Equiprobable Subcarrier Activation Method for OFDM With Index Modulation

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***Abstract*— Orthogonal frequency division multiplexing with index modulation (OFDM-IM) conveys additional information bits via the indices of active subcarriers. Consequently, the determination of the active subcarriers according to the incoming bits arises as a challenging problem. The existing solution resorts to the lexicographic ordering, which leads to a low implementation complexity but an unequal subcarrier activation probability. This letter proposes a distinct solution that allows the equiprobable activation of all OFDM-IM subcarriers with comparable implementation complexity. The signal-to-noise ratio gain achieved by the proposed solution is also analyzed. Computer simulations reveal the advantages of the proposed solution.**

***Index Terms*— OFDM, index modulation, subcarrier activation, combinatorial theory, cyclic shift.**

I. Introduction

THE potential of spatial modulation (SM) has been widely explored in the literature [1], [2]. Recently, the idea of SM is transplanted to OFDM subcarriers. Orthogonal frequency division multiplexing with index modulation (OFDM-IM) is thereby proposed, which conveys additional information through the indices of the active OFDM subcarriers [3]. Analyses from the perspective of bit error rate (BER) and achievable rate reveal that OFDM-IM can outperform traditional OFDM [3], [4]. The advantages of OFDM-IM have motivated a great deal of follow-up research more recently. In [5], the transmit diversity of OFDM-IM is improved from unity to two considering the coordinate interleaved orthogonal designs. Two new OFDM-IM systems are proposed in [6], including OFDM with generalized index modulation (OFDM-GIM) that allows a higher number of possible subcarrier activation patterns (SAPs) to be encoded, and OFDM with in-phase/quadrature index modulation (OFDM-I/Q-IM) that expands the index domain to include both the in-phase and quadrature dimensions [7]. In [8], OFDM-IM is further combined with the multi-input-multi-output (MIMO) transmission technique to form MIMO-OFDM-IM. More related studies on OFDM-IM can be found in the recent survey of [9].

In plain OFDM-IM and all of its variants mentioned above, the SAPs are related with the incoming bits according to the combinatorial theory to circumvent the possible storage problem as their size gets larger [3]. This one-to-one mapping method, which generates a lexicographic order, is relatively efficient; however, it does not necessarily ensure that each subcarrier is activated with an equal probability. The nonuniform subcarrier activation leads to the unequal protection of the transmitted bits, preventing the OFDM-IM related techniques from achieving their ultimate error performance. This letter aims to solve the aforementioned problems. Specifically, we propose a new mapping method called equiprobable subcarrier activation (ESA), which allows the activation of all subcarriers as equiprobably as possible. To apply the ESA method, the storage of a very small table is required at both the transmitter and the receiver. Based on the basic OFDM-IM framework, we further analyze the signal-to-noise ratio (SNR) gain of the ESA method over the existing one. Computer simulation results validate the analysis and show that ESA can provide up to 1.9dB SNR gain.[1]

II. System Model

Consider an OFDM system, which consists of L subcarriers. To implement OFDM-IM, the total subcarriers are divided into G groups, each containing $N = L/G$ subcarriers [3]. Here, either localized or interleaved manner can be employed for subcarrier grouping [10]. Without loss of generality, let us focus on the first group and index the subcarriers belonging to it from $\{1, \ldots, N\}$.

At the transmitter, a number of p incoming bits are fed into the first group and further divided into two parts. The first part, comprised of $p_1 = \lfloor \log_2 C(N, k) \rfloor$ bits, is used to activate k out of N subcarriers. The second part, comprised of $p_2 = k \log_2 M$ bits, is used to modulate k M-ary data symbols. Note that we may also merely rely on the subcarrier indices to convey information, in which case the activated subcarriers transmit 1s. In this letter, such special case is referred to as OFDM with subcarrier shift keying (OFDM-SSK).

