Supplementary Index Bit aided Transmit Diversity Scheme for Enhanced DCT-OFDM with Index Modulation

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Abstract—**In this paper, a novel supplementary index bit aided transmit diversity (SIB-TD) approach is proposed for an enhanced discrete cosine transform based orthogonal frequency division multiplexing with index modulation (EDCT-OFDM-IM) system. Specifically, conventional index modulation (IM) is employed for the first antenna, i.e., the main branch, and the non-activated subcarriers' indices in one index modulation group are utilized for the IM mapping on the same index modulation group for the second antenna, i.e., the diversity branch. Hence, each subcarrier can not be synchronously activated on the two antennas. Finally, the non-activated subcarriers of one antenna will transmit the same modulated symbols of the other antenna to exploit the diversity gain. At the receiver, a maximum likelihood group-wise receiver is also developed by detecting the main and diversity branches jointly. Simulation results demonstrate the superiority of the proposed scheme over both the conventional EDCT-OFDM-IM and the DFT-OFDM with Alamouti code either with or without IM respectively, even under the imperfect channel estimation.**

Index Terms—**Discrete Cosine Transform (DCT), Orthogonal Frequency Division Multiplexing (OFDM), Index Modulation (IM), Supplementary Index Bit Transmit Diversity (SIB-TD)**

I. INTRODUCTION

ORthogonal frequency division multiplexing (OFDM) with index modulation (OFDM-IM) [1]-[4], which employs the indices of active subcarriers as an additional means to convey information, constitutes an appealing multicarrier technique. Compared with conventional OFDM systems, it has been demonstrated in [4] that it can achieve a better bit error rate (BER) performance and lower peak-to-average power ratio (PAPR). Thanks to these merits, it shows promising potential in many scenarios, such as vehicle-to-vehicle (V2V), and even in 5G and beyond wireless communication systems [1] - [4]. However, in some contexts of cost or power limitation scenarios, e.g., in the machine-type communication, low cost and low power consumption at the end devices is the main focus and most critical demand. On one hand, real-valued DCT requires less implementation complexity than that for complex-valued DFT processing, and on the other hand, at the downlink where the network transmits and end devices receive, more sophisticated schemes may be exploited in the network due to the low cost and long battery life requirements at end devices, therefore, employing transmit diversity is desirable for downlink transmissions.

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To tackle this issue, an enhanced discrete cosine transform (DCT) based OFDM-IM (EDCT-OFDM-IM) scheme was proposed, where the operation of DFT is substituted by that of the low-complexity DCT, resulting in high energy efficiency. Both the theoretical and simulation results of [5] - [11] showed the superiority of the EDCT-OFDM-IM scheme over its DFT-OFDM-IM counterpart having the same transmit rate and the identical bandwidth in terms of BER performance. Due to these benefits, it becomes a promising solution for target applications, e.g., machine-type communication in the vertical market, where low power consumption is needed for the terminal. Moreover, since demands for low cost and low power consumption are mainly at the end devices , at the downlink where base station transmits and end devices receive, more sophisticated schemes, e.g., transmit diversity, may be exploited at the base station for downlink transmissions.

As a continuation of works carried out for single transmissions shown in [5], and to further improve the diversity gain of EDCT-OFDM-IM system, a novel supplementary index bit aided transmit diversity method is designed. Specifically, the conventional IM is invoked for the subcarriers of the first antenna and the non-activated indices of subcarriers are maintained for the IM design of the second antenna. Explicitly, we first mark the second antenna's subcarriers having the aforementioned non-activated indices and then IM is employed for the signed subcarriers with varied modulation schemes. Accordingly, at each time slot, the subcarriers of the two antennas can not be synchronously activated. Finally, the non-activated subcarriers of each antenna will transmit the same symbols of the alternative antenna to exploit the diversity gain. And a maximum likelihood group-wise receiver is designed by jointly detecting the subcarriers of the two antennas. Numerical results show that the proposed scheme is able to achieve significant diversity gains over the conventional EDCT-OFDM-IM.

