ROBUST BEAMFORMER AND ARTIFICIAL NOISES FOR MISO WIRETAP CHANNELS WITH MULTIPLE EAVESDROPPERS

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ABSTRACT

We consider MISO wiretap channels with multiple eavesdroppers. Under the deterministic channel uncertainties, the beamformer and the covariance of the artificial noise are jointly designed to minimize the transmit power subject to SINR constraints. Our design problems are resolved by using a semidefinite program, which can be numerically solved. Simulation results are provided to see the effects of channel uncertainties on the transmit power.

***Index Terms*—** Physical-layer secrecy, beamforming, artificial noise, eavesdropping, MISOME

1. INTRODUCTION

Wireless communications are susceptible to eavesdropping, since they can be inevitably overheard by eavesdroppers within a certain range in an open environment. When the channel state information (CSI) of the eavesdropper is known, secrecy is theoretically guaranteed if the communication rate between the transmitter and the legitimate receiver is lower than the so-called secrecy capacity [21]. With the development of multiple-input and multiple-output (MIMO) technologies, the importance of physical-layer secrecy has been re-acknowledged, since physical-layer secrecy may be enhanced by using multiple antenna systems. There have been many studies on physical-layer secrecy (see a detailed survey on physical-layer secrecy [14]).

For MIMO systems, secrecy capacity has been well studied in terms of information theoretical point of view [12, 16]. Based on secrecy capacity, the MISO wiretap channel with multiple eavesdroppers is studied in [7], while the MIMO wiretap channel with multiple eavesdroppers is in [8]. Even when the transmitter does not know locations and CSIs of eavesdroppers, physical-layer secrecy for MISO channels can be characterized [3]. The secrecy capacity for OFDM transmission over fading channels is investigated in [18].

Secrecy capacity has been investigated in [4] for systems where the transmitter equipped with multiple antennas sends secret information signals as well as interference signals to interfere the eavesdropper using eavesdropper's CSI. The interference signals are known as artificial noises, which are utilized to degrade the received signals of eavesdroppers. The transmission of interference signals is also called artificial-noise aided or assisted transmission.

In [19, 20], beamforming as well as broadcasting artificial noises have been utilized to improve communications security. In [10], the beamformer and the covariance of the artificial noise have been jointly optimized for MISO wiretap channels by using semidefinite programs (SDPs), where 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. Similarly, the optimal beamformer and covariance for colluding eavesdroppers has been designed in [17].

In practice, CSI has to be estimated. Since estimation errors of CSI are unavoidable, the effects of imperfect CSI should be studied. The outage of secrecy capacity is analyzed for MISO wiretap channels with imperfect eavesdropper's CSI [2]. Transmit beamforming with imperfect CSI has been proposed for MIMO wiretap channels [15]. An artificial-noise aided transmission with imperfect CSI has been developed for MISO wiretap channels with a single eavesdropper based on the worst-case secrecy rate maximization [6]. An artificial-noise assisted secure beamforming has been proposed in [11] for MISO wiretap channels with a single eavesdropper where the transmitter knows the statistics of the eavesdropper's channel. On the other hand, deterministic channel uncertainties have been introduced in the joint design of the beamformer and the covariance of the artificial noise to maximize the achievable secrecy rate under the sum power constraint [9].

This paper deals with MISO wiretap channels with multiple eavesdroppers. Under the deterministic channel uncertainties, the beamformer and the covariance of the artificial noise are jointly designed to minimize the transmit power that keeps the minimum SINR constraint for the communication to the legitimate receiver and the maximum allowable SINR constraint for the protection to the eavesdropping. We first formulate our design as an optimization problem. Then, we turn the original problem into an SDP, which can be numerically solved. Simulation results are provided to see the effects of channel uncertainties on the transmit power.

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. The signal $y_b(t)$ of the legitimate receiver is modeled as

$$y_b(t) = \boldsymbol{h}^{\mathcal{H}}\boldsymbol{x}(t) + n(t) \tag{1}$$

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 receive antenna can be expressed as

$$y_{e,m}(t) = \boldsymbol{g}_m^{\mathcal{H}}\boldsymbol{x}(t) + v_m(t), \quad m = 1, \dots, M \tag{2}$$

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 $n(t)$.