Assume that at a given time slot, the i-th SAP is selected by p_1 bits, where $i \in \{1, \ldots, 2^{p_1}\}$. The indices of the active

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[1] Upper and lower case boldface letters denote matrices and column vectors, respectively. Superscripts ^{T}, ^{H}, and $^{-1}$ stand for transpose, Hermitian transpose, and inversion operations, respectively. rank$\{\mathbf{X}\}$ returns the rank of matrix $\mathbf{X}$ and diag$\{\mathbf{x}\}$ creates a diagonal matrix whose diagonal elements are $\mathbf{x}$. The probability density function and the probability of an event are denoted by $f(\cdot)$ and $\Pr(\cdot)$, respectively. $\lfloor\cdot\rfloor$, $\lceil\cdot\rceil$, and $< \cdot >_N$ denote the floor, the ceiling, and the modulo N operations, respectively. $C(\cdot,\cdot)$ denotes the binomial coefficient. $\mathcal{X}$ denotes a normalized signal set of cardinality M.

subcarriers are thus included in the set $\boldsymbol{\beta}^{(i)}$, where $\boldsymbol{\beta}^{(i)} = \{\beta_1^{(i)}, \ldots, \beta_k^{(i)}\}$. The k modulated symbols selected by p_2 bits are denoted by $\{s_1, \ldots, s_k\}$, where $s_j \in \mathcal{X}$, $j = 1, \ldots, k$. Let $\mathbf{h} = [h_1, \ldots, h_N]^T$ represent the channel frequency response (CFR) vector of the first subcarrier group, whose elements are complex Gaussian distributed with zero mean and unit variance. Assuming perfect synchronization, the received signal at the l-th subcarrier can be written as

$$y_l = h_l x_l + n_l, \quad l = 1, \ldots, N \tag{1}$$

where $x_l = 0$ for $l \notin \boldsymbol{\beta}^{(i)}$, $x_{\beta_j^{(i)}} = s_j$ for $j = 1, \ldots, k$, and n_l denotes the noise term of variance N_0 at the l-th subcarrier.

From (1), the optimal maximum-likelihood (ML) detector can be readily derived as

$$\{\hat{i}, \hat{s}_1, \ldots, \hat{s}_k\} = \underset{\{i, s_1, \ldots, s_k\}}{\arg\min}\ \Phi \tag{2}$$

where $\Phi = \sum_{j=1}^{k} |y_{\beta_j^{(i)}} - h_{\beta_j^{(i)}} s_j|^2 + \sum_{j=1}^{N-k} |y_{\bar{\beta}_j^{(i)}}|^2$ and $\bar{\boldsymbol{\beta}}^{(i)}$ is the complement of $\boldsymbol{\beta}^{(i)}$. Alternative to the ML detection, the log-likelihood ratio (LLR) detector can be also used to decode the transmitted information while exhibiting much lower computational complexity [3]. It operates by first calculating the LLR values associated with all subcarriers from

$$LLR_l = \log \frac{\Pr(\mathrm{A}_l | y_l)}{\Pr(\bar{\mathrm{A}}_l | y_l)} = \log \frac{\Pr(\mathrm{A}_l)}{\Pr(\bar{\mathrm{A}}_l)} + \log \frac{f(y_l | \mathrm{A}_l)}{f(y_l | \bar{\mathrm{A}}_l)} \tag{3}$$

with $l = 1, \ldots, N$, where A_l represents the event that the l-th subcarrier is active while $\bar{\mathrm{A}}_l$ means just the opposite, and

$$\log \frac{f(y_l | \mathrm{A}_l)}{f(y_l | \bar{\mathrm{A}}_l)} = \frac{|y_l|^2}{N_0} + \log \frac{1}{M} \sum_{\tilde{s} \in \mathcal{X}} \frac{|y_l - h_l \tilde{s}|^2}{N_0}. \tag{4}$$

Then, it sorts all LLR values in descending order and picks out the first k ones, whose corresponding subcarriers are simply regarded to be active. Finally, the symbols carried on the determined active subcarriers are demodulated independently.

