Frequency domain oversampled OFDM with Index modulation for SFBC-based MIMO system in high-speed wireless communications

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Abstract

This paper presents a novel approach to the development of a multiple-input multiple-output (MIMO) orthogonal frequency division multiplexing with index modulation (OFDM-IM) using the space-frequency block code (SFBC) technique. The system, which offers a high transmission data rate and high performance in fast-fading channel conditions even with an oversampled OFDM-IM signal and imperfect channel estimation, represents a significant advancement in the field. The main focus of this paper is to analyze the performance of this novel SFBC-based MIMO system with an oversampled OFDM-IM signal in terms of the bit-error rate and throughput. Computer simulations unequivocally demonstrate that the performance of this system not only meets but surpasses that of conventional systems in high-speed wireless communications, marking a significant advancement in the field. The potential of the proposed method to inspire further research and development in the field is a key highlight, aiming to motivate the audience to explore new possibilities and contribute to the advancement of wireless communications.

Keywords: Multiple-input multiple-output, Oversampled OFDM, Index modulation, Space-frequency block code, Fast-fading channel estimation.

1. Introduction

A multiple-input and multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) system has already been provided for the 5G wireless communication networks [1-2] because it can support high transmission data rates and high performance with more network capacity. From the advantages of the OFDM with index modulation technique (OFDM-IM), which supports high spectral efficiency by saving energy in transmission signals, the OFDM-IM has a lot of attention in wireless communication fields [3-4]. In OFDM-IM, the information data is transmitted in terms of OFDM-IM subblock over all subcarriers in one OFDM-IM symbol. Each subblock includes the k_2-active subcarriers containing the M-ary modulated signal located by the following *IM* index [5] over k_1-available subcarriers. From such OFDM-IM structure, it can be called "(k_1,k_2)-OFDM-IM."

To increase the gain of transmit diversity, space-time block code (STBC) and space-frequency block code (SFBC) techniques are usually the options because they can provide a higher gain of transmit diversity relatively [6-7]. In the structure of STBC, in which two successive data symbols are encoded, the available subcarriers for data transmission would be smaller in the pilot-assisted channel estimation (PACE) structure [8]. Based on this limitation, SFBC is suitable for use because the imperfect channel state information (CSI) condition is considered. This means that the PACE method is assigned in this paper. In SFBC, the information signal is transmitted with combined coding and diversity gains. From all, the proposed system in this paper is called the SFBC-based MIMO-OFDM-IM system.

In [7], the SFBC MIMO-OFDM with Index Modulation (IM) method was proposed to enhance spectral efficiency and performance in wireless communication systems by using a Walsh code pilot to prevent pilot signal overlapping at the receiver. While this approach improves spectral efficiency and provides better BER performance, its effectiveness diminishes rapidly in fast-fading channels, leading to a decrease in throughput performance. Additionally, the oversampled OFDM-IM signal, which is commonly used in practical scenarios, was not considered in [7]. These issues are addressed and resolved in this paper.

This paper uses the SFBC-based MIMO-OFDM-IM system to work with the oversampled OFDM-IM signal in fast-fading wireless channels. In an oversampled OFDM-IM signal with adding zero-padding (ZP), the number of data subcarriers is shorter than the number of IFFT/FFT points in the OFDM-IM symbol to avoid the aliasing at the output of digital-to-analog (D/A) converter. Under imperfect CSI conditions in a fast-fading channel, the (a_T×a_R) channels over the OFDM-IM frame are required to estimate at the receiver by using the PACE method. To satisfy this requirement, the pilot needs to be assigned for all OFDM-IM symbols over the OFDM-IM frame. However, to estimate the (a_T×a_R) channels over the OFDM-IM frame while maintaining the high transmission data rate, the pilots are assigned in two dimensions (2D), including frequency axis and time axis; the pilots are scatteringly inserted among data subcarriers in OFDM-IM symbol for the frequency axis and added among OFDM-IM symbols over OFDM-IM frame for the time axis.

In the process of CE, the maximum-likelihood (ML) algorithm is applied to estimate the (a_T×a_R) channels in the frequency axis; the estimated (a_T×a_R) channels over the OFDM-IM symbol are obtained. Next, the cubic spline (CS) interpolation technique is used for calculating the (a_T×a_R) channels of the OFDM-IM symbol in the time axis. Finally, the CSIs of (a_T×a_R) channels over the OFDM-IM frame are known at the receiver. Employing such CSI knowledge in OFDM-IM detection [9] with SFBC's gain, the information data is correctly modulated.

In the main issues of this paper, there are various assumptions and improvements, which can be outlined as follows:

- The performance of an SFBC-based (a_T×a_R)-MIMO system with an oversampled OFDM-IM signal is assumed under the fast-fading environment, a condition commonly encountered in high-speed wireless communications where the signal experiences rapid fluctuations due to multipath propagation and other factors.
- This paper presents a proposed PACE method for the SFBC-based MIMO-OFDM-IM system. The proposed PACE method introduces the allocation of the 2-axis scattered pilot, including frequency and time axes, which are designed for the OFDM-IM symbol in the (a_T×a_R)-MIMO system with the evasion of the overlapping pilot signal at the receiver. The ML method is applied to estimate the channel response (CR) over the OFDM-IM symbol for the frequency axis, while the CS method is used to calculate the CR over the OFDM-IM frame for the time axis. This innovative method significantly improves the accuracy of channel estimation in fast-fading conditions.
- The comparison of the performances of the proposed SFBC-based (a_T×a_R)-MIMO system with oversampled OFDM-IM signal and the conventional systems is analyzed in terms of BER (bit-error rate) and TP (throughput). From the analysis, the proposed system demonstrates superior performance over the traditional system, providing a compelling argument for its adoption.

