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Secrecy Throughput Optimization for the WPCNs With Non-Linear EH Model

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ABSTRACT This paper investigates secure communication from a wireless-powered transmitter to a desired receiver with multiple eavesdroppers in the wireless powered communication networks (WPCNs). Considering the non-linear energy harvesting (EH) model, we propose a secure two-phase communication protocol with the help of a hybrid base station (HBS). First, in the power transfer (PT) phase, the HBS transfers wireless power to the transmitter. Then in the subsequent secure information transmission (SIT) phase, the transmitter sends the secret information using the energy harvested in the PT, under the protection of artificial noise (AN) generated by the HBS. First, based on this communication protocol, we maximize the secrecy throughput with perfect channel information state (CSI) under the transmit power constraint at the HBS. The secrecy throughput maximization (STM) problem is non-convex, and hence we reformulate it by exploiting the primal decomposition method (PDM) to obtain tractable forms. The PDM-based transmit scheme (PDM-TS) is proposed for the STM. In addition, considering the imperfect CSI of wiretap channel, we further design the robust transmit scheme for the worst-case secrecy throughput maximization (wSTM) problem. Since the wSTM shows high non-convexity, we extend the PDM by combining it with the S-procedure, and the PDM-based robust transmit scheme (PDM-RTS) is proposed for the wSTM. Finally, the numerical simulations are provided to show the effectiveness of the proposed transmit schemes.

INDEX TERMS Wireless powered communication networks (WPCNs), non-linear energy harvesting (EH) model, physical layer security (PLS), secrecy throughput, robust beamforming.

I. INTRODUCTION

A. BACKGROUND

The Internet of Things (IoT) has been recognized as a promising technique for the 5G communication, which can be applied in varied fields, e.g., manufacturing [1], precision agriculture [2] and the smart city [3]. The IoT networks usually contain large numbers of battery powered wireless devices for sensing, data processing and communication. Since these devices need to be replaced or recharged periodically due to the finite-capacity battery [4], the lifetime of IoT is limited.

Fortunately, the notion of energy harvesting (EH) has aroused an upsurge of interest as a promising resolution for prolonging the lifetime of IoT [5], [6]. Among the varied resources available for the EH, the radio-frequency (RF) wireless power can provide a sustainable power supply and is easy to be converted. Thus, the RF-EH paradigm has been shown to enhance the system energy efficiency in the energy-constraint wireless networks [7]–[9]. Currently, the RH-EH communication networks can be divided into two mainstreams: simultaneous wireless information and power transfer (SWIPT) networks and the wireless powered communication networks (WPCNs) [10]. For the WPCNs, the wireless power is first transferred to the EH nodes, then the EH nodes transmit information using the harvested energy.

Due to the broadcast nature of wireless medium, the communication security is another problem of serious concern. As an important supplement of traditional key-based cryptography method, the physical layer security (PLS) is promising for ensuring communication security [11]. Based on the information theory, the PLS has been derived by exploiting the uncertainty of wireless channel [12]. There have been several techniques for improving the performance of PLS. In [13] and [14], the secrecy rate was maximized by secure beamforming for the multiple-input single-output (MISO) and multiple-input multiple-output (MIMO) systems, respectively. The artificial noise (AN) was introduced to interfere the wiretap channel in [15], while the joint optimization of

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information signal and the AN was derived in [16] for saving transmit power. In [17], the assisting relay selection schemes were studied for enhancing secrecy performance using the Stackelberg Game.

Recently, employment of the PLS for the RH-EH wireless communication has attracted much attention. Assuming that the EH nodes may intercept the secret information, the optimal beamforming was designed to minimize the total transmit power for the SWIPT networks with secrecy constraints in [18]. In [19], a wireless-powered assisting jammer was employed to interfere the potential intercepting EH nodes. Reference [20] considered SWIPT relay networks and designed the optimal relaying strategies for the power splitting case and time switching case, respectively. In the WPCNs, reference [21] optimized the secrecy rate by using a hybrid access-point. Considering a wireless powered two-way relay networks, reference [22] maximized the secrecy rate with different secure relay protocols. With a hybrid base station (HBS) which can supply RF power and generate AN, reference [23] maximized the secrecy throughput of the WPCNs and obtained the closed-form optimal solution.

Generally, the secure transmit scheme design depends on the channel state information (CSI), which is not easy to be estimated perfectly in reality. Many works have managed to deal with the imperfect CSI. Assuming that the CSI errors lie in known sets of possible values, the worse-case secrecy performance was optimized in SWIPT networks [18], [24]. In [25]–[27], considering CSI errors submitting Gaussian distribution, the robust transmit schemes were investigated with secrecy outage probability constraints.

In the aforementioned research on secure communication in RH-EH systems, the linear EH model was taken, where the harvested power increases linearly with the input power. However, the recent research found that the linear EH model may be overly ideal and a new non-linear EH model was proposed in [28]. Based on this model, reference [29] and [30] investigated resource allocation schemes to minimize the transmit power with secrecy constraints. In [31], the energy efficiency was maximized for SWIPT networks based on non-linear EH model and it was found that the non-linear EH model can improve the secrecy performance.

B. RELATED WORK AND MOTIVATION

For secure communication in the WPCNs, the secrecy throughput was not studied in [21], [22], [30]. However, the transmission in the WPCNs is divided into two phases and the information is transferred only in the second phase. Hence, the research of maximizing secrecy throughput for the whole transmission slot in the WPCNs is meaningful and challenging. So far, a few works have concentrated on this topic for the WPCNs. In order to highlight the novelty of this paper, it is worthy to emphasize the following differences between this paper and existing works.

Firstly, the wireless powered transmitter is equipped with single antenna in [23], [32]–[36], and it will become difficult to harvest enough energy for information transmission when the RF power from the HBS is low. In addition, the single-antenna transmitter cannot conduct secure beamforming, which leads to poor secrecy performance especially for the case of multiple Eves. In 5G generation, the multi-antenna transmitter gets easily achievable, hence this paper investigates the secure transmission in the WPCNs with multi-antenna wireless powered transmitter.

Secondly, most of the existing works (see e.g., [23], [32]–[37]) only studied the secure transmission in the WPCNs with perfect CSI. However, it is hard to obtain the exact CSI in reality, especially when the Eves take passive intercepting method. Although the imperfect CSI was considered in [38] with single Eve, the AN was not employed therein. With the transmit scheme in [38], the secrecy performance will dramatically degrade when the number of Eves increases or the wiretap channels become superior to the main channel. To improve the secrecy performance for these cases, we employ the AN and design the AN signal in this paper.

Thirdly, although the AN was introduced in [23], [33]–[36], the AN signal was sub-optimal for the secure transmission. For instance, with single Eve, the simple zero-forcing (ZF) AN scheme was taken in [36], and with multiple Eves, the AN was generated in the null space of main channel in [23], [33], [34], [36]. As the introduction of AN may lead to extra power consumption, this paper designs the AN signal using the semi-definite programming (SDP) method to improve the secrecy performance and the power utilization efficiency simultaneously.

Fourthly, in most existing works related to the WPCNs, the traditional linear EH model was taken, which mismatches the practical circuit features and may result in misleading optimization solutions. To avoid the performance loss caused by the linear model, this paper considers the newly proposed non-linear EH model. Thus, our work is much closer to practical systems.

C. CONTRIBUTIONS

In this paper, we design the secure transmit schemes in the WPCNs which consist of a HBS, a wireless powered transmitter, a desired receiver and multiple passive Eves. Considering the non-linear EH model, the secrecy throughput is maximized for both the perfect CSI and imperfect CSI cases. The main contributions of this paper can be summarized as follows.

- We investigate the transmission in the WPCNs and design a secure two-phase communication protocol, including the power transfer (PT) and the secure information transmission (SIT). This protocol enables secure communication from the wireless powered transmitter to the receiver with the help the HBS. Especially, the newly proposed non-linear EH model is taken in this paper and the simulation results indicate the existence of optimal power supply of the HBS with this model.
- With the multiple-antenna wireless powered transmitter and AN generated by the HBS, we maximize the secrecy

throughput by jointly optimizing the RF signal covariance, the secret signal beamforming, the AN covariance and the time allocation between PT and SIT. The secure transmit schemes are designed for both the perfect CSI and imperfect CSI cases.