Fig. 1 depicts our system. We also call the basestation, the legitimate receiver, and the remaining receivers as Alice, Bob, and Eve, 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 that multiple Eves try to eavesdrop $s(t)$ by collecting their received signals as a vector defined as

$$\boldsymbol{y}_e(t) = [y_{e,1}(t), \dots, y_{e,M}(t)]^T = \boldsymbol{G}^{\mathcal{H}}\boldsymbol{x}(t) + \boldsymbol{v}(t) \tag{3}$$

where $\boldsymbol{G} = [\boldsymbol{g}_1, \dots, \boldsymbol{g}_M]$ and $\boldsymbol{v}(t) = [v_1(t), \dots, v_M(t)]^T$.

We assume that the basestation allows connections only from active terminals so that there is no possible inactive eavesdropper. We also assume that at most $N_t + 1$ receivers are allowed 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. Then, the transmitted signal vector can be expressed as

$$\boldsymbol{x}(t) = \boldsymbol{w}s(t) + \boldsymbol{z}(t) \tag{4}$$

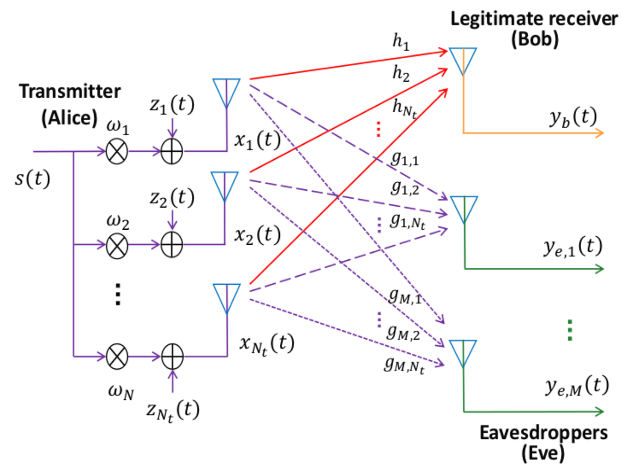

Fig. 1. System Diagram.

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

From (1) and (4), the SINR at Bob is found to be

$$\mathrm{SINR}_b(\boldsymbol{w}, \boldsymbol{\Sigma}) = \frac{|\boldsymbol{w}^{\mathcal{H}}\boldsymbol{h}|^2}{\boldsymbol{h}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{h} + \sigma_n^2}. \tag{5}$$

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 Eves and $\boldsymbol{D}^2 = \mathrm{diag}(\sigma_{v,1}^2, \dots, \sigma_{v,M}^2)$.

The optimal transmit beamforming vector $\boldsymbol{w}$ and covariance matrix $\boldsymbol{\Sigma}$ that minimize the transmit power under the constraints that the SINR of Bob is larger than or equal to the threshold γ_b and that the SINR of Eves is smaller than or equal to the threshold γ_{ce} can be obtained by solving the following convex optimization problem [17]:

$$\min_{\boldsymbol{W}, \boldsymbol{\Sigma}} \quad \mathrm{trace}\, \boldsymbol{W} + \mathrm{trace}\, \boldsymbol{\Sigma} \tag{7a}$$

$$\text{s.t.} \quad \frac{1}{\gamma_b}\mathrm{trace}\left(\boldsymbol{W}\boldsymbol{h}\boldsymbol{h}^{\mathcal{H}}\right) - \boldsymbol{h}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{h} \geq \sigma_n^2 \tag{7b}$$

$$\gamma_{ce}(\boldsymbol{G}^{\mathcal{H}}\boldsymbol{\Sigma}\boldsymbol{G} + \boldsymbol{D}^2) - \boldsymbol{G}^{\mathcal{H}}\boldsymbol{W}\boldsymbol{G} \succeq \boldsymbol{0} \tag{7c}$$

$$\boldsymbol{W} \succeq \boldsymbol{0}, \quad \boldsymbol{\Sigma} \succeq \boldsymbol{0} \tag{7d}$$

