Large-Scale MIMO Secure Transmission with Finite Alphabet Inputs

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Abstract—In this paper, we investigate secure transmission over the large-scale multiple-antenna wiretap channel with finite alphabet inputs. First, we show analytically that a generalized singular value decomposition (GSVD) based design, which is optimal for Gaussian inputs, may exhibit a severe performance loss for finite alphabet inputs in the high signal-to-noise ratio (SNR) regime. In light of this, we propose a novel Per-Group-GSVD (PG-GSVD) design which can effectively compensate the performance loss caused by the GSVD design. More importantly, the computational complexity of the PG-GSVD design is by orders of magnitude lower than that of the existing design for finite alphabet inputs in [1] while the resulting performance loss is minimal. Numerical results indicate that the proposed PG-GSVD design can be efficiently implemented in large-scale multiple-antenna systems and achieves significant performance gains compared to the GSVD design.

I. INTRODUCTION

Security is a critical issue for future 5G wireless networks. In today’s systems, the security provisioning relies on bit-level cryptographic techniques and associated processing techniques at various stages of the data protocol stack. However, these solutions have severe drawbacks and many weaknesses of standardized protection mechanisms for public wireless networks are well known; although enhanced ciphering and authentication protocols exist, they impose severe constraints and high additional costs for the users of public wireless networks. Therefore, new security approaches based on information theoretical considerations have been proposed and are collectively referred to as physical layer security [2–6].

Most existing work on physical layer security assumes that the input signals are Gaussian distributed. Although the Gaussian codebook has been proved to achieve the secrecy capacity of the Gaussian wiretap channel [3], the signals employed in practical communication systems are non-Gaussian and are often drawn from discrete constellations [7–10]. For the multiple-input, multiple-output, multiple antenna eavesdropper (MIMOME) wiretap channel with perfect channel state information (CSI) of both the desired user and the eavesdropper at the transmitter, a generalized singular value decomposition (GSVD) based precoding design was proposed to decouple the corresponding wiretap channel into independent parallel subchannels [11]. Then, the optimal power allocation policy across these subchannels was obtained by an iterative algorithm. However, the simulation results in [1] revealed that for finite alphabet inputs, the GSVD design is suboptimal. In fact, the iterative algorithm in [1] can significant improve the secrecy rate by directly optimizing the precoder matrix. Very recently, for the case when imperfect CSI of the eavesdropper is available at the transmitter, a secure transmission scheme was proposed in [12] based on the joint design of the transmit precoder matrix to improve the achievable rate of the desired user and the AN generation scheme to degrade the achievable rate of the eavesdropper. However, the computational complexities of the algorithms in [1] and [12] scale exponentially with the number of transmit antennas. Therefore, the algorithms in [1, 12] become intractable even for a moderate number of transmit antennas (e.g., eight).

In this paper, we investigate the secure transmission design for the large-scale MIMOME wiretap channel with finite alphabet inputs and perfect CSI of the desired user and the eavesdropper at the transmitter. The contributions of our paper are summarized as follows:

1) We derive an upper bound on the secrecy rate for finite alphabet inputs in the high SNR regime when the GSVD design is employed. The derived expression shows that, when \( N_t > N_1 \), in the high SNR regime, the GSVD design will result in at least \( (N_t - N_1) \log M \) b/s/Hz rate loss compared to the maximal rate for the MIMOME wiretap channel, where \( N_t \), \( N_1 \), and \( M \) denote the number of transmit antennas, the rank of the intended receiver’s channel, and the size of the input signal constellation set, respectively.

2) To tackle this issue, we propose a novel Per-Group-GSVD (PG-GSVD) design, which pairs different subchannels into different groups based on the GSVD structure. We prove that the proposed PG-GSVD design can eliminate the performance loss of the GSVD design with an order of magnitude lower computational complexity than the design in [1]. Accordingly, we propose an iterative algorithm based on the gradient descent method to optimize the secrecy rate.