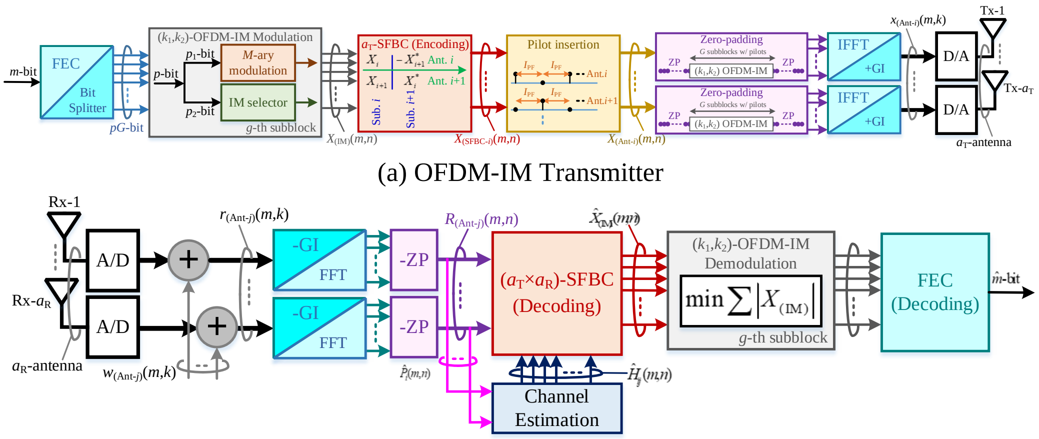
The rest of the paper is structured as follows: Section 2 presents the SFBC-based (a_T×a_R)-MIMO system model with an oversampled OFDM-IM signal. Section 3 introduces the proposed PACE method for the SFBC-based MIMO-OFDM-IM system with (a_T×a_R)-antenna under fast-fading channel conditions. In Section 4, the various computer simulation results are presented and discussed in detail, providing a comprehensive understanding of the system's performance. Finally, Section 5 summarizes the results and presents our future work.

2. SFBC-based (a_T×a_R)-MIMO system with oversampled OFDM-IM signal

Fig. 1 shows the block diagram of an SFBC-based MIMO system with a_T transmit antennas and a_R receive antennas using an oversampled OFDM-IM signal, which this paper considers. In **Fig. 1**, the m data bit sequences are encoded by FEC and split into G subblocks of (k_1,k_2)-OFDM-IM modulation at the transmitter. The input p bits (=p_1+p_2) of each subblock are divided into two parts; in *part* 1) p_1 bits ($= \lfloor log_2 C(k_1, k_2) \rfloor$) of p bits are input to M-ary modulation, where $\lfloor \cdot \rfloor$ is the floor operation, and in *part* 2) the remaining p_2 bits (=$k_2 \log_2 M$) are modulated by the location of the modulated data subcarrier of part 1, which are considered by the IM index selector, in which the k_2 active subcarrier's locations over k_1 subcarriers represent the IM index selector for each (k_1,k_2)-OFDM-IM subblock. It will be presented in Section 3 for more information on the OFDM-IM technique.

2.1 Proposed OFDM-IM with SFBC encoding

After OFDM-IM modulation, the OFDM-IM modulated signal $X_{(IM)}(n)$ is coded by SFBC encoding to increase the higher gain of transmit diversity, $X_{(SFBC\text{-}i)}(n)$ of the i-th transmit antenna (i=1, 2,…, a_T) can be expressed as.



(a) OFDM-IM Transmitter

(b) OFDM-IM Receiver

Fig. 1. Block diagram of proposed SFBC-based ($a_T \times a_R$)-MIMO system with oversampled OFDM-IM signal

$$\begin{bmatrix} X_{(\mathrm{SFBC}-i)}(n) & X_{(\mathrm{SFBC}-i)}(n+1) \\ X_{(\mathrm{SFBC}-i+1)}(n) & X_{(\mathrm{SFBC}-i+1)}(n+1) \end{bmatrix} = \begin{bmatrix} X_{(\mathrm{IM})}(n) & -X^{*}_{(\mathrm{IM})}(n+1) \\ X_{(\mathrm{IM})}(n+1) & X^{*}_{(\mathrm{IM})}(n) \end{bmatrix} \tag{1}$$

for n=0, 1,…, M, where M is the number of subcarriers. Next, the pilot $P_i(n)$ of the i-th transmit antenna is inserted scatteringly into G subblocks uniformly with pilot subcarrier intervals over M subcarriers, which can be given by.

$$X_{(\mathrm{Ant.}\text{-}i)}(n)=X_{(\mathrm{SFBC}\text{-}i)}(n)+P_i(n) \tag{2}$$

where $X_{(\mathrm{Ant.}\text{-}i)}(n)$ is the frequency domain signal, including the SFBC data signal and pilot signal of the i-th transmit antenna. After ZP adds that $X_{(\mathrm{Ant.}\text{-}i)}(n)$ signal at both ends to avoid the aliasing at the D/A output, it will then be performed by N-points IFFT, which can be expressed as.