- For the perfect CSI case, the secrecy throughput maximization (STM) is non-convex. Hence, we employ the primal decomposition method PDM) by divide the original problem into two simpler sub-problems, i.e., the beamforming design sub-problem (BFsP) and the time allocation sub-problem (TAsP). The PDM based transmit scheme (PDM-TS) is proposed for the STM.
- As an extended work, we study the worst-case secrecy throughput maximization (wSTM) problem when the CSI of wiretap channel is imperfect. This problem is highly non-convex, hence we extend the PDM by combining it with the S-procedure. The wSTM problem is divided into another two sub-problems, i.e., the robust beamforming design sub-problem (RBFsP) and the robust time allocation sub-problem (RTAsP). The PDM based robust transmit scheme (PDM-RTS) is proposed for the wSTM.

The organization of this paper is as follows. Section II introduces the system model and establishes the problem formulations. We solve the STM problem with perfect CSI in Section III, while deliver solution of the wSTM problem for the imperfect CSI case in Section IV. The simulation results and analysis are shown in Section V and the conclusion is drawn in Section VI.

The notation of this paper is as follows. Boldface lowercase and uppercase letters are used to denote vectors and matrices, respectively. $\mathbf{I}_n$ denotes the $n-by-n$ identity matrix, and 0 is a zero matrix. By $\mathbf{X} \succeq 0$, we mean that $\mathbf{X}$ is a Hermitian positive semidefinite matrix. The operators $(\cdot)^T$, $(\cdot)^H$, $Tr(\cdot)$ and $|\cdot|$ represent the transpose, Hermitian, trace and determinant operations, respectively. $\mathbb{R}^{m\times n}$ and $\mathbb{C}^{m\times n}$ stand for the set of matrices with real- and complex-valued entries. The symbol $E\{\cdot\}$ represents the statistical expectation of the argument and $[x]^+ = max(0, x)$. The Euclidean norm is denoted by $||\cdot||$ and $Re\{\cdot\}$ represents the real part of a complex value. $Diag(\mathbf{A}, \mathbf{B})$ represents a block diagonal matrix with the diagonal blocks with $\mathbf{A}$ and $\mathbf{B}$.

II. SYSTEM MODEL AND PROBLEM FORMULATIONS

A. SYSTEM MODEL

We investigate the secure communication in the WPCNs including a HBS, a transmitter, a desired receiver and K passive eavesdroppers (Eves). The antenna numbers of HBS and transmitter are N_h and N_t, respectively. Meanwhile the receiver and Eves are both single-antenna. It is assumed that the HBS provides constant power supply, while the transmitter utilizes the RF power harvested from HBS to transfer secret information.[1] The system model is shown in Fig.1.

[1] In this paper, we consider a simple scenario where the desired receiver and the Eves are located adjacently, hence these nodes can be clustered together referred as the receiving nodes.

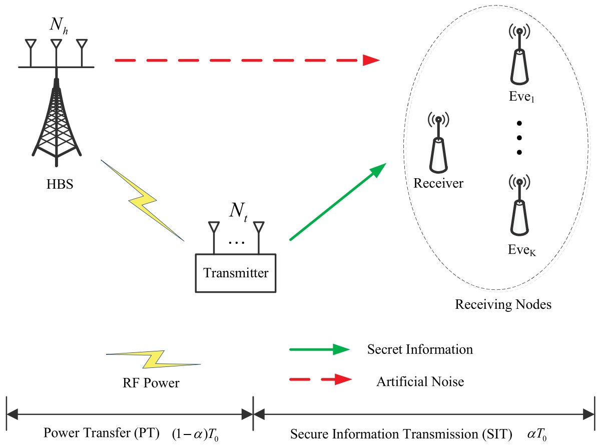

FIGURE 1. Illustration of the secure two-phase communication protocol in the WPCNs.

In this paper, the secure two-phase communication protocol is proposed for the WPCNs. The time for a whole transmission slot is assumed as T_0. Denote α as the time division ratio with $0 < \alpha < 1$. During the PT lasting for $(1-\alpha)T_0$, the HBS transmits RF power to the transmitter. Let $\mathbf{H}_P \in \mathbb{C}^{N_t\times N_h}$ represents the channel from the HBS to transmitter. The signal received at transmitter in the PT is

$$\mathbf{y}_t = \mathbf{H}_P\mathbf{v}_P + \mathbf{n}_t \tag{1}$$

where $\mathbf{v}_P \sim \mathcal{CN}(\mathbf{0}, \mathbf{Q}_P)$ denotes the RF signal from the HBS and $\mathbf{n}_t \sim \mathcal{CN}(\mathbf{0}, {\sigma_t}^2\mathbf{I})$ denotes the independent and identical distributed (i.i.d.) circular symmetric complex additive white Gaussian (AWGN) noises. Assuming the power supply of the HBS is P, it should be satisfied that $Tr(\mathbf{Q}_P) \leq P$. Let P_{rev} denote the received RF power from the HBS and P_{hst} denote the harvested power at transmitter. Most the existing works take the linear EH model, i.e.,

$$P_{hst} = \rho_l P_{rev} \tag{2}$$

where $0 < \rho_l < 1$ is the power conversion efficiency. However, this model is too ideal. A more practical non-linear EH model was proposed in [28], which was verified by measurement data. According to this model, the harvested power can be denoted as

$$P_{hst} = \frac{\Psi_E - M_E\Omega_E}{1-\Omega_E} \tag{3a}$$

$$\Psi_E = \frac{M_E}{1+e^{-a_E(\eta P_{rev}-b_E)}} \tag{3b}$$

$$\Omega_E = \frac{1}{1+e^{a_E b_E}} \tag{3c}$$

In (3b), $0 < \eta < 1$ denotes the division ratio of received power for information transmission.[2] M_E is a constant denoting the maximum harvested power at the transmitter when the EH circuit is saturated, a_E and b_E are parameters related to the detailed circuit specifications.

[2] In this paper, we assume that the transmitter also conducts information sensing and data processing, which requires $(1-\eta)P_{rec}$ of the received power.

During the subsequent SIT lasting for αT_0, the transmitter sends secret information to the desired receiver in the presence of K passive Eves. The channel from transmitter to receiver is $\mathbf{h}_r \in \mathbb{C}^{N_t \times 1}$, while the channel from transmitter to Eve k is $\mathbf{h}_k \in \mathbb{C}^{N_t \times 1}$, $k \in \{1, \ldots, K\}$. In this phase, the HBS generates AN to interfere the Eves for improving the secrecy performance. The channels from the HBS to receiver and Eve k are $g_r \in \mathbb{C}^{N_h \times 1}$ and $g_k \in \mathbb{C}^{N_h \times 1}$, respectively. Then, the received signals at receiver and Eve k can be denoted as

$$y_r = \mathbf{h}_r{}^H \mathbf{w} s + \mathbf{g}_r{}^H \mathbf{v}_J + n_r \tag{4a}$$
$$y_k = \mathbf{h}_k{}^H \mathbf{w} s + \mathbf{g}_k{}^H \mathbf{v}_J + n_k \tag{4b}$$

where s represents the secret information from the transmitter with $E\{|s|^2\} = 1$, and $\mathbf{w}$ is the corresponding beamforming vector. $\mathbf{v}_J \sim \mathcal{CN}(\mathbf{0}, \mathbf{Q}_J)$ represents the AN generated by the HBS which satisfies $Tr(\mathbf{Q}_J) \le P$. $n_r \in \mathbb{C}$ and $n_k \in \mathbb{C}$ represent the corresponding i.i.d. AWGN with variances $\sigma_r{}^2$ and $\sigma_k{}^2$.