3) Simulation results illustrate that the proposed designs are well suited for large-scale MIMO wiretap channels and achieve substantially higher secrecy rates than the GSVD design while requiring a much lower computational complexity than the precoder design in [1].
Notation: Vectors are denoted by lower-case bold-face letters; matrices are denoted by upper-case bold-face letters. Superscripts $(\cdot)^T$, $(\cdot)^*$, and $(\cdot)^H$ stand for the matrix transpose, conjugate, and conjugate-transpose operations, respectively. We use $\text{tr}(A)$ and $A^{-1}$ to denote the trace and the inverse of matrix $A$, respectively. $\perp$ denotes the orthogonal complement of a subspace. $\text{diag}\{b\}$ denotes a diagonal matrix with the elements of vector $b$ on its main diagonal. $\text{Diag}\{B\}$ denotes a diagonal matrix containing in the main diagonal the diagonal elements of matrix $B$. The $M \times M$ identity matrix is denoted by $\mathbf{I}_M$, and the all-zero $M \times N$ matrix and the all-zero $N \times 1$ vector are denoted by $\mathbf{0}$. The fields of complex numbers and real numbers are denoted by $\mathbb{C}$ and $\mathbb{R}$, respectively. $E[\cdot]$ denotes statistical expectation. $[A]_{mn}$ denotes the element in the $m$th row and $n$th column of matrix $A$. $[a]_m$ denotes the $m$th entry of vector $a$. We use $x \sim \mathcal{C}\mathcal{N}(0, \mathbf{R}_N)$ to denote a circularly symmetric complex Gaussian vector $x \in \mathbb{C}^{N \times 1}$ with zero mean and covariance matrix $\mathbf{R}_N$. $\text{null}(A)$ denotes the null space of matrix $A$.

II. SYSTEM MODEL

We study the MIMOME wiretap channel with a multiple-antenna transmitter (Alice), a multiple-antenna intended receiver (Bob), and a multiple-antenna eavesdropper (Eve), where the corresponding numbers of antennas are denoted by $N_t$, $N_r$, and $N_e$, respectively. The signals received at Bob and Eve are denoted by $y_b$ and $y_e$, respectively, and can be written as

$$y_b = \mathbf{H}_{ba} \mathbf{G} \mathbf{x}_a + \mathbf{n}_b$$

$$y_e = \mathbf{H}_{ea} \mathbf{G} \mathbf{x}_a + \mathbf{n}_e$$

where $\mathbf{x}_a = [x_1, x_2, \cdots, x_{N_t}]^T \in \mathbb{C}^{N_t \times 1}$ denotes the transmitted signal vector having zero mean and the identity matrix as covariance matrix, and $\mathbf{H}_{ba} \in \mathbb{C}^{N_r \times N_t}$ and $\mathbf{H}_{ea} \in \mathbb{C}^{N_e \times N_t}$ denote the channel matrices between Alice and Bob and between Alice and Eve, respectively. The complex independent identically distributed (i.i.d.) vectors $\mathbf{n}_b \sim \mathcal{C}\mathcal{N}(0, \mathbf{\Sigma}_b)$ and $\mathbf{n}_e \sim \mathcal{C}\mathcal{N}(0, \mathbf{\Sigma}_e)$ represent the channel noises at Bob and Eve, respectively. $\mathbf{G} \in \mathbb{C}^{N_t \times N_i}$ is a linear precoding matrix that has to be optimized for maximization of the secrecy rate. The precoding matrix has to satisfy the power constraint

$$\text{tr}\left\{\mathbb{E}[\mathbf{G} \mathbf{x}_a \mathbf{x}_a^H \mathbf{G}^H]\right\} = \text{tr}\left\{\mathbf{G} \mathbf{G}^H\right\} \leq P.$$  

In order to be able to present the basic idea behind PG-GSVD design in the simplest manner possible, we assume in this paper that the perfect instantaneous CSI of both the intended receiver and the eavesdropper are available at the transmitter in this paper. This assumption applies for the case where the transmitter intends to send a private message to a particular user in the system while regarding another user as eavesdropper, i.e., the eavesdropper is an idle user of the system [4, 5]. Based on the deterministic equivalent channel model for the large system limit derived in [13], the PG-GSVD design in this paper can be easily extended to the case where only the statistical CSI of the eavesdropper is available at the transmitter.

When the transmitter has perfect instantaneous knowledge of the eavesdropper’s channel, the achievable secrecy rate is given by [3]

$$C_{sec} = \max_{\text{tr}(\mathbf{G}\mathbf{G}^H) \leq P} R_{sec}(\mathbf{G})$$

$$R_{sec}(\mathbf{G}) = I(y_b; \mathbf{x}_a) - I(y_e; \mathbf{x}_a)$$

where $I(y; x)$ denotes the mutual information between input $x$ and output $y$.

The goal of this paper is to optimize the transmit precoding matrix $\mathbf{G}$ for maximization of the secrecy rate in (5) when the transmit symbols $\mathbf{x}_a$ are drawn from a discrete constellation set with $M$ equiprobable points such as $M$-ary quadrature amplitude modulation (QAM) and $N_t$ is large.