$$x_{(\mathrm{Ant.}-i)}(k) = \frac{1}{\sqrt{N}} \sum_{n=0}^{M-1} X_{(\mathrm{Ant.}-i)}(n) \cdot e^{j\frac{2\pi k}{N}(n+ZP)} \tag{3}$$

where $x_{(\mathrm{Ant.}\text{-}i)}(k)$ is the time domain signal at the k-th sample of the i-th transmit antenna. Finally, $x_{(\mathrm{Ant.}\text{-}i)}(k)$ with adding guard interval (GI) is transmitted to the receiver at the j-th received antenna (j=1, 2,…, a_R) via the multipath ($a_T \times a_R$)-fading channels, which can be defined by.

$$H_{ij}(n) = \sum_{n_{\mathrm{PATH}}=0}^{N_{\mathrm{PATH}}-1} \beta_{ij}(n_{\mathrm{PATH}}) \cdot e^{-j\frac{2\pi n_{\mathrm{PATH}}}{N}(n+ZP)} \tag{4}$$

where $H_{ij}(n)$ is the frequency domain CR of the i-th transmit and the j-th receive antennas. $B_{ij}(n_{\mathrm{PATH}})$ is the time domain CR at the n_{PATH}-th delay path, and N_{PATH} is the number of delay paths.

At the receiver side, at the j-th receive antenna, the received signal can be given after removing the GI and performed by N-points FFT.

$$R_{(\text{Ant.-}j)}(n)=H_{ij}(n)\, X_{(\text{Ant.-}i)}(n)+W_j(n) \tag{5}$$

where $R_{(\text{Ant.-}j)}(n)$ is the frequency domain received signal at the j-th receive antenna and $W_j(n)$ is the additional White Gaussian noise at the j-th receive antenna. For the received OFDM-IM pilot symbol signal, the locations of the pilot subcarrier sent from the i-th transmit antenna received at the j-th receive antenna is the received pilot $\hat{P}_i(n)$ which is first used for channel estimation and then on the frequency and time axes. From all, it will be presented in Section 3. Using the high accuracy of estimated CR, the gain of transmit diversity from SFBC decoding can be active. The SFBC decoding for decoding the received $R_{(\text{Ant.-}j)}(n)$ signals for two transmit antennas (a_T=2), which are considered in this paper, can be defined as the following equation.

$$X(n_e) = \frac{\left\{\sum_{j=1}^{a_R} R_j(n_e)H_{1j}^*(n_e) + R_j^*(n_o)H_{2j}(n_o)\right\}D(n') - \left\{\sum_{j=1}^{a_R} R_j(n_e)H_{2j}^*(n_e) - R_j^*(n_o)H_{1j}(n_o)\right\}B(n')}{A(n')D(n') - B(n')C(n')} \tag{6}$$

and

$$X(n_o) = \frac{\left\{\sum_{j=1}^{a_R} R_j(n_e)H_{2j}^*(n_e) - R_j^*(n_o)H_{1j}(n_o)\right\}A(n') - \left\{\sum_{j=1}^{a_R} R_j(n_e)H_{1j}^*(n_e) + R_j^*(n_o)H_{2j}(n_o)\right\}C(n')}{A(n')D(n') - B(n')C(n')} \tag{7}$$

for n_e=Even{n}, n_o=odd{n} and n'=0,1,…,M/2-1, where

$$A(n') = \sum_{j=1}^{a_R} \left\{\left|H_{1j}(n_e)\right|^2 + \left|H_{2j}(n_o)\right|^2\right\}$$

$$B(n') = \sum_{j=1}^{a_R} \left\{H_{1j}^*(n_e)H_{2j}(n_e) - H_{1j}^*(n_o)H_{2j}(n_o)\right\}$$

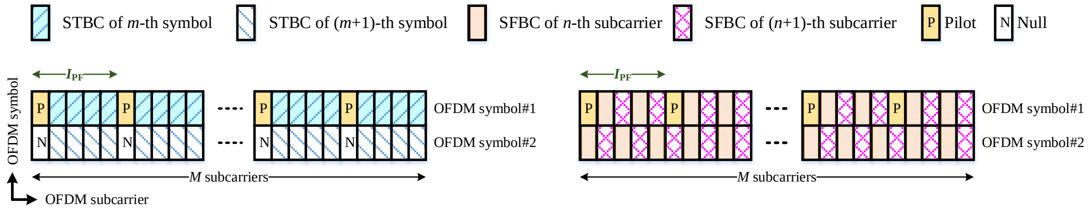
$$C(n') = \sum_{j=1}^{a_R} \left\{H_{1j}(n_e)H_{2j}^*(n_e) - H_{1j}(n_o)H_{2j}^*(n_o)\right\}$$

$$D(n') = \sum_{j=1}^{a_R} \left\{\left|H_{1j}(n_o)\right|^2 + \left|H_{2j}(n_e)\right|^2\right\}$$

Next, the SFBC decoded $\hat{X}_{IM}(n)$ signal is fed to the (k_1,k_2)-OFDM-IM demodulation, the ML detector [10]. Finally, the demodulated $\hat{m}$ data bit sequences are obtained after FEC decoding.

2.2 Transmission efficiencies of SFBC and STBC techniques

As mentioned earlier in the Introduction, the different structures of SFBC and STBC with pilot allocation are related to the system's transmission efficiency. The OFDM modulated signal X(m,n) at the m-th OFDM symbol with STBC encoding of the i-th transmit antenna (i=1, 2,…, a_T) can be expressed as.



(a) STBC encoding signal (b) SFBC encoding signal

Fig. 2. Pilot allocation for SFBC and STBC in two successive traditional OFDM symbols with M subcarriers.