and then putting $\boldsymbol{w} = \boldsymbol{W}^{\frac{1}{2}}$ with the optimal $\boldsymbol{W}$, where $\boldsymbol{A} \succeq \boldsymbol{B}$ means that $\boldsymbol{A} - \boldsymbol{B}$ is positive semidefinite. It should be noted that the optimal $\boldsymbol{W}$ is of rank one in theory, In practice, the channels have to estimated. Let us model $\boldsymbol{h}$ and $\boldsymbol{G}$ as

$$\boldsymbol{h} = \bar{\boldsymbol{h}} + \Delta\boldsymbol{h} \tag{8}$$
$$\boldsymbol{G} = \bar{\boldsymbol{G}} + \Delta\boldsymbol{G} \tag{9}$$

where $\bar{\boldsymbol{h}}$ and $\bar{\boldsymbol{G}}$ are known to Alice and $\Delta\boldsymbol{h}$ and $\Delta\boldsymbol{G}$ are unknown. As in [9], the uncertainties are assumed to be bounded such as

$$\|\Delta\boldsymbol{h}\|_2 = \|\boldsymbol{h} - \bar{\boldsymbol{h}}\|_2 \le \epsilon_b \tag{10}$$
$$\|\Delta\boldsymbol{G}\|_F = \|\boldsymbol{G} - \bar{\boldsymbol{G}}\|_F \le \epsilon_e \tag{11}$$

for some $\epsilon_b > 0$ and $\epsilon_e > 0$, where $\|\cdot\|_2$ and $\|\cdot\|_F$ denote the l_2 norm and the Frobenius norm. Our objective is to find the optimal $\boldsymbol{w}$ and $\boldsymbol{\Sigma}$ for this channel model.

3. DESIGN OF ROBUST BEAMFORMERS AND INTERFERENCE COVARIANCE

Let us define two sets as

$$\mathcal{B}_b = \{\boldsymbol{h} \mid \|\boldsymbol{h} - \bar{\boldsymbol{h}}\|_2 \le \epsilon_b\} \tag{12}$$
$$\mathcal{B}_e = \{\boldsymbol{G} \mid \|\boldsymbol{G} - \bar{\boldsymbol{G}}\|_F \le \epsilon_e\}. \tag{13}$$

We can design the robust beamformer and the interference covariance by minimizing (7a) subject to the constraints

$$\frac{1}{\gamma_b}\mathrm{trace}(\boldsymbol{W}\boldsymbol{h}\boldsymbol{h}^H) - \boldsymbol{h}^H\boldsymbol{\Sigma}\boldsymbol{h} \ge \sigma_n^2, \quad \boldsymbol{h} \in \mathcal{B}_b, \tag{14}$$

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

and (7d). Using the tricks in [9], let us turn this problem to a semidefinite program (SDP).

First, we utilize the $\mathcal{S}$-procedure [1, Appendix B.2]:

Lemma 1 *Suppose that*

$$f_k(\mathbf{x}) = \mathbf{x}^H\mathbf{A}_k\mathbf{x} + 2\mathrm{Re}\{\mathbf{b}_k^H\mathbf{x}\} + c_k \quad k = 1, 2 \tag{16}$$

where $\mathbf{A}_k$ is an $n \times n$ Hermite matrix, $\mathbf{b}_k$ is an $n \times 1$ complex vector, and c_k is a real number. The implication

$$f_1(\mathbf{x}) \le 0 \Longrightarrow f_2(\mathbf{x}) \le 0 \tag{17}$$

hold if and only if there exits a $\mu \ge 0$ such that

$$\mu\begin{bmatrix} \mathbf{A_1} & \mathbf{b_1} \\ \mathbf{b_1}^H & c_1 \end{bmatrix} - \begin{bmatrix} \mathbf{A_2} & \mathbf{b_2} \\ \mathbf{b_2}^H & c_2 \end{bmatrix} \succeq \mathbf{0}$$

Substituting $\boldsymbol{h} = \bar{\boldsymbol{h}} + \Delta\boldsymbol{h}$ into (14) leads to

$$\Delta\boldsymbol{h}^H(\boldsymbol{W} - \gamma_b\boldsymbol{\Sigma})\Delta\boldsymbol{h} + 2\mathrm{Re}\{\bar{\boldsymbol{h}}^H(\boldsymbol{W} - \gamma_b\boldsymbol{\Sigma})\Delta\boldsymbol{h}\} + \bar{\boldsymbol{h}}^H(\boldsymbol{W} - \gamma_b\boldsymbol{\Sigma})\bar{\boldsymbol{h}} - \gamma_b\sigma_n^2 \ge 0 \tag{18}$$

