Subcarrier Subset Selection Aided Transmit Precoding Achieves Full-Diversity in Index Modulation

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***Abstract*—Index modulation (IM) is a recently proposed multi-carrier transmission scheme, which conveys information both by conventional symbols as well as by the specific subcarrier activation patterns conveying them. However, an impediment of IM is that it lacks transmit diversity gain. In this paper, we circumvent this limitation by proposing a limited-feedback assisted IM transmission scheme. Specifically, Euclidean distance based subcarrier subset selection (ED-SSS) is proposed and its attainable transmit diversity order is shown to be $N_c - N_{IM} + 1$, where N_c is the total number of subcarriers in an IM block and N_{IM} is the number of subcarriers used for IM. Furthermore, the ED-SSS is shown to be amenable to low-complexity implementation owing to the orthogonality of its subcarriers. In order to attain the maximum transmit diversity order of N_c, ED-SSS is further extended with the aid of transmit precoding and its transmit diversity order is quantified. The proposed precoding assisted ED-SSS is shown to subsume several of the existing precoding aided IM transmission schemes. Simulation studies are conducted for validating our theoretical claims and also for quantifying the attainable performance gains of the proposed schemes. Specifically, at a BER of 10^{-3} an SNR gain as high as 8dB is observed in case of precoding when compared to its counterpart operating without precoding.**

***Index Terms*—Subcarrier subset selection, spatial modulation, index modulation, transmit diversity gain, Euclidean distance.**

I. Introduction

Spatial modulation (SM) [1]-[6] is a low-complexity multi-antenna transmission scheme capable of attaining a higher energy efficiency [4] than conventional multiple-input multiple-output (MIMO) transmission schemes [7], [8], [9]. Some of the key advantages of the SM system include: a) A single RF-chain at the transmitter [10], [11], resulting in low power dissipation; b) No inter-channel interference (ICI) at the receiver, yielding single-stream low maximum likelihood (ML) detection complexity [12], [13]. Furthermore, SM systems have also been extensively studied in frequency selective fading scenarios [14]-[16], where it was shown to attain a beneficial performance gain over the conventional MIMO schemes at moderate throughputs. Although, SM is known to offer several benefits over conventional MIMO schemes, it suffers from the lack of transmit diversity gain owing to having a single RF-chain at the transmitter. Several open and closed-loop techniques have been conceived for overcoming this limitation, which includes space-time coded SM transmission schemes [17]-[22], adaptive SM transmission schemes such as a) spatial and signal modulation order selection [23]; b) antenna subset selection [24]-[27] etc. Owing to the high diversity gain [26] of the Euclidean distance based antenna selection (EDAS) scheme [24], it has stimulated a significant research interests [28]-[33]. More specifically, a significant effort has been invested in reducing the computational complexity imposed by the EDAS [24], [28]-[32].

TABLE I
Comparison of various existing IM schemes.

	Description
Zheng [46]	Adaptive IM scheme is proposed, which selects the IM pattern consisting of a subcarrier subset and modulation order.
Choi [48]	Coded IM scheme is proposed, which employs outer code for improving the performance of transmitted index bits.
Zheng & Chan [49]	IM scheme is modified to utilize the silent subcarriers thereby improving the overall performance.
Gao *et al.* [50]	Precoded MIMO-OFDM IM scheme is proposed, which employs ZF based precoding and selects the active elements of the receiver side space-frequency blocks.
Zhang *et al.* [51]	Linear precoded IM is proposed, which aims to attain full transmit diversity by employing Vandermonde matrix based precoding.

A relative of the SM scheme, namely subcarrier index modulation (IM) [34]-[39] has also gained significant attraction owing to its advantages such as: a) ease of adoption in the existing practical systems; b) superior performance over the conventional orthogonal frequency division multiplexing (OFDM) at moderate throughputs [36]; c) better peak-to-average power ratio than the conventional OFDM system [40]. Owing to these benefits, IM has been envisioned to be the key technology for the next generation vehicular communications [41]. Analogous to the SM scheme, the IM scheme also suffers from the lack of transmit diversity gain. There are several recent advances in the literature that aim for improving the performance of the IM scheme [42]-[51]. Specifically, Basar

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[42] proposed to achieve a diversity order of two by invoking coordinate interleaving, where the real and imaginary parts of a pair of rotated modulated symbols are swapped before transmission. The associated constellation rotation angle was further optimized by Li *et al.* [43] based on closed-form analysis, where a precoding based method was also conceived for achieving a diversity order of two. In order to improve the bandwidth efficiency, Wen *et al.* [44] proposed permutation-based OFDM-IM, where multiple constellation modes were activated in turn in order to increase the number of transmission patterns. In [45], Dang *et al.* demonstrated that a diversity-throughput tradeoff may be achieved in OFDM-IM, where some subcarrier activation patterns were opportunistically disgarded for better performance. Zheng [46] studied the adaptive modulation order and subcarrier subset selection scheme, where the number of active subcarriers in each IM block is considered to be one. Note that no explicit diversity analysis was presented, although significant performance gains were demonstrated. In [47], Dang *et al.* further devised an adaptive OFDM-IM scheme for two-hop relaying, where the number of activated subcarriers as well as the associated mapping mechanism were dynamically selected based on the channel quality. In [48], Choi proposed a partially coded IM scheme in order to improve the detection performance of the bits transmitted over the subcarrier indices in the IM block, while transmitting uncoded bits over the conventional symbols. Although, this scheme improves the bit error ratio (BER) performance, it does not increase the overall diversity order of the system. Zheng and Chen [49] proposed an IM scheme that utilizes the inactive subcarriers in the IM block in order to improve the system performance. This scheme was shown to attain a transmit diversity order higher than or equal to two. As a further development, Gao *et al.* [50] proposed a precoding aided MIMO IM scheme, which eliminates ICI with the aid of a zero-forcing precoder. Although this scheme provides significant performance improvements over the conventional MIMO OFDM system, it requires full channel state information at the transmitter. Recently, Zhang *et al.* [51] proposed a linearly precoded IM system, which improves the performance of the conventional IM system by exploiting full transmit diversity. Table I summarizes a selection of the above mentioned key diversity-oriented OFDM-IM developments. Against this background, the following are the contributions of this paper:

1) We propose a Euclidean distance based subcarrier subset selection (ED-SSS) scheme, which is the first in the open literature that intrinsically amalgamates the limited-feedback-based subcarrier subset selection (SSS) and index modulation (IM). More explicitly, according to the SSS philosophy, a reduced number of N_{IM} out of N_c subcarriers are selected to be utilized based on their associated channel qualities. Furthermore, N_a out of N_{IM} are activated based on the IM mapping in order to transmit N_a independent complex-valued M-ary Phase Shift Keying (PSK) or Quadrature Amplitude Modulation (QAM) symbols. The proposed ED-SSS scheme has the following advantages. First of all, based on limited feedback of the channel quality, the seriously attenuated subcarriers are not used for data transmission. As a result, we prove that the ED-SSS inherently achieves a diversity order of $N_c - N_{IM} + 1$, which is higher than that of the diversity second-order solutions [42]-[49]. Secondly, we demonstrate that the ED-SSS scheme imposes a low subcarrier selection complexity, which is independent of the constellation size owing to the orthogonality of the subcarriers. This constitutes a key benefit of ED-SSS aided IM, which is not shared by the EDAS aided SM [24], since the spatial signatures are not orthogonal to each other.
2) In order to achieve the maximum attainable transmit diversity, we propose a transmit precoding (TPC) aided ED-SSS scheme, which subsumes the existing precoding based OFDM-IM schemes operating without SSS, such as that of [51]. Based on the classic upper bound of the symbol error probability, we prove that the proposed TPC aided ED-SSS scheme achieves the full diversity order of N_c. Moreover, the number of subcarrier subset combinations that have to be considered for achieving full diversity is also substantially reduced by the proposed design.

The remainder of the paper is organized as follows. The system model of IM and the ED-SSS are presented in Section II. The low-complexity ED-SSS and its diversity analysis as well as that of the TPC aided ED-SSS scheme are presented in Section III. Our simulation results and discussions are presented in Section IV, while Section V concludes the paper.