Considering the existence of K passive Eves without collusion, the secrecy rate can be calculated as [39]

$$R_s = [\log(1 + \frac{\mathbf{h}_r{}^H \mathbf{w}\mathbf{w}^H \mathbf{h}_r}{\mathbf{g}_r{}^H \mathbf{Q}_J \mathbf{g}_r + \sigma_r{}^2}) - \max_{1 \le k \le K} \log(1 + \frac{\mathbf{h}_k{}^H \mathbf{w}\mathbf{w}^H \mathbf{h}_k}{\mathbf{g}_k{}^H \mathbf{Q}_J \mathbf{g}_k + \sigma_k{}^2})]^+ \tag{5}$$

Assume slow fading channels which remain constant in the SIT. The secrecy throughput for a whole transmission slot can be represented as

$$T_s = \alpha T_0 R_s \tag{6}$$

B. PROBLEM FORMULATIONS

In this paper, we investigate the secrecy throughput in the WPCNs, i.e., the STM problem for the perfect CSI case and the wSTM problem for the imperfect CSI case. With perfect CSI, the STM is to maximize the secrecy throughput by jointly optimizing the RF signal covariance $\mathbf{Q}_P$ in PT, the secret signal beamforming $\mathbf{w}$, the AN covariance $\mathbf{Q}_J$ in SIT and the time division ratio α. That is

$$\mathcal{Q}_T: \quad \max_{\mathbf{Q}_P, \mathbf{w}, \mathbf{Q}_J, \alpha} T_s \tag{7a}$$
$$s.t.\ Tr(\mathbf{w}\mathbf{w}^H) \le \frac{(1-\alpha)P_{hst}}{\alpha}, \quad 0 < \alpha < 1 \tag{7b}$$
$$Tr(\mathbf{Q}_P) \le P, \quad Tr(\mathbf{Q}_J) \le P \tag{7c}$$
$$\mathbf{Q}_P \succeq \mathbf{0}, \quad \mathbf{Q}_J \succeq \mathbf{0} \tag{7d}$$

The constraints (7b) insures that the transmit power of the transmitter does not exceed its previously harvested power. The constraints (7c) implies the power supply of the HBS remains constant for the PT and SIT.

In practice, the CSI of wiretap channels is usually imperfect as the Eves take the passive intercepting method. In this paper, we take the deterministic model [18], [24] to describe the CSI errors. For Eve k, it is defined that

$$\mathbf{h}_k = \hat{\mathbf{h}}_k + \delta_{hk}$$
$$\mathbf{g}_k = \hat{\mathbf{g}}_k + \delta_{gk} \tag{8}$$

$\hat{\mathbf{h}}_k$ and $\hat{\mathbf{g}}_k$ are the estimated channels, while δ_{hk} and δ_{gk} denote the response CSI errors. It is set $||\delta_{hk}||^2 \le \varepsilon_{hk}$, $||\delta_{gk}||^2 \le \varepsilon_{gk}$, where ε_{hk}, ε_{gk} are known constants.

In this way, we design the robust transmit scheme to maximize worst-case secrecy throughput. The wSTM can be summarized as

$$\mathcal{Q}_{RT}: \quad \max_{\substack{\mathbf{Q}_P, \mathbf{w} \\ \mathbf{Q}_J, \alpha}} \min_{\delta_{hk}, \delta_{gk}} T_s \tag{9a}$$
$$s.t.\ Tr(\mathbf{w}\mathbf{w}^H) \le \frac{(1-\alpha)P_{hst}}{\alpha}, \quad 0 < \alpha < 1 \tag{9b}$$
$$Tr(\mathbf{Q}_P) \le P, \quad Tr(\mathbf{Q}_J) \le P \tag{9c}$$
$$||\delta_{hk}||^2 \le \varepsilon_{hk}, \quad ||\delta_{gk}||^2 \le \varepsilon_{gk}, \quad k \in \{1, \ldots, K\} \tag{9d}$$
$$\mathbf{Q}_P \succeq \mathbf{0}, \quad \mathbf{Q}_J \succeq \mathbf{0} \tag{9e}$$

In what follows, we first derive the solution of the STM with perfect CSI, then extend the work to exploit the wSTM with imperfect CSI.

III. TRANSMIT SCHEME DESIGN WITH PERFECT CSI

In this section, we design the PDM-TS for the STM by solving the problem $\mathcal{Q}_T$. To deal with the $\mathcal{Q}_T$, we first derive the optimal RF signal covariance $\mathbf{Q}_P$, then further design the secret signal beamforming $\mathbf{w}$, the AN covariance $\mathbf{Q}_J$ and the time division ratio α using the PDM.

A. RF SIGNAL DESIGN

It can be found the harvested power P_{hst} is only involved in (7b). Hence, we first design the RF signal to maximize the harvested power P_{hst} of the transmitter in this part. With the RF signal covariance $\mathbf{Q}_P$, the received RF power of the transmitter can be calculated as

$$P_{rev} = Tr(\mathbf{Q}_P \mathbf{H}_P{}^H \mathbf{H}_P) + N_t \sigma_t{}^2 \tag{10}$$

According to (3), P_{hst} increases monotonously with P_{rev}. Hence, the problem of maximizing P_{hst} is equivalent to maximizing the received RF power P_{rev} with the optimal $\mathbf{Q}_P$, that is

$$\mathcal{Q}_{RFB}: \quad \max_{\mathbf{Q}_P} Tr(\mathbf{Q}_P \mathbf{H}_P{}^H \mathbf{H}_P) + N_t \sigma_t{}^2 \tag{11a}$$
$$s.t.\ Tr(\mathbf{Q}_P) \le P \tag{11b}$$
$$\mathbf{Q}_P \succeq \mathbf{0} \tag{11c}$$

To solve problem $\mathcal{Q}_{RFB}$, we first introduce the following lemma.

Lemma 1 ([40]): Let A and B be two $N \times N$ positive semidefinite matrices, with eigenvalues $\alpha_1 \ge \cdots \ge \alpha_N$ and $\beta_1 \ge \cdots \ge \beta_N$, respectively. Then,

$$\sum_{i=1}^{N} \alpha_i \beta_{N-i+1} \le Tr(\mathbf{AB}) \le \sum_{i=1}^{N} \alpha_i \beta_i$$

Since $\mathbf{H}_P{}^H \mathbf{H}_P$ is positive semidefinite, we can perform the eigenvalue decomposition and obtain the result as

$\mathbf{H}_P{}^H\mathbf{H}_P = \mathbf{U}_{HP}\Lambda_{HP}\mathbf{U}_{HP}{}^H$. It is defined that $\Lambda_{HP} = diag(\lambda_1, \cdots, \lambda_{N_h})$, where $\lambda_i, i \in \{1, \ldots, N_h\}$ is the eigenvalue with $\lambda_1 \geq \cdots \geq \lambda_{N_h}$ and $\mathbf{U}_{HP}$ is the orthogonal matrix combined by corresponding eigenvectors. Assuming the eigenvalue of $\mathbf{Q}_P$ is $\mu_i, i \in \{1, \ldots, N_h\}$ with $\mu_1 \geq \cdots \geq \mu_{N_h}$ and $\sum_{i=1}^{N_h} \mu_i \leq P$, according to Lemma 1, we can get

$$Tr(\mathbf{Q}_P\mathbf{H}_P{}^H\mathbf{H}_P) + N_t\sigma_t{}^2 \leq \sum_{i=1}^{N_h} \lambda_i\mu_i + N_t\sigma_t{}^2 \quad (12)$$

where the equality holds when the eigenvalue decomposition of $\mathbf{Q}_P$ can be represented as $\mathbf{Q}_P = \mathbf{U}_{HP}\Lambda_{QP}\mathbf{U}_{HP}{}^H$ with $\Lambda_{QP} = diag(\mu_1, \cdots, \mu_{N_h})$. To maximize the received RF power P_{rev}, we set $\Lambda_{QP} = \Lambda_{QP\_m} = diag(P, 0, \cdots, 0)$. In this way, the maximum P_{rev} can be obtained as

$$P_{rev\_m} = P\lambda_1 + N_t\sigma_t{}^2 \quad (13)$$

and the corresponding RF signal covariance $\mathbf{Q}_{P\_m} = \mathbf{U}_{HP}\Lambda_{QP\_m}\mathbf{U}_{HP}{}^H$. The corresponding maximum harvested power P_{hst_m} can be obtained using (3).

