Robust Precoded MIMO-OFDM for Mobile Frequency-Selective Wireless Channels

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***Abstract*—We present a new multiple input multiple output (MIMO) technique, based on Walsh Hadamard Transform (WHT) precoding, to improve the robustness of space-frequency block coded orthogonal frequency division multiplexing (SFBC-OFDM) systems. The WHT is applied to the data symbols prior the Alamouti encoder at the transmitter and to the output of the Alamouti decoder at the receiver. The computational complexity of the proposed system is evaluated in terms of complex additions and multiplications where the numerical results show that the proposed system has a lower complexity as compared to other precoded OFDM systems. Moreover, the proposed system is highly robust to the channel time and frequency selectivity as compared to conventional SFBC, space time block coded (STBC) and other precoded OFDM systems.**

***Index Terms*—Precoding, OFDM, SFBC, STBC, Walsh.**

I. Introduction

ORTHOGONAL frequency division multiplexing (OFDM) has attracted significant attention in wireless communications because of its ability to convert severe frequency-selective fading channels into flat sub-channels. Furthermore, OFDM simplifies the receiver complexity because channel equalization can be performed using single-tap equalizers. Despite the fact that OFDM has some limitations such as high peak to average power ratio (PAPR) [1], owing to its remarkable advantages, OFDM has been adopted for several wireless communication standards such as LTE and LTE-A [2], and it is considered as a strong candidate for the fifth generation (5G) standard.

Despite the many advantages of OFDM, its bit error rate (BER) performance in fading channels is similar to single carrier systems. Therefore, incorporating diversity techniques is indispensable to combat time and frequency selectivity of the channel, and to improve the BER. In the literature, spacial diversity techniques, such as space time block codes (STBC), is considered as a pivotal tool to combat the consequences of deep fading on OFDM [3]. Moreover, the performance improvement can be gained using linear receivers similar to the ones used for single carrier systems [4], and hence, the additional complexity introduced is limited.

Although STBC-OFDM is attractive for several applications, using a large number of antennas might be infeasible for small size handheld devices. Consequently, other alternative diversity techniques have been considered in the literature. For example, precoded OFDM have demonstrated high robustness against the frequency selectivity of wireless channels [5], [6]. In such configurations, the frequency-domain symbols are applied to a particular unitary transform prior to the fast Fourier transform (FFT) to create frequency diversity [5], [6].

Despite the significant BER improvement that can be gained using precoding, Chung and Phoong [7] showed that combing spacial and frequency diversity is more beneficial than using any of them individually. Therefore, a precoded STBC-OFDM system was formed by combining Walsh Hadamard transform (WHT) with Alamouti STBC [7]. The numerical results reported in [7] confirm the robustness of the campsite diversity approach. However, the ultimate BER improvement is achieved only in static channels. In time varying channels, the BER suffers from severe degradation, which is typically manifested as high error floors. Furthermore, the precoding process increases the computational complexity of conventional OFDM due to the additional computations of the WHT. However, the computational complexity can be reduced by combining the WHT and FFT into a single transform [8], [6].

It is worth noting that low complexity STBC-OFDM can be achieved by assuming that the channel remains constant over two OFDM blocks. However, in time varying channels, the fixed channel assumption is no longer valid, which results in severe BER. In order to mitigate the adverse effects of time varying channels while using simple receivers, space frequency block coded OFDM (SFBC-OFDM) can be invoked [9], [10]. In SFBC, the space-time block is transmitted over two adjacent OFDM subcarriers. Consequently, SFBC-OFDM suffers from severe performance degradation in frequency-selective channels, where the channel response is not constant over adjacent subcarriers.

In the literature, several techniques have been proposed to improve the performance of STBC-OFDM and SFBC-OFDM systems over frequency-selective time-varying channels. For example, a precoded STBC-OFDM system is proposed in [11], which introduces only 3 dB improvement over STBC-OFDM systems at high signal to noise ratio (SNR) values. A space-time-frequency OFDM system is proposed in [12], however the performance improvement is limited. Al-Dweik *et al.* [13] proposed a novel precoded SFBC-OFDM system by forcing the channel frequency response to be equal over the SFBC block. The proposed system, denoted as channel matrix shaping (CMS), showed a significant performance improvement in time varying channels with severe frequency selectivity. However, BER reduction with low complexity can be achieved only using binary phase shift keying (BPSK) modulation.

Inspired by [7], we propose in this work a precoded SFBC (P-SFBC) system that combines both space and frequency diversity. Unlike the precoded system reported in [7], the P-SFBC is constructed in frequency domain, and hence it is more immune to the channel variations. Moreover, the complexity of the P-SFBC is less than [7] because we use $0.5N$-point instead of N-point WHT. The BER of the proposed system is evaluated under static, typical urban and bad urban channels. The computational complexity of the proposed system is estimated in terms of complex additions and multiplications, and the BER is compared to conventional SFBC, STBC, WHT-STBC [7] and CMS [13] systems.