III. LOW COMPLEXITY PRECODER DESIGN WITH INSTANTANEOUS CSI OF THE EAVESDROPPER

In this section, we first provide some useful definitions which will be used in the subsequent analysis. Then, we analyze the rate loss of the GSVD design [11] compared to the maximal rate for finite alphabet inputs in the high SNR regime. Finally, we propose a PG-GSVD precoder to compensate this performance loss with low complexity.

A. Some Useful Definitions

Let us introduce some useful definitions for the subsequent analysis.

**Definition 1:** Similar to [3, 11], we define the following subspaces

$$\mathcal{S}_{ba} = \text{null}(\mathbf{H}_{ba})^\perp \cap \text{null}(\mathbf{H}_{ea})$$

$$\mathcal{S}_{be} = \text{null}(\mathbf{H}_{ba})^\perp \cap \text{null}(\mathbf{H}_{ea})^\perp$$

$$\mathcal{S}_{ea} = \text{null}(\mathbf{H}_{ba}) \cap \text{null}(\mathbf{H}_{ea})^\perp$$

$$\mathcal{S}_n = \text{null}(\mathbf{H}_{ba}) \cap \text{null}(\mathbf{H}_{ea}).$$

Define $k = \text{rank}\left(\begin{bmatrix} \mathbf{H}_{ba}^H & \mathbf{H}_{ea}^H \end{bmatrix}^H\right)$ and hence $\text{dim}(\mathcal{S}_n) = N_t - k$. In addition, define $r = \text{dim}(\mathcal{S}_{ba})$ and $s = \text{dim}(\mathcal{S}_{be})$. Therefore, $\text{dim}(\mathcal{S}_n) = k - r - s$.

**Definition 2:** Following [3], we define the GSVD of the pair $(\mathbf{H}_{ba}, \mathbf{H}_{ea})$ as follows:

$$\mathbf{H}_{ba} = \mathbf{U}_{ba} \Sigma_{ba} \begin{bmatrix} \mathbf{\Omega}_{ba}^{-1} & 0 \end{bmatrix} \mathbf{U}_{ba}^H$$

$$\mathbf{H}_{ea} = \mathbf{U}_{ea} \Sigma_{ea} \begin{bmatrix} \mathbf{\Omega}_{ea}^{-1} & 0 \end{bmatrix} \mathbf{U}_{ea}^H$$

where $\mathbf{U}_{a} \in \mathbb{C}^{N_t \times N_t}, \mathbf{U}_{ba} \in \mathbb{C}^{N_r \times N_t}$, and $\mathbf{U}_{ea} \in \mathbb{C}^{N_e \times N_t}$ are unitary matrices. $\mathbf{\Omega}_{ba} \in \mathbb{C}^{k \times k}$ is a non-singular matrix with diagonal elements $\omega_i$, $i = 1, \ldots, k$. $\Sigma_{ba} \in \mathbb{C}^{N_t \times k}$ and $\Sigma_{ea} \in \mathbb{C}^{N_e \times k}$ can be expressed as

$$\Sigma_{ba} = \begin{bmatrix} \mathbf{U}_{ba} & \mathbf{0} \end{bmatrix} \begin{bmatrix} \mathbf{\Omega}_{ba}^{-1} & 0 \\ 0 & \mathbf{I}_{k-r} \end{bmatrix} \begin{bmatrix} \mathbf{U}_{ba}^H \\ \mathbf{0} \end{bmatrix}$$

$$\Sigma_{ea} = \begin{bmatrix} \mathbf{U}_{ea} & \mathbf{0} \end{bmatrix} \begin{bmatrix} \mathbf{\Omega}_{ea}^{-1} & 0 \\ 0 & \mathbf{I}_{r} \end{bmatrix} \begin{bmatrix} \mathbf{U}_{ea}^H \\ \mathbf{0} \end{bmatrix}$$
For finite alphabet signals is upper bounded by
\[
\hat{\Sigma}_{ea} = \begin{bmatrix}
I_{k-r-s} & s & r \\
0 & D_e & 0 \\
N_e - k + r & 0 & 0
\end{bmatrix}
\]
(9)
where \( D_b = \text{diag}([b_1, \ldots, b_s]) \in \mathbb{R}^{s \times s} \) and \( D_e = \text{diag}([e_1, \ldots, e_s]) \in \mathbb{R}^{s \times s} \) are diagonal matrices with real valued entries. The diagonal elements of \( D_b \) and \( D_e \) are ordered as follows:
\[
0 < b_1 \leq b_2 \leq \ldots \leq b_s < 1
\]
\[
1 > e_1 \geq e_2 \geq \ldots \geq e_s > 0
\]
and
\[
b_p^2 + e_p^2 = 1, \text{ for } p = 1, \ldots, s.
\]