Proposition 1: When the HBS transmits the RF signal with covariance $\mathbf{Q}_P = \mathbf{Q}_{P\_m}$, the harvested power of the transmitter is maximized as P_{hst_m} and the maximum secrecy throughput can be achieved under this condition.

Proof: Suppose the RF signal covariance $\mathbf{Q}_{P\_e} \neq \mathbf{Q}_{P\_m}$ with the corresponding harvested power P_{hst_e}, then it is satisfied $0 \leq P_{hst\_e} \leq P_{hst\_m}$ according to Lemma 1. Applying P_{hst_m} and P_{hst_e} into (7b), we can get the feasible domains of Q_T as Φ_m and Φ_e, respectively. Since $P_{hst\_e} \leq P_{hst\_m}$, it can be obtained $\Phi_e \subseteq \Phi_m$. So the solution of Q_T with P_{hst_e} is suboptimal and the Proposition 1 is proved.

With Proposition 1, the problem Q_T is transferred as

$$Q_{T1}: \quad \max_{\mathbf{w},\mathbf{Q}_J,\alpha} T_s \quad (14a)$$

$$s.t.\ Tr(\mathbf{w}\mathbf{w}^H) \leq \frac{(1-\alpha)P_{hst\_m}}{\alpha}, \quad 0 < \alpha < 1 \quad (14b)$$

$$Tr(\mathbf{Q}_J) \leq P, \quad \mathbf{Q}_J \succeq \mathbf{0} \quad (14c)$$

Actually, Q_{T1} is still non-convex due to the objective function in (14a). Hence, we further employ the PDM to deal with it. The PDM is based on decomposing the original complicated problem into several simple sub-problems controlled by a master problem and taking an iterative way to find the solution [41], [42]. In this paper, Q_{T1} is decomposed into two sub-problems, i.e., the BFsP and the TAsP. In the following, we will solve them respectively.

B. SOLUTION OF THE BFSP

In this part, the BFsP is involved, i.e., jointly optimizing the signal beamforming $\mathbf{w}$ and the AN covariance $\mathbf{Q}_J$ with a fixed time division ratio $\alpha = \alpha_0$. The main obstacles of solving Q_{T1} lie in (14a), which is non-convex. To deal with it, the sequential parametric convex approximation (SPCA) method [43] is employed.

The secrecy rate in (5) can be rewritten as

$$R_s = [R_m - \max_{1\leq k\leq K} R_{ek}]^+ \quad (15)$$

where

$$\begin{aligned} R_m &= \log(\mathbf{g}_r{}^H\mathbf{Q}_J\mathbf{g}_r + \sigma_r{}^2 + \mathbf{h}_r{}^H\mathbf{w}\mathbf{w}^H\mathbf{h}_r) \\ &\quad - \log(\mathbf{g}_r{}^H\mathbf{Q}_J\mathbf{g}_r + \sigma_r{}^2) \quad (16a) \\ R_{ek} &= \log(\mathbf{g}_k{}^H\mathbf{Q}_J\mathbf{g}_k + \sigma_k{}^2 + \mathbf{h}_k{}^H\mathbf{w}\mathbf{w}^H\mathbf{h}_k) \\ &\quad - \log(\mathbf{g}_k{}^H\mathbf{Q}_J\mathbf{g}_k + \sigma_k{}^2) \quad (16b) \end{aligned}$$

It can be found that the second term on the right hand side (RHS) of (16a) is convex and the first term on RHS of (16b) concave, which causes the (15) is non-convex and non-concave. To proceed, we exploit the first-order Taylor expansion and get the approximations of (16a) and (16b) as

$$\begin{aligned} \tilde{R}_m &= \log(\mathbf{g}_r{}^H\mathbf{Q}_J\mathbf{g}_r + \sigma_r{}^2 + \mathbf{h}_r{}^H\mathbf{W}\mathbf{h}_r) \\ &\quad - \log(\mathbf{g}_r{}^H\mathbf{Q}_{J0}\mathbf{g}_r + \sigma_r{}^2) - Tr \\ &\quad \times [\frac{\mathbf{g}_r\mathbf{g}_r{}^H}{\mathbf{g}_r{}^H\mathbf{Q}_{J0}\mathbf{g}_r + \sigma_r{}^2}(\mathbf{Q}_J - \mathbf{Q}_{J0})] \quad (17a) \\ \tilde{R}_{ek} &= \log(\mathbf{g}_k{}^H\mathbf{Q}_{J0}\mathbf{g}_k + \sigma_k{}^2 + \mathbf{h}_k{}^H\mathbf{W}_0\mathbf{h}_k) \\ &\quad + Tr[\frac{\mathbf{g}_k\mathbf{g}_k{}^H}{\mathbf{g}_k{}^H\mathbf{Q}_{J0}\mathbf{g}_k + \sigma_k{}^2 + \mathbf{h}_k{}^H\mathbf{W}_0\mathbf{h}_k}(\mathbf{Q}_J - \mathbf{Q}_{J0})] \\ &\quad + Tr[\frac{\mathbf{h}_k\mathbf{h}_k{}^H}{\mathbf{g}_k{}^H\mathbf{Q}_{J0}\mathbf{g}_k + \sigma_k{}^2 + \mathbf{h}_k{}^H\mathbf{W}_0\mathbf{h}_k}(\mathbf{W} - \mathbf{W}_0)] \\ &\quad - \log(\mathbf{g}_k{}^H\mathbf{Q}_J\mathbf{g}_k + \sigma_k{}^2) \quad (17b) \end{aligned}$$

where $\mathbf{W} = \mathbf{w}\mathbf{w}^H$ and $\mathbf{W}_0$, $\mathbf{Q}_{J0}$ are constant matrices. Then, the tractable approximation of problem Q_{T1} with $\alpha = \alpha_0$ can be represented as

$$\max_{\mathbf{W},\mathbf{Q}_J} (\tilde{R}_m - \max_{1\leq k\leq K} \tilde{R}_{ek}) \quad (18a)$$

$$s.t.\ Tr(\mathbf{W}) \leq \frac{(1-\alpha_0)P_{hst\_m}}{\alpha_0}, \quad \mathbf{W} \succeq \mathbf{0} \quad (18b)$$

$$Tr(\mathbf{Q}_J) \leq P, \quad \mathbf{Q}_J \succeq \mathbf{0} \quad (18c)$$

$$rank(\mathbf{W}) = 1 \quad (18d)$$

Introduce a slack variable ϕ, which denotes the upper bound of wiretapping rate with perfect CSI. Neglecting the rank constraint in (18d) by using the semidefinite relaxation (SDR), (18) can be re-expressed as

$$Q_{T1-AP}: \quad \max_{\mathbf{W},\mathbf{Q}_J,\phi} \tilde{R}_m - \phi \quad (19a)$$

$$s.t.\ \tilde{R}_{ek} \leq \phi, \quad k \in \{1, \ldots, K\} \quad (19b)$$

$$(18b), (18c) \quad (19c)$$

The problem Q_{T1-AP} is convex and can be solved efficiently (e.g. with CVX software [44]). Although neglecting the rank constraint, the following proposition implies that the accurate solution of (18) can still be obtained.

Proposition 2: If the SDR problem Q_{T1-AP} is feasible, the optimal solution $\mathbf{W}^*$ yields $rank(\mathbf{W}^*) = 1$.

The proof is shown in Appendix A.

Based on the SPCA method by iteratively solving the sequence of Q_{T1-AP}, the algorithm for solving the BFsP is summarized in Algorithm 1.

