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 over any time interval $[l, l+L_h]$, i.e., $g(l) \approx g(l+\tau)$, for $\tau \in [0, L_h]$. Hence, $y_{m,n}$ in (3) can be simplified to

$$y_{m,n} = H_{m,n} a_{m,n} + \underbrace{\sum_{(p,q)\neq(m,n)} H_{p,q} a_{p,q} \zeta_{m,n}^{p,q}}_{I_{m,n}} + \eta_{m,n}, \tag{4}$$

where $H_{m,n}$ denotes the channel frequency response at the (m,n) th lattice, and $I_{m,n}$ and $\eta_{m,n}$ are the intrinsic interference and noise terms, respectively.

The above FBMC formulation in SISO scenario may 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

$$y_{m,n}^k = \sum_{i=1}^{N_T} \left\{ H_{m,n}^{k,i} a_{m,n}^i + \sum_{(m,n)\neq(p,q)} H_{p,q}^{k,i} a_{p,q}^i \zeta_{m,n}^{p,q} \right\} + \eta_{m,n}^k, \tag{5}$$

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 component in the kth receive antenna.

2.2 Constraint Conditions

Preamble design in MIMO-FBMC system needs to satisfy several constraint conditions presented below.

(1): Compared with CP-OFDM system with complex-field orthogonality, the orthogonality of FBMC system only maintains in real-field. Recalling from PHYDYAS European project [9], one needs to deal with intrinsic imaginary interference from any lattice point $(m,n) \neq (p,q)$, even passing through a distortion-free

channel. Hence, it is ideal to have zero (or close to zero) $I_{m,n}$ in Eq. (3) by proper training design.
(2): To separate the transmission data from different antennas properly, the training set $S = \{S_1, S_2, \cdots, S_K\}$, each sequence having length L, is required to be a zero-correlation-zone (K, L, Z) sequence set [7], which is defined as follows.

$$\begin{array}{l} (1)\,,\ R_{S_\mu,S_\mu}(\tau) = 0,\ 1 \leq \mu \leq K \text{ and } 1 \leq |\tau| < Z; \\ (2)\,,\ R_{S_\mu,S_\nu}(\tau) = 0,\ \mu \neq \nu \text{ and } 0 \leq |\tau| < Z; \end{array} \tag{6}$$

where $R_{S_\mu,S_\nu}(\tau)$ denotes the periodic cross-correlation function between sequences S_μ and S_ν at time-shift τ (see [7] for detailed definition).
(3): To overcome the drawback of CTSD which requires two non-zero FBMC symbols for efficient training, it is ideal to design a ZCZ training set each sequence consists of purely real-valued elements in the frequency domain.

3 Efficient Preamble-Based Channel Estimation for MIMO-FBMC System

Motivated by the inefficient channel estimation approach above-mentioned and special attributes of the MIMO-FBMC system, our idea is to form orthogonal sequences set which has real-field orthogonality and interference self-cancellation capability.

3.1 Preamble Design of Real-Valued Orthogonal Sequences

Based on the constraint conditions stated in the above, we present below a ZCZ sequence set each sequence having purely real-valued elements in the frequency-domain:
STEP 1: Let $\mathbf{M} = [M(0), M(1), \cdots, M(N-1)]$ which is a real-valued unimodular sequence satisfying certain correlation properties. Consider the following sequence in the frequency-domain.

$$\begin{aligned} \mathbf{A}_1 &= [A_1(0), A_1(1), \cdots, A_1(2N-1)]_{1\times 2N} \\ &= [M(0), 0, M(1), 0, \cdots, M(N-1), 0]_{1\times 2N}. \end{aligned} \tag{7}$$

STEP 2: Apply IDFT to $\mathbf{A}_1$. Since $\mathbf{A}_1$ is obtained by inserting zeros into $\mathbf{M}$ in the frequency-domain, the time-domain sequence can be expressed as $\alpha_1 = [\alpha_1(0), \alpha_1(1), \cdots, \alpha_1(2N-1)]_{1\times 2N} = [\beta; \beta]$, where $\beta = [\alpha_1(0), \cdots, \alpha_1(N-1)]_{1\times N}$. Then, we have

$$\begin{aligned} \alpha_1(t) &= \frac{1}{\sqrt{2N}} \sum_{K=0}^{2N-1} A_1(k) \exp(\frac{j2\pi kt}{2N}) \\ &= \frac{1}{\sqrt{2N}} \sum_{K=0}^{N-1} M(k) \exp(\frac{j2\pi kt}{N}) = \frac{1}{\sqrt{2}} \mathrm{IDFT}(M). \end{aligned} \tag{8}$$

It can be easily proved that the periodic autocorrelation function (PACF) of α_1 is zero except at the time-shift of N (i.e., the middle time-shift position).