III. Proposed ESA Method

This letter focuses on the mapping of p_1 bits to the SAPs, which is called unranking, and the reverse operation, which is called ranking.

A. Combinatorial Method Revisited

The existing solution, which is called combinatorial method, generates lexicographically ordered sequences [3]. Assume that the elements of $\boldsymbol{\beta}^{(i)}$ are sorted in a strictly increasing order. Then, according to the combinatorial theory, we can build a one-to-one mapping which relates a natural number to k-combinations:

$$i = C(\beta_k^{(i)} - 1, k) + \cdots + C(\beta_1^{(i)} - 1, 1) + 1 \tag{5}$$

where $i \in \{1, \ldots, 2^{p_1}\}$. To implement ranking, we first substitute $\boldsymbol{\beta}^{(i)}$ into (5) and then map the integer i to p_1 bits. For unranking, we first map the p_1 bits to an integer i and then determine $\beta_k^{(i)}, \ldots, \beta_1^{(i)}$ successively. Specifically, we choose $\beta_k^{(i)}$ to be the maximal integer that satisfies $i \geq C(\beta_k^{(i)} - 1, k) + 1$ and $\beta_{k-1}^{(i)}$ to be the maximal integer satisfying $i \geq C(\beta_k^{(i)} - 1, k) + C(\beta_{k-1}^{(i)} - 1, k-1) + 1$ and so on. An example for $N = 8$ and $k = 2$ is given in Table I.

TABLE I
COMBINATORIAL METHOD VS. ESA METHOD ($N = 8$, $k = 2$)

Bits	Comb.	ESA	Bits	Comb.	ESA
[0 0 0 0]	$\{1,2\}$	$\{1,2\}$	[1 0 0 0]	$\{3,5\}$	$\{1,3\}$
[0 0 0 1]	$\{1,3\}$	$\{2,3\}$	[1 0 0 1]	$\{4,5\}$	$\{2,4\}$
[0 0 1 0]	$\{2,3\}$	$\{3,4\}$	[1 0 1 0]	$\{1,6\}$	$\{3,5\}$
[0 0 1 1]	$\{1,4\}$	$\{4,5\}$	[1 0 1 1]	$\{2,6\}$	$\{4,6\}$
[0 1 0 0]	$\{2,4\}$	$\{5,6\}$	[1 1 0 0]	$\{3,6\}$	$\{5,7\}$
[0 1 0 1]	$\{3,4\}$	$\{6,7\}$	[1 1 0 1]	$\{4,6\}$	$\{6,8\}$
[0 1 1 0]	$\{1,5\}$	$\{7,8\}$	[1 1 1 0]	$\{5,6\}$	$\{7,1\}$
[0 1 1 1]	$\{2,5\}$	$\{8,1\}$	[1 1 1 1]	$\{1,7\}$	$\{8,2\}$

B. Principle of ESA

The combinatorial method has some deficiencies. This can be observed from the example in Table I by examining the activation probabilities of all subcarriers, which are $\Pr(\mathrm{A}_1) = 3/8$, $\Pr(\mathrm{A}_2) = \Pr(\mathrm{A}_3) = \Pr(\mathrm{A}_4) = \Pr(\mathrm{A}_5) = \Pr(\mathrm{A}_6) = 5/16$, $\Pr(\mathrm{A}_7) = 1/16$, and $\Pr(\mathrm{A}_8) = 0$. This indicates that the lexicographically ordered subcarriers may be given totally different activation probabilities. However, since all subcarriers experience the same fading statistically, the mismatch will lead to a performance degradation to the system with ESA. This degradation as well as the identity of (3), which shows that the LLR detector can be further simplified for the case of the same prior likelihood value, motivates us to propose ESA.