Algorithm 1 SPCA Procedure for the BFsP

Initialize $\mathbf{W}_0, \mathbf{Q}_{J0}$.
repeat
  Solve Q_{T1-AP}, get the optimal solution $\mathbf{W}_t{}^*$ and $\mathbf{Q}_{J}{}_t^*$.
  Set $\mathbf{W}_0 = \mathbf{W}_t{}^*, \mathbf{Q}_{J0} = \mathbf{Q}_{J}{}_t^*$.
until $\tilde{R}_m - \phi$ in (19a) converges.
Output:
  The optimal solution $\tilde{\mathbf{W}}^*, \tilde{\mathbf{Q}}_J^*$ and corresponding secrecy throughput.

C. SOLUTION OF THE TASP

In this part, we derive the solution of the TAsP, i.e., the optimal time division ratio α for maximizing the secrecy throughput with the fixed $\tilde{\mathbf{W}}^*, \tilde{\mathbf{Q}}_J^*$. We first normalize the $\tilde{\mathbf{W}}^*, \tilde{\mathbf{Q}}_J^*$ with $\bar{\mathbf{W}} = \frac{\tilde{\mathbf{W}}^*}{Tr(\tilde{\mathbf{W}}^*)}$ and $\bar{\mathbf{Q}}_J = \frac{\tilde{\mathbf{Q}}_J^*}{Tr(\tilde{\mathbf{Q}}_J^*)}$ such that $Tr(\bar{\mathbf{W}})$=1 and $Tr(\bar{\mathbf{Q}}_J) = 1$. Then the secrecy throughput can be represented as

$$T_s = \alpha\left\{\log[1+C_r\frac{(1-\alpha)P_{hst\_m}}{\alpha}] - \max_{1\le k\le K}\log[1+C_k\frac{(1-\alpha)P_{hst\_m}}{\alpha}]\right\}^+, \quad 0<\alpha<1 \tag{20}$$

where

$$C_r = \frac{\mathbf{h}_r{}^H\bar{\mathbf{W}}\mathbf{h}_r}{\mathbf{g}_r{}^H\bar{\mathbf{Q}}_J\mathbf{g}_r + \sigma_r{}^2}$$
$$C_k = \frac{\mathbf{h}_k{}^H\bar{\mathbf{W}}\mathbf{h}_k}{\mathbf{g}_k{}^H\bar{\mathbf{Q}}_J\mathbf{g}_k + \sigma_k{}^2}$$

According to (20), the T_s is a function of α, which is shown non-concave. It is hard to derive the optimal value of (20) directly, hence we design an iterative approach to achieve an approximate optimal solution.

For Eve k, define the function

$$T_{s,k}(\alpha) = \alpha\left\{\log[1 + C_r\frac{(1-\alpha)P_{hst\_m}}{\alpha}] - \log[1 + C_k\frac{(1-\alpha)P_{hst\_m}}{\alpha}]\right\} \tag{21}$$

With function (21), we define the set $\Phi_k = \{\alpha \in R | T_{s,k}(\alpha) \ge t, t \ge 0\}$ and $\Phi = \bigcap_{k=1}^{K} \Phi_k$. Actually, Φ_k is the feasible region of α to ensure available secrecy throughput for Eve k. To solve $T_{s,k}(\alpha) \ge t$ is equal to solving the following inequality of α

$$\frac{1+C_r\frac{1-\alpha}{\alpha}P_{hst}}{1 + C_k\frac{1-\alpha}{\alpha}P_{hst}} \ge e^{\frac{t}{\alpha}} \tag{22}$$

Set $q = \frac{1}{\alpha}$ with $q > 1$, (22) can be transferred as

$$1+C_rP_{hst}q - C_rq - e^{tq}(1+C_kP_{hst}q - C_kP_{hst}) \ge 0 \tag{23}$$

To deal with (23), we give the following lemma.

Lemma 2: Let $f_k(q) = 1+C_rP_{hst}q - C_rq - e^{tq}(1+C_kP_{hst}q - C_kP_{hst})$, $f_k(q)$ is concave w.r.t. q for $q > 1$.

The proof is shown in Appendix B.

According to Lemma 2, the set Φ_k is convex. Hence, for the given secrecy throughput t, we can calculate the corresponding $\Phi_k, k \in \{1, \ldots, K\}$ (e.g., using numerical calculation method) as well as the set Φ. If $\Phi = \emptyset$, is implies the value of t is too large. Otherwise, we should increase t.

The algorithm for solving the TAsP is summarized in Algorithm 2.

Algorithm 2 Procedure of Solving the TAsP

Initialize t_{up}, t_{low}.
repeat
  Set $t = (t_{up} + t_{low})/2$.
  Get the sets $\Phi_k, \forall k$ and $\Phi = \bigcap_{k=1}^{K} \Phi_k$.
  If $\Phi = \emptyset$, $t_{up} = t$; else, $t_{low} = t$.
until $|t_{up} - t_{low}|$ converges.
Output:
  The optimal solution $\tilde{\alpha}^* = 1/\tilde{q}^*$ and corresponding secrecy throughput.

As mentioned above, we can obtain the PDM-TS for the STM by jointly optimizing the RF signal covariance $\mathbf{Q}_P$, the signal beamforming $\mathbf{w}$, the AN covariance $\mathbf{Q}_J$ and the time division ratio α. By iteratively solving the BFsP and the TAsP, the procedure of the PDM-TS design is summarized in Algorithm 3.

Algorithm 3 Procedure of the PDM-TS Design

Initialize α_0.
  Calculate the maximum harvested power P_{hst_m} with (13) and the corresponding $\mathbf{Q}_{P\_m}$.
repeat
  Solve the BFsP by Algorithm 1, obtain the optimal $\tilde{\mathbf{W}}^*, \tilde{\mathbf{Q}}_J^*$, calculate the maximum secrecy throughput T_s.
  Set $\bar{\mathbf{W}} = \frac{\tilde{\mathbf{W}}^*}{Tr(\tilde{\mathbf{W}}^*)}$ and $\bar{\mathbf{Q}}_J = \frac{\tilde{\mathbf{Q}}_J^*}{Tr(\tilde{\mathbf{Q}}_J^*)}$.
  Solve the TAsP by Algorithm 2, obtain the optimal $\tilde{\alpha}^*$ and the corresponding T_s.
  Set $\alpha_0 = \tilde{\alpha}^*$.
until T_s converges.
Output:
  The $\mathbf{Q}_{P\_m}$, the optimal solution $\mathbf{W}^*, \mathbf{Q}_J{}^*, \alpha^*$, as well as the corresponding T_s.

Converge analysis: With Algorithm 3, we obtain the optimal solution of problem Q_{RT}. We can prove it as following. Suppose in iteration i, the secrecy throughput obtained using Algorithm 1 and Algorithm 2 is T_{s1_i} and T_{s2_i}, respectively. It is satisfied that $T_{s1\_i} \le T_{s2\_i}$. For the next iteration $i+1$, it is satisfied that $T_{s2\_i} \le T_{s1\_i+1}$. In this way it is obtained that $T_{s1\_1} \le T_{s2\_1} \le T_{s1\_2} \le \cdots \le T_{s1\_i} \le T_{s2\_i}$, besides, the problem Q_{RT} is bounded. Hence, the Algorithm 3 will converge to the optimal solution.

IV. TRANSMIT SCHEME DESIGN WITH IMPERFECT CSI

In this section, we design the PDM-RTS for the wSTM by solving the problem Q_{RT}. Considering the Eves take passive intercepting method, we consider that the CSI of the wiretap channel is imperfect. In this case, the design of RF signal covariance $\mathbf{Q}_P$ is still the same with the derivation process in III. A. Hence, the wSTM can be represented as

$$Q_{RT1}: \quad \max_{\mathbf{w},\mathbf{Q}_J,\alpha} \min_{\delta_{hk},\delta_{gk}} T_s \tag{24a}$$

$$s.t.\ Tr(\mathbf{w}\mathbf{w}^H) \leq \frac{(1-\alpha)P_{hst\_m}}{\alpha}, \quad 0<\alpha<1 \tag{24b}$$

$$Tr(\mathbf{Q}_J) \leq P, \quad \mathbf{Q}_J \succeq \mathbf{0} \tag{24c}$$

$$||\delta_{hk}||^2 \leq \varepsilon_{hk}, ||\delta_{gk}||^2 \leq \varepsilon_{gk}, \quad k \in \{1,\ldots,K\} \tag{24d}$$

The problem Q_{RT1} is highly non-convex due to the objective function in (24a) and CSI errors constraint in (24d). Combined with the S-procedure, the PDM is extended for the robust transmit scheme design, and Q_{RT1} is divided into RBFsP and RTAsP. In the following, we will solve the two sub-problems, respectively.