The rest of this paper is organized as follows. Section II illustrates the precoded STBC system models. In Section III, the proposed P-SFBC system is presented. Simulation results are discussed in Section IV, and finally conclusion remarks are given in Section V.

In this work, the following notations are used. Uppercase boldface letters, such as $\mathbf{H}$, denote $N \times N$ matrices. Lowercase boldface letters, such as $\mathbf{p}$, denote row or column vectors with N elements. The complex conjugate and the transpose of $\mathbf{p}$ are denoted as $\check{\mathbf{p}}$ and $\mathbf{p}^T$, respectively. The expectation of a random variable x is denoted as E$[x]$.

II. Precoded STBC-OFDM System Model

Consider a precoded STBC-OFDM, denoted as WHT-STBC, that consists of two transmit and two receive antennas [7], [14]. The channel is assumed to be time invariant over two consecutive OFDM symbols, which corresponds to a quasi static fading channel. Two data sequences $\mathbf{a}^1$ and $\mathbf{a}^2$ each of which consists of N symbols are uniformly selected from a general Quadrature Amplitude Modulation (QAM) constellation. Then, each sequence is applied to an N-point WHT. Using vector notation, the WHT output can be expressed as

$$\mathbf{p}^i = \mathbf{T}_N \, \mathbf{a}^i, \;\; i = [1, 2] \tag{1}$$

where $\mathbf{T}_N$ denotes the $N \times N$ WHT matrix. The STBC encoder output $\mathbf{p}^1$ and $\mathbf{p}^2$ can be expressed as

$$\begin{bmatrix} \mathbf{p}^1 & \mathbf{p}^2 \end{bmatrix} \overset{STBC}{\rightarrow} \begin{bmatrix} \mathbf{p}^1 & -\check{\mathbf{p}}^2 \\ \mathbf{p}^2 & \check{\mathbf{p}}^1 \end{bmatrix} \triangleq \begin{bmatrix} \mathbf{u}^{1,1} & \mathbf{u}^{1,2} \\ \mathbf{u}^{2,1} & \mathbf{u}^{2,2} \end{bmatrix}$$

where $\mathbf{p}^i \in \mathbb{C}^{N\times 1}$ and $i = [1, 2]$, the inverted caret symbol denotes the complex conjugate and $\mathbf{u}^{i,\tau}$ denotes the transmitted block from antenna i at time period τ, $\{i, \tau\} \in \{1, 2\}$. Before transmission, each $\mathbf{u}^{i,\tau}$ block is OFDM modulated by applying an N-point inverse FFT (IFFT). The IFFT output can be written as $\mathbf{x}^{i,\tau}= \mathbf{F}^H\mathbf{u}^{i,\tau}$, where $\mathbf{F}^H$ is the Hermitian transpose of the normalized $N \times N$ FFT matrix $\mathbf{F}$. A cyclic prefix (CP) of P samples is added as a preamble at the beginning of each IFFT output $\mathbf{x}^{i,\tau}$ to compose a complete OFDM symbol with a total duration of T_t seconds and $N_\mathrm{t} = N+P$ samples. The four OFDM symbols, each of which consists of N_t samples, are upconverted to higher frequency bands and transmitted at time slot τ through antenna i.

The channels between transmit antenna $i \in \{1, 2\}$ and receive antenna $c \in \{1, 2\}$ are assumed to be independent Rayleigh fading channels. Each channel consists of $L_h^{i,c,\tau} + 1$ independent multipath components where each path has gain $h_l^{i,c,\tau}$ and delay $l^{i,c,\tau} \times T_\mathrm{s}$, where T_s is the sampling period and $l^{i,c,\tau} \in \left\{0, 1, ..., L_h^{i,c,\tau}\right\}$. At the receiver side, the received signals are downconverted to baseband with a sample period $T_\mathrm{s} = T_\mathrm{t}/N_\mathrm{t}$. Given that the system has perfect channel estimation and timing synchronization, the received signals at receiver c and time slot τ after removing the CP samples and applying FFT process can be expressed as

$$\mathbf{r}^{c,\tau} = \mathbf{H}^{1,c,\tau}\mathbf{u}^{1,\tau} + \mathbf{H}^{2,c,\tau}\mathbf{u}^{2,\tau} + \boldsymbol{\eta}^{c,\tau} \tag{2}$$

where $\boldsymbol{\eta}$ denotes the additive white Gaussian noise (AWGN) with zero-mean and variance $E[|\boldsymbol{\eta}|^2]$. If the channel is fixed during two OFDM symbols, the channel frequency response during each symbol can be represented as a diagonal matrix $\mathbf{H}$ =diag$([H_0, H_1, ..., H_{N-1}])$, where $H_k = \sum_{l=0}^{L_h} h_l e^{-j2\pi lk/N}$. Moreover, the received signals after FFT at the receivers' side can be expressed as

$$\mathbf{r}^{c,\tau} = \mathbf{H}^{1,c}\mathbf{u}^{1,\tau} + \mathbf{H}^{2,c}\mathbf{u}^{2,\tau} + \boldsymbol{\eta}^{c,\tau}. \tag{3}$$