B. Performance Loss of the GSVD Design

The precoding matrix for the GSVD design can be expressed as [11]
\[
G = U_a A P^\frac{1}{2}
\]
(10)
where \( P = \text{diag}(p_1, \ldots, p_{N_t}) \) represents a diagonal power allocation matrix and \( A \) is given by
\[
A = \begin{bmatrix}
k & N_t - k \\
N_t - k & 0 & 0
\end{bmatrix}
\]
(11)
For the GSVD precoder design in (10), the received signals \( y_b \) and \( y_e \) in (1) and (2) can be re-expressed as
\[
\tilde{y}_b = \hat{\Sigma}_{ba} \begin{bmatrix} k \\ N_t - k \end{bmatrix} \begin{bmatrix} 1 \\ 0 \\ 0 \\
0 & 0 & 0
\end{bmatrix} x_a + \tilde{n}_b
\]
(12)
\[
\tilde{y}_e = \hat{\Sigma}_{ea} \begin{bmatrix} k \\ N_t - k \end{bmatrix} \begin{bmatrix} 1 \\ 0 & 0 \\
0 & 0 & 0
\end{bmatrix} x_a + \tilde{n}_e
\]
(13)
where \( \tilde{y}_b = U^H_{ba} y_b, \tilde{y}_e = U^H_{ea} y_e, \tilde{n}_b = U^H_{ba} n_b, \) and \( \tilde{n}_e = U^H_{ea} n_e \).

Define \( N_1 = \text{rank}(H_{ba}) \) and \( N_2 = \text{rank}(H_{ea}) \). In the following theorem, we analyze the performance of the GSVD design for finite alphabet inputs in the high SNR regime.

Theorem 1: In the high SNR regime (\( P \to \infty \)), for the GSVD design in (10), the achievable secrecy rate \( R_{\text{sec, high}} \) for finite alphabet signals is upper bounded by
\[
R_{\text{sec, high}} \leq N_1 \log_2 M \text{ b/s/Hz.}
\]
(14)

Proof: See Appendix A.

C. PG-GSVD Precoder Design

As indicated in [18], in order to decouple the MIMO channels into \( N_t \) parallel subchannels, the MIMO channel matrix has to be an \( N_t \times N_t \) matrix. However, \( \Sigma_{ba} \) and \( \Sigma_{ea} \) in (12) and (13) are \( N_r \times N_t \) and \( N_e \times N_t \) matrices, respectively. As a result, we need to add to or remove from \( \tilde{y}_b, \Sigma_{ba}, \tilde{y}_e, \) and \( \Sigma_{ea} \) some zeros in (12) and (13). To this end, we define
\[
\hat{y}_b = \begin{bmatrix} k-r-s \\ N_t-k \end{bmatrix} \begin{bmatrix} 0 \\ \hat{y}_b^H \\
\hat{y}_e^H \\
0
\end{bmatrix},
\]
(15)
where \( \hat{y}_b \in \mathbb{C}^{(r+s) \times 1} \) is composed of the last \( r+s \) elements of \( \tilde{y}_b \). Furthermore, we define \( \omega = [\omega_1, \ldots, \omega_k] \in \mathbb{C}^{N_r \times 1} \), \( \hat{y}_e = [\hat{y}_e^H \ 0]^H \in \mathbb{C}^{N_e \times 1} \), \( \tilde{n}_b \sim \mathcal{CN}(0, \sigma_b^2 I_{N_r}) \), and \( \tilde{n}_e \sim \mathcal{CN}(0, \sigma_e^2 I_{N_e}) \). Define two diagonal matrices
\[
\hat{\Sigma}_{ba} = \begin{bmatrix} k-r-s \\ N_t-k \end{bmatrix} \begin{bmatrix} 0 \\ \hat{D}_b \\
0 & 0 \\
N_t-k & 0
\end{bmatrix}
\]
(16)
\[
\hat{\Sigma}_{ea} = \begin{bmatrix} k-r-s \\ N_t-k+r \end{bmatrix} \begin{bmatrix} 0 \\ \hat{D}_e \\
0 & 0 \\
N_t-k+r & 0
\end{bmatrix}
\]
(17)
where the elements of \( \hat{D}_b, \hat{R}_e, \hat{R}_b, \) and \( \hat{D}_e \) are obtained from the following two equations
\[
[\hat{\Sigma}_{ba}]_{i(i+k-r-s+i)} = [\Sigma_{ba}]_{i(i+k-r-s+i)} / \sqrt{\omega_i^2}, \quad i = 1, \ldots, s+r
\]
(18)
\[
\hat{\Sigma}_{ea} = [\Sigma_{ea}]_{i(i+k-r-s+i)} / \sqrt{\omega_i^2}, \quad i = 1, \ldots, k-r.
\]
(19)
We divide the transmit signal \( x_a \) into \( S \) streams and let \( N_s = N_t / S \). We define the set \( \{f_1, \ldots, f_N\} \) as a permutation of \( \{1, \ldots, N_t\} \), \( P_s \in \mathbb{C}^{N_t \times N_s} \) and \( V_s \in \mathbb{C}^{N_e \times N_s}, s = 1, \ldots, S \), denote a diagonal and a unitary matrix, respectively. \( V \in \mathbb{C}^{N_t \times N_e} \) denotes a unitary matrix. For the proposed PG-GSVD precoder, we set \( G \) as follows
\[
G = U_a A P^\frac{1}{2} V
\]
(20)
We set
\[
[\omega]_{l_i f_j} = [P_s]_{l_i f_j},
\]
(21)
where \( i = 1, \ldots, N_s \), \( s = 1, \ldots, S \), and \( j = (s-1)N_s + i \). Based on (20) and (21), the power constraint in (3) is equivalent to \( \sum_{s=1}^S \text{tr}(P_s) \leq P \).
Also, we set
\[
[V]_{l_i f_j} = \begin{cases} [V_s]_{mn} & \text{if } i = (s-1)N_s + m, \quad j = (s-1)N_s + n \\
0 & \text{otherwise}
\end{cases}
\]
For convenience, we assume \( N_e = N_t / S \) is an integer in this paper. If \( N_t / S \) is not an integer, we can easily obtain an integer \( N_e \) by adding zeros in (16) and (17).
Algorithm 1: Maximizing $R_{\text{sec}}(G)$ with respect to $P_s$ and $V_s$.