Notations: The uppercase boldface letters represent matrices and lowercase boldface letters represent vectors. The notations of $(\cdot)^T$ and $(\cdot)^H$ indicate the transpose and Hermitian transpose of a vector/matrix, respectively, while $|\cdot|$ represents the cardinality of a given set, or the magnitude of a complex quantity. The notations of $\|\cdot\|$ and $\|\cdot\|_F$ represent the two-norm of a vector and the Frobenious norm of a matrix, respectively. $\|\mathbf{x}\|_0$ represents the number of non-zero elements in $\mathbf{x}$. $\texttt{diag}(\mathbf{x})$ represents a diagonal matrix whose diagonal elements are given by the elements of $\mathbf{x}$. $\texttt{bdiag}(\mathbf{x}_1, \cdots, \mathbf{x}_N)$ forms a block-wise diagonal matrix based on the component submatrices $\{\mathbf{x}_n\}_{n=1}^N$. $\lfloor x \rfloor$ represents flooring a real-valued number x to the nearest integer. Given a set of indices I and an $n \times n$ matrix $\mathbf{A}$, $\mathbf{A}(I, I)$ represents a sub-matrix of $\mathbf{A}$ whose rows and columns are indexed by the elements of I and $\texttt{tr}(\mathbf{A})$ represents the trace of $\mathbf{A}$. $\binom{m}{n}$ represents the number of combinations of choosing n elements out of m elements. $\mathbb{R}$ and $\mathbb{C}$ represent the field of real and complex numbers, respectively. A circularly symmetric complex-valued Gaussian distribution with mean β and variance σ^2 is represented by $\mathcal{CN}(\beta, \sigma^2)$. Furthermore, $Q(y)$ represents the tail probability of standard normal distribution given by $\frac{1}{\sqrt{2\pi}} \int_y^\infty \exp\left(-\frac{u^2}{2}\right) du$. Expected value of a random variable X is denoted by $\texttt{E}(X)$.

II. PRELIMINARIES

A. Index Modulation

Consider a single input multiple output (SIMO) system having N_r receive antennas (RAs) and employing IM over N subcarriers. Let the total number of subcarriers be divided into P IM blocks each having N_c subcarriers [36]. Let furthermore the number of active subcarriers in each IM block be N_a. The total number of index activation patterns is given by $\binom{N_c}{N_a}$. The IM block in the frequency domain received over N_r RAs can be modeled as

$$\mathbf{y} = \sqrt{\frac{\rho}{N_a}}\mathbf{H}\mathbf{x} + \mathbf{n} \in \mathbb{C}^{N_r N_c}, \tag{1}$$

where ρ is the average signal-to-noise ratio (SNR), $\mathbf{H} = \mathtt{bdiag}(\mathbf{h}_1, \mathbf{h}_2, \ldots, \mathbf{h}_{N_c}) \in \mathbb{C}^{N_c N_r \times N_c}$ such that $\mathbf{h}_j \in \mathbb{C}^{N_r}$ is the channel at the j^{th} subcarrier of the IM block, $\mathbf{x} \in \mathbb{C}^{N_c}$ is the transmit IM vector with N_a non-zero elements chosen from an M-PSK/QAM signal set and $\mathbf{n}$ is the noise vector. The coefficients of $\mathbf{h}_j$ for $1 \leq j \leq N_c$ and those of $\mathbf{n}$ are modeled as $\mathcal{CN}(0,1)$. Note that the statistically independent channel coefficients over various subcarriers in an IM block can be attained by employing an interleaver [51]. The amount of information conveyed by each IM block is given by

$$R_{IM} = \left\lfloor \log_2 \binom{N_c}{N_a} \right\rfloor + N_a \log_2(M) \text{ bits}, \tag{2}$$

where M is the size of the QAM/PSK signal set employed. The ML solution assuming perfect channel state information at the receiver (CSIR) is given by

$$\hat{\mathbf{x}} = \arg\min_{\mathbf{x}\in\mathcal{X}} \|\mathbf{y} - \sqrt{\frac{\rho}{N_a}}\mathbf{H}\mathbf{x}\|^2, \tag{3}$$

where $\mathcal{X}$ is the set of all the legitimate transmit IM vectors.

B. Euclidean Distance based Subcarrier Subset Selection

Assuming that $N_{IM} < N_c$ subcarriers are used for IM, there can be a total of $n = \binom{N_c}{N_{IM}}$ possible choices. Let $\mathcal{I} = \{I_1, I_2, \ldots, I_n\}$ represent the set of enumerations of all possible combinations of N_{IM} subcarriers. Furthermore, let $\Delta\mathcal{X} = \{\mathbf{x}_1 - \mathbf{x}_2 \mid \mathbf{x}_1, \mathbf{x}_2 \in \mathcal{X}, \mathbf{x}_1 \neq \mathbf{x}_2\}$ and $\mathbf{H}_I$ represent a submatrix of $\mathbf{H}$, whose columns are indexed by the elements of I. The optimal choice of the subcarrier subset based on maximizing the Euclidean distance at the receiver is given by

$$I^* = \arg\max_{I\in\mathcal{I}} \min_{\mathbf{z}\in\Delta\mathcal{X}} \|\mathbf{H}_I \mathbf{z}\|^2. \tag{4}$$

Upon obtaining the optimal subcarrier subset I^*, the receiver signals the corresponding index to the transmitter over an error free feedback channel. The transmitter then activates the subcarriers indexed by I^* for data transmission. It is readily seen that the complexity involved in obtaining I^* in (4) is of the order $|\mathcal{I}||\Delta\mathcal{X}| = \binom{N_c}{N_{IM}} \left(\left\lfloor \binom{N_{IM}}{N_a} \right\rfloor M^{N_a} \right)^2$.

C. Precoding Aided Subcarrier Subset Selection

In case of TPC-aided ED-SSS, we have

$$I^* = \arg\max_{I\in\mathcal{I}} \min_{\mathbf{z}\in\Delta\mathcal{X}} \|\mathbf{H}_I \mathbf{P}\mathbf{z}\|^2, \tag{5}$$

where $\mathbf{P} \in \mathbb{C}^{N_{IM}\times N_{IM}}$ is the TPC matrix that spreads N_a symbols across all the N_{IM} subcarriers. The TPC matrix $\mathbf{P}$ has to satisfy $\mathbf{P}\mathbf{x} \neq \mathbf{0}$ for all $\mathbf{x} \in \Delta\mathcal{X}$ for unambiguous detection at the receiver. This can be readily ensured by considering a full-rank $\mathbf{P}$, whose null-space constitutes an empty set. Furthermore, the precoded IM symbols are transmitted over the specific subcarriers indexed by I^*, which enables attaining full transmit diversity, which is a key advantage of TPC, as discussed in detail in the next section.

III. MAIN RESULTS

In this section, we first discuss the low-complexity implementation of ED-SSS that exploits the orthogonality of the subcarriers. Furthermore, we quantify the attainable diversity gain of the ED-SSS scheme. Lastly, we discuss the TPC-aided ED-SSS, which overcomes the limitation of ED-SSS and attains full transmit as well as the receive diversity gains.

A. Low-Complexity ED-SSS

Definition 1: A regular M-QAM constellation is a Cartesian product of two PAM signal sets, say M_1-PAM and M_2-PAM, where $M = M_1 M_2$ and M_i-PAM $= \{-M_i+1, -M_i+3, \ldots, -1, 1, \ldots, M_i-3, M_i-1\}$.

Proposition 1: If $l_1, l_2, \ldots, l_{N_c}$ are the indices of the channel vectors over N_c subcarriers such that $\|\mathbf{h}_{l_1}\|^2 > \|\mathbf{h}_{l_2}\|^2 > \cdots > \|\mathbf{h}_{l_{N_c}}\|^2$ and a regular M-QAM constellation is employed over all the N_a active subcarriers, then the optimal subcarrier subset in (4) is given by $I^* = \{l_1, l_2, \ldots, l_{N_{IM}}\}$.

