Enhanced Alamouti Space-Time Block-Coding Transmission Based on a Developed OFDM System

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***Abstract*—This paper presents a new Alamouti space-time-block coding (STBC) orthogonal frequency division multiplexing (ST-OFDM) system. The transmitter of the proposed scheme is based on a low complexity X-transform as a modulation scheme instead of the inverse fast Fourier transform (IFFT). The proposed ST-X-OFDM scheme massively reduces the transmitter complexity, the effects of multipath channels, and the peak-to-average power ratio (PAPR). The complexity of the proposed scheme is evaluated and compared with the conventional ST-OFDM, showing that the proposed scheme is efficient, fast, and simple in implementation. An exact closed-form bit-error rate (BER) is derived for the proposed scheme over multipath channels and different kinds of modulation formats. Computer simulations corroborated by mathematical results prove that, over frequency-selective multipath channels and different mapping schemes, the proposed approach achieves significant signal-to-noise ratio (SNR) gain in comparison to the ST-OFDM system. The proposed ST-X-OFDM system adopts only a single, low complexity transform at the transmitter without precoding and achieves significant BER improvement over the conventional ST-OFDM system.**

***Keywords*—MIMO, OFDM, complexity, error probability, peak-to-average power ratio, multi-path fading channels.**

I. Introduction

Future generations of wireless communication systems are required to support high-data-rate applications [1]. The multiple-input multiple-output (MIMO) technique has shown a capability of achieving extraordinary throughput without extra power consumption or bandwidth expansion [2], [3].

MIMO systems rely on antenna diversity to mitigate the effect of multipath fading. MIMO systems that adopting Alamouti space-time block-coding (STBC) scheme has been shown to be an effective technique for wireless communications [4]. This is because of the attractive characteristics of the Alamouti STBC scheme for orthogonal codes, as it requires only simple linear processing at the receiver to recover the data symbols. To combat multipath delay spread in high-rate wireless systems, one efficient method is orthogonal frequency-division multiplexing (OFDM), which transforms a multipath channel to parallel independent flat-fading subchannels.

In wireless communications, the STBC technique and the OFDM have been embedded in a single modulation scheme called ST-OFDM [5]-[6]. The produced ST-OFDM approach was efficient in mitigating the effect of the ISI and enhancing the diversity of transmitted OFDM signal [7]-[8]. A further improvement in the ST-OFDM system has attracted the attention of many researchers in the last few years. Relating to our work, the approach in [9] demonstrated the advantages of using a unitary precoder in the ST-OFDM (UP-ST-OFDM), where it improved the transmission when the channel state information (CSI) is unknown at the transmitter. However, it is considerably complex as it requires at least two complex precoders as hardware, in addition to two inverse fast Fourier transforms (IFFTs) which perform the modulation scheme. More recently, the authors in [10] demonstrated the advantages of adopting a discrete Hartley transform (DHT) as a unitary precoder in the conventional OFDM system for peak-to-average power ratio (PAPR) reduction.

In this paper, we present a new efficient Alamouti ST-OFDM system with a simple structure and very low complexity transmitter. The transmitter of the proposed scheme utilizes a very low complexity **X**-transform instead of the conventional IFFT. The **X**-transform gathers the effects of the DHT and the FFT in a single, orthogonal, and very low complexity transform. Furthermore, a precise mathematical BER performance for the proposed ST-X-OFDM scheme is derived in this work for different kinds of transmission scenarios and over multipath channels. The proposed ST-X-OFDM system has the same single-tap equalizer complexity as the ST-OFDM, with significant reduction in transmitter complexity. The proposed ST-X-OFDM system exploits the channel diversity and achieves significant signal-to-noise ratio (SNR) gain over the conventional ST-OFDM system. From the PAPR point of view, the proposed scheme significantly reduces the PAPR of the ST-X-OFDM signal in comparison with the conventional ST-OFDM system.