1. Initialize $P_s$ with $\sum_{s=1}^{S} \text{tr} \left( P_s^{(0)} \right) = N_t$ and $V_s^{(0)}$ for $s = 1, \ldots, S$. Set $N_{\text{iter}}$ and $\varepsilon$ as the maximum iteration number and a threshold, respectively.
2. Initialize $R_{\text{sec}}(G)^{(1)}$ based on (27). Set counter $n = 1$.
3. Update $P_s^{(n)}$ for $s = 1, \ldots, S$ along the gradient decent direction $\nabla_{P_s} R(G)$.
4. Normalize $\sum_{s=1}^{S} \text{tr} \left( P_s^{(n)} \right) = P$.
5. Update $V_s^{(n)}$ for $s = 1, \ldots, S$ along the gradient decent direction $\nabla_{V_s} R(G)$.
6. Compute $R_{\text{sec}}(G)^{(n+1)}$ based on (27). If $R_{\text{sec}}(G)^{(n+1)} - R_{\text{sec}}(G)^{(n)} > \varepsilon$ and $n \leq N_{\text{iter}}$, set $n = n + 1$ and repeat Steps 3–5;
7. Compute $P$ and $V$ based on (21) and (22). Set $G = U_n A \hat{P}^2 V$, where $m = 1, \ldots, N_s$, $n = 1, \ldots, N_s$, $s = 1, \ldots, S$, $i = 1, \ldots, N_t$, and $j = 1, \ldots, N_l$. Finally, we let

$$[x_i]_{t_j} = [x_n]_{e_{t_j}}.$$ (23)

Based on (20)–(23) and a paring scheme $\{t_1, \ldots, t_N\}$, the equivalent received signals at Bob and Eve can be decoupled as follows

$$\hat{y}_b_{t_j} = \left[ \hat{\Sigma}_{ba} \right]_{t_j} \hat{x}_b_{t_j} + [\hat{n}_b]_{t_j},$$ (24)

$$\hat{y}_e_{t_j} = \left[ \hat{\Sigma}_{ea} \right]_{t_j} \hat{x}_e_{t_j} + [\hat{n}_e]_{t_j},$$ (25)

where

$$[\hat{x}]_{t_j} = \left[ P_s^{2} V_s x_s \right]_{i}.$$ (26)

for $i = 1, \ldots, N_s$, $s = 1, \ldots, S$, and $j = (s-1)N_s + i$. From (24) and (25), we observe that the transmit signal has been divided into $S$ independent groups. In each group, the equivalent signal dimension is $N_s \times 1$. We further define $\hat{y}_b_{t_j} = [y_b]_{s_j}$ and $\hat{y}_e_{t_j} = [y_e]_{s_j}$.

