OPTIMIZATION OF TRANSMIT SIGNALS TO INTERFERE EAVESDROPPING IN A WIRELESS LAN

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ABSTRACT

We consider physical-layer security of a wireless LAN where multiple receivers collude to eavesdrop the information from the basestation to the intended receiver. To enhance the physical-layer security, we design the interference signals to combat the eavesdropping. Our design problems are resolved using semidefinite relaxation problems, which can be numerically solved efficiently by the existing convex optimization solvers. Simulation results are provided to demonstrate the efficiency of the proposed designs.

Index Terms— physical-layer secrecy, beamforming, eavesdropping

1. INTRODUCTION

Wireless communications are susceptible to eavesdropping, since they can be inevitably overheard by eavesdroppers within a certain range in an open environment. One of the most commonly used methods to secure communications in the presence of third parties such as eavesdroppers is cryptography. However, the secret key distribution and management necessary for the encryption remain vulnerable to eavesdropping.

When the channel state information (CSI) of the eavesdropper is known, theoretically, secrecy is guaranteed if the communication rate between the transmitter and the legitimate receiver is lower than the so-called secrecy capacity, which is the maximum rate at which the transmitter can send the secret information to the legitimate receiver without being decoded by the eavesdropper.

In a wireless LAN, the basestation knows the CSIs of the active connected receivers. Once eavesdroppers' CSIs are available at the transmitter, one candidate to realize physical-layer secrecy is to employ a system in which the transmitter can degrade the quality of the received signals of eavesdroppers by sending interference signals, while keeping the quality of the received signals at the legitimate receiver by using their CSIs.

For MIMO systems, secrecy capacity has been well studied in terms of information theoretical point of view [7, 8, 10]. Secrecy capacity has been also investigated in [4] for systems where the transmitter equipped with multiple antennas sends secret information signals as well as interference signals. In theory, physical-layer secrecy is improved by using cooperative relays [2] and a separate transmitter that sends an interference signal [13]. Even when the transmitter does not know locations and CSIs of eavesdroppers, physical-layer secrecy has been characterized in [3] and beamforming as well as artificial noise broadcasting has been developed to increase communications security in [12]. In [6], a transmit beamforming with imperfect CSI has been proposed for MISO channels with direct transmission as well as cooperative jamming as a helper. However, the realization of physical-layer secrecy has not yet been fully established.

In [9], design problems of the signals transmitted from multiple antennas are proposed: The signal-to-interference-and-noise-ratios (SINR) of eavesdroppers are constrained to be low enough for decoding the secret information, while the SINR of the legitimate receiver is kept sufficiently large for decoding or is maximized under the maximum transmit power constraint. Moreover, [9] studies the case when multiple eavesdroppers cooperate with each other to form a joint received beamforming to improve their SINR, which can be mathematically equivalent to the case where one eavesdropper having multiple receive antennas forms an optimal receive beamforming. Beamformers to combat multiple colluding eavesdroppers are also studied in [11], where solutions are searched in a restricted space to utilize semidefinite programming (SDP). However, these beamformers for colluding eavesdropping are not optimal. In this paper, we design optimal beamformers against colluding users in a wireless LAN.

We first explain the communication scenario for our proposed design where SINR is utilized to evaluate the received signal level for legitimate receiver and other receivers. Since our design problems are not convex, we reformulate them as semidefinite relaxation (SDR) problems which can be proved to yield optimal solutions for the original problems. Numerical examples are provided to demonstrate the superior performance of our proposed designs over the SDR designs in [9] and the SDP design in [11].

2. SYSTEM MODEL AND PROBLEM STATEMENT

Let us consider a wireless LAN where the basestation has N_t transmit antennas and each mobile terminal in the wireless LAN has one antenna. For the simplicity of presentation, we assume that channels between the basestation and terminals are quasi-static flat fading.

Let $\boldsymbol{x}(t)$ be the transmitted signal vector at time t whose nth entry is the signal transmitted from the nth transmit antenna. Suppose that the basestation transmits secret information to a legitimate receiver. The signal $y_b(t)$ of the legitimate receiver is modeled as $y_b(t) = \boldsymbol{h}^{\mathcal{H}}\boldsymbol{x}(t) + n(t)$, where $\boldsymbol{h}$ is an $N_t \times 1$ channel vector, whose nth entry is the complex conjugate of the channel coefficient from the nth transmit antenna to the receiver, $()^{\mathcal{H}}$ stands for the complex conjugate transpose of a matrix or a vector, and $n(t)$ denotes an additive noise, which is assumed to be independent and identically distributed (i.i.d.) complex circular Gaussian with zero mean and variance σ_n^2.