The idea of ESA is to generate the SAPs in a cyclic manner. For better understanding, let us give an example. First, assume a legal SAP $\boldsymbol{\beta}^{(b_1)} = \{1, 2, \ldots, k\}$. Then, by performing cyclic shift on each element of $\boldsymbol{\beta}^{(b_1)}$ based on $\{1, 2, \ldots, N\}$, which is called column cyclic shift in this letter, we can obtain $N-1$ new patterns, which are $\{2, 3, \ldots, k+1\}$, $\{3, 4, \ldots, k+2\}, \ldots, \{N, 1, \ldots, k-1\}$. It is clear from the above N patterns that each subcarrier is activated with an equal probability, namely k/N. Similarly, performing column cyclic shifts to $\boldsymbol{\beta}^{(b_2)} = \{1, 2, \ldots, k-1, k+1\}$ for N times allows us to obtain another N different patterns, which also satisfy the ESA requirement. This process continues until $2^{\lfloor \log_2 C(N,k) \rfloor}$ different patterns are collected. It is worth noting that the above example only shows how to perform column cyclic shifts and ignores the problem of how to determine the basic patterns such as $\boldsymbol{\beta}^{(b_1)}$ and $\boldsymbol{\beta}^{(b_2)}$. The solution to this problem is given in the following. To begin with, let us define a so-called adjacent subcarrier distance vector (ASDV) as $\mathbf{d}^{(b)} = [d_1^{(b)}, \ldots, d_k^{(b)}]^T$ with $d_j^{(b)} = \langle \beta_{\langle j+1 \rangle_k}^{(b)} - \beta_j^{(b)} \rangle_N$, where $b \in \{1, \ldots, 2^{p_1}\}$ and $j \in \{1, \ldots, k\}$. For instance, in the above example $\mathbf{d}^{(b_1)} = [1, \ldots, 1, N+1-k]^T$, $\mathbf{d}^{(b_2)} = [1, \ldots, 1, 2, N-k]^T$, and all column cyclic shifts generated by $\boldsymbol{\beta}^{(b_1)}$ and $\boldsymbol{\beta}^{(b_2)}$ share the same $\mathbf{d}^{(b_1)}$ and $\mathbf{d}^{(b_2)}$, respectively. By definition, we can now conclude that all ASDVs associated with the basic patterns have the same l_1 norm, which equals N. As a double check, in the above example $\|\mathbf{d}^{(b_1)}\|_1 = \|\mathbf{d}^{(b_2)}\|_1 = N$. Therefore, based on this property, there will be $C(N-1, k-1)$ candidates for the basic patterns given N and k. However, we should be careful since those candidates include a couple of ones that are cyclicly shifted with each other, which will produce the same group of patterns. For example, for two ASDVs $[e, f, g]^T$ and $[g, e, f]^T$, where e, f, g are positive integers with $e + f + g = N$, their corresponding basic patterns are

Algorithm 1 ASDV Generation

1: Initialization: $\mathbf{d}^{(1)} = [1, \ldots, 1, N-k+1]^T$, $I_1 = N$, $b = 1$, $j = k - 1$
2: **while** $d_1^{(b)} \leq N/k$
3: $\quad b = b + 1$
4: $\quad d^{(b)}(1 : k-2) = d^{(b-1)}(1 : k-2)$
5: $\quad d_j^{(b)} = d_j^{(b-1)} + 1$
6: $\quad$ **for** $l = j + 1 : k - 1$, $d_l^{(b)} = 1$ **end for**
7: $\quad d_k^{(b)} = N - \sum_{j=1}^{k-1} d_j^{(b)}$
8: $\quad$ **if** $d_k^{(b)} < 1$, $j = j - 1$, Go to line 5 **end if**
9: $\quad$ Store $\mathbf{d}^{(b)}$ and calculate I_b if $\mathbf{d}^{(b)}$ is not a cyclic shift of $\mathbf{d}^{(1)}, \ldots, \mathbf{d}^{(b-1)}$.
10: **end while**
11: Output: $\mathbf{d}^{(1)}, \ldots, \mathbf{d}^{(\lceil T \rceil)}$, $I_1, \ldots, I_{\lceil T \rceil}$