Please refer to Appendix A for the proof. Let us recall that without exploiting the subcarrier orthogonality by Proposition 1, the subset selection of (4) requires the channel state information at the transmitter (CSIT) and has the selection complexity order of $|\mathcal{I}||\Delta\mathcal{X}| = \binom{N_c}{N_{IM}} \left(\left\lfloor \binom{N_{IM}}{N_a} \right\rfloor M^{N_a} \right)^2$. This is generally the case for the selection in the spatial domain [24], since the spatial signatures are not orthogonal to each other. The precision required for full CSI feedback depends on the target performance. By contrast, the proposed ED-SSS only requires the limited feedback of the channel qualities, where the sorting complexity is on the order of $N_c \log N_c$, when using algorithms such as Bubble sort, Timsort, Library sort [52]-[54], etc.

Bearing in mind that the PSK signalling has substantial advantages in terms of both BER performance [55],[56] and PAPR [57],[58] in OFDM-IM systems, we offer the following corollary based on Proposition 1:

Corollary 1: When M-PSK is employed in ED-SSS aided IM, the optimal subcarrier subset in (4) is also given by $I^* = \{l_1, l_2, \ldots, l_{N_{IM}}\}$ for $\|\mathbf{h}_{l_1}\|^2 > \|\mathbf{h}_{l_2}\|^2 > \cdots > \|\mathbf{h}_{l_{N_c}}\|^2$.

Let us consider the example of $M = 2$ and $|Q^c| = N_a$. We have $\min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2 |\mathbf{z}(i)|^2 = 4\|\mathbf{h}_{m_{N_{IM}}}\|^2$. When $|Q^c| = N_a - 1$, we have $\min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2 |\mathbf{z}(i)|^2 = \|\mathbf{h}_{m_{N_{IM}}-1}\|^2 + \|\mathbf{h}_{m_{N_{IM}}}\|^2$. When $|Q^c| = N_a - 2$, we have $\min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2 |\mathbf{z}(i)|^2 = \sum_{i=0}^{3} \|\mathbf{h}_{m_{N_{IM}}-i}\|^2$. Since $\sum_{i\in I} \|\mathbf{h}_i\|^2 |\mathbf{z}(i)|^2 = f(\|\mathbf{h}_{m_1}\|^2, \|\mathbf{h}_{m_2}\|^2, \cdots, \|\mathbf{h}_{m_{N_{IM}}}\|^2)$ is a linear function for various values of $|Q^c|$, the subset $I^* = \{l_1, l_2, \ldots, l_{N_{IM}}\}$ is the solution of the optimization problem

$\max_{I\in\mathcal{I}} \min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2|\mathbf{z}(i)|^2$. Similar arguments also hold for the case of $M > 2$.

B. Diversity Analysis of ED-SSS

The system model in (1) can be equivalently written as

$$\mathbf{Y} = \sqrt{\frac{\rho}{N_a}}\mathbf{H}_s\mathbf{X} + \mathbf{N} \in \mathbb{C}^{N_r\times N_c}, \tag{6}$$

where $\mathbf{H}_s = [\mathbf{h}_1, \mathbf{h}_2, \ldots, \mathbf{h}_{N_c}] \in \mathbb{C}^{N_r\times N_c}$, $\mathbf{X} = \texttt{diag}(\mathbf{x}) \in \mathbb{C}^{N_c\times N_c}$ and $\mathbf{N}$ is the noise matrix. Let G_i represent the set of IM indices associated with the subcarrier subset I_i for $1 \leq i \leq n$. Let the transmit codebook associated with I_i be

$$\mathcal{C}_i = \left\{ \texttt{diag}\left(\sum_{k\in J} s_k\mathbf{e}_k\right) \;\middle|\; s_k \in M-\text{QAM}, J\in G_i \right\}. \tag{7}$$

Let the set of difference matrices associated with the codebook $\mathcal{C}_i$ be

$$\Delta\mathcal{C}_i = \{\mathbf{X} - \mathbf{X}' \mid \mathbf{X}, \mathbf{X}' \in \mathcal{C}_i, \mathbf{X} \neq \mathbf{X}'\}. \tag{8}$$

Furthermore, let $\mathbf{X}_{min}(k) = \arg\min_{\mathbf{X}\in\Delta\mathcal{C}_k} \|\mathbf{H}_s\mathbf{X}\|_F^2$, $\mathbf{X}_{min} = [\mathbf{X}_{min}(1), \mathbf{X}_{min}(2), \ldots, \mathbf{X}_{min}(n)] \in \mathbb{C}^{N_c\times nN_c}$ and the index of the optimal subcarrier subset $k^* = \arg\max_l \{\alpha_l\}_{l=1}^n$, where $\alpha_l = \min_{\mathbf{X}\in\Delta\mathcal{C}_l} \|\mathbf{H}_s\mathbf{X}\|_F^2$.

Proposition 2: The union bound on the symbol error probability of the ED-SSS is given by

$$\mathtt{P_e} \leq \frac{|\mathcal{C}_{k^*}| - 1}{2}\left(\frac{\rho\lambda^*}{4nN_a}\right)^{-qN_r}, \tag{9}$$

where $q = \texttt{rank}(\mathbf{X}_{min})$ and λ^* is the smallest non-zero Eigenvalue of $\mathbf{X}_{min}\mathbf{X}_{min}^H$.

Proof: Given the channel realization $\mathbf{H}$, the pairwise error probability (PEP) between any two distinct transmit signals indexed by m_1, m_2 in $\mathcal{C}_{k^*}$ is given by

$$\texttt{PEP}(\mathbf{X}_{m_1} \to \mathbf{X}_{m_2}|\mathbf{H}_s) = Q\left(\sqrt{\frac{\rho}{N_a}}\frac{\|\mathbf{H}_s(\mathbf{X}_{m_1} - \mathbf{X}_{m_2})\|_F}{2}\right), \tag{10}$$

$$\leq \frac{1}{2}\exp\left(\frac{-\rho\|\mathbf{H}_s(\mathbf{X}_{m_1} - \mathbf{X}_{m_2})\|_F^2}{4N_a}\right). \tag{11}$$

We note that Craig's formula was invoked for evaluating the exact expression of the Q-function in [59]. Nonetheless, in this work, we are mainly concerned with determining the diversity order with the aid of the classic upper bound expression of the PEP following the derivations in [60], [61]. As a result, we have:

$$\|\mathbf{H}_s(\mathbf{X}_{m_1} - \mathbf{X}_{m_2})\|_F^2 \geq \|\mathbf{H}_s\mathbf{X}_{min}(k^*)\|_F^2 \geq \frac{\|\mathbf{H}_s\mathbf{X}_{min}\|_F^2}{n}. \tag{12}$$

Furthermore, we have $\|\mathbf{H}_s\mathbf{X}_{min}\|_F^2 = \texttt{tr}(\mathbf{H}_s\mathbf{X}_{min}\mathbf{X}_{min}^H\mathbf{H}_s^H) = \texttt{tr}(\bar{\mathbf{H}}_s\mathbf{D}_{min}\bar{\mathbf{H}}_s^H)$, where $\mathbf{X}_{min}\mathbf{X}_{min}^H = \mathbf{U}\mathbf{D}_{min}\mathbf{U}^H$ and $\bar{\mathbf{H}}_s = \mathbf{H}_s\mathbf{U}$. Thus, we have $\|\mathbf{H}_s\mathbf{X}_{min}\|_F^2 = \sum_{i=1}^{N_r}\sum_{j=1}^{q}\lambda_j|\bar{h}_s(i,j)|^2 \geq \lambda^*\sum_{i=1}^{N_r}\sum_{j=1}^{q}|\bar{h}_s(i,j)|^2$, where $q = \texttt{rank}(\mathbf{X}_{min})$ and λ^* is the smallest non-zero Eigenvalue of $\mathbf{X}_{min}\mathbf{X}_{min}^H$. From the above equations, we arrive at

$$\texttt{PEP}(\mathbf{X}_{m_1} \to \mathbf{X}_{m_2}|\mathbf{H}_s) \leq \frac{1}{2}\exp\left(\frac{-\rho\lambda^*}{4nN_a}\sum_{i=1}^{N_r}\sum_{j=1}^{q}|\bar{h}_s(i,j)|^2\right). \tag{13}$$