II. Proposed System Analysis

The proposed ST-X-OFDM system is shown in Fig. 1 where the Alamouti STBC scheme with two transmit antennas and M_R receive antennas is used. It is seen from Fig. 1 that the **X**-transform is used in the transmitter as an OFDM modulation scheme instead of the conventional inverse FFT. The binary data is first mapped into complex modulated symbols such as M-array quadrature phase shift keying (M-PSK) and M-array quadrature-amplitude-modulation (M-QAM) constellations. These data symbols, $\left\{S_n^{(b)}\right\}_{n=0}^{N-1}$, are then organised in blocks of size N at each transmitter branch, $(b = 1, 2)$, for modulation. Each information symbol is an $N \times 1$ information

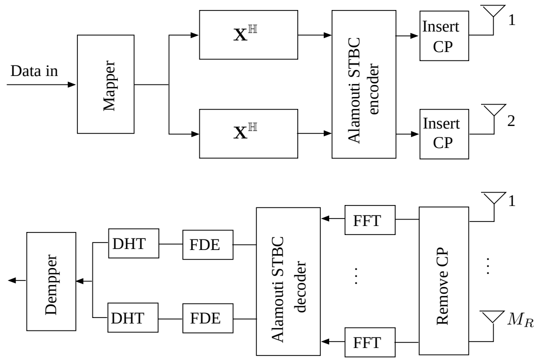


Fig. 1. Transmitter and receiver block diagrams of the proposed scheme.

vector. Each vector is then processed by the DHT as follows:

$$r_n^{(b)} = \sum_{m=0}^{N-1} a_{n,m} S_m^{(b)}, \qquad (b = 1, 2), \tag{1}$$

where $\mathbf{A}_n$ is the n_{th} row of the DHT matrix, $\mathbf{A}$, and $a_{n,m}$ denotes the n_{th} row and m_{th} column of the matrix $\mathbf{A}$. Alamouti STBC encoder is used to encode $\mathbf{r}^{(1)}$ and $\mathbf{r}^{(2)}$ as:

$$\begin{bmatrix} \mathbf{r}^{(1)} \\ \mathbf{r}^{(2)} \end{bmatrix}, \quad (2) \qquad \begin{bmatrix} -\mathbf{r}^{(2)^*} \\ \mathbf{r}^{(1)^*} \end{bmatrix}, \quad (3)$$

where $(.)^*$ denotes the complex conjugate operation. The resulting vectors is then processed by the IFFT as:

$$\mathbf{s}^{(b)} = \mathbf{F}^{\mathbb{H}} \mathbf{r}^{(b)} \tag{4a}$$

$$\mathbf{s}^{(b)^\dagger} = \mathbf{F}^{\mathbb{H}} \mathbf{r}^{(b)^*}, \tag{4b}$$

where $\mathbf{s}^{(b)^\dagger} = \left[s_0^{(b)^*}, s_{N-1}^{(b)^*}, \cdots, s_2^{(b)^*}, s_1^{(b)^*} \right]^T$, $\mathbf{F}$ is the normalized FFT matrix, and $(.)^{\mathbb{H}}$ is the Hermitian operation. In order to reduce the transmitter complexity by combining the DHT and the IFFT transforms into single low complexity unitary transform, the Alamouti STBC encoder is performed after the IFFT as:

$$\mathbf{s}^{(b)} = \mathbf{F}^{\mathbb{H}} \mathbf{A} \mathbf{S}^{(b)}, \tag{5}$$

and then the Alamouti STBC encoder is performed on the vectors $\mathbf{s}^{(b)}$, $(b = 1, 2)$, as:

$$\begin{bmatrix} \mathbf{s}^{(1)} \\ \mathbf{s}^{(2)} \end{bmatrix}, \quad (6) \qquad \begin{bmatrix} -\mathbf{s}^{(2)^\dagger} \\ \mathbf{s}^{(1)^\dagger} \end{bmatrix}, \quad (7)$$

A. *X-Transmitter Based Transmitter*

The complexity of the DHT in addition to that of the IFFT is very high. This is especially considerable when the information symbols are drawn from a complex constellation where, as the DHT is real transform, the DHT process needs to be performed twice one for real and the other for imaginary part of the complex symbols. However, the transmitter complexity can be reduced by merging the DHT and the IFFT in a single, fast, and efficient transform as follows:

$$s_k^{(b)} = \frac{1}{N} \sum_{n=0}^{N-1} \sum_{m=0}^{N-1} S_m^{(b)} \left[\cos\left(\frac{2\pi nm}{N}\right) + \sin\left(\frac{2\pi nm}{N}\right) \right] e^{j\frac{2\pi nk}{N}}. \tag{8}$$