However, the received signal in (3) can be expanded to

$$\begin{aligned} \mathbf{r}^{1,1} &= \mathbf{H}^{1,1}\mathbf{p}^1 + \mathbf{H}^{2,1}\mathbf{p}^2 + \boldsymbol{\eta}^{1,1} \\ \mathbf{r}^{2,1} &= \mathbf{H}^{1,2}\mathbf{p}^1 + \mathbf{H}^{2,2}\mathbf{p}^2 + \boldsymbol{\eta}^{2,1} \\ \mathbf{r}^{1,2} &= \mathbf{H}^{2,1}\check{\mathbf{p}}^1 - \mathbf{H}^{1,1}\check{\mathbf{p}}^2 + \boldsymbol{\eta}^{1,2} \\ \mathbf{r}^{2,2} &= \mathbf{H}^{2,2}\check{\mathbf{p}}^1 - \mathbf{H}^{1,2}\check{\mathbf{p}}^2 + \boldsymbol{\eta}^{2,2}. \end{aligned} \tag{4}$$

Therefore, we can write the kth element of $\mathbf{r}^{c,\tau}$ as

$$\mathbf{v}_k = \boldsymbol{\mathcal{D}}_k\mathbf{p}_k + \boldsymbol{\eta}_k \tag{5}$$

where $\mathbf{v}_k = \begin{bmatrix} r_k^{1,1} & r_k^{2,1} & \check{r}_k^{1,2} & \check{r}_k^{2,2} \end{bmatrix}^T$, $\mathbf{p}_k = \begin{bmatrix} p_k^1 & p_k^2 \end{bmatrix}^T$ and

$$\boldsymbol{\mathcal{D}}_k = \begin{bmatrix} H_k^{1,1} & H_k^{2,1} \\ H_k^{1,2} & H_k^{2,2} \\ \check{H}_k^{2,1} & -\check{H}_k^{1,1} \\ \check{H}_k^{2,2} & -\check{H}_k^{1,2} \end{bmatrix}. \tag{6}$$

It is worth noting that the properties of $\boldsymbol{\eta}$ remain the same with the conjugation and negation processes, therefore, the same notation is used. The precoded data described by the vector $\mathbf{q}_k$ can be recovered at the receiver by computing [4]

$$\mathbf{q}_k = \left[\boldsymbol{\mathcal{D}}_k^H\boldsymbol{\mathcal{D}}_k\right]_k^{-1} \boldsymbol{\mathcal{D}}_k^H\mathbf{v}_k \tag{7}$$

where $\mathbf{q}_k \in \mathbb{C}^{2\times 1}$. The decision variables $\mathbf{s}^1$ and $\mathbf{s}^2$ can be obtained as

$$\mathbf{s}^c = \mathbf{T}_N^{-1}\mathbf{d}^c \quad , c = [1,\, 2] \tag{8}$$

where $\mathbf{d}^c \in \mathbb{C}^{N\times 1}$.

When linear receivers are considered, precoded OFDM systems perform very well in frequency-selective channels due to the frequency diversity introduced by the precoding process. However, precoding does not provide any advantage in time varying channels. Consequently, combining precoding with STBC may suffer from severe BER degradation in fast fading channels. Alternatively, combing precoding with SFBC should offer a superior BER performance in time varying channels.

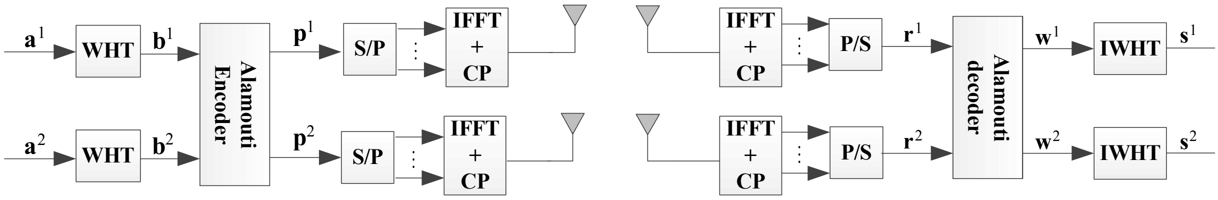

Fig. 1. Block diagram of the proposed P-SFBC system.