A. SOLUTION OF THE RBFSP

In this part, the RBFsP is involved, i.e., jointly optimizing the $\mathbf{w}$, $\mathbf{Q}_J$ with fixed $\alpha = \alpha_0$ for the imperfect CSI. Introducing a slack variable φ which denotes the upper bound of wiretapping rate with imperfect CSI, the optimization problem can be represented as

Q_{RT1S} :

$$\max_{\mathbf{W},\mathbf{Q}_J,\alpha} \min_{\delta_{hk},\delta_{gk}} \log(1+\frac{\mathbf{h}_r{}^H\mathbf{W}\mathbf{h}_r}{\mathbf{g}_r{}^H\mathbf{Q}_J\mathbf{g}_r+\sigma_r{}^2})-\varphi \tag{25a}$$

$$s.t.\ \log[1+\frac{(\hat{\mathbf{h}}_k+\delta_{hk})^H\mathbf{W}(\hat{\mathbf{h}}_k+\delta_{hk})}{(\hat{\mathbf{g}}_k+\delta_{gk})^H\mathbf{Q}_J(\hat{\mathbf{g}}_k+\delta_{gk})+\sigma_k{}^2}] \leq \varphi, \quad k\in\{1,\ldots,K\} \tag{25b}$$

$$Tr(\mathbf{W}) \leq \frac{(1-\alpha_0)P_{hst\_m}}{\alpha_0}, \quad \mathbf{W}\succeq\mathbf{0} \tag{25c}$$

$$Tr(\mathbf{Q}_J) \leq P, \quad \mathbf{Q}_J\succeq\mathbf{0} \tag{25d}$$

$$||\delta_{hk}||^2 \leq \varepsilon_{hk}, ||\delta_{gk}||^2 \leq \varepsilon_{gk}, \quad k\in\{1,\ldots,K\} \tag{25e}$$

$$rank(\mathbf{W}) = 1 \tag{25f}$$

with $\mathbf{W} = \mathbf{w}\mathbf{w}^H$. Due to the errors constraint in (25e), the method in III. B cannot be applied for the problem Q_{RT1S} directly. To proceed, we design a bilevel quick search (BQS) method [45]. Taking a nested manner, the BQS can be divided into the inner problem and outer problem. For the inner problem, we fix the variable $\varphi = \varphi_f$ (i.e., guaranteeing the worst-case wiretapping rate no more than φ_f) and solve the corresponding Q_{RT1S}. Let $H(\varphi_f)$ denote the optimal solution of the inner problem. Then the outer problem is to find the optimal φ_f for maximizing the objective function of Q_{RT1S}, which can be equivalently solved by

$$\max_{\varphi_f} H(\varphi_f) \tag{26}$$

In the following, we will derive the inner problem first. With $\varphi = \varphi_f$, (25b) can be rewritten as

$$\begin{aligned}&\delta_{hk}{}^H\mathbf{W}\delta_{hk} - \delta_{gk}{}^H(e^{\varphi_f}-1)\mathbf{Q}_J\delta_{gk}\\&\quad+2Re\{\hat{\mathbf{h}}_k^H\mathbf{W}\delta_{hk} - \hat{\mathbf{g}}_k^H(e^{\varphi_f}-1)\mathbf{Q}_J\delta_{gk}\}\\&\quad+\hat{\mathbf{h}}_k^H\mathbf{W}\hat{\mathbf{h}}_k - \hat{\mathbf{g}}_k^H(e^{\varphi_f}-1)\mathbf{Q}_J\hat{\mathbf{g}}_k-(e^{\varphi_f}-1)\sigma_k{}^2 \leq 0\end{aligned} \tag{27}$$

Setting $\delta_k = [\delta_{hk}{}^H, \delta_{gk}{}^H]^H$, $\Xi = diag[\mathbf{W}, -(e^{\varphi_f}-1)\mathbf{Q}_J]$ and $\mathbf{e}_k = [\hat{\mathbf{h}}_k^H, \hat{\mathbf{g}}_k^H]^H$, (27) can be re-expressed as

$$\delta_k{}^H\Xi\delta_k + 2Re\{\mathbf{e}_k{}^H\Xi\delta_k\} + \mathbf{e}_k{}^H\Xi\mathbf{e}_k - (e^{\varphi_f}-1)\sigma_k{}^2 \leq 0 \tag{28}$$

To deal with (28), we introduce the following lemma.

Lemma 3 (S-Procedure [46]): Let $f_k(\mathbf{x}), k = 1, 2$ be defined as

$$f_k(\mathbf{x}) = \mathbf{x}^H\mathbf{A}_k\mathbf{x} + 2Re\{\mathbf{b}_k{}^H\mathbf{x}\} + c_k$$

where $\mathbf{A}_k = \mathbf{A}_k{}^H \in \mathbb{C}^{n\times n}$, $\mathbf{b}_k \in \mathbb{C}^n$ and $c_k \in \mathbb{R}$. Then, the implication $f_1(\mathbf{x}) \leq 0 \Rightarrow f_2(\mathbf{x}) \leq 0$ hold if and only if there exist $\mu \geq 0$ such that

$$\mu\begin{bmatrix}\mathbf{A}_1 & \mathbf{b}_1\\ \mathbf{b}_1{}^H & c_1\end{bmatrix} - \begin{bmatrix}\mathbf{A}_2 & \mathbf{b}_2\\ \mathbf{b}_2{}^H & c_2\end{bmatrix} \succeq \mathbf{0}$$

Provided there exists a point $\widehat{\mathbf{x}}$ with $f_k(\widehat{\mathbf{x}}) > 0$.

According to (25e), it can be obtained

$$\mathbf{e}_k{}^H\mathbf{e}_k = \delta_{hk}{}^H\delta_{hk} + \delta_{gk}{}^H\delta_{gk} \leq \varepsilon_{hk} + \varepsilon_{gk} \tag{29}$$

Based on Lemma 3, (28) holds with (29) is equal to $\exists u_k \geq 0$ and it is satisfied

$$u_k\begin{bmatrix}\mathbf{I} & \mathbf{0}\\ \mathbf{0} & -\varepsilon_{hk}-\varepsilon_{gk}\end{bmatrix} - \begin{bmatrix}\Xi & \Xi\mathbf{e}_k\\ \mathbf{e}_k{}^H\Xi & \mathbf{e}_k{}^H\Xi\mathbf{e}_k-(e^{\varphi_f}-1)\sigma_k{}^2\end{bmatrix} \succeq \mathbf{0} \tag{30}$$

Using the procedure above, (25b) and (25e) are transferred into (30), which is tractable Linear Matrix Inequalities (LMIs).