From the trigonometric identities, (8) can be written as

$$s_k^{(b)} = \frac{1}{2N} \left(\sum_{m=0}^{N-1} S_m^{(b)} \sum_{n=0}^{N-1} \left[\cos\left(\frac{2\pi n}{N}(m-k)\right) + \cos\left(\frac{2\pi n}{N}(m+k)\right) \right] + j \left[\cos\left(\frac{2\pi n}{N}(m-k)\right) - \cos\left(\frac{2\pi n}{N}(m+k)\right) \right] \right). \tag{9}$$

In more expressive form, (9) can be rewritten as

$$s_k^{(b)} = \sum_{m=0}^{N-1} S_m^{(b)} X_{km}^{\mathbb{H}}, \quad (k = 0, 1, 2, ..., N-1), \tag{10}$$

where $X_{km}^{\mathbb{H}}$ is the k_{th}, $(0 \leq k \leq N-1)$, row and the m_{th}, $(0 \leq m \leq N-1)$ column element of the $\mathbf{X}^{\mathbb{H}}$ (inverse $\mathbf{X}$ transform (IXT)).

From (9) one can notice the following; $X_{k,m}^{\mathbb{H}} = 1$ when $k = m = 1$ or $k = m = \frac{N}{2}$, $X_{k,m}^{\mathbb{H}} = \frac{1}{2} + j\frac{1}{2}$ when $k = m$ and $X_{k,m}^{\mathbb{H}} = \frac{1}{2} - j\frac{1}{2}$ when $k = N - m$ and $X_{km}^{\mathbb{H}} = 0$ elsewhere. Thus, (5) can be written as

$$\mathbf{s}^{(b)} = \mathbf{X}^{\mathbb{H}} \mathbf{S}^{(b)}. \tag{11}$$

The $\mathbf{X}$-transform of size N includes $\frac{N}{2}$ butterflies, each butterfly is shown in Fig. 2.

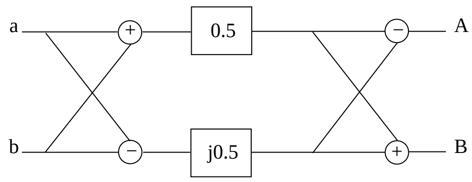


Fig. 2. Basic butterfly (unit) of **X**-transform.

It is clear from Fig. 2 that the **X**-transform involves very few additions and no multiplication at all.

B. *Transmission over Frequency-Selective Multipath Channels*

To alleviate the ISI, a cyclic prefix (CP) of length must be no less than the maximum delay of the multipath channel is appended to each OFDM block of length N before the transmission. At a given OFDM symbol period T_s, the two signals $\mathbf{s}^{(1)}$ and $\mathbf{s}^{(2)}$ are simultaneously transmitted from antenna 1 and antenna 2, respectively. However, during the next symbol time, $-\mathbf{s}^{(2)^\dagger}$ and $\mathbf{s}^{(1)^\dagger}$ are simultaneously transmitted from antennas 1 and 2, respectively. The transmitted signal from the b_{th} antenna passes through a multipath channel ($h_l^{(b)} \neq 0, \forall \;\; 0 < l < L$) of $L + 1$ taps and is corrupted by additive white Gaussian noise (AWGN) before it arrives at the receiver. The channels $h_l^{(1)}$ and $h_l^{(2)}$ are assumed, but not required, to have the same number of taps. Assuming that the channels are statistically independent and invariant during two consecutive OFDM symbols. Therefore, at the m_rth receiver antenna, the consecutive received signals at time t and $t+T_s$ can be written as:

$$y_k^{(m_r,1)} = \sum_{l=0}^{L} h_l^{(1,m_r)} s_{k-l}^{(1)} + \sum_{l=0}^{L} h_l^{(2,m_r)} s_{k-l}^{(2)} + w_k^{(m_r,1)}, \quad (12)$$

$$y_k^{(m_r,2)} = -\sum_{l=0}^{L} h_l^{(1,m_r)} {s_{k-l}^{(2)}}^{\dagger} + \sum_{l=0}^{L} h_l^{(2,m_r)} {s_{k-l}^{(1)}}^{\dagger} + w_k^{(m_r,2)}, \quad (13)$$