Algorithm 2 Unranking

1: Initialization: Transfer p_1 bits into integer i, $b = 1$, $i' = i$
2: **while** $i' > I_b$, $i' = i' - I_b$, $b = b + 1$ **end while**
3: $\beta_1^{(i)} = i'$
4: **for** $j = 2$ to k
5: $\quad \beta_j^{(i)} = \beta_{j-1}^{(i)} + d_{j-1}^{(b)}$
6: $\quad$ **if** $\beta_j^{(i)} > N$, $\beta_j^{(i)} = \beta_j^{(i)} - N$ **end if**
7: **end for**
8: Output: $\boldsymbol{\beta}^{(i)}$

$\{1, 1+e, 1+e+f\}$ and $\{1, 1+g, 1+g+e\}$, respectively, where the latter is the g-times column cyclic shifted version of the former. Since there are at most $k - 1$ distinct cyclic shifts for a pattern, the number of valid candidates for the basic patterns approximates to $T = C(N-1, k-1)/k$.

From above, we summarize the generation of ASDVs in Algorithm 1. Since all ASDVs have the same l_1 norm of N, we only consider the case that the first element of any ASDV is no larger than N/k, as reflected in line 2. With all ASDVs and the numbers of column cyclic shifts they can generate, denoted by $I_1, \ldots, I_{\lceil T \rceil}$, we can therefore proceed to perform ranking and unranking.

1) Unranking: To determine the SAP according to the corrsponding p_1 bits, we first figure out which basic pattern it belongs to and then decide how many column cyclic shifts it should perform. This process is detailed in Algorithm 2, where the "while" loop determines the basic pattern, the variable i' stores the number of column cyclic shifts, and the "for" loop performs the column cyclic shifts.

2) Ranking: Given a SAP $\boldsymbol{\beta}^{(\hat{i})}$, the basic pattern it belongs to can be directly determined according to the definition of ASDV and the number of column cyclic shifts it results in can be also directly inferred from its first subcarrier index. The entire process is presented in Algorithm 3.

According to the aforementioned principle, we also present the results of the ESA method for $N = 8$ and $k = 2$ in Table I. It is observed that all subcarriers are activated with an equal probability of $1/4$ for this example.

C. Complexity Analysis

From above, we see that both the ranking and unranking operations involve very simple mathematical operations and

Algorithm 3 Ranking

1: Initialization: $b' = 1, \upsilon = 0$
2: Calculate $\mathbf{d}^{(b)}$ from $\boldsymbol{\beta}^{(\hat{i})}$ according to ASDV definition.
3: **while** $\mathbf{d}^{(b)} \neq \mathbf{d}^{(b')}$, $\upsilon = \upsilon + I_{b'}$, $b' = b' + 1$ **end while**
4: $\upsilon = \upsilon + \beta_1^{(\hat{i})}$
5: Output: The binary representation of υ

the complexity mainly arises from the generation of ASDVs. Fortunately, the ASDVs need to be prepared only once prior to the signal transmission and reception provided that they can be saved by both the transmitter and the receiver. Since the total ASDVs just include roughly $C(N-1, k-1)$ indices, this storage requirement is easy to satisfy in today's systems.

D. SNR Gain Analysis

Assume two transmitted signal vectors $\mathbf{x} = [x_1, \ldots, x_N]^T$ and $\hat{\mathbf{x}} = [\hat{x}_1, \ldots, \hat{x}_N]^T$, and that the number of bits carried by them in difference is $e(\mathbf{x}, \hat{\mathbf{x}})$. Denote $\boldsymbol{\varepsilon}$ as an index set whose elements indicate the locations associated with the non-zero values of $\mathbf{x} - \hat{\mathbf{x}}$, $\tilde{\mathbf{h}}$ as the effective CFR vector whose entries are $\{h_t, t \in \boldsymbol{\varepsilon}\}$, and $\mathbf{X}$ as a diagonal matrix of dimensions $|\boldsymbol{\varepsilon}| \times |\boldsymbol{\varepsilon}|$ whose diagonal elements are the non-zero values of $\mathbf{x}-\hat{\mathbf{x}}$. At high SNR, the BER of $\mathbf{x}-\hat{\mathbf{x}}$. At high SNR, the BER of OFDM-IM systems can be asymptotically approximated as [3]