Taking expectation over $\mathbf{H}_s$ of (6), we have

$$\texttt{PEP}(\mathbf{X}_{m_1} \to \mathbf{X}_{m_2}) \leq \frac{1}{2}\prod_{i=1}^{N_r}\prod_{j=1}^{q}\mathtt{E}\left(\exp\left(\frac{-\rho\lambda^*}{4nN_a}|\bar{h}_s(i,j)|^2\right)\right), \tag{14}$$

$$\leq \frac{1}{2}\prod_{i=1}^{N_r}\prod_{j=1}^{q}\left(1 + \frac{\rho\lambda^*}{4nN_a}\right)^{-1}. \tag{15}$$

At high SNRs, we have $\frac{\rho\lambda^*}{4nN_a} \gg 1$ and hence

$$\texttt{PEP}(\mathbf{X}_{m_1} \to \mathbf{X}_{m_2}) \leq \frac{1}{2}\left(\frac{\rho\lambda^*}{4nN_a}\right)^{-qN_r}. \tag{16}$$

Thus, the union bound on the symbol error probability is given by

$$\mathtt{P_e} \leq \frac{1}{|\mathcal{C}_{k^*}|}\sum_{\mathbf{X}_{m_1}\in\mathcal{C}_{k^*}}\;\sum_{\mathbf{X}_{m_1}\neq\mathbf{X}_{m_2}\in\mathcal{C}_{k^*}}\texttt{PEP}(\mathbf{X}_{m_1} \to \mathbf{X}_{m_2}), \tag{17}$$

$$\leq \frac{|\mathcal{C}_{k^*}| - 1}{2}\left(\frac{\rho\lambda^*}{4nN_a}\right)^{-qN_r}. \tag{18}$$

■

It is evident from (18) that the attainable transmit diversity order of ED-SSS is given by $q = \texttt{rank}(\mathbf{X}_{min})$. The following proposition states that the transmit diversity order of ED-SSS is $N_c - N_{IM} + 1$.

Proposition 3: For any given channel realization $\mathbf{H}$, we have $\texttt{rank}(\mathbf{X}_{min}) = N_c - N_{IM} + 1$ and hence the transmit diversity order of ED-SSS is $N_c - N_{IM} + 1$.

Proof: Let the set of indices $l_1^{(k)}, l_2^{(k)}, \ldots, l_{N_{IM}}^{(k)}$ associated with the codebook $\mathcal{C}_k$ is such that $\|\mathbf{h}_{l_1^{(k)}}\|^2 > \|\mathbf{h}_{l_2^{(k)}}\|^2 > \cdots > \|\mathbf{h}_{l_{N_{IM}}^{(k)}}\|^2$. From Proposition 1, we have $\mathbf{X}_{min}(k) = 4\|\mathbf{h}_{l_{N_{IM}}^{(k)}}\|^2\mathbf{e}_{l_{N_{IM}}^{(k)}}\mathbf{e}_{l_{N_{IM}}^{(k)}}^T$, which is a rank-1 matrix. The index $l_{N_{IM}}^{(k)}$ appears in $\binom{N_c-1}{N_{IM}-1}$ combinations out of n combinations. Thus, $\binom{N_c-1}{N_{IM}-1}$ $\mathbf{X}_{min}(i)$ matrices in $\mathbf{X}_{min}$ belong to the same subspace constituting a rank-1 matrix. Similarly, the next $\binom{N_c-2}{N_{IM}-1}$ combinations also constitute a set of rank-1 matrices which results in rank-2 $\mathbf{X}_{min}$. Since there can be $N_c - N_{IM} + 1$ partitions of $\mathbf{X}_{min}$, each constituting a rank-1 matrix and linearly independent of the matrices from other partitions, we have $\texttt{rank}(\mathbf{X}_{min}) = N_c - N_{IM} + 1$. ■

The following examples illustrate the above concepts.

Example 1: Let $N_c = 3, N_{IM} = 2$ and $N_r = 1$. Without loss of generality, let $|h_1|^2 > |h_2|^2 > |h_3|^2$ and

$$\mathcal{I} = \left\{\underbrace{\{1,2\}}_{I_1}, \underbrace{\{2,3\}}_{I_2}, \underbrace{\{1,3\}}_{I_3}\right\}. \tag{19}$$

Then, we have

$$\mathbf{X}_{min} = [\mathbf{X}_{min}(1)\ \mathbf{X}_{min}(2)\ \mathbf{X}_{min}(3)], \tag{20}$$

where

$$\mathbf{X}_{min}(i) = \begin{bmatrix} 0 & 0 & 0 \\ 0 & 0 & 0 \\ 0 & 0 & * \end{bmatrix}, \tag{21}$$

for $i \in \{2, 3\}$ and

$$\mathbf{X}_{min}(1) = \begin{bmatrix} 0 & 0 & 0 \\ 0 & * & 0 \\ 0 & 0 & 0 \end{bmatrix}. \tag{22}$$

It is then readily seen that $\mathtt{rank}(\mathbf{X}_{min}) = 2$.

Example 2: Let $N_c = 4, N_{IM} = 3$ and $N_r = 1$. Without loss of generality, let $|h_1|^2 > |h_2|^2 > |h_3|^2 > |h_4|^2$ and

$$\mathcal{I} = \left\{ \underbrace{\{1,2,4\}}_{I_1}, \underbrace{\{1,3,4\}}_{I_2}, \underbrace{\{2,3,4\}}_{I_3}, \underbrace{\{1,2,3\}}_{I_4} \right\}. \tag{23}$$

Then, we have

$$\mathbf{X}_{min} = [\mathbf{X}_{min}(1)\ \mathbf{X}_{min}(2)\ \mathbf{X}_{min}(3)\ \mathbf{X}_{min}(4)], \tag{24}$$

where

$$\mathbf{X}_{min}(i) = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & * \end{bmatrix}, \tag{25}$$

for $1 \leq i \leq 3$ and

$$\mathbf{X}_{min}(4) = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & * & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix}. \tag{26}$$

Observe that we have $\mathtt{rank}(\mathbf{X}_{min}) = 2$.

C. Attaining Full Diversity with Precoding

Let the transmit codebook associated with I_i when employing TPC be

$$\mathcal{C}'_i = \left\{ \mathtt{diag}\left(\mathbf{P}_i \sum_{k \in J} s_k \mathbf{e}_k \right) \;\middle|\; s_k \in M-\mathrm{QAM}, J \in G_i \right\}, \tag{27}$$

where $\mathbf{P}_i \in \mathbb{C}^{N_c \times N_c}$ is the TPC matrix associated with the index subset I_i, i.e. $\mathbf{P}_i(I_i, I_i) = \mathbf{P}$ and zeros elsewhere. Let the set of difference matrices associated with the codebook $\mathcal{C}'_i$ be

$$\Delta\mathcal{C}'_i = \{\mathbf{X} - \mathbf{X}' \mid \mathbf{X}, \mathbf{X}' \in \mathcal{C}'_i, \mathbf{X} \neq \mathbf{X}'\}. \tag{28}$$

Furthermore, let $\mathbf{X}'_{min}(k) = \arg\min_{\mathbf{X} \in \Delta\mathcal{C}'_k} \|\mathbf{H}_s\mathbf{X}\|_F^2$, $\mathbf{X}'_{min} = [\mathbf{X}'_{min}(1), \mathbf{X}'_{min}(2), \ldots, \mathbf{X}'_{min}(n)] \in \mathbb{C}^{N_c \times nN_c}$ and the index of the optimal subcarrier subset be $k^* = \arg\max_l \{\beta_l\}_{l=1}^n$, where $\beta_l = \min_{\mathbf{X} \in \Delta\mathcal{C}'_l} \|\mathbf{H}_s\mathbf{X}\|_F^2$.

Proposition 4: For any given channel realization $\mathbf{H}$, the transmit diversity order of TPC-aided ED-SSS is N_c, when $\|\mathbf{P}\mathbf{x}\|_0 = N_{IM}\ \forall\ \mathbf{x} \in \Delta\mathcal{X}$.

TABLE II
COMPARISON OF ED-SSS WITH THE EXISTING SCHEMES.