where w_k is the k_{th} sample of the vector $\mathbf{w} = [w_0, w_1, \cdots, w_{N-1}]^T$, which represents the AWGN samples. The AWGN samples are independent normally distributed random variables with zero mean and variance $\sigma_n^2 = E\left\{|w_k|^2\right\}$, and $E\{.\}$ denotes the expectation operation. The received signal y_k is then processed by the FFT transform after discarding the CP, hence, the produced signals can be written as:

$$Y_n^{(m_r,1)} = r_n^{(1)} H_n^{(1,m_r)} + r_n^{(2)} H_n^{(2,m_r)} + \Omega_n^{(m_r,1)}, \quad (14)$$

$$Y_n^{(m_r,2)^*} = r_n^{(1)} H_n^{(2,m_r)^*} - r_n^{(2)} H_n^{(1,m_r)^*} + \Omega_n^{(m_r,2)^*}, \quad (15)$$

where Ω_n is the frequency domain representation of the AWGN (w_n) and $H_n = \sum_{l=0}^{L} h_l e^{-j\frac{2\pi nl}{N}}$, $(0 \le n \le N-1)$ is the channel transfer function at the n_{th} index. To generalise (14) and (15) to all $n = 0, 1, ..., N-1$, we use the following notations: $\mathbf{h}^{(1)} = \text{diag}(H_n^{(1,m_r)})_{n=0}^{N-1}$, $\mathbf{h}^{(2)} = \text{diag}(H_n^{(2,m_r)})_{n=0}^{N-1}$, $\mathbf{Y}^{(1)} = (Y_n^{(1)})_{n=0}^{N-1}$, $\mathbf{Y}^{(2)} = (Y_n^{(2)})_{n=0}^{N-1}$, $\mathbf{\Omega}^{(1,m_r)} = (\Omega_n^{(1,m_r)})_{n=0}^{N-1}$. Therefore, (14) and (15) can be written as

$$\mathbf{Y}^{(1,m_r)} = \mathbf{h}^{(1,m_r)}\mathbf{r}^{(1)} + \mathbf{h}^{(2,m_r)}\mathbf{r}^{(2)} + \mathbf{\Omega}^{(1,m_r)}, \quad (16a)$$

$$\mathbf{Y}^{(2,m_r)^*} = \mathbf{h}^{(2,m_r)^*}\mathbf{r}^{(1)} - \mathbf{h}^{(1,m_r)^*}\mathbf{r}^{(2)} + \mathbf{\Omega}^{(2,m_r)^*} \quad (16b)$$

In matrix form, (16b) can be written as

$$\mathbf{Y} = \mathbf{H}\mathbf{r} + \mathbf{\Omega}, \quad (17)$$

where $\boldsymbol{Y} = \left[\mathbf{Y}^{(1,1)}, \mathbf{Y}^{(1,2)^*}, \cdots, \mathbf{Y}^{(M_R,1)}, \mathbf{Y}^{(M_R,2)^*}\right]^T$, $\mathbf{r} = \left[\mathbf{r}^{(1)}, \mathbf{r}^{(2)}\right]^T$ and $\mathbf{\Omega} = \left[\mathbf{\Omega}^{(1,1)}, \mathbf{\Omega}^{(1,2)^*}, \cdots, \mathbf{\Omega}^{(M_R,1)}, \mathbf{\Omega}^{(M_R,2)^*}\right]^T$, and the channel matrix is given as

$$\mathbf{H} = \begin{bmatrix} \mathbf{h}^{(1,1)} & \mathbf{h}^{(2,1)} \\ \mathbf{h}^{(2,1)^*} & -\mathbf{h}^{(1,1)^*} \\ \vdots & \vdots \\ \mathbf{h}^{(1,m_r)} & \mathbf{h}^{(2,m_r)} \\ \mathbf{h}^{(2,m_r)^*} & -\mathbf{h}^{(1,1)^*} \end{bmatrix}. \quad (18)$$