The remaining receiver in the wireless LAN can overhear the secret information from transmitter to the legitimate receiver. Let us assume that there are M receivers in addition to the legitimate receiver. The signal at the mth receiver can be expressed as

$$y_{e,m}(t) = \boldsymbol{g}_m^{\mathcal{H}}\boldsymbol{x}(t) + v_m(t), \quad m = 1, \ldots, M \tag{1}$$

where $\boldsymbol{g}_m$ is an $N_t \times 1$ channel vector, whose nth entry is the complex conjugate of the channel coefficient from the nth transmit antenna to the mth receiver, and the additive noise $v_m(t)$ at the mth receiver is i.i.d. complex circular Gaussian with zero mean and non-zero variance $\sigma_{v,m}^2 > 0$. We assume that $\{v_m(t)\}_{m=1}^{M}$ are independent of each other and of $n(t)$.

Following the convention, we also call the basestation, the legitimate receiver, and the remaining receivers as Alice, Bob, and Eves, respectively.

Let the secret information data that Alice wants to inform only to Bob be $s(t)$, which is assumed to have zero mean and unit variance. Suppose Eves try to eavesdrop $s(t)$ by collecting their received signals as a vector defined as

$$\boldsymbol{y}_e(t) \; = \; [y_{e,1}(t), \ldots, y_{e,M}(t)]^T = \boldsymbol{G}^{\mathcal{H}}\boldsymbol{x}(t) + \boldsymbol{v}(t), \tag{2}$$

where $\boldsymbol{G} = [\boldsymbol{g}_1, \ldots, \boldsymbol{g}_M]$ and $\boldsymbol{v}(t) = [v_1(t), \ldots, v_M(t)]^T$. It should be remarked that the same model can be obtained if there is one eavesdropper that has M receive antennas.

We assume that there is no inactive eavesdropper and there are at most $N_t + 1$ receivers in the wireless LAN. To improve the signal-to-interference-and-noise-ratio (SINR) at Bob, Alice utilizes transmit beamforming. At the same time, to interfere the eavesdropping, Alice sends the interference signal $z_n(t)$ from her nth transmit antenna. This is the so-called artificial noise (AN) aided (transmit) beamforming [4], whose transmitted signal vector can be expressed as

$$\boldsymbol{x}(t) = \boldsymbol{w}s(t) + \boldsymbol{z}(t) \tag{3}$$

where the nth entry of $\boldsymbol{w}$ denotes the weight at the nth transmit antenna and the interference noise vector $\boldsymbol{z}(t)$ is given by $\boldsymbol{z}(t) = [z_1(t), \ldots, z_{N_t}(t)]^T$. We assume that $\boldsymbol{z}(t)$ is i.i.d. circular Gaussian with zero mean and covariance matrix $\boldsymbol{\Sigma}$, which is positive semidefinite.

From (3), the SINR at Bob is found to be

$$\mathrm{SINR}_b(\boldsymbol{w}, \boldsymbol{\Sigma}) = \frac{|\boldsymbol{w}^{\mathcal{H}}\boldsymbol{h}|^2}{\boldsymbol{h}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{h} + \sigma_n^2}. \tag{4}$$

Likewise, the SINR at the antenna of each Eve can be expressed as

$$\mathrm{SINR}_{e,m}(\boldsymbol{w}, \boldsymbol{\Sigma}) = \frac{\boldsymbol{g}_m^{\mathcal{H}}\boldsymbol{w}\boldsymbol{w}^{\mathcal{H}}\boldsymbol{g}_m}{\boldsymbol{g}_m^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{g}_m + \sigma_{v,m}^2}, \quad m = 1, \ldots, M. \tag{5}$$