II. SYSTEM MODEL OF EDCT-OFDM-IM WITH SIB-TD

Since there are both inactive and active sub-carriers in one group with index modulation, by nature these inactive

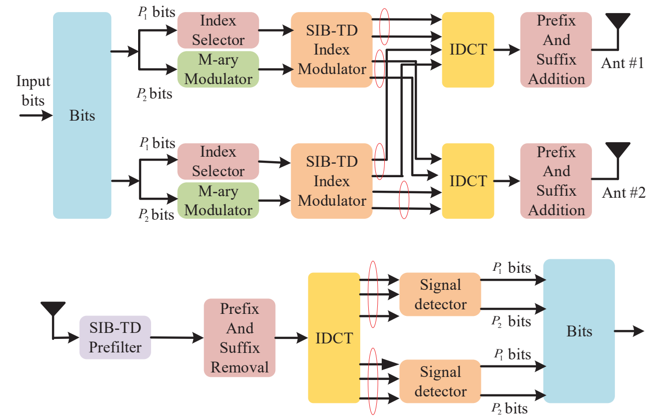

Fig. 1. System model of SIB-TD for EDCT-OFDM-IM.

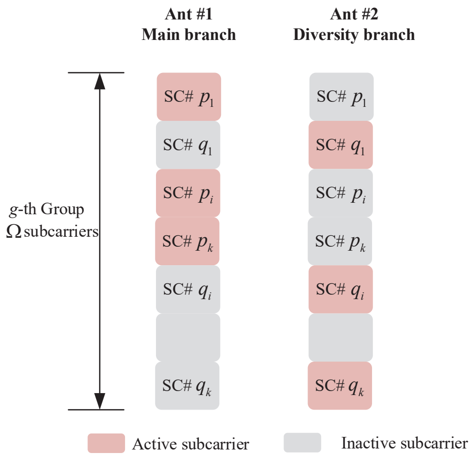

Fig. 2. SIB-TD index modulator.

subcarriers in one antenna can be active in the other antenna at the same time, such that the interference between two antennas is avoided, resulting in a so-called supplementary index bits transmit diversity (SIB-TD). For an index bits vector in the index modulation module for one antenna, the index bits vector in the other antenna is not the same as its counterpart in the first antenna, but the index bits vector with the largest Euclidian distance in the codebook to the index bits vector in the first antenna, a.k.a., Supplementary Index Bits. Fig. 1 illustrates a SIB-TD system for EDCT-OFDM-IM.

A. SIB-TD Transmitter

A SIB-TD based EDCT-OFDM-IM transmitter consists of two branches, i.e., a main branch and a diversity branch with each branch intended for each of the two antennas. Both branches share the same set of input information bit stream, which is partitioned into G groups, and every group corresponds to Ω subcarriers where the total number subcarriers is $N = G \times \Omega$. The index modulation cannot be performed across all subcarriers due to complexity constraints, hence the total number of subcarriers can be partitioned into a number of small groups. And in each group there are B bits equally which are further split into two parts, one part P_1 bits input to the index selector module, and P_2 bits input to the ordinary M-*ary* ASK modulator. Then the SIB-TD index modulator generates two sets of signals to be input to IDCT modules in the main and diversity branch respectively based on the outputs from both the index selector and the ordinary modulator. As shown in Figure 2, for each of G groups, the SIB-TD index modulator selects K subcarriers out of Ω subcarriers according to the index selector, and these K active subcarriers transmit M-*ary* modulated symbols for the main branch, i.e., Antenna #1, and at the same time, the SIB-TD index modulator selects another K active subcarriers for the diversity branch, i.e., Antenna #2, and these K active subcarriers at the diversity branch are not overlapped with the K active subcarriers at the main branch, which implies a condition between K and the group size Ω: $2K \leq \Omega$. If $2K < \Omega$, then we can choose K out of $\Omega - K$ subcarriers at the second antenna. As such, the index of different choices will convey additional bits. Actually the term SIB-TD comes from this non-overlapped, or supplementary active subcarriers of the two branches. Note that $P_1 = \lfloor \log_2 C_\Omega^K \rfloor$, and $P_2 = K \log_2 M$.