At the receiver side, the frequency domain signal, $\mathbf{Y}$, is then multiplied by the Hermitian conjugate of the channel matrix $\mathbf{H}^{\mathbb{H}}$ to simplify the equalization, as follows

$$\tilde{\mathbf{Y}} = \mathbf{H}^{\mathbb{H}}\mathbf{Y}, \quad (19)$$

which can be rewritten as

$$\tilde{\mathbf{Y}}^{(b)} = \left(\sum_{t=1}^{2}\sum_{m_r=1}^{M_R} |\mathbf{h}^{(t,m_r)}|^2\right)\mathbf{r}^b + \overline{\mathbf{\Omega}}^{(b)}, \quad (20)$$

where $\overline{\mathbf{\Omega}}^{(b)} = \mathbf{H}_b^{\mathbb{H}}\mathbf{\Omega}, b = 1, 2$ and $\mathbf{H}_b^{\mathbb{H}}$ is the b_{th} row of the matrix $\mathbf{H}^{\mathbb{H}}$. Let $\Sigma_n = \left(\sum_{t=1}^{2}\sum_{m_r=1}^{M_R} |H_n^{(t,m_r)}|^2\right)$ The channel equalization is then performed in frequency domain, after the FFT.

The compensation for the channel effects on the received signal can be achieved by the zero-forcing (ZF) equalizer by simply dividing each individual sample of the received vector, $\tilde{Y}_n^{(b,m)}$ in (20) by the corresponding value of the channel transfer function (Σ_n) as:

$$\hat{r}_n^{(b)} = \frac{\tilde{Y}_n^{(1)}}{\Sigma_n} = r_n^{(b)} + \xi_n^{(b)}, \quad (21)$$

where $\xi_n^{(b)} = \frac{\overline{\Omega_n}^{(b)}}{\Sigma_n}$ represent the amplified AWGN noise part from. Substituting (1) into (21) yields

$$\hat{r}_n^{(b)} = \sum_{m=0}^{N-1} a_{n,m} S_m^{(b)} + \xi_n^{(b)}. \quad (22)$$

After removing the channel gain, the equalized signal, $\hat{r}_n^{(b)}$, is then transformed by the matrix $\mathbf{A}$ as follows

$$q_i^{(b)} = \sum_{n=0}^{N-1} a_{i,n}\hat{r}_n^{(b)} \qquad (i = 0, 1, 2, ..., N-1). \quad (23)$$

Substituting (22) into (23) yields

$$q_i^{(b)} = \sum_{n=0}^{N-1} a_{i,n}\left(\sum_{m=0}^{N-1} a_{n,m}S_m^{(b)}\right) + \sum_{n=0}^{N-1} a_{i,n}\xi_n^{(b)}. \quad (24)$$

Since the DHT is an orthogonal transform, $a_{i,n} \times a_{n,m}$ equal to 1 only when $m = i$ and zero elsewhere, the first term of (24) equals to S_i, hence, (24) can be written as

$$q_i^{(b)} = S_i^{(b)} + \hat{\xi}_i^{(b)}, \qquad (b = 1, 2), \quad (25)$$

where $\hat{\xi}_i^{(b)} = \sum_{n=0}^{N-1} a_{i,n}\xi_n^{(b)}$. The difference between the transmitted and received signals, $e_i^{(b)} = q_i^{(b)} - S_i^{(b)}$, is the error signal, it can be written as

$$e_i^{(b)} = \sum_{n=0}^{N-1} a_{i,n}\xi_n^{(b)}, \qquad (b = 1, 2). \quad (26)$$

The noise power, $\mathcal{P}_{n_i} = E[|e_i^{(b)}|^2]$, is given as:

$$\mathcal{P}_{n_i} = \sigma_v^2 \sum_{n=0}^{N-1} |a_{i,n}|^2 \frac{1}{\Sigma_n}. \quad (27)$$

Thus, the signal-to-noise ratio, $SNR_i = \frac{E[|S_i|^2]}{E[|e_i|^2]} = \frac{E_s}{\mathcal{P}_{n_i}}$, where E_s is the useful symbol power, will be given as

$$SNR_i = \frac{\gamma_s}{\sum_{n=0}^{N-1} |a_{i,n}|^2 \frac{1}{\Sigma_n}}, \quad (28)$$

where $\gamma_s = \frac{E_s}{\sigma_v^2}$ is the signal power (in terms of symbol) to noise ratio. The BER of the individual subchannels are respectively given as in [11] as follows:

$$P_e^{M-PSK} = \frac{\mu}{mN} \sum_{i=0}^{N-1} Q\left(\sqrt{2SNR_i}\sin(\frac{\pi}{M})\right), \quad (29)$$

TABLE I
TRANSMITTER COMPLEXITY OF THE PROPOSED ST-X-OFDM AND THE CONVENTIONAL ST-OFDM.