If Eves utilize the maximum SINR receive beamforming vector, then SINR of Eves can be improved such that

$$\mathrm{SINR}_{ce}(\boldsymbol{w}, \boldsymbol{\Sigma}) = \max_{\boldsymbol{r} \neq \boldsymbol{0}} \frac{\boldsymbol{r}^{\mathcal{H}}\boldsymbol{G}^{\mathcal{H}}\boldsymbol{w}\boldsymbol{w}^{\mathcal{H}}\boldsymbol{G}\boldsymbol{r}}{\boldsymbol{r}^{\mathcal{H}}(\boldsymbol{G}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{G} + \boldsymbol{D}^2)\boldsymbol{r}} \tag{6}$$

where $\boldsymbol{r}$ denotes the receive beamforming weight at the antennas of Eve and

$$\boldsymbol{D}^2 = \mathrm{diag}(\sigma_{v,1}^2, \ldots, \sigma_{v,M}^2). \tag{7}$$

We would like to optimize the transmit beamforming vector $\boldsymbol{w}$ and covariance matrix $\boldsymbol{\Sigma}$ so that Bob's and Eve's SINRs satisfy certain expected thresholds under the condition that $\boldsymbol{h}$ and $\boldsymbol{G}$ are available at the transmitter as formulated in [11]. The first one is minimizing the total transmit power:

Problem 1 Design $\boldsymbol{w}$ and $\boldsymbol{\Sigma}$ that minimize the transmit power subject to the constraints that the SINR of Bob is larger than or equal to the threshold γ_b and that the SINR of Eve is smaller than or equal to the threshold γ_{ce}.

$$\min_{\boldsymbol{w}, \boldsymbol{\Sigma}} \quad ||\boldsymbol{w}||^2 + \mathrm{trace}\,\boldsymbol{\Sigma} \tag{8a}$$

$$\text{s.t.} \quad \mathrm{SINR}_b(\boldsymbol{w}, \boldsymbol{\Sigma}) \geq \gamma_b \tag{8b}$$

$$\mathrm{SINR}_{ce}(\boldsymbol{w}, \boldsymbol{\Sigma}) \leq \gamma_{ce} \tag{8c}$$

$$\boldsymbol{\Sigma} \succeq 0, \tag{8d}$$

where $\boldsymbol{A} \succeq \boldsymbol{B}$ means that $\boldsymbol{A} - \boldsymbol{B}$ is positive semidefinite.

The second one is maximizing Bob's SINR:

Problem 2 Design $\boldsymbol{w}$ and $\boldsymbol{\Sigma}$ that maximize Bob's SINR subject to the constraints that the SINR of Eve is smaller than or equal to the threshold γ_{ce} and that the maximum transmit power is not more than P_{max}.

$$\max_{\boldsymbol{w}, \boldsymbol{\Sigma}} \quad \mathrm{SINR}_b(\boldsymbol{w}, \boldsymbol{\Sigma}) \tag{9a}$$

$$\text{s.t.} \quad ||\boldsymbol{w}||^2 + \mathrm{trace}\,\boldsymbol{\Sigma} \leq P_{max} \tag{9b}$$

$$\mathrm{SINR}_{ce}(\boldsymbol{w}, \boldsymbol{\Sigma}) \leq \gamma_{ce} \tag{9c}$$

$$\boldsymbol{\Sigma} \succeq 0. \tag{9d}$$

Similar problems have been studied in [9], where each SINR of Eve's receive antenna is constrained to be less than or equal to a threshold γ_e, that is,

$$\mathrm{SINR}_{e,m}(\boldsymbol{w},\boldsymbol{\Sigma}) \le \gamma_e, \quad m = 1,\ldots,M. \tag{10}$$

However, since $\mathrm{SINR}_{ce}(\boldsymbol{w},\boldsymbol{\Sigma})$ is an upper bound of SINR attained by eavesdropping, we do not require the constraints (10) imposed by the design in [9].

3. DESIGN WITH SEMIDEFINITE RELAXATION

Without loss of generality, we assume that $\boldsymbol{G}$ has full column rank. We do not consider the case that one receiver is located at the same place with the legitimate receiver and then $\boldsymbol{h}$ is in the column space of $\boldsymbol{G}$. Thus we assume that $\boldsymbol{h}$ is not in the column space of $\boldsymbol{G}$. Otherwise, SINR_{ce} is always not smaller than SINR_b.

Since (8b) is not convex, original problems are not convex. Thus, it is not easy to directly solve the problems. To enable optimal designs by convex optimization, we resort to semidefinite relaxation (SDR) of the original problems.