III. Precoded-SFBC System Model

Generally speaking, SFBC is more robust than STBC in time varying channels. Consequently, the same trend is expected in precoded SFBC and STBC systems as well. An SFBC-OFDM system that consists of two transmit and two receive antennas is considered. Consider two $N/2$ data sequences $\mathbf{a}^1$ and $\mathbf{a}^2$, each of which is applied to an $\frac{N}{2} \times \frac{N}{2}$ WHT. The data symbols in $\mathbf{a}^1$ and $\mathbf{a}^2$ are selected uniformly from a general QAM constellation. The WHT output vector is given by

$$\mathbf{b}^i = \mathbf{T}_{N/2}\, \mathbf{a}^i,\ i = [1,\, 2]\,. \tag{9}$$

The two $N/2$ data sequences $\mathbf{b}^i = \left[b_0^i,\, b_1^i,\, ...,\, b_{N/2-1}^i\right]^T$, $i = [1,\, 2]$, are applied to a standard Alamouti encoder where the precoded symbols are mapped as

$$\left[b_k^1\ b_k^2\right] \overset{STBC}{\rightarrow} \begin{bmatrix} b_k^1 & -\breve{b}_k^2 \\ b_k^2 & \breve{b}_k^1 \end{bmatrix},\quad k = 0,\, 1,\, ...,\, N/2-1. \tag{10}$$

Each row in (10) is buffered to form an N-symbols sequence,

$$\begin{aligned} \mathbf{p}^1 &= \left[b_0^1,\ -\breve{b}_0^2,\, b_1^1,\ -\breve{b}_1^2,\, ...,\, b_{N/2-1}^1,\ -\breve{b}_{N/2-1}^2\right] \\ \mathbf{p}^2 &= \left[b_0^2,\, \breve{b}_0^1,\, b_1^2,\, \breve{b}_1^1,\, ...,\, b_{N/2-1}^2,\, \breve{b}_{N/2-1}^1\right]. \end{aligned} \tag{11}$$

The sequences $\mathbf{p}^1$ and $\mathbf{p}^2$ are applied to a conventional OFDM modulator to generate the time domain samples. Using the same procedure and channel model used in the WHT-STBC system, the received signals at receiver c after discarding the CP and applying FFT process can be expressed as

$$\mathbf{r}^c = \mathbf{H}^{1,c}\mathbf{p}^1 + \mathbf{H}^{2,c}\mathbf{p}^2 + \boldsymbol{\eta}^c. \tag{12}$$

Since the matrices $\mathbf{H}^{i,c}$ are diagonal, each block of two adjacent elements in $\mathbf{r}^c$ have the form given by (12). Accordingly, the FFT output at positions k and $k{+}1$, $k = 0, 2, ..., N{-}2$ after substituting $p_k^1 = b_k^1$, $p_k^2 = b_k^2$, $p_{k+1}^1 = -\breve{b}_k^2$, and $p_{k+1}^2 = \breve{b}_k^1$ is given by

$$r_k^c \;=\; H_k^{1,c} b_k^1 + H_k^{2,c} b_k^2 + \eta_k^c \tag{13}$$

$$r_{k+1}^c \;=\; -H_{k+1}^{1,c} \breve{b}_k^2 + H_{k+1}^{2,c} \breve{b}_k^1 + \eta_{k+1}^c. \tag{14}$$

Using the assumption that the channel is constant over at least two adjacent subcarriers, i.e. $H_k^{1,c} = H_{k+1}^{1,c}$. The FFT output can be written as

$$\mathbf{v}_k = \boldsymbol{\mathcal{D}}_\mathbf{k} \mathbf{b}_k + \boldsymbol{\eta}_k \tag{15}$$

where $\mathbf{v}_k = \left[\ r_k^1\quad r_k^2\quad \breve{r}_{k+1}^1\quad \breve{r}_{k+1}^2\ \right]^T$, $\mathbf{b}_k = \begin{bmatrix} b_k^1 \\ b_k^2 \end{bmatrix}$ and

$$\boldsymbol{\mathcal{D}}_k = \begin{bmatrix} H_k^{1,1} & H_k^{2,1} \\ H_k^{1,2} & H_k^{2,2} \\ \breve{H}_k^{2,1} & -\breve{H}_k^{1,1} \\ \breve{H}_k^{2,2} & -\breve{H}_k^{1,2} \end{bmatrix}. \tag{16}$$

Then, the Alamouti decoder output can be calculated as

$$\mathbf{z}_k = \left[\boldsymbol{\mathcal{D}}_k^H \boldsymbol{\mathcal{D}}_k\right]^{-1} \boldsymbol{\mathcal{D}}_k^H \mathbf{v}_k. \tag{17}$$

where $\mathbf{z}_k = \left[\ b_k^1\quad b_k^2\ \right]^T$. Finally, the two $N/2$ symbols sequences $\mathbf{w}^c = \left[b_k^c,\, b_{k+1}^c,\, ...,\, b_{k+N/2-1}^c\right]^T$, $c = [1, 2]$, are applied to the IWHT. The decision variables $\mathbf{s}^1$ and $\mathbf{s}^2$ can be expressed as