N	ST-X-OFDM R_{O}	ST-OFDM R_{O}
64	496	3840
128	1008	8960
256	2032	20480
512	4080	46080
1024	8176	102400

$$P_e^{M-QAM} = \frac{4 - 2^{(2-\frac{m}{2})}}{mN} \sum_{i=0}^{N-1} Q\left(\sqrt{\frac{3SNR_i}{M-1}}\right). \quad (30)$$

In (29) and (30), μ denotes the number of nearest neighbours signal points, M is the level on constellation and $m = \log_2 M$ represents the number of bits in each data symbol, $Q(x)$ denotes the Q-function of x and SNR_m is the signal power-to-noise ratio which is given in (28). The exact BER calculation of the proposed scheme is obtained for M-PSK and M-QAM by substituting the specific parameters that assigned to M-PSK and M-QAM and substituting (28) into (29) and (30).

III. COMPLEXITY CALCULATIONS

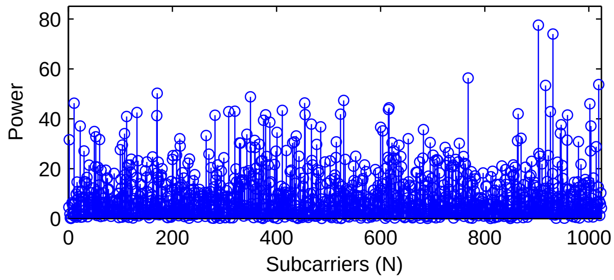
In this section, the arithmetic operations of the proposed ST-X-OFDM system are calculated and compared with the conventional ST-OFDM, based on a single butterfly algorithm.

A. *X-Transform*

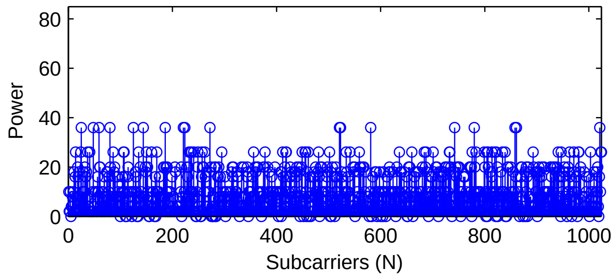
The number of butterflies in the structure of **X**-transform is $\frac{1}{2}(N-2)$, each butterfly is shown in Fig. 2. It can be observed from Fig. 2 that each butterfly, ignoring the scaling factor, involves 4 complex additions (C_{A}) that gives 8 real operations, R_{A}. Then the overall complexity is $R_{\mathrm{A}} = 4(N-2)$. As shown in Fig.1, twice the complexity of **X**-transform is required at the transmitter of the proposed ST-X-OFDM system, hence, the overall complexity is equal to $8(N-2)$ real additions.

B. *ST-OFDM*

There are a lot of different algorithms to implement the FFT. In this paper, we evaluate the complexity of the FFT based on single butterfly algorithm and 4/2 implementation where each complex multiplication involves 4 real multiplications (R_{M}) and two real additions (R_{A}) and each complex addition is equivalent to two real additions. The arithmetic complexity of the FFT is given as: $R_{\mathrm{M}} = 2N\log_2 N$ and $R_{\mathrm{A}} = 3N\log_2 N$. For ST-OFDM system, two FFTs are required at the transmitter, Hence the total transmitter complexity of the ST-OFDM system is $R_{\mathrm{O}} = 10N\log_2 N$. It is obvious from table I that the proposed scheme has enormously reduces the transmitter complexity. The complexity at the receiver of the conventional ST-OFDM is fewer than the complexity at the receiver of the proposed system, however, the price is the significant BER performance improvement and huge reduction in the transmitter complexity and the PAPR of the transmitted signal when the proposed scheme is used.