With $\boldsymbol{W} = \boldsymbol{w}\boldsymbol{w}^{\mathcal{H}}$, (8b) can be expressed as

$$\frac{1}{\gamma_b}\,\mathrm{trace}\left(\boldsymbol{W}\boldsymbol{h}\boldsymbol{h}^{\mathcal{H}}\right) - \boldsymbol{h}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{h} \ge \sigma_n^2, \boldsymbol{W} \succeq 0, \mathrm{rank}(\boldsymbol{W}) = 1. \tag{11}$$

In general, SDR ignores some constraints to make the optimization problem to be semidefinite. Here we remove the rank condition $\mathrm{rank}(\boldsymbol{W}) = 1$ to obtain the constraint

$$\frac{1}{\gamma_b}\,\mathrm{trace}\left(\boldsymbol{W}\boldsymbol{h}\boldsymbol{h}^{\mathcal{H}}\right) - \boldsymbol{h}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{h} \ge \sigma_n^2, \boldsymbol{W} \succeq 0, \tag{12}$$

which is convex in the design parameters $\boldsymbol{\Sigma}$ and $\boldsymbol{W}$. Similarly, with $\boldsymbol{W} = \boldsymbol{w}\boldsymbol{w}^{\mathcal{H}}$, (8c) can be rewritten as

$$\boldsymbol{r}^{\mathcal{H}}\boldsymbol{V}\boldsymbol{r} \succeq 0, \forall \boldsymbol{r} \neq 0, \;\; \boldsymbol{W} \succeq 0, \;\; \mathrm{rank}(\boldsymbol{W}) = 1, \tag{13}$$

where

$$\boldsymbol{V} = \gamma_{ce}(\boldsymbol{G}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{G} + \boldsymbol{D}^2) - \boldsymbol{G}^{\mathcal{H}}\boldsymbol{W}\boldsymbol{G}. \tag{14}$$

The necessary and sufficient condition for (13) is that $\boldsymbol{V}$ is non-negative definite. Without the rank condition $\mathrm{rank}(\boldsymbol{W}) = 1$, we have the following convex constraints:

$$\gamma_{ce}(\boldsymbol{G}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{G} + \boldsymbol{D}^2) - \boldsymbol{G}^{\mathcal{H}}\boldsymbol{W}\boldsymbol{G} \succeq 0, \quad \boldsymbol{W} \succeq 0. \tag{15}$$

Thus, our relaxed problem can be expressed as

$$\begin{aligned}
\min_{\boldsymbol{W},\boldsymbol{\Sigma}} \quad & \mathrm{trace}\,\boldsymbol{W} + \mathrm{trace}\,\boldsymbol{\Sigma} && \text{(16a)}\\
\text{s.t.} \quad & \frac{1}{\gamma_b}\,\mathrm{trace}\left(\boldsymbol{W}\boldsymbol{h}\boldsymbol{h}^{\mathcal{H}}\right) - \boldsymbol{h}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{h} \ge \sigma_n^2 && \text{(16b)}\\
& \gamma_{ce}(\boldsymbol{G}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{G} + \boldsymbol{D}^2) - \boldsymbol{G}^{\mathcal{H}}\boldsymbol{W}\boldsymbol{G} \succeq 0 && \text{(16c)}\\
& \boldsymbol{\Sigma} \succeq 0, \quad \boldsymbol{W} \succeq 0. && \text{(16d)}
\end{aligned}$$

This relaxed problem is a semidefinite program, whose optimal solution can be obtained by existing packages, e.g., CVX [5] .

In general, SDR gives an approximate solution of the original problem. Importantly, we can prove that if the SDR problem is feasible, then its optimum $\boldsymbol{W}$ is of rank one, which means that the optimal solution of the SDR problem is indeed the optimal solution of the original problem. However, we omit the proof for the lack of space. The same goes for the SDR problem for **Problem 2** as shown below.

By putting $\boldsymbol{W} = \boldsymbol{w}\boldsymbol{w}^{\mathcal{H}}$ and ignoring the rank condition, the SDR problem for **Problem 2** is given by

$$\begin{aligned}
\max_{\boldsymbol{W},\boldsymbol{\Sigma}} \quad & \frac{\boldsymbol{h}^{\mathcal{H}}\boldsymbol{W}\boldsymbol{h}}{\boldsymbol{h}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{h} + \sigma_n^2} && \text{(17a)}\\
\text{s.t.} \quad & \mathrm{trace}\,\boldsymbol{W} + \mathrm{trace}\,\boldsymbol{\Sigma} \le P_{max} && \text{(17b)}\\
& \gamma_{ce}(\boldsymbol{G}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{G} + \boldsymbol{D}^2) - \boldsymbol{G}^{\mathcal{H}}\boldsymbol{W}\boldsymbol{G} \succeq 0 && \text{(17c)}\\
& \boldsymbol{\Sigma} \succeq 0 \quad \boldsymbol{W} \succeq 0. && \text{(17d)}
\end{aligned}$$

Since the objective function is quasi-convex, the SDR problem is a quasi-convex problem. We borrow the idea to reformulate our quasi-convex problem to a convex semidefinite programming SDP from [9] by using the Charnes-Cooper transformation [1].