$$P_e \cong \frac{1}{p2^p} \sum_{\mathbf{x}} \sum_{\hat{\mathbf{x}}} \left(\frac{4^r}{12} + \frac{3^r}{4} \right) \left(\frac{k}{N} \right)^r \prod_{\xi=1}^{r} \lambda_\xi^{-1} \gamma^{-r} e(\mathbf{x}, \hat{\mathbf{x}})$$
$$\cong \gamma^{-d_{min}} \frac{4^{d_{min}} + 3^{d_{min}+1}}{12p2^p(N/k)^{d_{min}}} \sum_{d_{\min}} \prod_{\xi=1}^{d_{min}} \lambda_\xi^{-1} e(\mathbf{x}, \hat{\mathbf{x}}) \tag{6}$$

where $r = \text{rank}\{\mathbf{X}^H \mathbf{X} \mathbf{C}_{\tilde{\mathbf{h}}}\}$, $d_{min} = \min r$ is the diversity order achieved by the system, $\sum_{d_{min}}$ represents the sum operation applying to all $\mathbf{x}$ and $\hat{\mathbf{x}}$ conditioned on $r = d_{min}$, $\lambda_\xi, \xi = 1, \ldots, r$, are the non-zero eigenvalues of $\mathbf{X}^H \mathbf{X} \mathbf{C}_{\tilde{\mathbf{h}}}$, and $\gamma = k/(NN_0)$ is average received SNR.

By definition, the coding gain can be derived from (6) as

$$G_c = \left(\frac{4^{d_{\min}} + 3^{d_{\min}+1}}{12p2^p(N/k)^{d_{min}}} \sum_{d_{\min}} \prod_{\xi=1}^{d_{\min}} \lambda_\xi^{-1} e(\mathbf{x}, \hat{\mathbf{x}}) \right)^{-\frac{1}{d_{\min}}} \tag{7}$$

which satisfies $P_e \cong (G_c \gamma)^{-d_{min}}$. From (7), the SNR gain of the ESA method over the combinatorial method can be readily expressed in terms of decibels as

$$G_{snr} = 10 \log_{10}(G_c^{ESA}/G_c^{COM}) \tag{8}$$

where G_c^{ESA} and G_c^{COM} denote the resulting coding gains of the ESA and the combinatorial methods, respectively. From (7) and (8), we see that the SNR gain originates from the selection of $\mathbf{x}$ and $\hat{\mathbf{x}}$, which determines both λ_ξ and $e(\mathbf{x}, \hat{\mathbf{x}})$.

IV. Simulation Results and Analysis

We conduct computer simulations to examine the BER performance of the proposed ESA method. The system parameters are chosen as $L = 128$ and $N = 8$. The wireless channel is modeled by an uniform power delay profile with 10 taps.

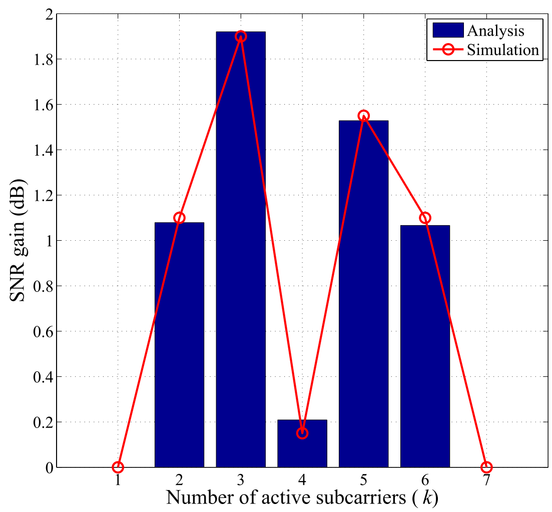

Fig. 1. The relationship between the SNR gain and the number of active subcarriers for OFDM-SSK with $N = 8$.