Consider the following time-shifted version of α_1 which is defined as

$$\alpha_2 = T^{N/2}(\alpha_1) = [\alpha_1(3N/2), \alpha_1(3N/2+1), \cdots, \alpha_1(3N/2-1)]_{1\times 2N}, \tag{9}$$

where $T^{\tau}(\cdot)$ denotes the right-cyclically shifted version of $(\cdot)$ for τ positions. Note that a time domain shift corresponds to a phase rotation in the frequency domain. It can be easily proved that the corresponding frequency domain sequence is $\mathbf{A}_2 = [A_2(0), A_2(1), \cdots, A_2(2N-1)]_{1\times 2N}$, where

$$A_2(k) = \exp(-\frac{j\pi k}{2})A_1(k) = \begin{cases} (-1)^{k/2}A_1(k),\ k \text{ is even;} \\ 0,\ \text{otherwise} \end{cases} \tag{10}$$

In this way, two columns of real-valued training sequences have been obtained.
STEP 3: Choose another seed sequence $\mathbf{M}'$ satisfying certain correlation properties as that of $\mathbf{M}$. By STEP 1, we obtain the third real-valued training sequence $\mathbf{A}_3$ whose time-domain dual is expressed as α_3.
STEP 4: Following STEP 2, α_4 is obtained by applying right-cyclic shift of $N/2$) to $\mathbf{A}_3$. Therefore, four ZCZ training sequences can be obtained. Next, we verify the orthogonality of these sequences.

Let us recall the original sequence $\mathbf{M}$ and the newly designed sequence $\mathbf{A}_1$. Since the frequency-domain preamble sequences are obtained by inserting zeros, the time-domain sequence α_1 is formed by cascading two identical sequences β, i.e. $\alpha_1 = [\beta; \beta]$.

$$R_{\alpha_1,\alpha_1}(\tau) = \sum_{n=0}^{2L-1} \alpha_1(n)(\alpha_1(n+\tau))^* = 2\sum_{n=0}^{L-1} \beta(n)(\beta(n+\tau))^* = 2R_{\beta,\beta}(\tau) \tag{11}$$

So, if $\mathbf{M}$ is a ZCZ sequence of length N, α_1 will be a ZCZ sequence of length $2N$. By calculating their related properties, we assert that $\{\alpha_1, \alpha_2, \alpha_3, \alpha_4\}$ is a $(4, 2N, N/2)$ ZCZ sequences set.

$$\begin{array}{ll} (1) & R_{\alpha_i,\alpha_i}(\tau) = 0,\ 1 \le i \le 4 \text{ and } 1 \le |\tau| < N \\ (2) & R_{\alpha_i,\alpha_j}(\tau) = 0,\ i \ne j \text{ and } 0 \le |\tau| < N/2 \end{array} \tag{12}$$

In this paper, we select the first two columns of a Hadamard matrix as "seed" sequences $\mathbf{M}$ and $\mathbf{M}'$ to generate our training sequences.

3.2 Efficient Channel Estimation

Once N_T sequences are generated, they will be simultaneously transmitted over N_T transmit antennas. Our proposed method suppresses the ISI by inserting G zero symbols as shown in Fig. 1. Because of zero insertion in the training sequences, ICI can be substantially suppressed. In addition, the proposed method takes up only one non-zero FBMC symbol (in contrast to two non-zero FBMC