Then, the constraint (14) is equivalent to

$$\Delta\boldsymbol{h}^H\Delta\boldsymbol{h} \le \epsilon_b^2 \Longrightarrow \Delta\boldsymbol{h}^H\boldsymbol{W}_b\Delta\boldsymbol{h} + 2\mathrm{Re}\{\bar{\boldsymbol{h}}^H\boldsymbol{W}_b\Delta\boldsymbol{h}\} + \bar{\boldsymbol{h}}^H\boldsymbol{W}_b\bar{\boldsymbol{h}} - \gamma_b\sigma_n^2 \ge 0 \tag{19}$$

where $\boldsymbol{W}_b = \boldsymbol{W} - \gamma_b\boldsymbol{\Sigma}$.

From Lemma 1 with $\mathbf{x} = \Delta\boldsymbol{h}$, $\mathbf{A_1} = \boldsymbol{I}$, $\mathbf{b_1} = \mathbf{0}$, $c_1 = -\epsilon_b^2$, $\mathbf{A_2} = -\boldsymbol{W}_b$, $\mathbf{b_2} = -\boldsymbol{W}_b\bar{\boldsymbol{h}}$, and $c_2 = -\gamma_b\sigma_n^2 + \bar{\boldsymbol{h}}^H\boldsymbol{W}_b\bar{\boldsymbol{h}}$, we find that (19) holds if and only if for $\lambda_b \ge 0$, $\boldsymbol{T}_b(\boldsymbol{W}, \boldsymbol{\Sigma}) \succeq \mathbf{0}$, where

$$\boldsymbol{T}_b(\boldsymbol{W}, \boldsymbol{\Sigma}) = \begin{bmatrix} \lambda_b\boldsymbol{I}_{N_t} + \boldsymbol{W}_b & \boldsymbol{W}_b\bar{\boldsymbol{h}} \\ \bar{\boldsymbol{h}}^H\boldsymbol{W}_b & -\lambda_b\epsilon_b^2 - \gamma_b\sigma_n^2 + \bar{\boldsymbol{h}}^H\boldsymbol{W}_b\bar{\boldsymbol{h}} \end{bmatrix}. \tag{20}$$

Therefore, in place of (14), we can use the linear matrix inequality (LMI) constraint, which is convex in our design parameters $\boldsymbol{W}$ and $\boldsymbol{\Sigma}$.

Similarly, the constraint (15) can be transformed into an LMI based on the following lemma [13]:

Lemma 2 *Let*

$$f(\mathbf{X}) = \mathbf{X^H A X} + \mathbf{X^H B} + \mathbf{B^H X} + \mathbf{C}, \quad \mathbf{D} \succeq \mathbf{0}. \tag{21}$$

Then,

$$f(\mathbf{X}) \succeq \mathbf{0}, \forall\mathbf{X} \in \{\mathbf{X} | \mathrm{trace}(\mathbf{DXX})^H \le 1\} \tag{22}$$

holds if and only if

$$\begin{bmatrix} \mathbf{C} & \mathbf{B}^H \\ \mathbf{B} & \mathbf{A} \end{bmatrix} - t\begin{bmatrix} \mathbf{I} & \mathbf{0} \\ \mathbf{0} & -\mathbf{D} \end{bmatrix} \succeq \mathbf{0} \quad \textit{for some} \;\; t \ge 0 \tag{23}$$

We substitute $\boldsymbol{G} = \bar{\boldsymbol{G}} + \Delta\boldsymbol{G}$ into (15) to obtain

$$\Delta\boldsymbol{G}^H\boldsymbol{W}_e\Delta\boldsymbol{G} + \Delta\boldsymbol{G}^H\boldsymbol{W}_e\bar{\boldsymbol{G}} + \bar{\boldsymbol{G}}^H\boldsymbol{W}_e\Delta\boldsymbol{G} + \bar{\boldsymbol{G}}^H\boldsymbol{W}_e\bar{\boldsymbol{G}} + \gamma_{ce}\boldsymbol{D}^2 \preceq \mathbf{0} \tag{24}$$

where $\boldsymbol{W}_e = \boldsymbol{W} - \gamma_{ce}\boldsymbol{\Sigma}$.