The output of each index modulator in the main branch can be denoted by $\mathbf{x}_g^{(M)} = [x_{(g,0)}^{(M)}, x_{(g,1)}^{(M)}, \ldots, x_{(g,\Omega-1)}^{(M)}]^T$, $g = 0, 1, \ldots, G-1$, which consists of $\Omega - K$ zeros. Then the data symbol vector taken by the IDCT module at the main branch is expressed as:

$$\mathbf{X}^{(M)} = [X_0^{(M)}, X_1^{(M)}, \ldots, X_{N-1}^{(M)}] \tag{1}$$

$$= [(\mathbf{x}_0^{(M)})^T, (\mathbf{x}_1^{(M)})^T, \ldots, (\mathbf{x}_{G-1}^{(M)})^T]^T. \tag{2}$$

And similar expression for the diversity branch:

$$\mathbf{X}^{(D)} = [X_0^{(D)}, X_1^{(D)}, \ldots, X_{N-1}^{(D)}] \tag{3}$$

$$= [(\mathbf{x}_0^{(D)})^T, (\mathbf{x}_1^{(D)})^T, \ldots, (\mathbf{x}_{G-1}^{(D)})^T]^T. \tag{4}$$

Similar to the single antenna transmission case, we can assume the symmetric extended prefix and suffix have the same length v without loss of generality in order to have a circular property and further enable single-tap equalizer together with prefiltering at the receiver. Therefore, the total length of one transmitted OFDM symbol in the time domain is $L_1 = N + 2v$ and the transmitted signal vector of the main branches and diversity branch are represented by

$$\begin{cases} \mathbf{u}^{(M)} = \mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(M)} \\ \mathbf{u}^{(D)} = \mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(D)} \end{cases}, \tag{5}$$

where $\mathbf{D} \in \mathbb{R}^{N \times N}$ is a power normalised type-II DCT matrix [7], and $\mathbf{T}_{PS} = [\mathbf{J}_v, \mathbf{0}_{v\times(N-v)}; \mathbf{I}_N; \mathbf{0}_{v\times(N-v)}, \mathbf{J}_v]$ is the $L_1 \times N$ matrix that adds the suffix and prefix at both ends of a data symbol block, where $\mathbf{J}_v$ and $\mathbf{I}_N$are the reversal matrix and identity matrix respectively.

B. SIB-TD Receiver

The vector for a L-tap Rayleigh fading channel from the main branch and diversity branch to the receiver antenna can be denoted by $\mathbf{h}^{(M)} = [h_1^{(M)}, h_2^{(M)}, \ldots, h_L^{(M)}]$, $\mathbf{h}^{(D)} = [h_1^{(D)}, h_2^{(D)}, \ldots, h_L^{(D)}]$ respectively. Correspondingly, the multipath channels $\mathbf{H}^{(M)} \in \mathbb{R}^{L_1\times L_1}$, $\mathbf{H}^{(D)} \in \mathbb{R}^{L_1\times L_1}$ in a matrix form can be represented as Toeplitz matrices whose first row and first column being $[h_L^{(M)}, h_{L-1}^{(M)}, \ldots, h_1^{(M)}, \mathbf{0}_{1\times(L_1-L)}]$ and $[h_L^{(M)}, \mathbf{0}_{1\times(L_1-1)}]^T$

for the main branch, $[h_L^{(D)}, h_{L-1}^{(D)}, \ldots, h_1^{(D)}, \mathbf{0}_{1\times(L_1-L)}]$ and $[h_L^{(D)}, \mathbf{0}_{1\times(L_1-1)}]^T$ for the diversity branch respectively.

Then the signal at the receive antenna can be expressed as:

$$\begin{aligned}\mathbf{r} &= \mathbf{H}^{(M)}\mathbf{u}^{(M)} + \mathbf{H}^{(D)}\mathbf{u}^{(D)} + \mathbf{n} \\ &= \mathbf{H}^{(M)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(M)} + \mathbf{H}^{(D)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(D)} + \mathbf{n}.\end{aligned} \tag{6}$$

where $\mathbf{n}$ is additive noise with a covariance matrix $\mathbf{R}_n = \sigma^2\mathbf{I}_{L1}$. As shown in Figure (1), prefiltering for SIB-TD is performed for the main and diversity branch respectively which enables a single-tap equalizer at the receiver together with symmetric prefix and suffix extension, similar to the single transmit antenna case for each branch, where the matrix form of the pre-filter for the main branch $\mathbf{P}^{(M)} \in \mathbb{R}^{L_1\times L_1}$ is represented by a Toeplitz matrix whose first row and column is $[h_L^{(M)}, \mathbf{0}_{1\times(L_1-1)}]$ and $[h_L^{(M)}, h_{L-1}^{(M)}, \ldots, h_1^{(M)}, \mathbf{0}_{1\times(L_1-L)}]^T$, respectively, where the matrix form of the pre-filter for the diversity branch $\mathbf{P}^{(D)} \in \mathbb{R}^{L_1\times L_1}$ is also represented by a Toeplitz matrix whose first row and column is $[h_L^{(D)}, \mathbf{0}_{1\times(L_1-1)}]$ and $[h_L^{(D)}, h_{L-1}^{(D)}, \ldots, h_1^{(D)}, \mathbf{0}_{1\times(L_1-L)}]^T$, respectively.