For (25a) which is non-convex, we employ the first-order Taylor expansion and get

$$\begin{aligned}\tilde{R}_r =& \log(\mathbf{g}_r{}^H\mathbf{Q}_J\mathbf{g}_r+\sigma_r{}^2+\mathbf{h}_r{}^H\mathbf{W}\mathbf{h}_r)-\varphi_f\\&-\log(\mathbf{g}_r{}^H\mathbf{Q}_{J0}\mathbf{g}_r+\sigma_r{}^2)-Tr\\&\times[\frac{\mathbf{g}_r\mathbf{g}_r{}^H}{\mathbf{g}_r{}^H\mathbf{Q}_{J0}\mathbf{g}_r+\sigma_r{}^2}(\mathbf{Q}_J-\mathbf{Q}_{J0})]\end{aligned} \tag{31}$$

where $\mathbf{Q}_{J0}$ is a constant matrix. Notice that (31) is concave w.r.t. $\mathbf{W}$ and $\mathbf{Q}_J$. Then the tractable approximation of inner problem is represented as

$$Q_{RT1S\_ai}: \quad \max_{\mathbf{W},\mathbf{Q}_J} \tilde{R}_r \tag{32a}$$

$$s.t.\ (30),\ k\in\{1,\ldots,K\} \tag{32b}$$

$$(25c), (25d) \tag{32c}$$

In the problem Q_{RT1S_ai}, we relax the rank constraint in (25f). Yet, the following proposition shows the rank-one property of the optimal solution.

Proposition 3: If the SDR problem Q_{RT1S_ai} is feasible, the optimal solution $\mathbf{W}^*$ yields $rank(\mathbf{W}^*) = 1$.

The proof is shown in Appendix A.

Then the inner problem can be solved using the SPCA method by iteratively solving the sequence of problem Q_{RT1S_ai}. For the outer problem, we can solve it using one dimensional search (e.g., golden research) to find the optimal φ_f.

The algorithm for solving the RBFsP is summarized in Algorithm 4.

Algorithm 4 BQS Procedure for the RBFsP

Initialize φ^{low} and φ^{up}.
repeat
  Set $\varphi_1 = 0.382(\varphi^{up} - \varphi^{low})$, $\varphi_2 = 0.618(\varphi^{up} - \varphi^{low})$.
  for $i = 1, 2$ **do**
    Initialization Q_{J0}.
    repeat
      Calculate the optimal solution of Q_{RT1S_ai} with $\varphi_f = \varphi_i$, get the optimal solution $\mathbf{W}_t^*$ and $\mathbf{Q}_{J_t}^*$.
      Set $\mathbf{Q}_{J0} = \mathbf{Q}_{J_t}^*$.
    until $\tilde{R}_r$ in (32a) converges.
    Get the response maximum secrecy rate $\tilde{R}_{r\_i}^*$, $i \in \{1, 2\}$.
  end for
  If $\tilde{R}_{r\_1}^* > \tilde{R}_{r\_2}^*$, set $\varphi^{up} = \varphi_2$; else set $\varphi^{low} = \varphi_1$.
until $|\varphi^{up} - \varphi^{low}|$ converges.
Output:
  The optimal solution $\tilde{\mathbf{W}}_R^*$, $\tilde{\mathbf{Q}}_{JR}^*$ and secrecy rate $\tilde{R}^* = (\tilde{R}_{r\_1}^* + \tilde{R}_{r\_2}^*)/2$.

B. SOLUTION OF THE RTASP

In this part, we derive the solution of the RTAsP, i.e., the optimal time division ratio α with the given $\tilde{\mathbf{W}}_R^*$, $\tilde{\mathbf{Q}}_{JR}^*$ for the imperfect CSI. Setting $\bar{\mathbf{W}}_R = \frac{\tilde{\mathbf{W}}_R^*}{Tr(\tilde{\mathbf{W}}_R^*)}$, $\bar{\mathbf{Q}}_{JR} = \frac{\tilde{\mathbf{Q}}_{JR}^*}{Tr(\tilde{\mathbf{Q}}_{JR}^*)}$, the secrecy throughput can be represented as

$$\tilde{T}_s = \alpha\{\log[1+\tilde{C}_r \frac{(1-\alpha)P_{hst\_m}}{\alpha}] - \max_{1\le k\le K} \log[1+\tilde{C}_k \frac{(1-\alpha)P_{hst\_m}}{\alpha}]\}, \quad 0 < \alpha < 1 \tag{33}$$

where

$$\tilde{C}_r = \frac{\mathbf{h}_r{}^H \bar{\mathbf{W}}_R \mathbf{h}_r}{\mathbf{g}_r{}^H \bar{\mathbf{Q}}_{JR} \mathbf{g}_r + \sigma_r{}^2}$$

$$\tilde{C}_k = \frac{(\hat{\mathbf{h}}_k + \delta_{hk})^H \bar{\mathbf{W}}_R (\hat{\mathbf{h}}_k + \delta_{hk})}{(\hat{\mathbf{g}}_k + \delta_{gk})^H \bar{\mathbf{Q}}_{JR} (\hat{\mathbf{g}}_k + \delta_{gk}) + \sigma_k{}^2}$$

We aim to maximize the worst-case secrecy throughput. According to (33), the worst case happens when the achievable rate of wiretap channel is maximized, which means the corresponding $\tilde{C}_k$ of Eve k is maximized. Since $\tilde{C}_k$ and α are uncoupled, we calculate the maximum $\tilde{C}_k$ first.

Proposition 4: Considering the CSI errors with $||\delta_{hk}||^2 \le \varepsilon_{hk}$, $||\delta_{gk}||^2 \le \varepsilon_{gk}$, it holds that $\tilde{C}_k \le \tilde{C}_{k\max} = \frac{c_{hk}}{c_{gk}}$, where

$$\begin{aligned} c_{hk} \triangleq\ & \max_{\mathbf{L}_{hk}} Tr(\mathbf{L}_{hk}\mathbf{Q}_{1hk}) && (34a)\\ & s.t.\ Tr(\mathbf{L}_{hk}\mathbf{Q}_{2h}) \le \varepsilon_{hk} && (34b)\\ & \quad Tr(\mathbf{L}_{hk}\mathbf{Q}_{3h}) = 1 && (34c)\\ & \quad \mathbf{L}_{hk} \succeq \mathbf{0}, \mathbf{L}_{hk} \in \mathbb{C}^{(N_s+1)\times(N_s+1)} && (34d) \end{aligned}$$

and

$$\begin{aligned} c_{gk} \triangleq\ & \min_{\mathbf{L}_{gk}} Tr(\mathbf{L}_{gk}\mathbf{Q}_{1gk}) && (35a)\\ & s.t.\ Tr(\mathbf{L}_{gk}\mathbf{Q}_{2g}) \le \varepsilon_{gk} && (35b)\\ & \quad Tr(\mathbf{L}_{gk}\mathbf{Q}_{3g}) = 1 && (35c)\\ & \quad \mathbf{L}_{gk} \succeq \mathbf{0}, \mathbf{L}_{gk} \in \mathbb{C}^{(N_h+1)\times(N_h+1)} && (35d) \end{aligned}$$

with the expressions of $\mathbf{Q}_{1hk} = \begin{bmatrix} \bar{\mathbf{W}}_R & \bar{\mathbf{W}}_R\hat{\mathbf{h}}_k \\ \hat{\mathbf{h}}_k^H\bar{\mathbf{W}}_R & \hat{\mathbf{h}}_k^H\bar{\mathbf{W}}_R\hat{\mathbf{h}}_k \end{bmatrix}$, $\mathbf{Q}_{2h} = diag(\mathbf{I}_{N_t}, 0)$, $\mathbf{Q}_{3h} = diag(\mathbf{0}_{N_t}, 1)$ and $\mathbf{Q}_{1gk} = \begin{bmatrix} \bar{\mathbf{Q}}_{JR} & \bar{\mathbf{Q}}_{JR}\hat{\mathbf{g}}_k \\ \hat{\mathbf{g}}_k^H\bar{\mathbf{Q}}_{JR} & \hat{\mathbf{g}}_k^H\bar{\mathbf{Q}}_{JR}\hat{\mathbf{g}}_k \end{bmatrix}$, $\mathbf{Q}_{2g} = diag(\mathbf{I}_{N_h}, 0)$, $\mathbf{Q}_{3g} = diag(\mathbf{0}_{N_h}, 1)$.

The proof is shown in Appendix C.