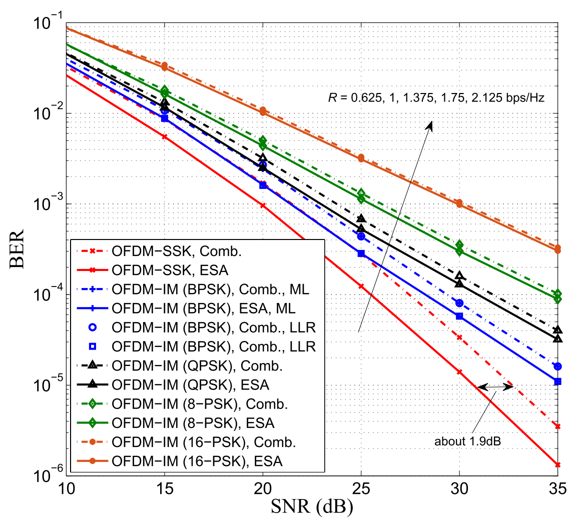

Fig. 2. Combinatorial method vs. ESA method in terms of BER for OFDM-IM with BPSK, QPSK, 8-PSK, and 16-PSK modulations and OFDM-SSK with $N = 8$ and $k = 3$.

To completely isolate the effect of ESA, we first consider the OFDM-SSK system. By varying the number of active subcarriers k, we obtain the SNR gains in Fig. 1, where the simulation results are obtained at a BER value of 10^{-5}. One can observe that an up to 1.9dB SNR gain is approximately achieved for $k = 3$, in which case the spectral efficiency is $R = 0.625$ bps/Hz. For $k = 1$ and 7, no gain is attained since in both cases, all eight possible subcarrier activation patterns are used for indexing. Actually, since the ratio of the illegal subcarrier activation patterns is just about 8.6%, a very small SNR gain, about 0.2dB, is obtained for $k = 4$. Finally, it should be noted that since the exact SNR gain depends on the specific probability distribution of the active subcarriers, asymmetric SNR gains are obtained.

We then consider the OFDM-IM system with BPSK, QPSK, 8-PSK, and 16-PSK modulations, which correspond to spectral efficiencies of $R = 1, 1.375, 1.75, 2.125$ bps/Hz, respectively, to examine the effect of ESA in the presence of modulated symbols. The comparison results for $k = 3$ are shown in Fig. 2, where all analytical upper bounds are removed for clarity. The LLR detector exhibits nearly the same performance as the ML detector; therefore, we only present an example for BPSK modulation in the figure. It can be seen from Fig. 2 that the SNR gains for the considered scenarios are approximately 1.3dB, 0.8dB, 0.6dB, and 0.3dB, respectively. The decreasing SNR gain can be understood since the error contributed by modulation bits becomes dominant at high SNR and the ratio of the number of bits carried by the modulated symbols to that of the active subcarrier indices becomes higher as the modulation order increases. However, as the OFDM-IM scheme is beneficial for transmission rates below 2 bps/Hz [4], [9], the ESA method is favorable in practice.

It is worth noting that in the above simulations, the localized grouping has been assumed. Although results are not shown in this letter, one may also observe similar phenomena when employing the interleaved grouping that can further improve the performance of OFDM-IM/OFDM-SSK [10].

V. Conclusions and Remarks

In this letter, we have addressed the problem of unbalanced subcarrier activation in the existing OFDM-IM systems and proposed an ESA method as a solution. It has been shown that the ESA method outperforms the existing combinatorial method while exhibiting similar implementation complexity.

The ESA method can be also applied to OFDM-GIM and OFDM-I/Q-IM systems to obtain improved performance, since the former can be regarded as a multiple mixed OFDM-IM system with different subcarrier activation strategies while the latter is a double mixed OFDM-IM system operating on the real and imaginary parts of the signals. Moreover, due to its nature, the ESA method also applies to any other systems with index modulation, such as spatial modulated MIMO systems.

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