Thus for each branch, the received frequency domain vector can be rewritten as the following equation after performing sequentially pre-filtering, DCT demultiplexing and removing guard sequence:

$$\begin{cases}\mathbf{z}^{(M)} = \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(M)}\mathbf{r} \\ \mathbf{z}^{(D)} = \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(D)}\mathbf{r}.\end{cases} \tag{7}$$

The matrix $\mathbf{R}_{PS} = [\mathbf{0}_{N\times v}, \mathbf{I}_N, \mathbf{0}_{N\times v}]$ represents the removal of the guard sequence. $\mathbf{n} \in \mathbb{R}^{L_1\times 1}$ is the additive white Gaussian noise (AWGN) noise vector and each of its elements follows Gaussian distribution with mean 0 and variance N_0. Then the above equation can be further expanded as

$$\begin{cases}\mathbf{z}^{(M)} = \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(M)}(\mathbf{H}^{(M)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(M)} \\ \quad + \mathbf{H}^{(D)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(D)} + \mathbf{n}) \\ \quad = \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(M)}\mathbf{H}^{(M)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(M)} \\ \quad + \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(M)}\mathbf{H}^{(D)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(D)} + \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(M)}\mathbf{n} \\ \mathbf{z}^{(D)} = \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(D)}(\mathbf{H}^{(M)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(M)} \\ \quad + \mathbf{H}^{(D)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(D)} + \mathbf{n}) \\ \quad = \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(D)}\mathbf{H}^{(D)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(D)} \\ \quad + \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(D)}\mathbf{H}^{(M)}\mathbf{T}_{PS}\mathbf{D}^H\mathbf{x}^{(M)} + \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(D)}\mathbf{n}\end{cases}. \tag{8}$$

The selection of $\mathbf{T}_{PS}$ and $\mathbf{R}_{PS}$ diagonalizes the following parts:

- $\mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(M)}\mathbf{H}^{(M)}\mathbf{T}_{PS}\mathbf{D}^H \longrightarrow \mathbf{H}_{eff}^{(M)} \in \mathbb{R}^{N\times N}$ and the diagonal elements can be expressed as $diag(\mathbf{H}_{eff}^{(M)}) = [H_0^{(M)}, H_1^{(M)}, \ldots, H_{N-1}^{(M)}]$
- $\mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(D)}\mathbf{H}^{(D)}\mathbf{T}_{PS}\mathbf{D}^H \longrightarrow \mathbf{H}_{eff}^{(D)} \in \mathbb{R}^{N\times N}$ and the diagonal elements can be expressed as $diag(\mathbf{H}_{eff}^{(D)}) = [H_0^{(D)}, H_1^{(D)}, \ldots, H_{N-1}^{(D)}]$

For the other items in the equation, $\mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(M)}\mathbf{H}^{(D)}\mathbf{T}_{PS}\mathbf{D}^H \longrightarrow \mathbf{O}^{(M)} \in \mathbb{R}^{N\times N}$ represents the cross term from the diversity branch to the main branch, and $\mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(D)}\mathbf{H}^{(M)}\mathbf{T}_{PS}\mathbf{D}^H \longrightarrow \mathbf{O}^{(D)} \in \mathbb{R}^{N\times N}$ represents the cross term from the main branch to the diversity branch. Assuming the radio channels from two branches are independent and have the same power delay profile (PDP), these two cross terms cannot be diagonalized as in the case of the single antenna transmission.