	Channel signatures	**Complexity**	**Diversity order**
GSM/SM with EDAS [24], [27]	Non-orthogonal	$\mathcal{O}(M^{N_a})$	$N_c - N_{IM} + 1$
ED-SSS	Orthogonal	Independent of M	$N_c - N_{IM} + 1$
ED-SSS with precoding	Non-orthogonal	$\mathcal{O}(M^{N_a})$	N_c

Proof: From Proposition 2, it is sufficient to show that $\mathtt{rank}(\mathbf{X}'_{min}) = N_c$ in order to prove that the transmit diversity order of TPC-aided ED-SSS is N_c. Note that each $\mathbf{X}'_{min}(k)$ is a diagonal matrix with N_{IM} non-zero elements at locations indexed by I_k due to the condition that $\|\mathbf{P}\mathbf{x}\|_0 = N_{IM}$ $\forall\ \mathbf{x} \in \Delta\mathcal{X}$. Since there exists $I'_1, I'_2, \ldots, I'_k$ $(k \leq n)$ such that $\cup_{i=1}^k I'_i = \{1, 2, \ldots, N_c\}$, we have $\mathtt{rank}(\mathbf{X}'_{min}) = N_c$. ■

Corollary 1: In an IM system employing TPC-aided ED-SSS, where the TPC $\mathbf{P}$ satisfies the condition $\|\mathbf{P}\mathbf{x}\|_0 = N_{IM}$ $\forall\ \mathbf{x} \in \Delta\mathcal{X}$, no more than $\lceil \frac{N_c}{N_{IM}} \rceil$ subcarrier subsets have to be considered for obtaining the optimal subcarrier subset that achieves full transmit diversity gain.

When $N_c = N_{IM}$, having a single subcarrier subset suffices in order to attain full transmit diversity gain. This case corresponds to the linearly precoded IM scheme of [51].

Example 3: Let $N_c = 4, N_{IM} = 3$, $N_r = 1$ and

$$\mathcal{I} = \left\{ \underbrace{\{1,2,4\}}_{I_1}, \underbrace{\{1,3,4\}}_{I_2}, \underbrace{\{2,3,4\}}_{I_3}, \underbrace{\{1,2,3\}}_{I_4} \right\} \tag{29}$$

as in the previous example. Then, we have

$$\mathbf{X}'_{min} = [\mathbf{X}'_{min}(1)\ \mathbf{X}'_{min}(2)\ \mathbf{X}'_{min}(3)\ \mathbf{X}'_{min}(4)], \tag{30}$$

where

$$\mathbf{X}_{min}(1) = \begin{bmatrix} * & 0 & 0 & 0 \\ 0 & * & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & * \end{bmatrix}, \mathbf{X}_{min}(2) = \begin{bmatrix} * & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \\ 0 & 0 & * & 0 \\ 0 & 0 & 0 & * \end{bmatrix}, \tag{31}$$

$$\mathbf{X}_{min}(3) = \begin{bmatrix} 0 & 0 & 0 & 0 \\ 0 & * & 0 & 0 \\ 0 & 0 & * & 0 \\ 0 & 0 & 0 & * \end{bmatrix}, \mathbf{X}_{min}(4) = \begin{bmatrix} * & 0 & 0 & 0 \\ 0 & * & 0 & 0 \\ 0 & 0 & * & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix}. \tag{32}$$

It is then readily seen that $\mathtt{rank}(\mathbf{X}_{min}) = 4$. Furthermore, it is straightforward to see that any two distinct subcarrier subsets I'_1, I'_2 yield $I'_1 \cup I'_2 = \{1, 2, 3, 4\}$. Thus, any two distinct subcarrier subset combinations yield full diversity gain.

Table II compares some of the characteristics of the ED-SSS with that of the existing schemes. Note that although TPC in ED-SSS allows achieving full diversity gain, it imposes high computational complexity owing to the non-orthogonal channel signatures.

D. Performance under Correlated Subcarriers

So far we have assumed that the channel coefficients across subcarriers are uncorrelated and statistically independent. Although this condition can be approached by employing a long frequency-domain interleaver [51], the validity of this assumption may become eroded, when we have a limited the number of multipath components. This renders the subcarriers in an IM block correlated. Thus, it is worth extending our analysis to correlated channel conditions. Let the correlated channel over the subcarriers be modeled as $\mathbf{H}_s\mathbf{T}^{1/2}$, where $\mathbf{T} \in \mathbb{C}^{N_c \times N_c}$ is the subcarrier correlation matrix.

Proposition 5: The union bound on the symbol error probability of the ED-SSS operating under correlated channel condition is given by

$$\mathtt{P_e} \leq \frac{|\mathcal{C}_{k^*}|-1}{2}\left(\frac{\rho\bar{\lambda}^*}{4nN_a}\right)^{-\bar{q}N_r}, \qquad (33)$$

where $\bar{q} = \mathtt{rank}(\mathbf{X}_{min}\mathbf{X}_{min}^H\mathbf{T})$ and $\bar{\lambda}^*$ is the smallest non-zero Eigenvalue of $\mathbf{X}_{min}\mathbf{X}_{min}^H\mathbf{T}$.

The proof is along the same lines as that of Proposition 2, hence it is omitted. It is evident from Proposition 5 that a full-rank $\mathbf{T}$ is sufficient for attaining the diversity gain of $N_c - N_{IM} + 1$ and N_c in case of ED-SSS and TPC-aided ED-SSS, respectively.

IV. Simulation Studies

Simulation scenario: While evaluating a BER of 10^{-t}, we have employed at least 10^{t+1} bits in all our simulations. The receiver is assumed to have perfect CSI and employ ML decoding in all the schemes considered. Furthermore, we note that the TPC matrix introduced in Sec. II.C may invoke any of the richly documented historical full-rank constructions in the space-time domain such as the lattice-based [62]-[64], field extension aided [65] and division algebras assited [66]-[68] constructions. However, it was astutely pointed out in [66] that the TPC matrices that achieve the optimal diversity-multiplexing tradeoff do not always lead to the best BER performance. Given that the existing TPC solutions conceived for OFDM-IM [50],[51],[43] are not generally applicable, based on our simulations, we opt for using the following rotation matrices contructed based on rotated $\mathbf{Z}^n$ lattices [62]-[64]. For the IM system associated with $N_{IM} = 3$, we have

$$\mathbf{P} = \begin{bmatrix} -0.3279 & -0.5910 & -0.7369 \\ -0.7369 & -0.3279 & 0.5910 \\ -0.5910 & 0.7369 & -0.3279 \end{bmatrix}, \qquad (34)$$

and for $N_{IM} = 4$, we have

$$\mathbf{P} = \begin{bmatrix} -0.3663 & -0.7677 & 0.4230 & 0.3120 \\ -0.2264 & -0.4744 & -0.6845 & -0.5049 \\ -0.4744 & 0.2264 & -0.5049 & 0.6845 \\ -0.7677 & 0.3663 & 0.3120 & -0.4230 \end{bmatrix}. \qquad (35)$$

The correlated subcarriers are simulated by assuming $\mathbf{T}(i,j) = \zeta^{|i-j|}$ in [64], where $0 \leq \zeta \leq 1$ models the correlation coefficient. In our simulation studies, we consider two values of $\zeta \in \{0.1, 0.9\}$, where 0.1 corresponds to a fairly uncorrelated channel scenario and 0.9 corresponds to a high-correlation scenario.

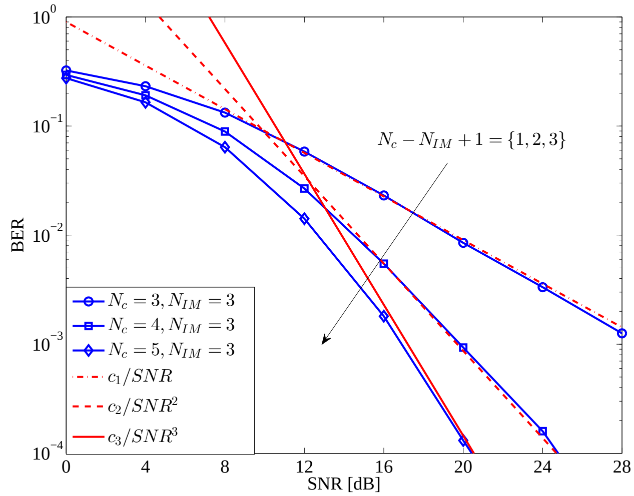

Fig. 1. BER performance of ED-SSS in an IM system having $N_{IM} = 3$, $N_a = 2$, $N_r = 1$, $N_c \in \{3,4,5\}$ and employing 4-QAM signal set. The reference curves c_k/SNR^k for $k \in \{1,2,3\}$ are provided to illustrate the attainable diversity order.