Table 1. Transmission efficiency of SFBC and STBC techniques with pilot allocation

Techniques	**Available subcarriers (M_A)**	**Transmission efficiency over 2M**	**%**
No pilot	2M	1	100%
SFBC	$M(2I_{PF}$-2)/I_{PF}	1-0.5R	87.5%
STBC	$M(2I_{PF}$-1)/I_{PF}	1-R	75.0%

Note: R=1/I_{PF}

$$\begin{bmatrix} X_{(\mathrm{STBC}-i)}(m,n) & X_{(\mathrm{STBC}-i)}(m+1,n) \\ X_{(\mathrm{STBC}-i+1)}(m,n) & X_{(\mathrm{STBC}-i+1)}(m+1,n) \end{bmatrix} = \begin{bmatrix} X(m,n) & -X^*(m+1,n) \\ X(m+1,n) & X^*(m,n) \end{bmatrix} \quad (8)$$

for m=1, 2,…, N_L, where N_L is the number of OFDM symbols in one OFDM frame. From (1) and (8), it can be seen that two consecutive OFDM data subcarriers are encoded by SFBC encoding; in contrast with STBC encoding, the whole OFDM data subcarriers of two successive OFDM data symbols are encoded. This means that two consecutive OFDM data symbols are robust and reliable in SFBC encoding, as they are independent. Therefore, the pilot allocation is assigned under the condition of the structures of STBC and SFBC techniques, as shown in **Fig. 2**. **Fig. 2** compares pilot allocation for SFBC and STBC in two successive traditional OFDM symbols with M subcarriers. In **Fig. 2**, it can be seen that the subcarriers assigned for the pilot in the OFDM symbol#1 of the OFDM symbol#2 are null in the STBC technique. At the same time, in the SFBC technique, all subcarriers over OFDM symbol#2 are available for data transmission. The transmission efficiencies of STBC and SFBC with pilot allocation assigned in **Fig. 2** can be calculated for I_{PF}=4, as shown in **Table 1**.

The results in **Table 1** show that the SFBC technique's transmission efficiency surpasses that of the STBC technique. This superiority and advantage of the SFBC technique are vital reasons for its consideration in this paper.

3. Proposed PACE method for practical SFBC-based (a_T×a_R)-MIMO-OFDM-IM system

Firstly, the definition of a practical OFDM-IM system is a system with an oversampled OFDM-IM signal in which M subcarriers are shorter than N-points IFFT/FFT ($M<N$). For this section, we propose a PACE method including the 2-axis scattered pilot assignments with ML-based CE and CS-based CE methods in two axes for practical SFBC-based (a_T×a_R)-MIMO-OFDM-IM system over time-varying fading channels with high-speed wireless communication.

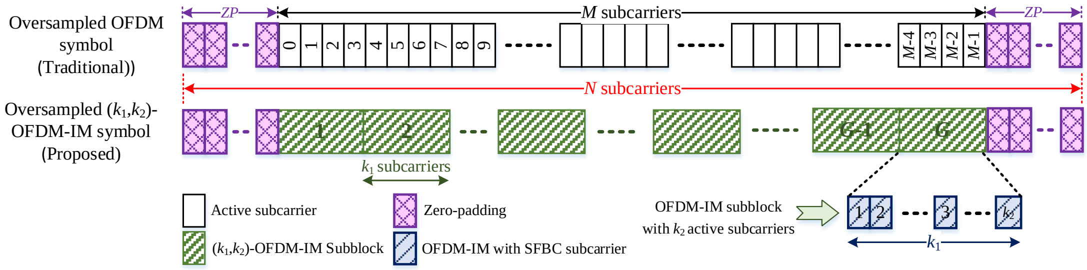


Fig. 3. Proposal of oversampled (k_1,k_2)-OFDM-IM symbols with SFBC in the frequency domain.

Table 2. A look-up table example for (4,2)-OFDM-IM

Bits (p_1=2 bits)	Pointers	g-th Subblock
"00"	1,2	$[k_2^1 \quad k_2^2 \quad 0 \quad 0]^T$
"01"	2,3	$[0 \quad k_2^1 \quad k_2^2 \quad 0]^T$
"10"	3,4	$[0 \quad 0 \quad k_2^1 \quad k_2^2]^T$
"11"	1,4	$[k_2^1 \quad 0 \quad 0 \quad k_2^2]^T$

Note: $[]^T$ is transpose operation

3.1 Proposed frequency domain oversampling for zero-padding OFDM-IM transmission signal

As mentioned in the Introduction, the oversampled signal is considered unavoidably to avoid the aliasing effect at the D/A output occurring in the practical system. For this reason, we propose the (k_1,k_2)-OFDM-IM symbol by oversampling for a practical SFBC-based (a_T×a_R)-MIMO-OFDM-IM system, as shown in **Fig. 3**. In **Fig. 3**, the proposed frequency domain structure of oversampled (k_1,k_2)-OFDM-IM symbol indicates almost the same as the traditional oversampled OFDM symbol with M subcarriers. At both ends of G subblocks are added by zero-padding with ZP subcarriers. In (k_1,k_2)-OFDM-IM symbol, G subblocks over M subcarriers can be given by.

$$G = \frac{M}{k_1} \tag{9}$$

Each p-bit groups are mapped to the g-th subblock (g=1, 2,…, G) with length k_1 subcarriers. Contrary to the traditional OFDM symbol, p-bit groups are not only mapped by M-ary modulation but also by the pointers of the active subcarriers, which can give an example of the look-up table shown in **Table 2**. The look-up table is designed to be utilized at the transmitter and is known at the receiver. This table equips the following pointers for p_1-bit in the g-th subblock. **Table 2** shows the example for (4,2)-OFDM-IM with p_1=2 bits. From the table, the size of possible realization (s_{PR}) can be calculated by.