$$\mathbf{s}^c = T_{N/2}\mathbf{w}^c,\ \ c = [1,\, 2]\,. \tag{18}$$

IV. Complexity Analysis

The computational complexity of the proposed P-SFBC and WHT-STBC systems [7] can be estimated and compared in terms of number of complex additions/subtractions (A_c) and complex multiplications (M_c) needed at the transmitter and receiver, respectively. The proposed system consists of two $\frac{N}{2} \times \frac{N}{2}$ WHT followed by two N- points IFFT operations at the transmitter as shown in Fig. 1. The $\frac{N}{2} \times \frac{N}{2}$ WHT applied to N symbols should be computed twice, each of which requires $\frac{N}{2}\log_2\frac{N}{2}$ complex additions. The WHT unitary matrix elements are either 1 or -1, which leads to the fact that WHT transform does not need any multiplication operations. Therefore, the total complexity for the $\frac{N}{2} \times \frac{N}{2}$ WHT is $N\log_2\frac{N}{2}$ complex additions. On the other hand, each N-point IFFT at the transmitter requires $\frac{N}{2}\log_2 N$ complex multiplications and $N\log_2 N$ complex additions. At the receiver side, the proposed system consists of two N-point FFT and two $\frac{N}{2} \times \frac{N}{2}$ IWHT with a complexity equal to the IFFT and WHT at the transmitter, respectively. The total number of complex additions and complex multiplications for the proposed system are $A_c = 8N\log_2 N - 4N$ and $M_c = 2N\log_2 N$, respectively.

The number of complex additions and multiplications of WHT-STBC system [7] consists of two N-point IFFT and two $N \times N$ WHT at the transmitter as well as two N-point FFT and $N \times N$ IWHT at the receiver. The $N \times N$ WHT applied to N symbols requires $N\log_2 N$ complex additions only. Thus, the total number of complex additions and complex multiplications are $A_c = 8N\log_2 N$ and $M_c = 2N\log_2 N$, respectively. Therefore, the proposed system and WHT-STBC system have identical number of complex multiplications, but the complex additions are less by $4N$.

TABLE I
SAMPLE DELAYS AND GAINS OF 6-TAP TYPICAL AND BAD URBAN CHANNEL MODELS.

	Typical Urban		Bad Urban	
Tap	Delay (sample)	Power (dB)	Delay (sample)	Power (dB)
1	0	-3	0	-2.5
2	0	0	1	0
3	1	-2	2	-3
4	3	-6	3	-5
5	5	-8	10	-2
6	10	-10	13	-4

V. Numerical Results

The proposed system model is evaluated using LTE downlink physical layer parameters [2]. The LTE downlink system is based on orthogonal frequency division multiple access (OFDMA), which has similar characteristics to OFDM. Single user OFDMA is considered in this work with $N = 128$ subcarriers, which corresponds to the data part of the OFDM symbol with a duration of $T_{\mathrm{u}} = 66.7\mu s$ and subcarrier spacing of 15 kHz. The number of CP samples is $P = 32$, with a duration of $T_{\mathrm{cp}} = 16.67\mu s$. The data symbols are drawn from Quadrature Phase Shift Keying (QPSK) constellation and the total OFDM symbol period is $T_{\mathrm{t}} = 83.37\mu s$. To cover a wide range of operating scenarios, two channel models are considered with different Doppler spreads, namely, the 6-tap typical (TU6) and bad urban (BU6) [15]. The TU6 and BU6 are considered as mild and moderate frequency-selective fading channels, respectively. The normalized time delays and attenuations of the considered channels are given in Table 1. The root mean-squared delay spreads $\sigma(\tau)$ for the considered channels are $\sigma(\tau) = 1.4309\ \mu s$ and $9.9097\ \mu s$, respectively. The multipath components' gains are generated as complex independent Gaussian random variables and the fading is generated using the Jakes' Doppler spectrum model. The BER of the proposed system is compared to conventional STBC, SFBC, WHT-STBC [7] and CMS systems [13]. The results are presented in terms of the normalized maximum Doppler shift $F_{\mathrm{d}} = f_{\mathrm{d}} T_{\mathrm{u}}$. The channel estimation is performed as described in [13], i.e., only the diagonal elements of the channel elements are known perfectly while all other off-diagonal elements are assumed to be zeros. The uncoded BER is the key performance indicator used for comparisons with other systems.

The BER of the considered systems in static ($F_{\mathrm{d}} = 0$) TU6 and BU6 channels is shown in Fig. 2. As it can be noted from the figure, the precoded STBC-OFDM [7] offers the lowest BER due to the composite frequency and space diversity. Moreover, the static channel perfectly satisfies the condition for the linear receiver to deliver the minimum BER. The STBC provided only spacial diversity and hence its BER is worse than [7]. It is worth noting that the precoded STBC and conventional STBC are independent of the channel frequency selectivity, and hence, they offer the same BER under the TU6 and BU6 channels. The P-SFBC experienced some BER degradation as compared to [7] because linear receivers are not optimum for SFBC configuration in frequency-selective channels. It is worth noting that CMS and SFBC offer equivalent performance because CMS is not optimized for QPSK modulation and the channel selectivity is not severe.