A. Validation of Transmit Diversity Order of ED-SSS

In order to validate the theoretical claims concerning the achievable diversity gain of ED-SSS, let us consider an IM system having $N_{IM} = 3$, $N_a = 2$, $N_r = 1$, $N_c \in \{3,4,5\}$ and employing 4-QAM signal set. Fig. 1 depicts the attainable BER performance of ED-SSS for various values of N_c as well as the reference curves c_k/SNR^k for $k \in \{1,2,3\}$. The constants associated with the reference curves are given by $c_1 = 0.9$, $c_2 = 8.7$ and $c_3 = 145$. It is evident from Fig. 1 that the ED-SSS attains a transmit diversity order of $N_c - N_{IM} + 1$. Fig. 2 depicts the attainable BER performance of ED-SSS in the aforementioned system when employing $N_r = 2$. The reference curves c'_k/SNR^k for $k \in \{2,4,6\}$ are also provided for illustrating the attainable diversity order. The constants associated with the reference curves are given by $c'_1 = 2.6$, $c'_2 = 185$ and $c'_3 = 29500$. It is evident from Fig. 2 that ED-SSS attains a total diversity order of $N_r(N_c - N_{IM} + 1) = 2(N_c - N_{IM} + 1)$. Fig. 3 compares the attainable BER performance of ED-SSS in an IM system having $N_{IM} = 4$, $N_a = 2$, $N_r \in \{1,2\}$, $N_c \in \{4,5,6\}$ and employing 4-QAM signal set. Observe from Fig. 3 that ED-SSS attains higher diversity gain when N_c is increased from 4 to 6.

B. Performance of ED-SSS with Various Signal Sets

Fig. 4 depicts the BER performance of ED-SSS in an IM system having $N_{IM} = 3$, $N_a = 2$, $N_r = 1$, $N_c \in \{3,4\}$ and employing BPSK, 4-QAM and 8-QAM signal sets. It is seen from Fig. 4 that ED-SSS attains a higher diversity order for all the signal sets considered, when N_c is increased. Similarly, Fig. 5 depicts the BER performance of ED-SSS in an IM system having $N_{IM} = 4$, $N_a = 2$, $N_r = 1$, $N_c \in \{4,5\}$ and employing BPSK, 4-QAM and 8-QAM signal sets. Again, it is observed that ED-SSS attains a higher diversity order for

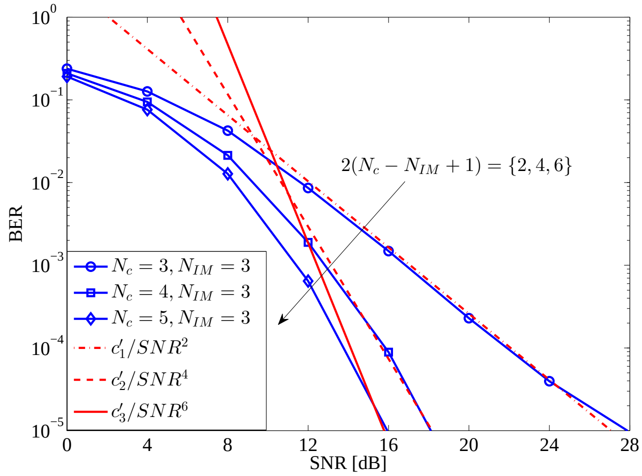

Fig. 2. BER performance of ED-SSS in an IM system having $N_{IM} = 3$, $N_a = 2$, $N_r = 2$, $N_c \in \{3, 4, 5\}$ and employing 4-QAM signal set. The reference curves c'_k/SNR^{2k} for $k \in \{1, 2, 3\}$ are provided to illustrate the attainable diversity order.

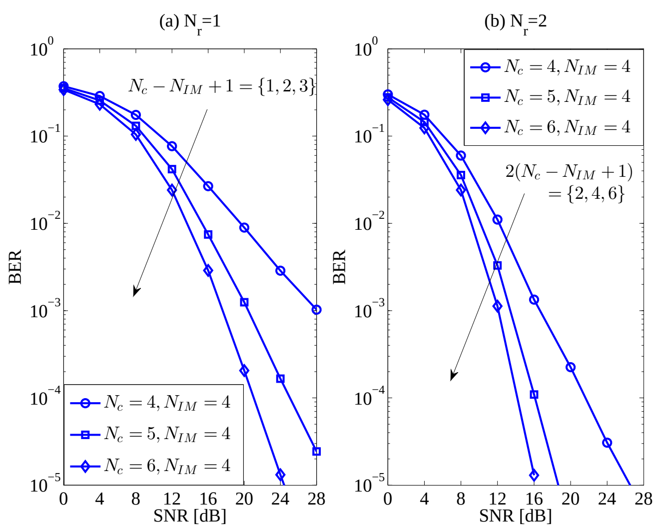

Fig. 3. BER performance of ED-SSS in an IM system having $N_{IM} = 4$, $N_a = 2$, $N_r \in \{1, 2\}$, $N_c \in \{4, 5, 6\}$ and employing 4-QAM signal set. Plot (a) corresponds to $N_r = 1$ and plot (b) corresponds to $N_r = 2$.

all the signal sets considered, when N_c is increased. Thus, we conclude that our results concerning the attainable diversity order of ED-SSS stands validated.

C. Performance of ED-SSS with Precoding

Fig. 6 compares the BER performance of TPC-aided ED-SSS in an IM system having $N_{IM} = 3$, $N_a = 2$, $N_r \in \{1, 2\}$, $N_c \in \{3, 4, 5\}$ and employing a 4-QAM signal set. It is evident from Fig. 6 that there is a significant performance improvement, when N_c is increased. Similarly, Fig. 7 compares the BER performance of TPC-aided ED-SSS in an IM system having $N_{IM} = 4$, $N_a = 2$, $N_r \in \{1, 2\}$, $N_c \in \{4, 5, 6\}$ and employing a 4-QAM signal set. Fig. 8 compares the BER performance of TPC-aided ED-SSS to that

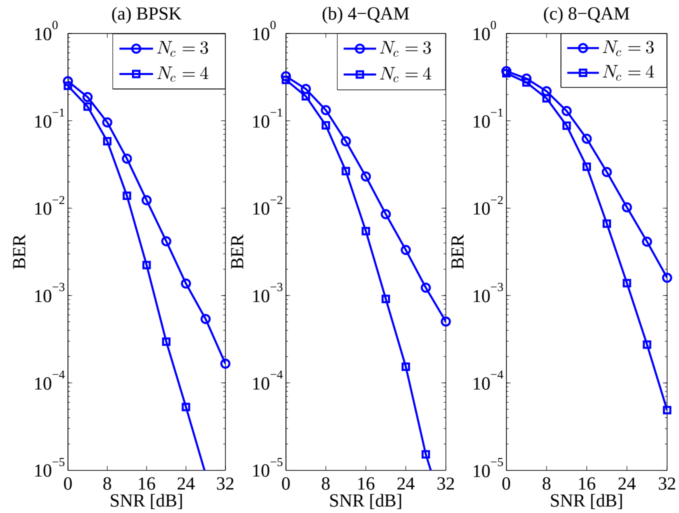

Fig. 4. BER performance of ED-SSS in an IM system having $N_{IM} = 3$, $N_a = 2$, $N_r = 1$, $N_c \in \{3, 4\}$ and employing BPSK, 4-QAM and 8-QAM signal sets. Plot (a), plot (b), and plot (c) correspond to BPSK, 4-QAM, and 8-QAM signal sets, respectively.