$$s_{\mathrm{PR}} = 2^{p_1} \tag{10}$$

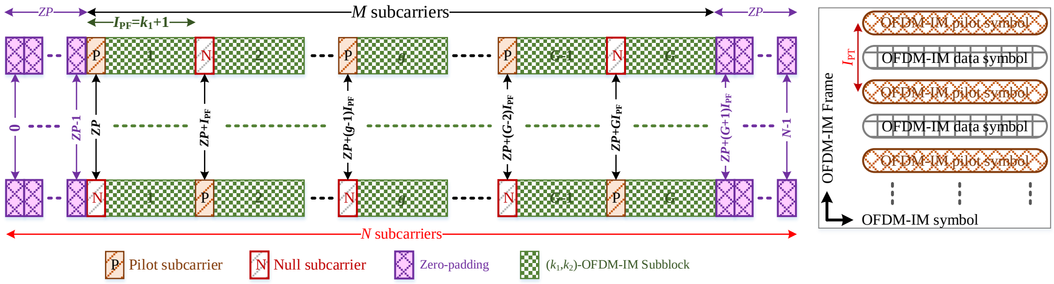


Fig. 4. Proposed 2D-scattered pilot assignment for SFBC-based MIMO system with oversampled OFDM-IM signal.

From (10), there are four possible realizations (s_{PR}=4) corresponding to four symbols by two bits (p_1=2 bits). From the proposed oversampled (k_1,k_2)-OFDM-IM symbols in the frequency domain, as shown in **Fig. 3**, it can be defined as.

$$X_{(\mathrm{Ant.}-i)}(m) = \begin{cases} \mathrm{OFDM-IM\ subblock}, n = ZP, ZP+1, \dots, ZP+M-1 \\ \mathrm{Zero-pading} \qquad\qquad , otherwise \end{cases} \tag{11}$$

where ZP is the number of zero-padding on each side, and M is the number of subcarriers.

3.2 Proposed PACE method with 2D-scattered pilot

The CSI of the fast-fading (a_T×a_R) channel must be known by estimating at the receiver. This can be achieved by the pilot signals that are received and assigned to the transmitter. In this study, this paper proposes a 2D-scattered pilot assignment for estimation of the CSI of the (a_T×a_R)-channel under fast-fading channel condition as shown in **Fig. 4**. The proposed 2D-scattered pilot assignment is considered under the condition of the oversampled OFDM-IM transmission as shown in **Fig. 3**. To maintain the high data rate, the scattered-pilot signal is assigned for some OFDM-IM symbols over OFDM-IM frame. Such OFDM-IM symbol, including the pilot, is called the OFDM-IM pilot symbol; in contrast to the other, it is called the OFDM-IM data symbol in this paper.

Fig. 4 shows the OFDM-IM frame with the proposed 2D-scattered pilot assignment in which OFDM-IM symbols adding the scattered-pilot with the interval of the pilot in frequency axis (I_{PF}) are cyclically inserted with the interval of the pilot in time axis (I_{PT}) over OFDM-IM frame which can be given by.

$$X_{(\mathrm{Ant.}-i)}(m) = \begin{cases} X_{(\mathrm{Ant.}-i)}\big|_{in\ (2)} & , m = (n_{\mathrm{PT}}-1)I_{\mathrm{PT}}+1 \\ X_{(\mathrm{SFBC}-i)} & , otherwise \end{cases} \tag{12}$$

for m=1, 2, …, N_S and n_{PT}=1, 2, …, N_S/I_{PT}, where N_S is the number of OFDM-IM symbols in one OFDM-IM frame.

In the OFDM-IM pilot symbol (m=(n_{PT}-1)I_{PT}+1)) given in (12), pilot subcarriers and nulls are assigned to avoid the overlapped received pilot signal at each receive antenna of the receiver. The pilot is activated in the different subcarrier's location for the transmit antenna, while the location of the pilot used for the other transmit antennas is null. From **Fig. 4**, the pilot and null are added uniformly with I_{PF} intervals among G subblocks over M subcarriers. The space of the subcarrier between pilot (P) and null (N) is related to the k_1-length of each subblock, which can be given by.

$$I_{PF}=k_1+1 \tag{13}$$

This paper assumes two transmit antennas (a_T=2) for the system. The assignment of scattered-pilot, including P-pilot and N-null over OFDM-IM subblocks in the oversampled OFDM-IM symbol, as shown in **Fig. 4**, can be expressed as follows.

- At transmit antenna 1 (i=1),

$$P_{\mathrm{Ant}-1}(n)=\begin{cases}\mathrm{P-pilot} & ,n=ZP+(g_{\mathrm{odd}}-1)I_{\mathrm{PF}}\\ \mathrm{N-null} & ,n=ZP+(g_{\mathrm{even}}-1)I_{\mathrm{PF}}\\ \mathrm{OFDM-IM\ subblock} & ,otherwise\end{cases} \tag{14}$$

and

- At transmit antenna 2 (i=2),

$$P_{\mathrm{Ant}-2}(n)=\begin{cases}\mathrm{P-pilot} & ,n=ZP+(g_{\mathrm{even}}-1)I_{\mathrm{PF}}\\ \mathrm{N-null} & ,n=ZP+(g_{\mathrm{odd}}-1)I_{\mathrm{PF}}\\ \mathrm{OFDM-IM\ subblock} & ,otherwise\end{cases} \tag{15}$$

for g=1, 2, …, G where g_{odd} and g_{even} are the odd and even of g numbers, respectively. Using the proposed 2D-scattered pilot assignment, the CSI can be estimated independently at each receive antenna without the collision of the sent pilot signal from two transmit antennas of the transmitter.