(a) Power in OFDM signal.



(b) Power in X-OFDM signal.

Fig. 3. Power in OFDM signal (a) the conventional system and (b) the proposed system.

IV. PEAK-TO-AVERAGE POWER RATIO (PAPR)

The high peak-value of the OFDM signal is considered a major problem in the OFDM systems [12]. The **X**-transform reserves the average power of the signal that is transformed by the **X**-transform exactly as the DFT. However, it can enormously reduce the peak power of the OFDM signal as it hugely reduces the number of additions of data symbols that performs each OFDM sample from N in the case of conventional OFDM to only two in the case of the proposed ST-X-OFDM scheme. This is clearly shown in Fig. 3, where it can be observed that for the case of the system that based on the conventional DFT, the instantaneous peak power can approach more than twice that of the proposed ST-X-OFDM scheme, as shown in Figs 3(a) and 3(b). In other words, the **X**-transform can significantly reduce the PAPR of the OFDM system naturally, without using any PAPR reduction technique that is indispensable in the conventional OFDM system.

V. SIMULATION RESULTS AND DISCUSSION

The mapping schemes that were used in our simulation are QPSK and 16-QAM modulations, with the total number of modulated symbols being 2048; hence, 1024 symbols are assigned to each antenna and ZF equalizer is used. International telecommunication union (ITU) pedestrian B channel is used.

Figs. 4 and 5 show the theoretical and simulation BER performance of the proposed ST-X-OFDM and the conventional ST-OFDM systems for the QPSK and the 16-QAM modulations respectively. In each figure, two transmit antennas and one receive antenna (2×1), and two transmit antennas and two receive antennas (2×2) are performed.

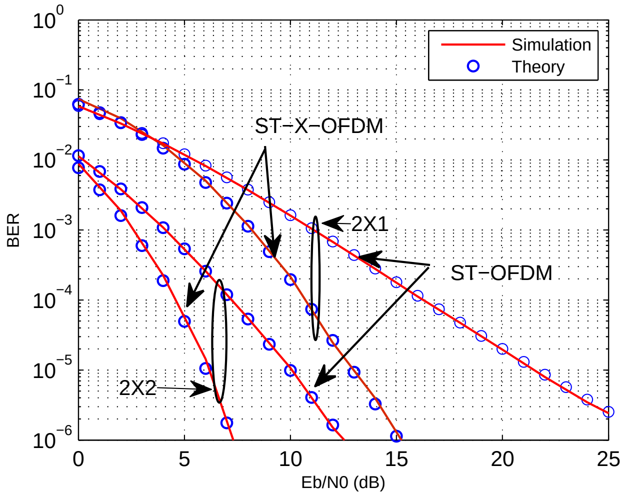


Fig. 4. BER performance of the proposed ST-X-OFDM and ST-OFDM systems over ITU pedestrian B channel model, ZF detection and QPSK modulation format.

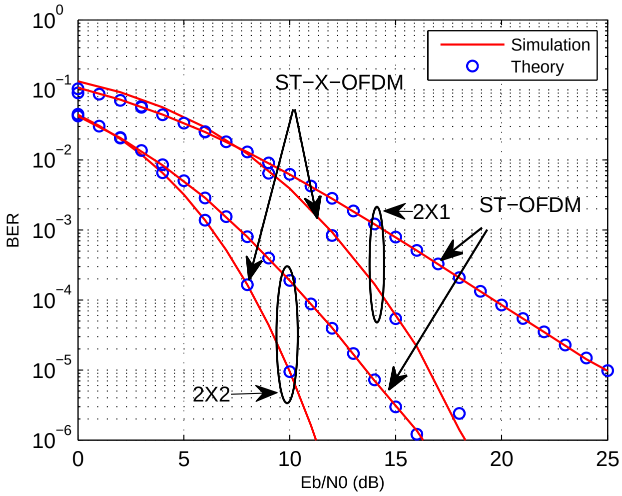


Fig. 5. BER performance of the proposed ST-X-OFDM and ST-OFDM systems over ITU pedestrian B channel model, ZF detection and 16-QAM modulation format.