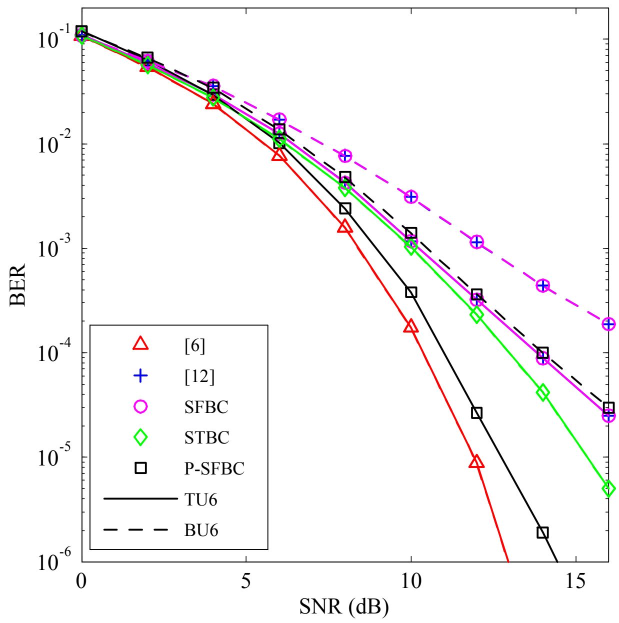

Fig. 2. BER performance of the proposed P-SFBC, SFBC, STBC and CMS systems over TU6 and BU6 channels where $F_{\mathrm{d}} = 0$.

Fig. 3 shows the BER of SISO, WHT-STBC and P-SFBC over time-varying TU6 channel using Doppler spread f_d of 750 Hz and 1500 Hz, which corresponds to normalized Doppler values $F_{\mathrm{d}} = 0.05$ and 0.1, respectively. Assuming a carrier frequency of 5 GHz, the considered speeds are about 160 and 320 km/hr. As it can be noted from the figure, the BER of [7] substantially increases as a function of F_{d}, while the proposed P-SFBC demonstrates high robustness against the channel time variations compared to the WHT-STBC system. The BER of the P-SFBC system shows negligible BER degradation at low Doppler values ($F_{\mathrm{d}} = 0.05$) while the degradation is slightly higher at high Doppler value ($F_{\mathrm{d}} = 0.1$). Such degradation is a result of the inter carrier interference (ICI) introduced by the Doppler spread. Unlike the P-SFBC, the severe BER degradation of the WHT-STBC system in time varying channels is caused by the ICI and the poor performance of linear receivers in such scenarios. It is worth noting that the BER degradation of the P-SFBC is significantly less than the BER degradation of the WHT-STBC, where the P-SFBC outperforms the WHT-STBC system by more than of 4 dB at BER of 10^{-4}. The figure also presents the BER of SISO systems for comparison purposes.

The BER of the P-SFBC is further compared with the performance of the CMS, SFBC and STBC systems over time-varying TU6 channel as shown in Fig. 4. It can be observed that the P-SFBC system outperforms all the other considered systems.

The BER of the considered systems is evaluated over time varying TU6 and BU6 channels for $F_{\mathrm{d}} = 0$ as shown in Fig. 5.

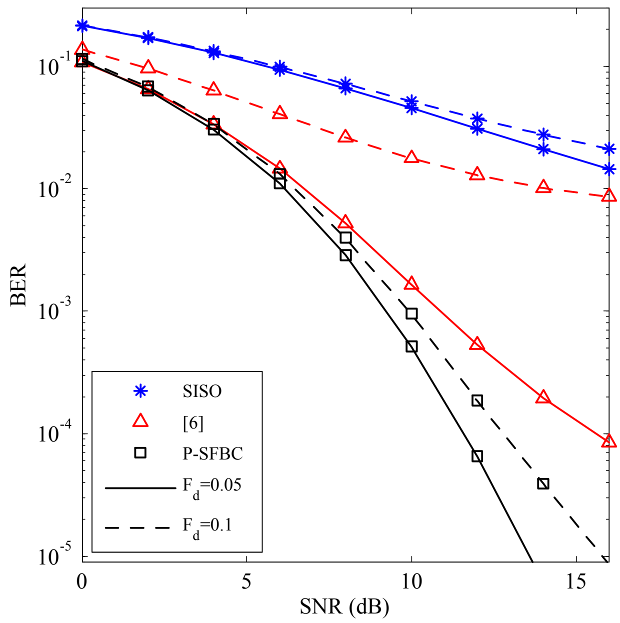

Fig. 3. BER performance of SISO, P-SFBC and WHT-STBC systems over TU6 channel using $F_{\mathrm{d}} = 0.05$ and 0.1.