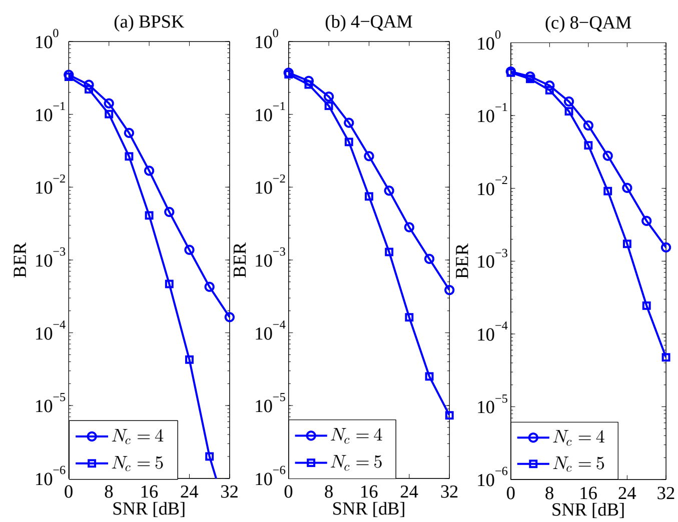

Fig. 5. BER performance of ED-SSS in an IM system having $N_{IM} = 4$, $N_a = 2$, $N_r = 1$, $N_c \in \{4, 5\}$ and employing BPSK, 4-QAM and 8-QAM signal sets. Plot (a), plot (b), and plot (c) correspond to BPSK, 4-QAM, and 8-QAM signal sets, respectively.

without precoding in an IM system having $N_{IM} = 3$, $N_a = 2$, $N_r = 1$, $N_c \in \{3, 5\}$ and employing a 4-QAM signal set. It is seen in Fig. 8 that TPC provides significant performance improvements over its counterpart operating without TPC. Specifically, at a BER of 10^{-3} an SNR gain of about 8dB is observed, when $N_c = 3$ and $N_r = 1$. Furthermore, at a BER of 10^{-4}, an SNR gain of about 5dB is recorded for $N_c = 3$ and $N_r = 2$. When $N_c = 5$, we see that TPC does not yield significant performance gains when $N_{IM} = 3$. Thus, we may resort to ED-SSS dispensing with precoding when N_c is large, whose complexity is significantly lower than that of its counterpart with precoding. However, when $N_{IM} = 4$

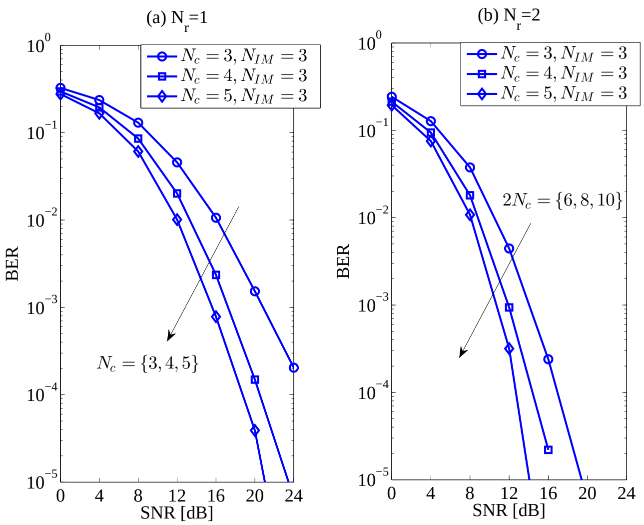

Fig. 6. BER performance of ED-SSS with precoding in an IM system having $N_{IM} = 3$, $N_a = 2$, $N_c \in \{3, 4, 5\}$ and employing 4-QAM signal set. Plot (a) and plot (b) correspond to $N_r = 1$ and $N_r = 2$, respectively.

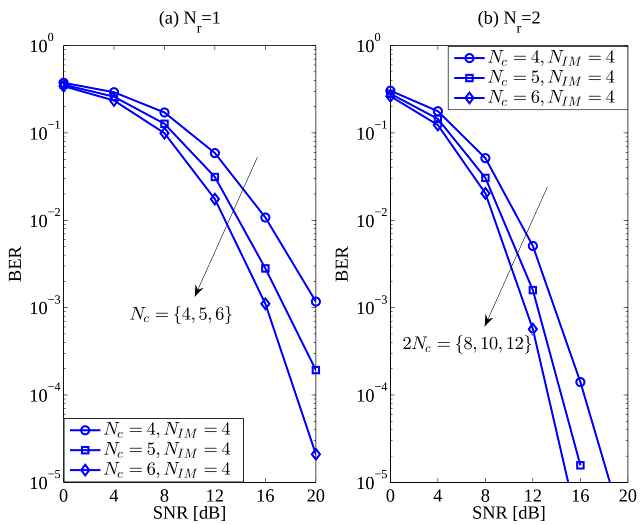

Fig. 7. BER performance of ED-SSS with precoding in an IM system having $N_{IM} = 4$, $N_a = 2$, $N_c \in \{4, 5, 6\}$ and employing 4-QAM signal set. Plot (a) and plot (b) correspond to $N_r = 1$ and $N_r = 2$, respectively.

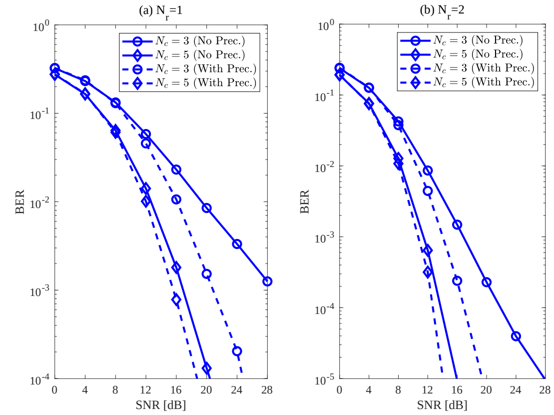

Fig. 8. Comparison of BER performance of ED-SSS with precoding with that without precoding in an IM system having $N_{IM} = 3$, $N_a = 2$, $N_r = 1$, $N_c \in \{3, 5\}$ and employing 4-QAM signal set. Plot (a) and plot (b) correspond to $N_r = 1$ and $N_r = 2$, respectively.

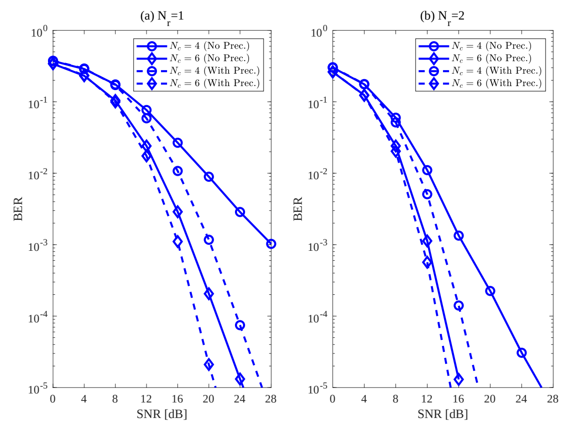

Fig. 9. Comparison of BER performance of ED-SSS with precoding with that without precoding in an IM system having $N_{IM} = 4$, $N_a = 2$, $N_r = 1$, $N_c \in \{4, 6\}$ and employing 4-QAM signal set. Plot (a) and plot (b) correspond to $N_r = 1$ and $N_r = 2$, respectively.

we see that TPC does yield a significant performance gain. Specifically, at a BER of 10^{-5}, we see an SNR gain of about 3dB, which is evident from Fig. 9(a).

D. Performance under Correlated Subcarriers

Fig. 10 compares the BER performance of ED-SSS in an IM system having $(N_c, N_{IM}) \in \{(5, 3), (6, 4)\}$, $N_a = 2$, $N_r = 1$, and employing a 4-QAM signal set, when operating in a correlated channel having $\zeta \in \{0.1, 0.9\}$. Specifically, Fig. 10(a) and Fig. 10(b) correspond to $(N_c, N_{IM}) = (5, 3)$ and $(N_c, N_{IM}) = (6, 4)$, respectively. Observe from both the plots that the channel's correlation imposes significant performance degradations. Specifically, an SNR degradation of about 3dB is observed at a BER of about 10^{-4}, when the correlation factor ζ is increased from 0.1 to 0.9. Similar observations also hold for the TPC-aided ED-SSS, which is evident from Fig. 11. Although there is a gradual performance degradation when the channel's correlation factor is increased, it is evident from both Fig. 10 and Fig. 11 that the proposed schemes retains the diversity gain guaranteed by our theoretical analysis.