It can be seen from Figs. 4 and 5 that the theoretical results are undistinguishable from the simulation results. It can also be observed from Figs. 4 and 5 that for all the transmission scenarios, the proposed ST-X-OFDM system is superior to the conventional ST-OFDM scheme. For the case of the QPSK, it can be observed from Fig. 4 that the proposed ST-X-OFDM achieves about 9 dB SNR, per bit (E_b/N_0), gain over the conventional ST-OFDM at 10^{-5} BER. This SNR gain is about 10 dB in the case of the 16-QAM as shown in Fig. 5. Thus, the proposed scheme not only massively reduces the transmitter complexity and the PAPR, but also achieves significant SNR gain in comparison with the ST-OFDM system.

VI. Conclusions

In this paper, a new Alamouti ST-OFDM system based on an efficient **X**-transform is presented. The **X**-transform is a single orthogonal transform that combines the effects of the DHT and the DFT with a huge reduction in the arithmetic operations. An exact close-form expression for the BER performance of the proposed ST-X-OFDM over multipath channels has also been derived in this work. The proposed ST-X-OFDM scheme has a very simple transmitter structure and showed significant transmission performance improvement in comparison with the conventional ST-OFDM system. It has been shown by computer simulation over frequency-selective multipath channels that the proposed ST-X-OFDM achieves valuable SNR gain and significant PAPR reduction in comparison with the conventional ST-OFDM. This improvement is attributed to the fact that the **X**-transform distributes each information symbol more widely across the whole spectrum than the DFT, as the former has the effects of two orthogonal transforms, hence, increasing the diversity of the transmitted signal.

References

[1] K. Wen-Yi and M. Fitz, Design and analysis of transmitter diversity using intentional frequency offset for wireless communications, IEEE Trans. Veh. Technol., vol. 46, pp. 871881, Nov. 1997.

[2] D. Dao and C. Tellambura, Optimal rotations for quasi-orthogonal STBC with two-dimensional constellations, in IEEE Global Telecommunications Conference (GLOBECOM), vol. 4, Dec. 2005, pp. 15.

[3] D. Senaratne and C. Tellambura, ”Spatial Multipath Resolution with Space Time Block Codes,” IEEE Wireless Commun. Lett., vol. 1, pp. 249-252, 2012.

[4] S. Alamouti, A simple transmit diversity technique for wireless communications, IEEE J. Sel. Areas Commun., vol. 16, pp. 14511458, Oct. 1998.

[5] A. Scaglione, P. Stocia, S. Barbarossa, G. Giannakis, and H. Sampath, Optimal designs for space-time linear precoders and decoders, IEEE Trans. Signal Process., vol. 50, no. 5, pp. 10511064, May 2002.

[6] K. Eunseok and H. Daesik, A robust STBC-Based transmit diversity scheme for OFDM systems over spatially transmit correlated fading channels, IEEE Trans. Veh. Technol., vol. 56, pp. 984991, Mar. 2007.

[7] N. Al-Dhahir, ”Single-carrier frequency-domain equalization for space-time block-coded transmissions over frequency-selective fading channels,” Communications Letters, IEEE, vol. 5, pp. 304-306, 2001.

[8] L. Jong-Ho and K. Seong-Cheol, Efficient ISI cancellation for STBC-OFDM systems using successive interference cancellation, IEEE Commun. Lett., vol. 10, pp. 629631, 2006.

[9] C. Yuan-Hwui and P. See-May, Unitary precoders for ST-OFDM systems using alamouti STBC, IEEE Transactions on circuits and Systems I: Regular Papers, vol. 55, pp. 28602869, 2008.

[10] I. Baig and V. Jeoti, ”PAPR analysis of DHT-precoded OFDM system for M-QAM,” in International Conference on Intelligent and Advanced Systems (ICIAS), 2010, pp. 1-4.

[11] J.G. Proakis and M. Salehi, ”Digital Communications”, McGraw-Hill Companies, Inc, 1221 Avenue of the Americas, New York, 2008.

[12] H.A. Leftah and S. Boussakta, ”Efficient coded DCT-OFDM system utilizing Walsh-Hadamard Transform,” in IEEE Wireless Telecommunications Symposium (WTS), 2012, pp. 1-5.