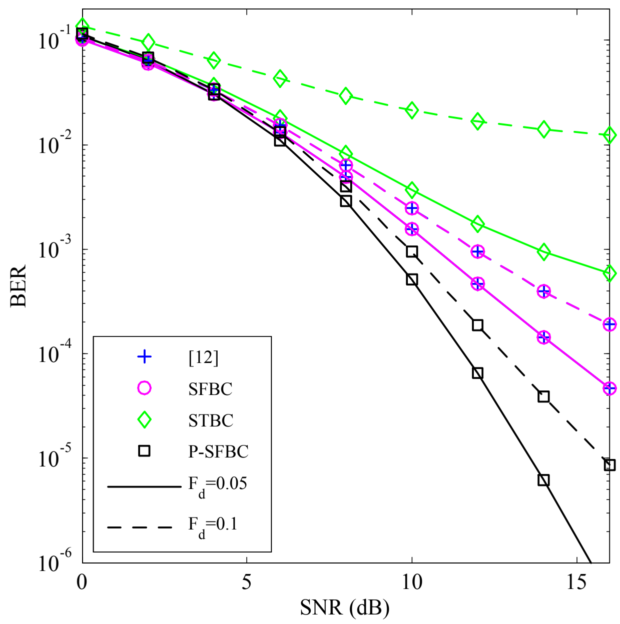

Fig. 4. BER performance of the proposed P-SFBC, SFBC, STBC, CMS and WHT-STBC systems over TU6 channel using $F_{\mathrm{d}} = 0.05$ and 0.1.

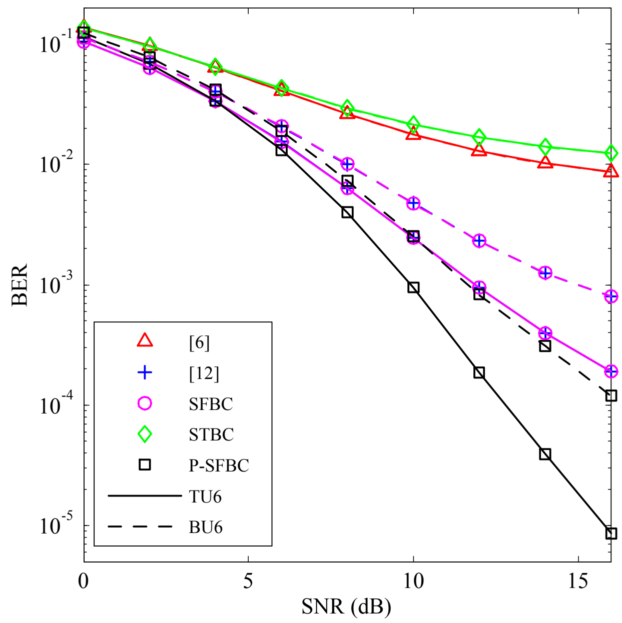

Fig. 5. BER of the considered systems over TU6 and BU6 channel for $F_{\mathrm{d}} = 0.1$.

It can be noted that the BER of the proposed P-SFBC, SFBC, and CMS systems degrades with the frequency selectivity of the channel. However, the P-SFBC system still outperforms all the other considered systems. The figure also confirms the independency of STBC and WHT-STBC systems from the frequency selectivity of the channel.

Fig. 6 presents the BER versus F_{d} for the P-SFBC, CMS, STBC, SFBC and WHT-STBC systems over TU6 channel where $SNR = 10\ dB$. It can be noted that the P-SFBC system outperforms all other considered systems for $F_{\mathrm{d}} \gtrsim 0.023$. For F_{d} values less than 0.023, the WHT-STBC outperforms the other systems due to the frequency diversity achieved by the spreading of the WHT matrix. However, its BER increases aggressively for $F_{\mathrm{d}} \gtrsim 0.023$ compared to the BER of P-SFBC and SFBC systems, which degrades slightly by increasing F_{d}. Moreover, the SFBC outperforms STBC and WHT-STBC systems at $F_{\mathrm{d}} \gtrsim 0.02$ and $F_{\mathrm{d}} \gtrsim 0.049$, respectively. It is worth noting that both STBC and WHT-STBC systems show similar BER at $F_{\mathrm{d}} \gtrsim 0.1$.

VI. Conclusion

In this paper, a novel precoded SFBC system was proposed, in order to overcome the adverse effects of the channel mobility in SFBC systems. The BER of all considered systems was evaluated over TU6 and BU6 channel models with various F_{d} values. The simulation results revealed that the system is robust to the time and frequency selectivity of the channel, and hence it renders itself as an efficient alternative for the conventional and state-of-the-art systems such as the WHT-STBC. Although the WHT-STBC system outperformed the P-SFBC system at static and low mobility conditions, both systems offered low BER values because the ICI is negligible