Applying Lemma 2 for $\mathbf{X} = \Delta\boldsymbol{G}$, $\mathbf{A} = -\boldsymbol{W}_e$, $\mathbf{B} = -\boldsymbol{W}_e\bar{\boldsymbol{G}}$, $\mathbf{C} = -\bar{\boldsymbol{G}}^H\boldsymbol{W}_e\bar{\boldsymbol{G}} + \gamma_{ce}\boldsymbol{D}^2$, and $\mathbf{D} = \epsilon_e^{-2}\boldsymbol{I}$, we find that (15) holds if and only if for $t \ge 0, \boldsymbol{T}_e(\boldsymbol{W}, \boldsymbol{\Sigma}, t) \succeq \mathbf{0}$, where

$$\boldsymbol{T}_e(\boldsymbol{W}, \boldsymbol{\Sigma}, t) = \begin{bmatrix} \bar{\boldsymbol{G}}^H\boldsymbol{W}_e\bar{\boldsymbol{G}} + (\gamma_{ce}\boldsymbol{D}^2 - t\boldsymbol{I}) & \bar{\boldsymbol{G}}^H\boldsymbol{W}_e \\ \boldsymbol{W}_e\bar{\boldsymbol{G}} & \boldsymbol{W}_e + \frac{t}{\epsilon_e^2}\boldsymbol{I} \end{bmatrix}. \tag{25}$$

From the above results, our design problem can be cast into the following SDP:

$$\min_{\boldsymbol{\omega}, \boldsymbol{\Sigma}, \lambda_b, t} \;\; \mathrm{trace}\boldsymbol{W} + \mathrm{trace}\boldsymbol{\Sigma} \tag{26a}$$
$$s.t. \boldsymbol{T}_b(\boldsymbol{W}, \boldsymbol{\Sigma}, \lambda_b) \succeq \mathbf{0} \tag{26b}$$
$$\boldsymbol{T}_e(\boldsymbol{W}, \boldsymbol{\Sigma}, t) \succeq \mathbf{0} \tag{26c}$$
$$\boldsymbol{W} \succeq \mathbf{0}, \boldsymbol{\Sigma} \succeq \mathbf{0}, \lambda_b \ge 0, t \ge 0 \tag{26d}$$

One can show that if the problem is feasible, the optimal $\boldsymbol{W}$ is of rank 1. The optimal solution can be numerically obtained by existing optimization packages, e.g., CVX [5].

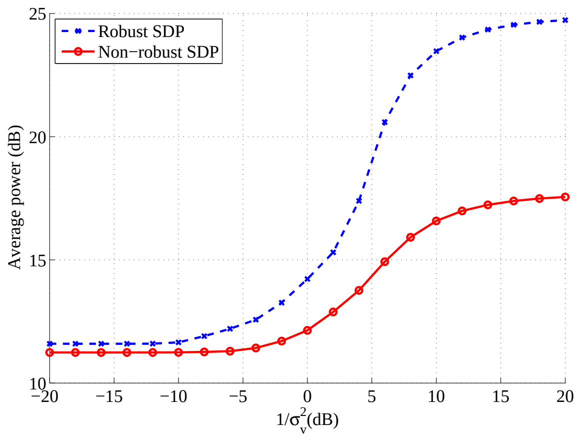

Fig. 2. Average transmit powers by robust design and for non-robust design against $1/\sigma_v^2$ ($N_t = 4$,$M = 3$, $\gamma_b = 10$dB, $\gamma_{ce} = 5$dB, $\alpha_b = 0.03$, $\alpha_e = 0.1$)