Eq. (8) can be rewritten as

$$\begin{cases}\mathbf{z}^{(M)} = \mathbf{H}_{eff}^{(M)}\mathbf{x}^{(M)} + \mathbf{O}^{(M)}\mathbf{x}^{(D)} + \mathbf{G}_{eff}^{(M)}\mathbf{n} \\ \mathbf{z}^{(D)} = \mathbf{H}_{eff}^{(D)}\mathbf{x}^{(D)} + \mathbf{O}^{(D)}\mathbf{x}^{(M)} + \mathbf{G}_{eff}^{(D)}\mathbf{n},\end{cases} \tag{9}$$

where $\mathbf{G}_{eff}^{(M)} = \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(M)}$,$\mathbf{G}_{eff}^{(D)} = \mathbf{D}\mathbf{R}_{PS}\mathbf{P}^{(D)}$ represents the effective noise correlation matrix. The equation can be reformed as

$$\begin{pmatrix}\mathbf{z}^{(M)} \\ \mathbf{z}^{(D)}\end{pmatrix} = \begin{pmatrix}\mathbf{H}_{eff}^{(M)} & \mathbf{O}^{(M)} \\ \mathbf{O}^{(D)} & \mathbf{H}_{eff}^{(D)}\end{pmatrix}\begin{pmatrix}\mathbf{x}^{(M)} \\ \mathbf{x}^{(D)}\end{pmatrix} + \begin{pmatrix}\mathbf{G}_{eff}^{(M)} \\ \mathbf{G}_{eff}^{(D)}\end{pmatrix}\mathbf{n}. \tag{10}$$

Let

$$\begin{cases}\mathbf{z} = \begin{pmatrix}\mathbf{z}^{(M)} \\ \mathbf{z}^{(D)}\end{pmatrix} \\ \mathbf{x} = \begin{pmatrix}\mathbf{x}^{(M)} \\ \mathbf{x}^{(D)}\end{pmatrix} \\ \mathbf{H} = \begin{pmatrix}\mathbf{H}_{eff}^{(M)} & \mathbf{O}^{(M)} \\ \mathbf{O}^{(D)} & \mathbf{H}_{eff}^{(D)}\end{pmatrix} \\ \tilde{\mathbf{n}} = \begin{pmatrix}\mathbf{G}_{eff}^{(M)} \\ \mathbf{G}_{eff}^{(D)}\end{pmatrix}\mathbf{n}\end{cases}, \tag{11}$$

Eq. (10) can thus be expressed as

$$\mathbf{z} = \mathbf{H}\mathbf{x} + \tilde{\mathbf{n}}. \tag{12}$$

The group-wise form of Eq. (12) can be expressed for the g-th group in the main and diversity branch as:

$$\begin{cases}\mathbf{z}_g^{(M)} = \mathbf{H}_g^{(M)}\mathbf{x}_g^{(M)} + \mathbf{O}_{g,g}^{(M)}\mathbf{x}_g^{(D)} \\ \qquad + \sum\limits_{q=1, q\neq g}^{G} \mathbf{O}_{g,q}^{(M)}\mathbf{x}_q^{(D)} + \mathbf{G}_{eff,g}^{(M)}\mathbf{n} \\ \mathbf{z}_g^{(D)} = \mathbf{H}_g^{(D)}\mathbf{x}_g^{(D)} + \mathbf{O}_{g,g}^{(D)}\mathbf{x}_g^{(M)} \\ \qquad + \sum\limits_{q=1, q\neq g}^{G} \mathbf{O}_{g,q}^{(D)}\mathbf{x}_q^{(M)} + \mathbf{G}_{eff,g}^{(D)}\mathbf{n} \qquad g = 1, \ldots, G,\end{cases} \tag{13}$$

where $\mathbf{z}_g^{(M)}$ is the g-th group of the main branch DCT output:

$$\mathbf{z}_g^{(M)} = \begin{pmatrix} z_{(g-1)\Omega+1}^{(M)} \\ \vdots \\ z_{(g-1)\Omega+\Omega}^{(M)}\end{pmatrix}_{\Omega\times 1}, \tag{14}$$

$\mathbf{z}_g^{(D)}$ is the g-th group of the diversity branch DCT output:

$$\mathbf{z}_g^{(D)} = \begin{pmatrix} z_{(g-1)\Omega+1}^{(D)} \\ \vdots \\ \mathbf{z}_{(g-1)\Omega+\Omega}^{(D)}\end{pmatrix}_{\Omega\times 1}, \tag{15}$$