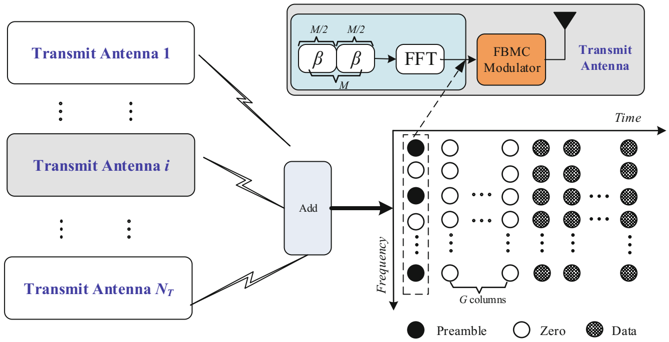


Fig. 1. Proposed preamble structure in MIMO-FBMC system with N_T transmit antenna.

training symbols in CTSD [7]) and therefore, about half training overhead can be saved.

For an $N_T \times N_R$ MIMO-FBMC system, there are $N_T N_R$ independent channels to be measured, in which 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 the channel is quasi-static. The key idea is to use the real-valued orthogonal sequences to estimate the time-domain channel coefficients. The channel impulse response can be expressed as

$$\mathbf{h}_{i,k} = [h_{i,k}(0), h_{i,k}(1), \cdots, h_{i,k}(L-1)]^T, \tag{13}$$

$\mathbf{h}_{i,k}$ represents the channel impulse response vector from the i-th ($i = 1, 2, \cdots, N_T$) transmit antenna to the k-th ($k = 1, 2, \cdots, N_R$) receive antenna. Similar to SISO-FBMC system, the demodulation of the received signal at the (m, n)th time-frequency lattice in the kth receive antenna associated to $\mathbf{A}_i$ can be written as

$$y_k(m) \approx \sum_{i=1}^{N_T} \sum_{\tau=0}^{L_h-1} A_i(m) e^{-j2\pi m\tau/M} h_{i,k}(\tau) + \eta_k(m). \tag{14}$$

For the ease of expression, let $\mathbf{W}_M = e^{-j2\pi/M}$. Note that $A_i(m) = \sum_{n=0}^{M-1} \alpha_i(n) \mathbf{W}_M^{m,n}$, and let $n' = n + \tau$. Then (14) can be rewritten as (15), where $(\cdot)_M$ represents the modulo M operation.

$$y_k(m) = \sum_{\tau=0}^{L_h-1} \left(\sum_{i=1}^{N_T} \left(\sum_{n'=0}^{M-1} \alpha_i(n'-\tau)_M W_M^{mn'} \right) h_{i,k}(\tau) \right) + \eta_k(m). \tag{15}$$

The above formula can be written in matrix form as $\mathbf{y}_k = \mathbf{WSh}_k + \eta_k$, where $\mathbf{S} = [\mathbf{S}_1, \mathbf{S}_2, \cdots, \mathbf{S}_{N_T}]^T$, $\mathbf{h}_k = [h_{1,k}, h_{2,k}, \cdots, h_{N_T,k}]^T$,

$$\mathbf{S}_i = \begin{bmatrix} \alpha_i(0) & \alpha_i(M-1) & \cdots & \alpha_i(M-L_h+1) \\ \alpha_i(1) & \alpha_i(0) & \cdots & \alpha_i(M-L_h+2) \\ \vdots & \vdots & \ddots & \vdots \\ \alpha_i(M-1) & \alpha_i(M-2) & \cdots & \alpha_i(M-L_h) \end{bmatrix}. \tag{16}$$

Here, $\mathbf{S}_i, i = 1, 2, \cdots, N_T$ are $M \times L_h$ matrices shown as (16), where the matrix $\mathbf{W}$ is an FFT matrix. Applying IFFT to $\mathbf{y}_k$, we obtain

$$\mathbf{r}_k = \frac{1}{M}\mathbf{W}^H \mathbf{y}_k = \mathbf{Sh}_k + \mathbf{w}_k, \tag{17}$$

where

$$\mathbf{w}_k = [\omega_k(0), \omega_k(1), \cdots, \omega_k(M-1)]^T,\ \omega_k(n) = \tfrac{1}{M}\sum_{m=0}^{M-1} \eta_k(m) e^{j2\pi mn/M}. \tag{18}$$