Based on (24) and (25), the secrecy rate in (5) can be expressed as

$$R_{\text{sec}}(G) = \sum_{s=1}^{S} \left( I(y_{b,s}; x_s) - I(y_{e,s}; x_s) \right).$$ (27)

The gradients of $I(y_{b,s}; x_s)$ and $I(y_{e,s}; x_s)$ with respect to $P_s$ and $V_s$ can be found in [19, Eq. (22)], based on which an iterative algorithm can be derived for maximizing $R_{\text{sec}}(G)$, as given in Algorithm 1.

For the precoder design with finite alphabet inputs, the computational complexity is mainly dominated by the required number of additions in calculating the mutual information and the MSE matrix when $N_t$ is large [15]. Considering the decoupled structure in (27), the computational complexity of Algorithm 1 grows linearly with $SM^{2N_s}$. However, the computational complexity of the algorithm in [1] scales linearly with $M^{2N_t}$. We observe that for large-scale MIMO systems when $N_t$ is large, the computational complexity of Algorithm 1 is by orders of magnitude lower than the algorithm in [1].

### Table I: Number of additions required for calculating the mutual information and the MSE matrix for the system considered in Figure 1.

| $4 \times 3 \times 2$ | BPSK | QPSK |
|----------------------|------|------|
| GSVD                 | 8    | 16   |
| Algorithm 1          | 32   | 512  |
| Algorithm 1 in [1]   | 256  | 65536|

### Table II: Number of additions required for calculating the mutual information and the MSE matrix for the system considered in Figure 2.

| $64 \times 48 \times 48$ | BPSK | QPSK |
|--------------------------|------|------|
| GSVD                     | 128  | 256  |
| Algorithm 1              | 512  | 8192 |
| Algorithm 1 in [1]       | 3.04e+038 | 1.13e+071 |

For the PG-GSVD design in (20), we have the following theorem.

**Theorem 2:** If the inequality $(k - N_2)N_s \geq N_t$, then we can always find a permutation $\{\ell_1, \ldots, \ell_N\}$ for the PG-GSVD design in (20), which achieves $R_{\text{sec,high}} = N_t \log_2 M$ b/s/Hz in the high SNR regime.

**Proof:** See Appendix B.

The algorithm in [1] is equivalent to setting $N_s = N_t$ in Algorithm 1. Therefore, as long as $k - N_2 \neq 0$, it can compensate the performance loss of the GSVD design and achieve the saturation rate $N_t \log_2 M$ b/s/Hz in the high SNR regime, as shown in [1, Figs. 1, 2]. However, in this case, the computational complexity of the algorithm in [1] grows exponentially with $N_t$. This is prohibitive in large-scale MIMO systems. For typical large-scale MIMO systems, we have $N_t > N_2$ [14, 15], which implies $k - N_2 \neq 0$. As a result, by properly choosing $N_s$, we can reach a favorable trade-off between complexity and secrecy rate performance.

### IV. Numerical Results

We set $\sigma_g = \sigma_e$ and define $SNR = P/(N_s \sigma_e^2)$. Furthermore, we use $N_t \times N_r \times N_e$ to denote the simulated wiretap channel.

#### A. Scenarios with Instantaneous CSI of the Eavesdropper

In this subsection, the elements of $H_{ba}$ and $H_{ea}$ are generated independently and randomly. Tables I and II compare the computational complexities of the different schemes for the systems considered in Figures 1 and 2, respectively.

Figure 1 plots the secrecy rate for the $4 \times 3 \times 2$ wiretap channel for different precoder designs and different modulation schemes for $N_s = 2$. We observe from Figure 1 that Algorithm 1 achieves a similar performance as the precoder design in [1] but with orders of magnitude lower computational complexity as indicated in Table I. Both designs achieve the maximal rate $N_t \log_2 M$ b/s/Hz in the high SNR regime as indicated by Theorem 2. In contrast, the GSVD design yields an obvious rate loss in the high SNR regime. For the channels of Bob and Eve, we have $D_{b,1} = 0.57$, $D_{e,1} = 0.81$. As explained in Example 1, the GSVD design sets $p_1 = p_2 = 0$ in this case. Therefore, the GSVD design suffers from a 2 log$_2 M$ exponent.
expression reveals that the GSVD design may lead to a serious performance loss. Based on this, we proposed a PG-GSVD design to overcome the negative properties of the GSVD design while retaining an affordable computational complexity for large-scale MIMO systems. Simulations indicated that the proposed design performs well in large-scale MIMOME wiretap channels and achieves substantial secrecy rate gains compared to the GSVD design for finite alphabet inputs.