Let us define

$$\eta = \frac{1}{\boldsymbol{h}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{h} + \sigma_n^2}. \tag{18}$$

In place of $\boldsymbol{W}$ and $\boldsymbol{\Sigma}$, we express the SDR problem (17) with $\bar{\boldsymbol{W}}$ and $\bar{\boldsymbol{\Sigma}}$ defined as

$$\bar{\boldsymbol{W}} = \eta\boldsymbol{W}, \quad \bar{\boldsymbol{\Sigma}} = \eta\boldsymbol{\Sigma}. \tag{19}$$

Then, the SDR problem (17) can be expressed as

$$\begin{aligned}
\max_{\bar{\boldsymbol{W}},\bar{\boldsymbol{\Sigma}},\eta} \quad & \boldsymbol{h}^{\mathcal{H}}\bar{\boldsymbol{W}}\boldsymbol{h} && \text{(20a)}\\
\text{s.t.} \quad & \boldsymbol{h}^{\mathcal{H}}\bar{\boldsymbol{\Sigma}}\boldsymbol{h} + \eta\sigma_n^2 = 1 && \text{(20b)}\\
& \mathrm{trace}\,\bar{\boldsymbol{W}} + \mathrm{trace}\,\bar{\boldsymbol{\Sigma}} \le \eta P_{max} && \text{(20c)}\\
& \gamma_{ce}(\boldsymbol{G}^{\mathcal{H}}\bar{\boldsymbol{\Sigma}}\boldsymbol{G} + \eta\boldsymbol{D}^2) - \boldsymbol{G}^{\mathcal{H}}\bar{\boldsymbol{W}}\boldsymbol{G} \succeq 0 && \text{(20d)}\\
& \bar{\boldsymbol{\Sigma}} \succeq 0, \quad \bar{\boldsymbol{W}} \succeq 0, \quad \eta \ge 0. && \text{(20e)}
\end{aligned}$$

This SDR problem is found to be a semidefinite program whose optimal solution can be obtained numerically. Since $\eta^\star \neq 0$ from (20c), the optimal solution for (17) can be obtained by solving the SDR problem above.

4. SIMULATION RESULTS

Our proposed designs are compared with the SDR designs developed in [9] and the SDP design in [11] by numerical simulations.

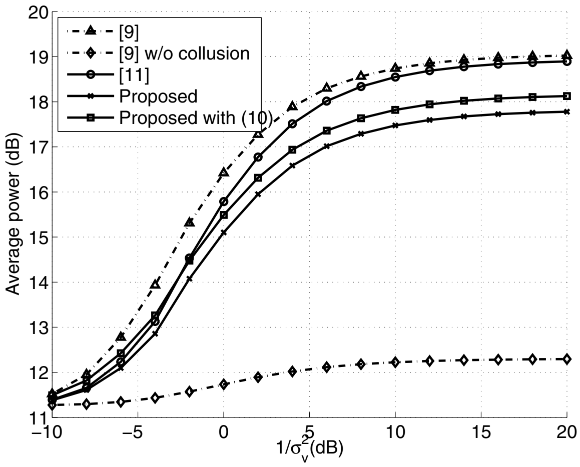

Fig. 1. Average transmit power by the proposed design (with $\times$), the proposed design with constraints (10) (with $\square$), the SDP design in [11] (with $\circ$), and the SDR design in [9] with/without collusion (with $\triangle$ and $\diamond$), for $N_t = 4$, $M = 3$, $\gamma_b = 10$dB, and $\gamma_{ce} = 5$dB.