Since in general the rank of S is equal to the number of columns for $M > L_h$, the linear least square channel estimator for the kth receive antenna is shown as

$$\tilde{\mathbf{h}}_k = [\tilde{h}_{1,k}, \tilde{h}_{2,k}, \cdots, \tilde{h}_{N_T,k}]^T = (\mathbf{S}^H\mathbf{S})\mathbf{S}^H \mathbf{r}_k. \tag{19}$$

If the noise terms satisfy zero-mean white normal distribution, one can see that the channel estimator is unbiased, which means $E[\tilde{h}_k] = h_k$. And thus the channel estimation MSE can be defined as

$$MSE = \mathrm{E}\left[(h_k - \tilde{h}_k)^H (h_k - \tilde{h}_k)\right] = 2\sigma^2 \mathrm{Tr}((\mathbf{S}^H\mathbf{S})^{-1}), \tag{20}$$

where $\mathrm{E}(\cdot)$ denotes the expectation and $\mathrm{Tr}(\cdot)$ denotes the matrix trace operation.

4 Simulation Result

In this section, we calculate the spectrum efficiency of different systems and evaluate the performance of our proposed channel estimation method in terms of mean square error (MSE) of channel estimation.

4.1 Comparison of Spectrum Efficiency

The normalized spectrum efficiency γ of MIMO-FBMC system compared with ideal MIMO-OFDM system with no overhead (i.e., no guard time interval and no frequency-domain training symbols) are

$$\gamma_{FBMC} = \tfrac{N_D}{N_D + N_P} \times 100\% \quad \gamma_{OFDM} = \tfrac{M}{N_{CP} + M} \times 100\%, \tag{21}$$

where N_D and N_P denote the number of data and training symbols, and N_{CP} denotes the length of the cyclic prefix, respectively.

For a MIMO-OFDM system, every distinct ZCZ sequence is sent in a different transmit antenna and superimposed within the same time-frequency resource, leading to $N_P = 1$ complex-valued symbols. The main wasting of spectrum lies in CP (1/4 or 1/8 of a symbol). In contrast, a non-CP MIMO-FBMC system relies on well-localized time-frequency pulse shaping filter to suppress the effect of dispersive channel. Both the IAM and ICM require $N_T(G+1)$ real-valued symbols. Obviously, an increasing N_T will lower the system spectrum efficiency.

Our earlier proposed CTSD is a code–division multiplexing based channel estimation whose training overhead has nothing to do with the number of N_T. But its preamble data has complex-value in the frequency domain and this means it requires at least two FBMC symbols as well as $3G$ columns of zeros. In this paper, our proposed method only requires one FBMC training symbol and G zero symbols. Because of this, our proposed channel estimation scheme owns higher spectrum efficiency, as shown in Table 1.

Table 1. Spectrum efficiency comparison ($N_D = 50$)

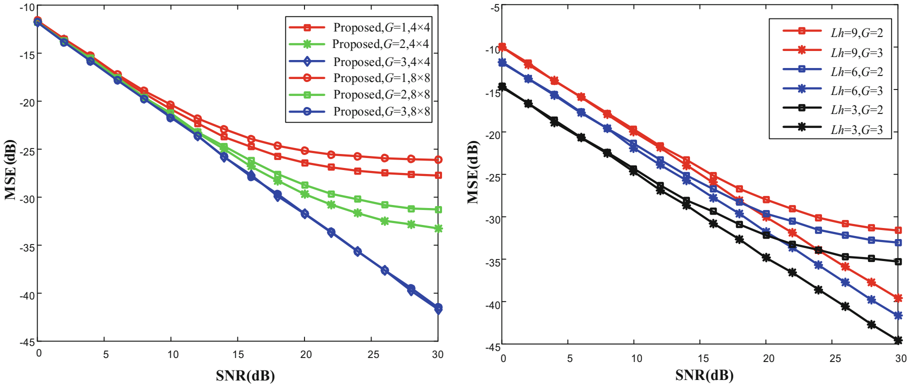
	MIMO-OFDM		MIMO-FBMC ($G = 3$)		
	CP = 1/8	CP = 1/4	IAM & ICM	CTSD	Proposed
$N_T = 2$	88.9%	80%	92.6%	90.1%	96.2%
$N_T = 4$	88.9%	80%	86.2%	90.1%	96.2%
$N_T = 8$	88.9%	80%	75.8%	90.1%	96.2%

4.2 Numerical Simulations

MIMO-FBMC systems employing the pulse shaping filter in the PHYDYAS project with 256 subcarriers are considered. Moreover, multipath channels are adopted, each having L_h sample-paced fading coefficients, in which the channel tap coefficients are assumed to be independent complex Gaussian random variables with uniformly distributed phases and Rayleigh distributed envelopes. The transmitted power from each transmit antenna is assumed to be same.