$\mathbf{H}_g^{(M)}$ is the g-th sub-diagonal matrix of the effective channel $\mathbf{H}_{eff}^{(M)}$:

$$\mathbf{H}_g^{(M)} = \begin{pmatrix} H_{(g-1)\Omega+1}^{(M)} & 0 & \ldots & 0 \\ \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \ldots & H_{(g-1)\Omega+\Omega}^{(M)}\end{pmatrix}_{\Omega\times\Omega}, \tag{16}$$

$\mathbf{H}_g^{(D)}$ is the g-th sub-diagonal matrix of the effective channel $\mathbf{H}_{eff}^{(D)}$:

$$\mathbf{H}_g^{(D)} = \begin{pmatrix} H_{(g-1)\Omega+1}^{(D)} & 0 & \dots & 0 \\ \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \dots & H_{(g-1)\Omega+\Omega}^{(D)} \end{pmatrix}_{\Omega\times\Omega}, \tag{17}$$

$\mathbf{O}_g^{(M)}$ is the sub-matrix of $\mathbf{O}^{(M)}$ with rows from $(g-1)\Omega+1$ to $g\Omega$:

$$\mathbf{O}_g^{(M)} = \begin{pmatrix} O_{(g-1)\Omega+1,1}^{(M)} & O_{(g-1)\Omega+1,2}^{(M)} & \cdots & O_{(g-1)\Omega+1,N}^{(M)} \\ \vdots & \vdots & \vdots & \vdots \\ O_{(g-1)\Omega+\Omega,1}^{(M)} & O_{(g-1)\Omega+\Omega,2}^{(M)} & \cdots & O_{(g-1)\Omega+\Omega,N}^{(M)} \end{pmatrix}_{\Omega\times N} \tag{18}$$

Furthermore, $\mathbf{O}_g^{(M)}$ is partitioned into G square matrices labelled as $\mathbf{O}_{g,q}^{(M)}$, $q = 1, ..., G$, and $\sum_{q=1, q\neq g}^{G} \mathbf{O}_{g,q}^{(M)} \mathbf{x}_q^{(D)}$ is the inter-branch inter-group interference in the main branch, and $\sum_{q=1, q\neq g}^{G} \mathbf{O}_q^{(D)} \mathbf{x}_q^{(M)}$ is the inter-branch inter-group interference in the diversity branch. Let $\mathbf{z}_g = [z_{(g-1)\Omega+1}^{(M)}, \cdots z_{(g-1)\Omega+\Omega}^{(M)}, z_{(g-1)\Omega+1}^{(D)} \cdots \mathbf{z}_{(g-1)\Omega+\Omega}^{(D)}]_{2\Omega\times 1}^T$, and

$$\begin{cases} \mathbf{H}_g = \begin{pmatrix} \mathbf{H}_g^{(M)} & \mathbf{O}_{g,g}^{(M)} \\ \mathbf{O}_{g,g}^{(D)} & \mathbf{H}_g^{(D)} \end{pmatrix}_{2\Omega\times 2\Omega} \\ \hat{\mathbf{n}}_g = \begin{pmatrix} \sum_{q=1, q\neq g}^{G} \mathbf{O}_{g,q}^{(M)} \mathbf{x}_q^{(D)} + \mathbf{G}_{eff,g}^{(M)} \mathbf{n} \\ \sum_{q=1, q\neq g}^{G} \mathbf{O}_q^{(D)} \mathbf{x}_q^{(M)} + \mathbf{G}_{eff,g}^{(D)} \mathbf{n} \end{pmatrix} \end{cases}, \tag{19}$$

And $\mathbf{x}_g^{(M)}$ represents the output symbols after SIB-TD index modulator at the main branch, and all possible $\mathbf{x}_g^{(M)}$ constitutes the main branch codebook $\Lambda^{(M)}$, and similarly $\mathbf{x}_g^{(D)}$ and $\Lambda^{(D)}$ for the diversity branch. And let $\mathbf{x}_g = [\mathbf{x}_g^{(M)} \quad \mathbf{x}_g^{(D)}]^T$ then the group-wise detection problem is denoted as:

$$\mathbf{z}_g = \mathbf{H}_g \mathbf{x}_g + \hat{\mathbf{n}}_g. \tag{20}$$

Upon introducing composite variables, codebook and matrices as shown above, the joint detection and demodulation for the receiver of EDCT-OFDM-IM with SIB-TD turns into a small-scaled group-wise classical detection problem. This implies that a composite codebook $\Lambda = [\Lambda^{(M)} \quad \Lambda^{(D)}]^T$ is used in the group-wise detection. By cancelling the cross terms in $\hat{\mathbf{n}_g}$, then a maximum-likelihood (ML) detection can be further applied to Eq. (20) similar to the single transmit antenna case. And we can formulate the optimum maximum likelihood detector for the g-th group accordingly. If denoting by $z_{\epsilon+g\Omega} = z_{(g,\epsilon)}$, $H_{\epsilon+g\Omega}^{(M)} = H_{(g,\epsilon)}^{(M)}$, $H_{\epsilon+g\Omega}^{(D)} = H_{(g,\epsilon)}^{(D)}$, $O_{\epsilon+g\Omega}^{(M)} = O_{(g,\epsilon)}^{(M)}$, $O_{\epsilon+g\Omega}^{(D)} = O_{(g,\epsilon)}^{(D)}$, $v_{\epsilon+g\Omega}^{(M)} = v_{(g,\epsilon)}^{(M)}$, and $v_{\epsilon+g\Omega}^{(D)} = v_{(g,\epsilon)}^{(D)}$, where ϵ is the offset in terms of number of subcarriers to the first subcarrier within group g, i.e., $\epsilon \in 0, 1, 2, ..., \Omega-1$, then the estimate of the transmit symbols in the g-th group can be expressed as:

$$\mathbf{x}_t = \arg\min_{\mathbf{x}_g \in \Lambda} \sum_{\epsilon=0}^{\Omega-1} \Big(\frac{|z_{(g,\epsilon)} - H_{(g,\epsilon)}^{(M)} x_{(g,\epsilon)}^{(M)} - O_{(g,\epsilon)}^{(M)} x_{(g,\epsilon)}^{(D)}|^2}{v_{(g,\epsilon)}^{(M)}} \tag{21}$$

$$+ \frac{|z_{(g+G,\epsilon)} - H_{(g,\epsilon)}^{(D)} x_{(g,\epsilon)}^{(D)} - O_{(g,\epsilon)}^{(D)} x_{(g,\epsilon)}^{(M)}|^2}{v_{(g,\epsilon)}^{(D)}} \Big). \tag{22}$$

TABLE I
SETUP FOR SIB-TD AND DFT-OFDM-IM WITH ALAMOUTI CODE

Parameter	EDCT-OFDM-IM with SIB-TD	DFT-OFDM-IM with Alamouti code
Spectrum Efficiency	1bps/Hz, 1.5 bps/Hz	1.5 bps/Hz
DCT size (N)	256	128
suffix length (v)	16	-
prefix length (v)	16	
channel model	Rayleigh with L taps	
tap number (L)	10	
Power Delay Profile	uniform distribution	
modulation scheme	BPSK, 4ASK	QPSK
group size (Ω)	4	
active SC number (K)	2	
Index bits in each group(P_1)	2	2
· Modulation bits in each group(P_2)	2, 4	4

III. NUMERICAL RESULTS

In this section, simulation results are carried out over Rayleigh fading channel with both imperfect and ideal channel state information. The simulation setup is illustrated in detail in Table I, where the spectrum efficiency (SE) is calculated as

$$SE = \frac{\lfloor \log_2 C_\Omega^K \rfloor + K \log_2 M}{\Omega} \tag{23}$$

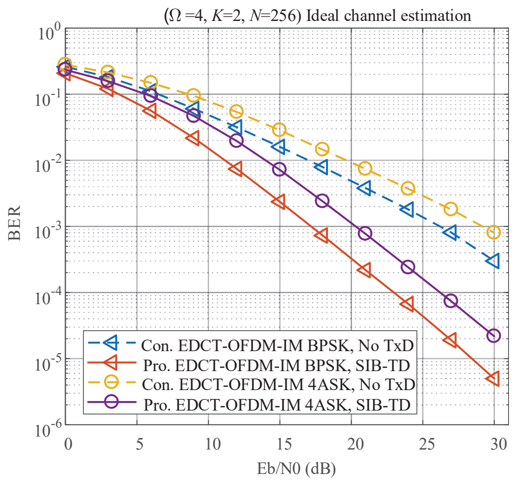

Fig. 3. Performance comparison between SIB-TD and no TxD under ideal channel estimation.