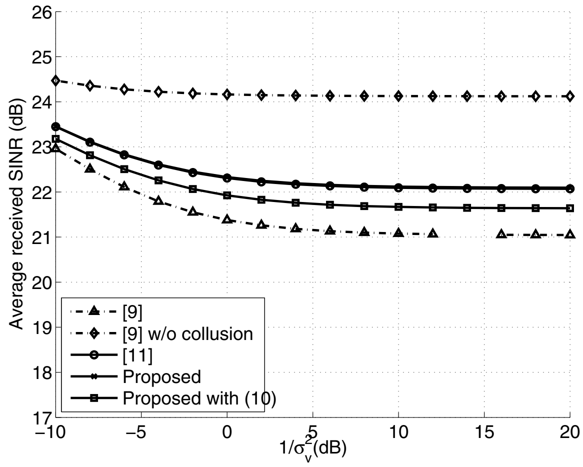

Fig. 2. Average received SINR of Bob by the proposed design (with $\times$), the proposed design with constraints (10) (with $\square$), the SDP design in [11] (with $\circ$), and the SDR design in [9] with/without collusion (with $\triangle$ and $\diamond$), for $N_t = 4$, $M = 3$, $\gamma_{ce} = 5$dB, and $P_{max} = 25$dB.

For comparison, we design the beamformer and the covariance of interference signals with/without collusion based on the algorithms in [9]. For the case of collusion, we set the same γ_{ce} as the proposed design in **Problem 1**. This means that the threshold $\gamma_{e,m}$, defined as Eq. (10) in [9], of the SDR problem is set to be γ_{ce}/M. Also, we set the same power limit P_{max} for **Problem 2**.

We also compute the optimal beamformer and covariance of interference signals for our problems with the constraints (10). Since the constraints (10) are convex, the relaxed problems with additional constraints are still convex.

As proved in [9, Prop.1], solving the SDR problems with the instantaneous channel $\boldsymbol{h}$ leads to the exact solution in theory. However, it is necessary to relax the rank of a matrix variable when the problem is solved numerically. Thus, the numerically obtained matrix variable has to be approximated to a matrix of rank one, which degrades the objectives. For our comparisons, we compare our objective values with the objective values of the SDR problems before relaxation. This implies that the values of relaxed problems are performance limits.

The channels state information vector $\boldsymbol{h}$ and $\{\boldsymbol{g}_m\}_{m=1}^M$ are assumed to be known exactly at the transmitter and randomly generated such that they are i.i.d. complex Gaussian with zero mean and covariance matrix $\boldsymbol{I}_{N_t}/N_t$, where $\boldsymbol{I}_{N_t}$ is an identity matrix of size $N_t \times N_t$. Bob's noise power is $\sigma_n^2 = 0$ dB, while Eve's noise power at each receive antenna is set as $\sigma_{v,m}^2 = \sigma_v^2$ for each $m \in [1, M]$. CVX [5], a package for specifying and solving convex programs, is utilized to numerically solve the optimization problems. The results are averaged over 10^3 channel realizations.

This example demonstrates the power efficiency of the proposed design in **Problem 1**. Fig. 1 compares the average transmit powers obtained by the proposed design (with $\times$), the proposed design with constraints (10) (with $\square$), the SDP design in [11] (with $\circ$), and the SDR design in [9] with/without collusion (with $\triangle$ and $\diamond$), for different noise power $1/\sigma_v^2$ at Eve, where $N_t = 4$, $M = 3$, $\gamma_b = 10$dB, and $\gamma_{ce} = 5$dB. Note that $1/\sigma_v^2$ physically means the overhearing ability of Eve, where a large $1/\sigma_v^2$ means strong overhearing ability and vice versa. The results for $1/\sigma_v^2 < -10$dB are omitted, since the average transmit powers of different designs are quite similar. This is simply because the channel noises of Eve is large enough so that the transmitter does not need to spend extra power for weak Eve's SINR.

It is reasonable that the SDR without collusion in [9] attains the minimum power, since it does not consider any collusion. As can be seen, our proposed design attains the smallest average transmit power among the methods for collusion, that is, our design exhibits the best performance. Comparing the results ($\square$) of our SDR with additional constraints and the results ($\triangle$) of the SDR design in [9] for collusion, one can conclude that the referenced SDR design fails to obtain optimal solutions.

Fig. 2 compares Bob's average SINR by the proposed designs and the referenced SDR design for different noise power $1/\sigma_v^2$ at Eve. As we can see, no matter how much the noise power of Eve is, the proposed design always obtains the largest SINR at Bob among the other designs for collusion.

Acknowledgment This work was partly supported by JSPS KAKENHI Grant Number 70273919.

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