From (5), the received pilot signal at the pilot's location (n_{Pil}) as given by (14) and (15) is rewritten by.

$$R_{(\mathrm{Ant.}\text{-}j)}(n_{\mathrm{Pil}})=H_{ij}(n_{\mathrm{Pil}})\,P_{(\mathrm{Ant.}\text{-}i)}(n_{\mathrm{Pil}})+W_j(n_{\mathrm{Pil}}) \tag{16}$$

where

$$n_{\mathrm{Pil}}=\begin{cases}ZP+(g_{\mathrm{odd}}-1)I_{\mathrm{PF}} & ,i=1\\ ZP+(g_{\mathrm{even}}-1)I_{\mathrm{PF}} & ,i=2\end{cases} \tag{17}$$

Divided (16) by known $P_{(\mathrm{Ant.}\text{-}i)}$ given by (14) and (15), the frequency domain CRs $H_{ij}(n_{Pil})$ at the location of n_{Pil} of OFDM-IM symbol are firstly estimated by.

$$\hat{H}_{ij}(n_{\mathrm{Pil}})=\frac{R_{(\mathrm{Ant}-j)}(n_{\mathrm{Pil}})}{P_{(\mathrm{Ant}-i)}(n_{\mathrm{Pil}})}=H_{ij}(n_{\mathrm{Pil}})+\frac{W_j(n_{\mathrm{Pil}})}{P_{(\mathrm{Ant}-i)}(n_{\mathrm{Pil}})} \tag{18}$$

where $W_j(n_{\text{Pil}})$ is an uncorrelated Gaussian variable, the estimated time-domain CR as given by (18) over OFDM-IM symbol can be calculated by applying the ML method, which can be provided by.

$$L_{ML}\left[\hat{\beta}_{ij}(n_{\text{PATH}})\right] = \min_{\hat{\beta}_{ij}(n_{\text{PATH}})} \left[\left|H_{ij}(n_{\text{Pil}})\big|_{(4)} - \hat{H}_{ij}(n_{\text{Pil}})\big|_{(18)}\right|^2\right] \tag{19}$$

for $\hat{\beta}_{ij}(n_{\text{PATH}}) = [\hat{\beta}_{ij}(0), \ldots, \hat{\beta}_{ij}(N_{\text{GI}} - 1)]$. Taking the partial derivative of (19) with $\hat{\beta}^*_{ij}(n_{\text{PATH}})$, and let it be equal to zero, where * represents conjugate operation, which can be expressed as.

$$\begin{aligned}&\sum_{n_g=0}^{N_{g_{\text{odd}}}\, or\, N_{g_{\text{even}}}-1} \sum_{n_{\text{GI}}=0}^{N_{\text{GI}}-1} \hat{H}_{ij}(n_{\text{Pil,Ant.}-i})\big|_{(18)} \cdot e^{j\frac{2\pi n_{\text{GI}} n_g}{N}} \\ &= \sum_{n_g=0}^{N_{g_{\text{odd}}}\, or\, N_{g_{\text{even}}}-1} \sum_{n_{\text{PATH}}=0}^{N_{\text{GI}}-1} \sum_{n_{\text{GI}}=0}^{N_{\text{GI}}-1} \hat{\beta}_{ij}(n_{\text{PATH}}) \cdot e^{-j\frac{2\pi n_g}{N}(n_{\text{PATH}} - n_{\text{GI}})}\end{aligned} \tag{20}$$

where $N_{g_{\text{odd}}}$ and $N_{g_{\text{even}}}$ are the number of g_{odd} and g_{even}, respectively. From (20), the optimization of $\hat{\beta}_{ij}(n_{\text{PATH}})$ is linear, the Moore-Penrose matrix inversion (†) can be applied to solve (20), which can be expressed as.

$$\boldsymbol{\beta}_{ij} = \mathbf{e}^{\dagger} \mathbf{H}_{ij}\big|_{(18)} \tag{21}$$

where $\boldsymbol{\beta}_{ij}$ is the matrix of time domain CR $\hat{\beta}_{ij}(n_{\text{PATH}})$ at the n_{PATH}-th delay path with ($N_{\text{GI}}\times 1$)-size, $\mathbf{H}_{ij}$ is the matrix of the frequency domain CRs $\hat{H}_{ij}(n_{\text{Pil}})$ estimated at the location of n_{Pil} as given in (18) with ($N_{g_{\text{odd}}}$ or $N_{g_{\text{even}}}\times 1$)-size and **e** is the matrix of the exponential term in (20) with ($N_{g_{\text{odd}}}$ or $N_{g_{\text{even}}}\times N_{\text{GI}}$)-size which can be given by.

$$\mathbf{e} = \sum_{n_g=0}^{N_{g_{\text{odd}}}\, or\, N_{g_{\text{even}}}-1} \sum_{n_{\text{GI}}=0}^{N_{\text{GI}}-1} e^{j\frac{2\pi n_{\text{GI}} n_g}{N}} \tag{22}$$