APPENDIX A

PROOF OF THEOREM 1

Based on (6) and (12), \( I(y_b; x_a) \) in (5) for the GSVD design becomes

\[
I(y_b; x_a) = \sum_{i=1}^{s} I(b_i^2 p_{k-r-s+i}) + \sum_{i=1}^{r} I(p_{k-s+i})
\]

(28)

where \( I(\gamma) = I(x; \sqrt{T}x + n) \). Therefore, for \( P \to \infty \), we obtain

\[
\lim_{P \to \infty} I(y_b; x_a) \leq (s + r) \log_2(M).
\]

(29)

According to Inclusion–Exclusion Principle [20], we know

\[
dim(S_{ba}) + dim(S_{be}) = dim(S_{ba} \cup S_{be}) - dim(S_{ba} \cap S_{be}).
\]

(30)

For the subspaces \( S_{ba} \) and \( S_{be} \), we have

\[
S_{ba} \cap S_{be} = \left( \text{null}(H_{ba})^\perp \cap \text{null}(H_{ea}) \right) \cap \left( \text{null}(H_{ba})^\perp \cap \text{null}(H_{ea}) \right)
\]

(31a)

\[
= \left( \text{null}(H_{ba})^\perp \cap \text{null}(H_{ea}) \right) \cap \left( \text{null}(H_{ea}) \cap \text{null}(H_{ea}) \right)
\]

(31b)

\[
= \text{null}(H_{ba})^\perp \cap \left( \text{null}(H_{ea}) \cap \text{null}(H_{ea}) \right)
\]

(31c)

\[
= \emptyset
\]

(31d)

where (31b) and (31c) are obtained based on the properties of intersections [21].

Also, we have

\[
S_{ba} \cup S_{be} = \left( \text{null}(H_{ba})^\perp \cap \text{null}(H_{ea}) \right) \cup \left( \text{null}(H_{ba})^\perp \cap \text{null}(H_{ea}) \right)
\]

(32a)

\[
= \text{null}(H_{ba})^\perp \cap \left( \text{null}(H_{ea}) \cup \text{null}(H_{ea}) \right)
\]

(32b)

\[
= \text{null}(H_{ba})^\perp
\]

(32c)

where (32b) and (32c) are obtained based on the Distributive Law of sets [21] and the Rank–Nullity Theorem [22], respectively.

From (30)–(32), we obtain

\[
s + r = \dim(S_{ba}) + \dim(S_{be}) = \dim(\text{null}(H_{ba})^\perp).
\]

(33)
Assuming \( \mathbf{v}_i \in \mathbb{C}^{N_t \times 1} \) and \( \mathbf{u}_j \in \mathbb{C}^{N_r \times 1} \) are the \( N_t \) left and \( N_r \) right singular vectors of \( \mathbf{H}_{ba} \), respectively, \( i = 1, \ldots, N_t \), \( j = 1, \ldots, N_r \), \( \mathbf{H}_{ba} \) can be written as
\[
\mathbf{H}_{ba} = \sum_{i=1}^{N_t} \lambda_i \mathbf{u}_i \mathbf{v}_i^H
\]
(34)
where \( \lambda_i \) is the singular value of \( \mathbf{H}_{ea} \). For \( N_1 < N_t \), we have
\[
\text{null} (\mathbf{H}_{ba}) = \sum_{i=N_1+1}^{N_t} \omega_i \mathbf{v}_i \mathbf{v}_i^H
\]
(35)
where \( \omega_i \) denotes an arbitrary non-zero complex value, \( i = 1, \ldots, N_t \). Based on the property of the orthogonal complement of a subspace [23], we obtain
\[
\left( \text{null} (\mathbf{H}_{ba}) \right) ^\perp = \left( \sum_{i=N_1+1}^{N_t} \omega_i \mathbf{v}_i \mathbf{v}_i^H \right) ^\perp
\]
(36a)
\[
= \text{null} \left( \sum_{i=N_1+1}^{N_t} \omega_i \mathbf{v}_i \mathbf{v}_i^H \right)
\]
(36b)
\[
= \sum_{i=1}^{N_t} \omega_i \mathbf{v}_i \mathbf{v}_i^H.
\]
(36c)
Therefore, we have
\[
\dim \left( \text{null} (\mathbf{H}_{ba}) \right) ^\perp = N_1.
\]
(37)
For \( N_1 = N_t \), \( \text{null} (\mathbf{H}_{ea}) = \emptyset \), and we obtain
\[
\dim \left( \text{null} (\mathbf{H}_{ba}) \right) ^\perp = N_t.
\]
(38)
Combining (5), (29), (33), (37), and (38) completes the proof.