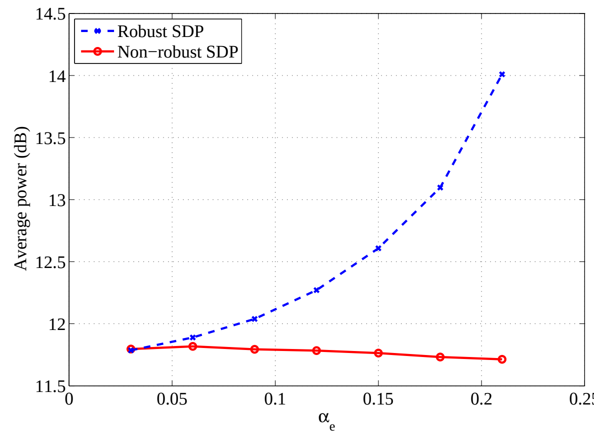

Fig. 3. Average transmit powers by our robust design and by non-robust design against α_e ($N_t = 4$, $M = 3$, $\sigma_v^2 = 0$dB, $\gamma_b = 10$dB, $\gamma_{ce} = 5$dB, $\alpha_b = 0.03$)

4. SIMULATION RESULTS

To see the effects of channel uncertainties on the transmit power, we compare our proposed design with the non-robust design proposed in [17]. The beamformer and the covariance of interference signals of the proposed design are obtained from solving (26), while the beamformer and the covariance of interference signals of the non-robust design are from (7). 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.

Let us clarify the difference between the two designs. To deal with channel deterministic channel uncertainties, the proposed design has additional constraints. Thus, it is obvious that the proposed design requires more transmit power if the problem is feasible and its feasibility is degraded compared to the non-robust design at the expense of its robustness.

The channels vector $\boldsymbol{h}$ and $\{\boldsymbol{g}_m\}_{m=1}^M$ are 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 Eves' noise power at each receive antenna is set as $\sigma_{v,m}^2 = \sigma_v^2$ for each $m \in [1, M]$.

We normalize the channel uncertainty bounds ϵ_e and ϵ_e (cf. (10) and (11)) by the norms of the known parts of channels as $\alpha_e = \frac{\epsilon_e}{\sqrt{E\{\|\bar{G}\|_F^2\}}}$ and $\alpha_b = \frac{\epsilon_b}{\sqrt{E\{\|\bar{h}\|^2\}}}$ where $E\{\cdot\}$ denotes the expectation operator. These are used to control the channel uncertainties in our simulations.

Fig. 2 compares the average transmit power by our robust design and the average transmit power by the non-robust design for $N_t = 4$, $M = 3$, $\gamma_b = 10$dB, $\gamma_{ce} = 5$dB, $\alpha_b = 0.03$, and $\alpha_e = 0.1$ as a function of $1/\sigma_v^2$. As the noise variance σ_v^2 at Eve decreases, $1/\sigma_v^2$ increases. Thus, $1/\sigma_v^2$ can be considered as a measure of the overhearing ability of Eves, where a large $1/\sigma_v^2$ means strong overhearing ability and vice versa.

As the condition of the eavesdropping improves, more power is required for both designs. It can been observed that the impact of the overhearing ability of Eves increases a lot between $1/\sigma_v^2 = 0$dB and $1/\sigma_v^2 = 10$dB and is saturated at high values of $1/\sigma_v^2$.

Fig. 3 depicts the average transmit powers by our robust design and by the non-robust design for $N_t = 4$, $M = 3$, $\sigma_v^2 = 0$dB, $\gamma_b = 10$dB, $\gamma_{ce} = 5$dB, and $\alpha_b = 0.03$ as a function of the Eves' channel uncertainty α_e, where the infeasible cases are excluded. As α_e increases, the uncertainty of channels to Eves increases, that is, the amount of information about eavesdroppers of Alice decreases. The transmit power of our robust design is an increasing function of the uncertainty α_e, while the transmit power of the non-robust design remains. For these setting, it can be confirmed that if the problem is feasible, then we can keep the SINR at Alice and the SINR at Eves even for relatively large uncertainty.

5. CONCLUSIONS

Taking account into the channel uncertainties, we have jointly designed robust beamformer and interference signals for MISO wiretap channels with multiple eavesdroppers, that keeps the SINR of the legitimate receiver and constrains the SINR of eavesdroppers. Our original design problem is reformulated as an SDP to be numerically solved. The effects of channel uncertainties on the design have been verified by simulation results.

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