Table 3. Simulation parameters

Parameters	**Information**
Modulation for Data and Pilot	QPSK with coherent demodulation
N-point FFT/IFFT	256
ZP-padding	48
M subcarriers	N-2ZP
(k_1,k_2)-OFDM-IM symbol	(4,2)-OFDM-IM
N_{GI} samples	24
(I_{PT},I_{PF})-interval	(4,k_1+1)
Occupied bandwidth (MHz)	5
Radiofrequency (GHz)	5

FEC	
Encoding	Convolutional with ½-code rate and 7-constraint length
Decoding	Viterbi (Hard decision)
Interleaver	Random
Packet length (bit/packet)	512
Time-varying fading channel model	
Channel	Rayleigh fading
Profile	-1dB-decaying with Exponential
N_{PATH} delay paths	8
$f_D T_S$	0.05 (Corresponding to approx. 300km/h)
Channel estimation Techniques	
Frequency axis	ML method
Time axis	CS method

It can be seen in the result in (22) that it can be calculated once in advance. For this reason, the computational complexity of the CE process would be decreased. Finally, the frequency domain CR $\hat{H}_{ij}(n)$ over OFDM-IM symbol can be obtained by performing FFT to the results in (21).

As the fast-fading channel condition is assumed in this paper, only the frequency domain CRs $\hat{H}_{ij}(n)$ over OFDM-IM symbol is not enough for data demodulation to be correct. The frequency domain CRs $\hat{H}_{ij}(n)$ over OFDM-IM symbols in one OFDM-IM frame is essential to correct data demodulation. For this fact, The frequency domain CRs $\hat{H}_{ij}(n)$ over one OFDM-IM frame are estimated by the CS method, which interpolates the frequency domain CRs $\hat{H}_{ij}(n)$ over OFDM-IM symbol with I_{PT}.

Employing such $\hat{H}_{ij}(n)$ in SFBC decoding, as given in (6) and (7), the high gain of transmit diversity is due to the active SFBC technique, enhancing system performance relatively.

4. Simulation results

This section showcases the superior performance of the proposed method through computer simulations, evaluated in terms of bit-error rate (BER) and throughput (THP) using parameters which are listed in **Table 3**. The proposed SFBC-based MIMO system with an oversampled OFDM-IM signal, featuring two transmit (a_T=2) and two and four receive (a_R=2 and 4) antennas, is evaluated under the Rayleigh higher time-varying fading channel with 8-delay paths and -1dB-decaying. (4,2)-OFDM-IM with QPSK modulation for data and pilot, N=256 with ZP=48, and N_{GI}=24 samples. To validate the potential of the proposed method, we compare it with the conventional ($a_T \times a_R$)-transmit-diversity called Tx-Div($a_T \times a_R$) and the conventional Tx-Div($a_T \times a_R$) with Maximum Combiner Ratio (MRC) called MRC($a_T \times a_R$) methods. The results of these comparisons unequivocally confirm the superiority of our proposed system. This superiority is not just a marginal improvement but a significant leap forward in high-speed wireless communications, marking an essential advancement in the field and leaving no doubt about its potential.

Fig. 5 shows the BER performance of the proposed SFBC method of using (2×2) and (2×4) antennas compared with the conventional methods versus the carrier-to-noise power ration (CNR) at $f_D T_S$→0.05 (corresponding to approx.300km/h), where f_D and T_S are the Doppler frequency (Hz) and the symbol time duration (sec), respectively. The results in **Fig. 5** show the BER of Conv.Tx-Div(2×2) method is the worst, which cannot be implemented for the system under fast-fading channel conditions. The result of conventional Tx-Div with the MRC method is better and much improved than Conv—Tx-Div (2×2)'s, especially when employing the receive antenna by four. Due to the gain of the MRC technique, however, the proposed SFBC method performs much better BER than the Conv.MRC method because the SFBC technique provides a higher transmit diversity gain, especially with more receive antenna.

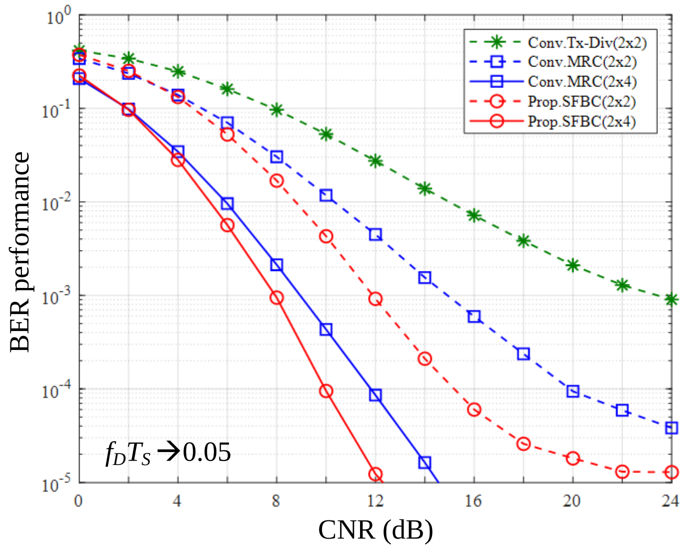


Fig. 5. BER versus CNR at $f_D T_S \rightarrow 0.05$.

Fig. 6 presents the results in terms of BER versus $f_D T_S$ at CNR=10dB. From **Fig. 6**, it can be observed that the proposed SFBC method can achieve much better BER when taking the receive antenna by four than the others from slow-fading to fast-fading. For the results in **Figs. 5** and **6**, the gain of SFBC decoding can be active by employing the estimated channel obtained by the accurate CE method under fast-fading channel conditions.