Figure 2 shows the MSE performance of the proposed channel estimation method with different guard interval G, antenna number N_T and multipath channels number L_h. It is shown that increasing N_T and L_h will lower the estimation performance. The MSE floors at high SNR region for $G = 1, 2$ appearing in our proposed method are mainly caused by the residual ISI from neighboring training- or data- symbols and interference from quasi-orthogonal sequences. But when $G = 3$ increases, the MSE floor at high SNR region disappears.

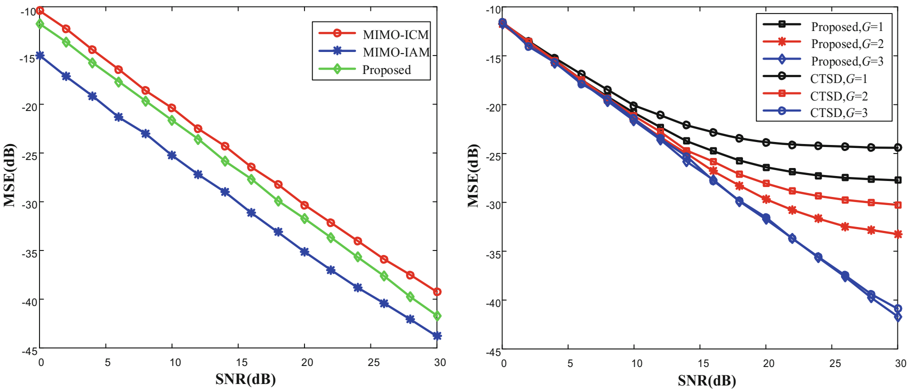
Figure 3(a) compares our proposed method with IAM and ICM. Note that training symbols in IAM or ICM are sent in turn over different transmit antennas and thus the MSE for each method is equivalent to that in SISO scenario. One can see that, the performance of our proposed method is better than ICM. IAM shows a better performance because of the interference approximation and more preamble resourse.



(a) $N_T = 4, 8$ (b) $L_h = 3, 6, 9$

Fig. 2. MSE performance comparison of the proposed method in different situations.

Figure 3(b) compares our proposed channel estimation with CTSD. For both methods, the average power of the training sequences (preamble symbol and the guard interval) is one. According to the simulation result, the MSE performance of our proposed method is better than CTSD. That's because there are two real-valued symbols in CTSD to be calculated and thus larger amount of interference incurred, but the proposed method has only one.



(a) compare with IAM,ICM (b) compare with CTSD

Fig. 3. MSE performance comparison between our proposed method and the former methods.

5 Conclusion

In this paper, we have presented an efficient channel estimation scheme in MIMO-FBMC systems. Our proposed training scheme is based on "real-valued orthogonal sequences" and allows simultaneous transmission of preambles over one FBMC symbol, regardless of the number of transmit antennas. Compared to our previously proposed CTSD training scheme in [7], the overall training overhead is reduced by half. Numerical simulation results have shown that our proposed training scheme strikes a good tradeoff between channel estimation performance and training overhead.

Acknowledgement. The work of S. Hu is jointly supported by the MOST Program of International S&T Cooperation (Grant No. 2016YFE0123200), Science and Technology on Electronic Information Control Laboratory (Grant No. 6142105040103) and Fundamental Research Funds for the Central Universities (Grant No. ZYGX2015J012). The work of Z. Liu and S. Hu was supported in part by National Natural Science Foundation of China (Grant No. 61750110527, Grant No. 61471100/61701503), Research Fund for International Young Scientists. The work of Z. Liu and P. Xiao was also supported in part by European Commission under the 5GPPP H2020 Clear5G Project (Grant No. 761745). The support of members at 5GIC of the University of Surrey (www.surrey.ac.uk/5gic) during this work is also acknowledged.

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