Moreover, we also simulate the impacts of imperfect channel estimation, which is modelled by adding noise to the ideal channel knowledge as shown in the below equation:

$$\tilde{H}_g = H_g + H_e^{(g)} \tag{24}$$

where $H_e^{(g)} \in CN(0, \sigma_e^2)$ and σ_e^2 corresponds to accuracy of pragmatic channel estimates.

Fig. 3 and 4 compare the performance of the SIB-TD based EDCT-OFDM-IM system to that of conventional EDCT-OFDM-IM for BPSK and 4ASK under ideal and imperfect channel estimation respectively. A 5dB gain of the proposed SIB-TD over the single transmission scheme can be observed for 4ASK and BPSK at the target BER 10^{-2} even under the

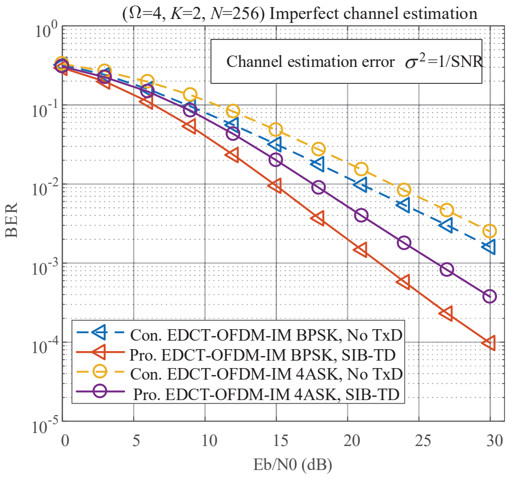

Fig. 4. Performance comparison between SIB-TD and no TxD under imperfect channel estimation.

TABLE II
SETUP FOR DFT-OFDM WITH ALAMOUTI CODE

Parameter	DFT-OFDM, Alamouti code
Spectrum Efficiency	1.0 bps/Hz
DFT size (N)	128
prefix length (v)	16
channel model	Rayleigh with L taps
tap number (L)	10
Power Delay Profile	uniform distribution

imperfect channel estimates, while a larger gain for the ideal channel knowledge. It is worth mentioning that the performance comparison has been conducted under the practical assumption that channel estimation error variance is inversely proportional to SNR, manifesting the practicality and viability of the proposed scheme.

To provide further insights, Fig. 5 compares the performance of the proposed scheme to that of Alamouti code [12] based DFT-OFDM and DFT-OFDM-IM schemes under the same spectrum efficiency of 1 bps/Hz and 1.5 bps/Hz, respectively. Ideal channel state information is assumed and the setups are shown in Table I and II in details. It is shown that the proposed DCT-OFDM based SIB-TD scheme is capable of providing performance gains over the Alamouti code based DFT-OFDM and DFT-OFDM-IM counterparts, especially for the Alamouti code based DFT-OFDM scheme.

IV. CONCLUSION

In this paper, a novel transmit diversity method with SIM concept has been proposed for an EDCT-OFDM-IM system, where the active sub-carriers indices of the two antennas are different at the common time slots. By designing different codebooks of the two antennas, the non-activated subcarrier of each antenna can transmit the same symbol of the other antenna to obtain diversity gain. A maximum likelihood group-wise receiver was designed by jointly detecting the symbols of the two antennas. Simulation results showed that the proposed scheme can achieve significant performance gains over both its conventional EDCT-OFDM-IM counterpart and the DFT-OFDM with Alamouti code, no matter with or without index

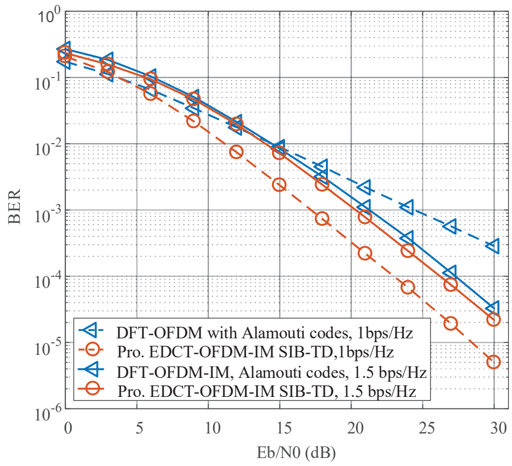

Fig. 5. Performance comparison of the proposed scheme to that of Alamouti code based DFT-OFDM schemes.

modulation.

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