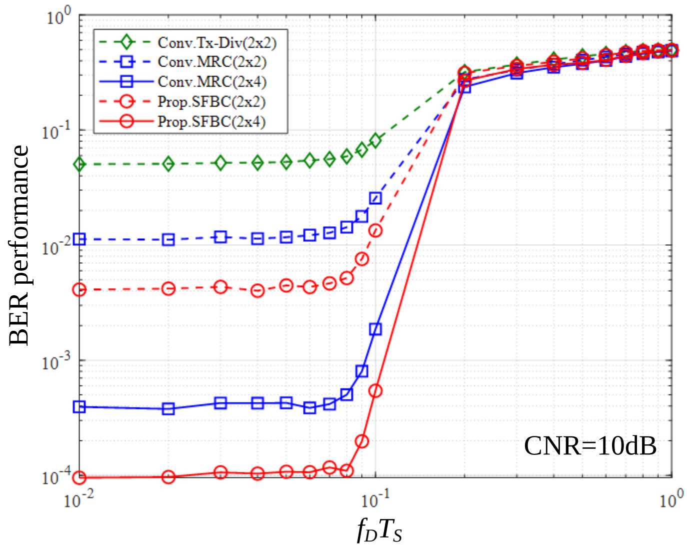


Fig. 6. BER versus $f_D T_S$ at CNR=10dB.

Fig. 7 shows the THP performance versus CNR at $f_D T_S \rightarrow 0.05$ (corresponding to approx.300km/h). At CNR=10dB, the THP of the proposed SFBC method of using (2×4) antennas is approximately 0.38bps/Hz, 0.38bps/Hz, 0.06bps/Hz, and 0.34bps/Hz improved for the Conv.Tx-Div(2×2), Conv.MRC(2×2), Conv.MRC(2×4), and Prop.SFBC(2×2), respectively, because of the effective transmit diversity technique with more receive antennas. For the results of THP in **Fig. 7**, which the Prop.SFBC(2×4) method gains approximately 0.06bps/Hz and 0.34bps/Hz more than Conv.MRC(2×4) and Prop.SFBC(2×2) methods at CNR=10dB mean that the Prop.SFBC(2×4) method can improve approximately 0.3Mbps and 1.7Mbps of transmission data rate over the Conv.MRC(2×4) and Prop.SFBC(2×2) methods, respectively, at 5MHz of occupied bandwidth.

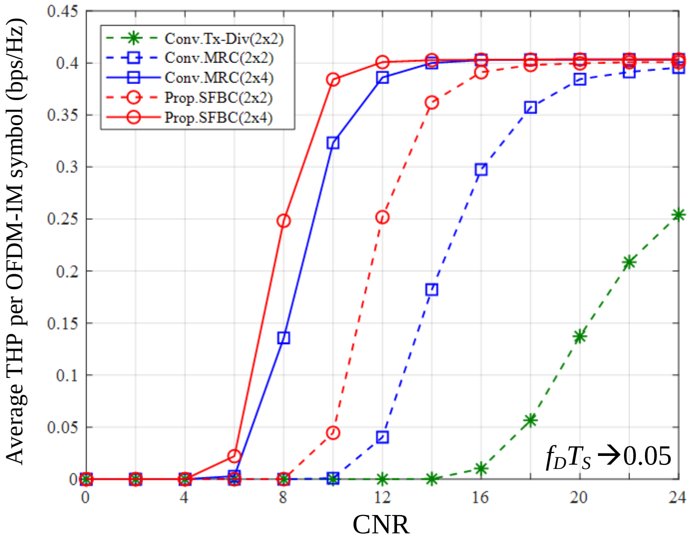


Fig. 7. THP versus CNR at $f_D T_S \rightarrow 0.05$.

5. Conclusion

This paper proposes a novel approach to developing an SFBC-based ($a_T \times a_R$)-MIMO system with an oversampled OFDM-IM signal. The proposed system includes the proposed PACE method consisting of the 2D-scattered pilot assignment to estimate the CSI of the ($a_T \times a_R$)-channel under fast-fading channel conditions. The ML-based CE method is used to calculate the channel over the OFDM-IM symbol with IPF in the frequency axis, and the CS-based CE method is used to estimate the channel over the OFDM-IM frame with IPT on the time axis. The BER performance of the proposed SFBC-based ($a_T \times a_R$)-MIMO system with an oversampled OFDM-IM signal was evaluated by computer simulation and compared to the conventional Tx-Div($a_T \times a_R$) and MRC($a_T \times a_R$) methods. Our results unequivocally demonstrate that the proposed SFBC($a_T \times a_R$) method significantly outperforms the conventional methods in terms of BER, especially when using four transmit antennas in the fast-fading channel. Furthermore, the proposed SFBC (2×4) method shows an approximate improvement of 0.3Mbps and 1.7Mbps in transmission data rates over the Conv.MRC(2×4) and the proposed SFBC(2×2) methods at 5MHz occupied bandwidth, respectively. These results confirm the effectiveness of the proposed methods in enhancing BER with a high data rate in fast-fading channel conditions, even with an oversampled OFDM-IM signal and imperfect CSI. In future work, we plan to explore deep learning-based channel estimation techniques to further enhance the performance of the proposed method in massive MIMO configurations, particularly for high-speed mobile scenarios. This will involve addressing challenges such as rapid channel variations and scalability to large antenna arrays, aiming to improve reliability and efficiency under practical deployment conditions.

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