**APPENDIX B**

**PROOF OF THEOREM 2**

The key idea of achieving the maximal rate \( N_t \log M \) b/s/Hz in the high SNR regime is to guarantee that all \( N_t \) signals can be received by Bob but not by Eve. To achieve this, \( N_s \) signals are combined into a group and transmitted along the subchannels \( \mathbf{R}_t \) in (16). As a result, we need to analyze the dimension of \( \mathcal{S}_{ba} \).

Based on the Inclusion–Exclusion Principle [20], we have
\[
\dim (\mathcal{S}_{ba}) + \dim (\mathcal{S}_n) = \dim (\mathcal{S}_{ba} \cup \mathcal{S}_n) - \dim (\mathcal{S}_{ba} \cap \mathcal{S}_n).
\]
(39)
Following similar steps as in (31) and (32), we obtain
\[
\mathcal{S}_{ba} \cap \mathcal{S}_n = \left( \text{null} (\mathbf{H}_{ba}) ^\perp \cap \text{null} (\mathbf{H}_{ea}) \right) \cap \left( \text{null} (\mathbf{H}_{ba}) \cap \text{null} (\mathbf{H}_{ea}) \right)
\]
(40a)
\[
= \left( \text{null} (\mathbf{H}_{ea}) \cap \text{null} (\mathbf{H}_{ba}) ^\perp \right) \cap \left( \text{null} (\mathbf{H}_{ba}) \cap \text{null} (\mathbf{H}_{ea}) \right)
\]
(40b)
\[
= \text{null} (\mathbf{H}_{ea}) \cap \left( \text{null} (\mathbf{H}_{ba}) ^\perp \cap \text{null} (\mathbf{H}_{ba}) \right) \cap \text{null} (\mathbf{H}_{ea})
\]
(40c)
\[
= \emptyset
\]
(40d)
and
\[
\mathcal{S}_{ba} \cup \mathcal{S}_n = \left( \text{null} (\mathbf{H}_{ba}) ^\perp \cap \text{null} (\mathbf{H}_{ea}) \right) \cup \left( \text{null} (\mathbf{H}_{ba}) \cap \text{null} (\mathbf{H}_{ea}) \right)
\]
(41a)
\[
= \text{null} (\mathbf{H}_{ea}) \cap \left( \text{null} (\mathbf{H}_{ba}) ^\perp \cup \text{null} (\mathbf{H}_{ea}) \right)
\]
(41b)
\[
= \text{null} (\mathbf{H}_{ea}).
\]
(41c)
Since \( \text{rank} (\mathbf{H}_{ea}) = N_2 \), we have \( \dim (\text{null} (\mathbf{H}_{ea})) = N_t - N_2 \). Then, based on (39), (40d), (41c), we obtain
\[
N_1 - k = N_t - N_2.
\]
(42)
From (42), we know \( r = k - N_2 \).

When \( (k - N_2) N_s \geq N_t \), we design the PG-GSVD precoder in (20) as follows. We set
\[
P = \begin{pmatrix}
0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 \\
0 & 0 & \text{diag} (p_1, \ldots, p_r) & 0 \\
0 & 0 & 0 & 0 \\
\end{pmatrix}.
\]
(43)
Also, we select a pairing scheme \( \{\ell_1, \ldots, \ell_{N_t}\} \) in (21) satisfying
\[
[\mathbf{P} \mathbf{s}]_{ii} = \begin{cases}
0 & \text{if } 1 \leq i \leq N_s - 1 \\
p_i & \text{if } i = N_s
\end{cases}
\]
(44)
for \( s = 1, \ldots, S \), \( i = 1, \ldots, N_s \), and \( j = 1, \ldots, r \).

Based on the design in (43) and (44), in the high SNR regime, we have
\[
I (y_{bs}; x_s) \rightarrow_{N_t \rightarrow \infty} N_s \log M
\]
(45)
\[
I (y_{es}; x_s) = 0.
\]
(46)
Substituting (45) and (46) into (27) completes the proof.

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