V. Conclusions

We have considered a subcarrier IM aided system and proposed Euclidean distance based subcarrier subset selection in order to increase the diversity gain. We have quantified the attainable transmit diversity order of ED-SSS and shown that to be amenable to low-complexity implementation. Furthermore, we have extended ED-SSS with the aid of precoding, which was shown to overcome the limitation of ED-SSS by attaining full transmit diversity gain. Simulation results were presented to validate the theoretical claims and also to demonstrate the performance gains of the proposed schemes.

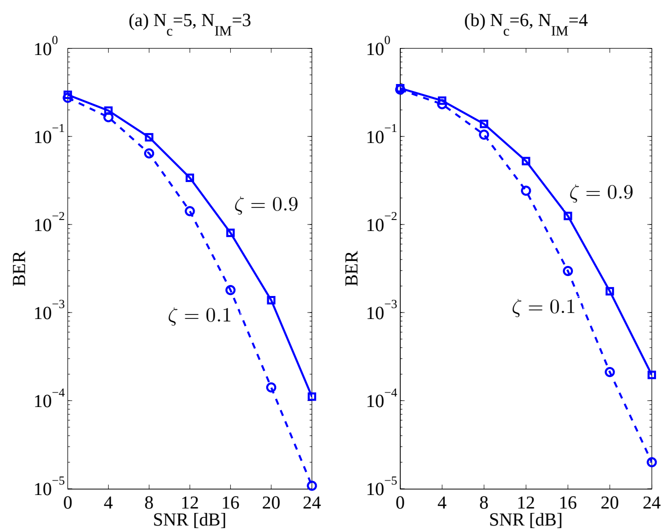

Fig. 10. Comparison of BER performance of ED-SSS in an IM system having $N_{IM} \in \{3, 4\}$, $N_c \in \{5, 6\}$, $N_a = 2$, $N_r = 1$, and employing 4-QAM signal set while operating in a correlated channel having $\zeta \in \{0.1, 0.9\}$.

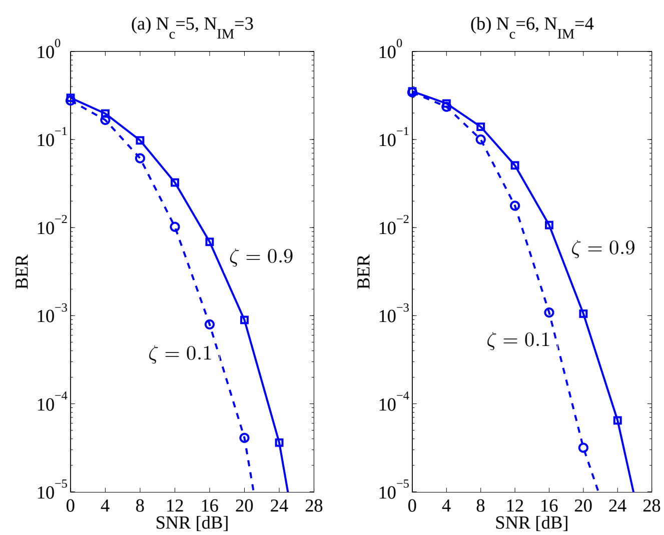

Fig. 11. Comparison of BER performance of ED-SSS with precoding in an IM system having $N_{IM} \in \{3, 4\}$, $N_c \in \{5, 6\}$, $N_a = 2$, $N_r = 1$, and employing 4-QAM signal set while operating in a correlated channel having $\zeta \in \{0.1, 0.9\}$.

VI. Appendix A

Proof of Proposition 1

Proof: The problem of finding the optimal subcarrier subset in (4) can be equivalently written as

$$I^* = \arg\max_{I\in\mathcal{I}} \min_{\mathbf{z}\in\Delta\mathcal{X}} \|\mathbf{H}_I\mathbf{z}\|^2, \quad (36)$$

$$= \arg\max_{I\in\mathcal{I}} \min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\mathbf{z}(i)\|^2, \quad (37)$$

$$= \arg\max_{I\in\mathcal{I}} \min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2|\mathbf{z}(i)|^2. \quad (38)$$

Any $\mathbf{z} \in \Delta\mathcal{X}$ can be expressed as $\mathbf{z} = \sum_{i=1}^{N_a} s_i\mathbf{e}_{j_i} - \sum_{i=1}^{N_a} s'_i\mathbf{e}_{k_i}$, where $Q_1 = \{j_1, j_2, \ldots, j_{N_a}\}$ and $Q_2 = \{k_1, k_2, \ldots, k_{N_a}\}$ are the support sets of the transmit vectors associated with $\mathbf{z}$, s_i and s'_i are M-QAM symbols. Let $Q^c = Q_1 \cap Q_2$, $Q_1^d = Q_1 \backslash Q^c$ and $Q_2^d = Q_2 \backslash Q^c$. Note that $0 \leq |Q^c| \leq N_a$. Furthermore, we have

$$|\mathbf{z}(i)|^2 = \begin{cases} |s_i - s'_i|^2, & i \in Q^c \\ |s_i|^2, & i \in Q_1^d \\ |s'_i|^2, & i \in Q_2^d \end{cases}. \quad (39)$$

Given $I \in \mathcal{I}$, we have $\min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2|\mathbf{z}(i)|^2 \equiv \min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in Q^c} \|\mathbf{h}_i\|^2|s_i - s'_i|^2 + \sum_{i\in Q_1^d} \|\mathbf{h}_i\|^2|s_i|^2 + \sum_{i\in Q_2^d} \|\mathbf{h}_i\|^2|s'_i|^2$. Let the elements of I be arranged as $\{m_1, m_2, \ldots, m_{N_{IM}}\}$ such that $\|\mathbf{h}_{m_1}\|^2 > \|\mathbf{h}_{m_2}\|^2 > \cdots > \|\mathbf{h}_{m_{N_{IM}}}\|^2$. When $|Q^c| = N_a$, we have $|Q_1^d| = |Q_2^d| = 0$ and $\min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2|\mathbf{z}(i)|^2 = 4\|\mathbf{h}_{m_{N_{IM}}}\|^2$, which follows from the fact that $\min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} |\mathbf{z}(i)|^2 = 4$ when a regular M-QAM constellation is employed. When $|Q^c| = N_a - 1$, we have $|Q_1^d| = |Q_2^d| = 1$ and $\min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2|\mathbf{z}(i)|^2 = 2\|\mathbf{h}_{m_{N_{IM}-1}}\|^2 + 2\|\mathbf{h}_{m_{N_{IM}}}\|^2 > 4\|\mathbf{h}_{m_{N_{IM}}}\|^2$. It is straightforward to show that for $|Q^c| < N_a - 1$, we have $\min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2|\mathbf{z}(i)|^2 > 4\|\mathbf{h}_{m_{N_{IM}}}\|^2$. Thus, when employing a regular M-QAM, we have $\min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\|^2|\mathbf{z}(i)|^2 = 4\|\mathbf{h}_{m_{N_{IM}}}\|^2$. Thus, the solution to the problem of $\max_{I\in\mathcal{I}} \min_{\mathbf{z}\in\Delta\mathcal{X}} \sum_{i\in I} \|\mathbf{h}_i\mathbf{z}(i)\|^2$ can be obtained by successively eliminating the subsets that are associated with the subcarriers $l_{N_c}, l_{N_c-1}, \ldots, l_{N_{IM}+1}$, which yields $I^* = \{l_1, l_2, \ldots, l_{N_{IM}}\}$. ■

VII. Acknowledgements

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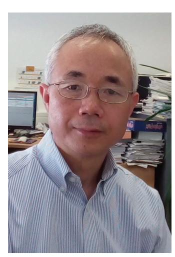

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