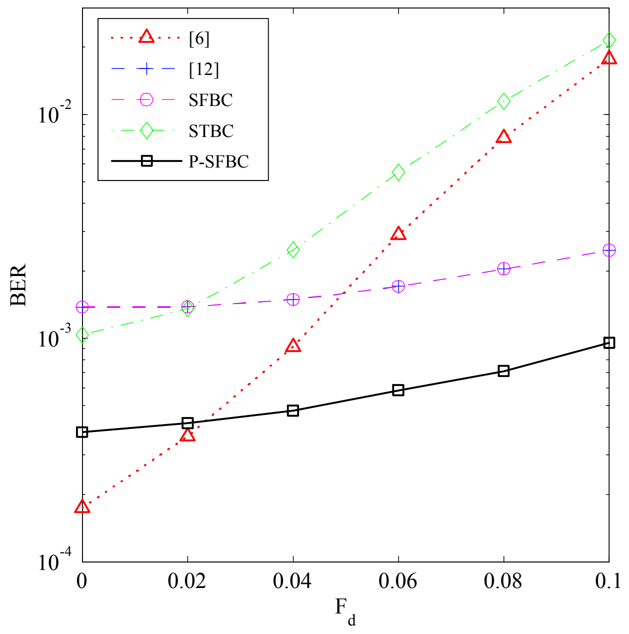

Fig. 6. BER versus the normalized Doppler spread at $SNR = 10\ dB$.

at such Doppler values. The computational complexity of the proposed system was evaluated in terms of complex additions and multiplications to be compared to the complexity of the 2×2 version of WHT-STBC system [7]. The proposed system introduced identical number of complex multiplications and lower number of complex additions compared to the WHT-STBC system due to the use of $\frac{N}{2} \times \frac{N}{2}$ WHT.

References

[1] E. Al-Dalakta, A. Al-Dweik, A. Hazmi, C. Tsimenidis and B. Sharif, “PAPR Reduction using Maximum Cross Correlation,” *IEEE Commun. Lett.*, vol. 16, no.12, pp. 2032-2035, Dec. 2012.

[2] Physical Layer Aspects, Evolved Universal Terrestrial Radio Access (E-UTeA), Third-Generation Partnership Project TS 25.814, 2006-09.

[3] M. Uysal, N. Al-Dhahir and C. Georghiades, "A space-time block-coded OFDM scheme for unknown frequency-selective fading channels," *IEEE Commun. Lett.*, vol. 5, no. 10, pp. 393-395, Oct. 2001.

[4] S. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE J. Sel. Areas Commun.*, vol. 16, pp. 1451-1458, Oct. 1998.

[5] S. Wang, S. Zhu, and G. Zhang, “A Walsh–Hadamard coded spectral efficient full frequency diversity OFDM system,” *IEEE Trans. Commun.*, vol. 58, no. 1, pp. 28–34, Jan. 2010.

[6] M. Ahmed, S. Boussakta, B. Sharif and C. Tsimenidis, "OFDM based on low complexity transform to increase multipath resilience and reduce PAPR," *IEEE Trans. Signal Process*, vol. 59, no. 12, pp. 5994-6007, Dec. 2011.

[7] Y. Chung and S. Phoong, “Unitary precoders for ST-OFDM systems using Alamouti STBC,” *IEEE Trans. Circuits Syst. I*, vol. 55, pp. 2860-2869, Oct. 2008.

[8] S. Boussakta and A. G. J. Holt, ”Fast algorithm for calculation of both Walsh Hadamard and Fourier transforms (FWFTs),” *Electron. Lett.*, vol. 25, pp. 1352-1354, 1989.

[9] C. Li, S. Wang and K. Chan, "Low complexity transmitter architectures for SFBC MIMO-OFDM systems," *IEEE Trans. Commun.*, vol. 60, no. 6, pp. 1712-1718, June 2012.

[10] A. Omri, R. Hamila, A. Hazmi, R. Bouallegue and A. Al-Dweik, “Enhanced Alamouti decoding scheme for DVB-T2 systems in SFN channels,” *in Proc. IEEE PIMRC*, Toronto, Canada, pp. 2046-2050, Sep. 2011.

[11] L. C. Tran, X. Huang, E. Dutkiewicz, and J. Chicharo, “STC-MIMO block spread OFDM in frequency selective rayleigh fading channels,” *in Proc. IEEE Symp. Computers Commun.*, 2006, pp. 277–280, 2006.

[12] K. Lee, Y. Kim, and J. Kang, “A novel orthogonal space–time–frequency block code for OFDM systems,” *IEEE Commun. Lett.*, vol. 13, no. 9, pp. 652–654, Sep. 2009.

[13] A. Al-Dweik, F. Kalbat, S. Muhaidat, O. Filio, and S. M. Ali, “Robust MIMO-OFDM system for frequency-selective mobile wireless channels,” *IEEE Trans. Veh. Technol.*,vol. 64, no. 5, pp. 1739-1749, May 2015.

[14] W. Yan, Z. Lei, and S. Sun, “Performance of Wash-Hadamard transformed STBC OFDM system,” *in Proc. IEEE Veh. Technol. Conf.*, vol. 2, pp. 738–741, May 2004.

[15] K. Fazel and S. Kaiser, *Multi-carrier and spread spectrum systems*, Second Edition, WILEY, 2008.