For Eve k, build the function with $\tilde{C}_k = \tilde{C}_{k\max}$ as

$$\tilde{T}_{s,k}(\alpha) = \alpha\{\log[1 + \tilde{C}_r \frac{(1-\alpha)P_{hst\_m}}{\alpha}] - \log[1 + \tilde{C}_{k\max} \frac{(1-\alpha)P_{hst\_m}}{\alpha}]\} \tag{36}$$

Define the sets $\tilde{\Phi}_k = \{\alpha \in R | \tilde{T}_{s,k}(\alpha) \ge t, t \ge 0\}$ and $\tilde{\Phi} = \bigcap_{k=1}^{K} \tilde{\Phi}_k$. Similar to the analysis in III. C, it can be verified that the set $\tilde{\Phi}_k$ is convex. Hence, we can calculate the corresponding $\tilde{\Phi}_k, k \in \{1, \ldots, K\}$ and the $\tilde{\Phi}$ using the similar method in III. C. The algorithm for solving the RTAsP is summarized in Algorithm 5.

Algorithm 5 Procedure of Solving the RTAsP

Initialize t_{up}, t_{low}.
repeat
  Set $t = (t_{up} + t_{low})/2$.
  Get the sets $\tilde{\Phi}_k, \forall k$ and $\tilde{\Phi} = \bigcap_{k=1}^{K} \tilde{\Phi}_k$.
  If $\tilde{\Phi} = \emptyset$, $t_{up} = t$; else, $t_{low} = t$.
until $|t_{up} - t_{low}|$ converges.
Output:
  The optimal solution $\tilde{\alpha}^* = 1/\tilde{q}^*$ and corresponding secrecy throughput.

As mentioned above, we can obtain the PDM-RTS for the wSTM. By iteratively solving the RBFsP and the RTAsP, the procedure of the PDM-RTS design is summarized in Algorithm 6.

Algorithm 6 Procedure of the PDM-RTS Design

Initialize α_0.

Calculate the maximum harvested power P_{hst_m} with (13) and the corresponding $\mathbf{Q}_{P\_m}$.

repeat

Solve the RBFsP by Algorithm 4, obtain the optimal $\tilde{\mathbf{W}}_R^*$, $\tilde{\mathbf{Q}}_{RJ}^*$, calculate the maximum secrecy throughput T_s.

Set $\bar{\mathbf{W}}_R = \frac{\tilde{\mathbf{W}}_R^*}{Tr(\tilde{\mathbf{W}}_R^*)}$ and $\bar{\mathbf{Q}}_{JR} = \frac{\tilde{\mathbf{Q}}_{JR}^*}{Tr(\tilde{\mathbf{Q}}_{JR}^*)}$.

Solve the RTAsP by Algorithm 5, obtain the optimal $\tilde{\alpha}^*$ and the corresponding T_s.

Set $\alpha_0 = \tilde{\alpha}^*$.

until T_s converges.

Output:

The $\mathbf{Q}_{P\_m}$, the optimal solution $\mathbf{W}_R{}^*$, $\mathbf{Q}_{RJ}{}^*$, α^*, as well as the corresponding T_s.

It should be noticed that the Algorithm 4 for solving the RBFsP is actually suboptimal using the BQS, hence the Algorithm 6 will converge to a suboptimal solution.

V. SIMULATION RESULTS AND ANALYSIS

In this section, the numerical simulation results are provided to validate the proposed transmit schemes. It is assumed that the transmitter is located on the line connecting the HBS and the receiving nodes. The distances from the HBS to transmitter, from transmitter to the receiving nodes and from HBS to receiving nodes are d_1, d_2 and D, respectively. It is satisfied that $D = d_1 + d_2$. The far-field WPCNs is considered with $d_1 \geq 1$, $d_2 \geq 1$ and $D = 10m$ [6].

The simplified distance-dependent path loss model is taken in this paper [21], [32]. All the entries of the channels are assumed as i.i.d. complex Gaussian random variables with zero mean and variances d^{-n}, where d denotes the distance between the two nodes and $n = 2$. Without loss of generality, the following setting is assumed, unless specified. Eves number $K = 3$, normalized transmission slot $T_0 = 1$, the noise power $\sigma_t{}^2 = \sigma_r{}^2 = \sigma_k{}^2 = -30dBm, k \in \{1, 2, 3\}$ [30]. The secrecy throughput is calculated for a whole transmission slot. The parameters of the non-linear EH model are $a_E = 150$, $b_E = 0.0014$, $M_E = 24mW$ and $\eta = 0.5$ [30]. For the CSI errors in (8), we define $\varepsilon_{hk} = u_h||\hat{\mathbf{h}}_k||^2$, $\varepsilon_{gk} = u_g||\hat{\mathbf{g}}_k||^2$ for $k \in \{1, 2, 3\}$, in which the parameters u_h and u_g denote the uncertainty of wiretap channel estimation [10], [25]. For comparison, we introduce the following schemes.

- AN in the null space of main channel (null-AN): The AN generated by HBS in the SIT lies in the null space of main channel $\mathbf{h}_r$ [23], [33], [34], [36].
- Transmit scheme without AN (no-AN): In the SIT, the HBS generate no AN [38].
- Transmit scheme with antenna selection (AS): In the SIT, the transmitter selects an optimal antenna to send secret information [37].

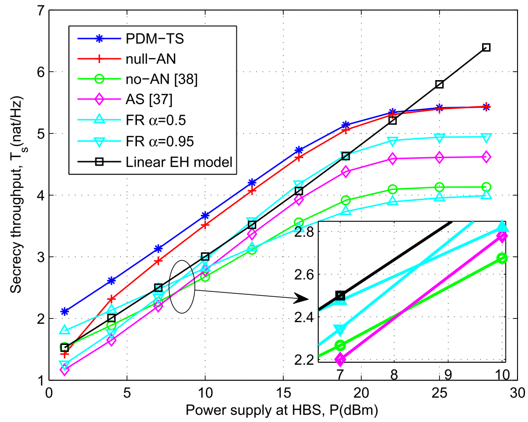

FIGURE 2. Secrecy throughput T_s versus the power supply at HBS P with perfect CSI.

- Fixed time division ratio scheme (FR): Set the time division ratio α as fixed.

We first exploit the secrecy performance with perfect CSI. The distance from transmitter to the receiving nodes is fixed as $d_1 = 5m$ with $N_h = N_t = 4$. Fig. 2 illustrates the secrecy throughput T_s as a function of the power supply at HBS P for different transmit schemes. Overall, the higher P leads to better secrecy performance. Compared with null-AN scheme, the PDM-TS almost shows no superiority when P is high. This is because with sufficient power supply, the advantage of well-designed AN using Algorithm 1 is not obvious. But the PDM-TS shows impressive advantage over the no-AN scheme, e.g., the gap is 0.5 nat/Hz when $P = 1dBm$ and increases to about 1.4 nat/Hz when $P = 28dBm$. The antenna selection is a sub-optimal diversity method, hence the AS scheme shows worse secrecy performance compared to the proposed PDM-TS with the gap about 0.9 nat/Hz. It also demonstrates that the FR scheme leads to secrecy performance degradation for both the low time division ratio $\alpha = 0.5$ and high ratio $\alpha = 0.95$. In fact, the optimal α is a function of CSI and power supply P, hence the fixed α cannot guarantee the optimum. We also illustrate the secrecy performance of transmit scheme with the linear EH model, where the energy conversation efficiency is set as $\rho_l = 0.5$. It can be found that the secrecy throughput T_s with linear EH model increases almost linearly, while T_s with non-linear EH model becomes saturated for high power supply P. This difference coincides with the RF harvesting characteristics of these two models shown in [28].

Fig. 3 shows the secrecy throughput T_s versus the distance d_1 for different antenna numbers of the HBS and the transmitter with $P = 10dBm$. It can be found that the T_s first decreases with d_1, for the reason that the RF power harvested by the transmitter becomes less. When the transmitter moves towards the receiving nodes with $d_1 \geq 5$, the T_s increases with d_1. This indicates that the influence of reducing path loss of information transmission in the SIT becomes more obvious. In addition we can find that increasing the antenna numbers of the HBS and transmitter is helpful for increasing

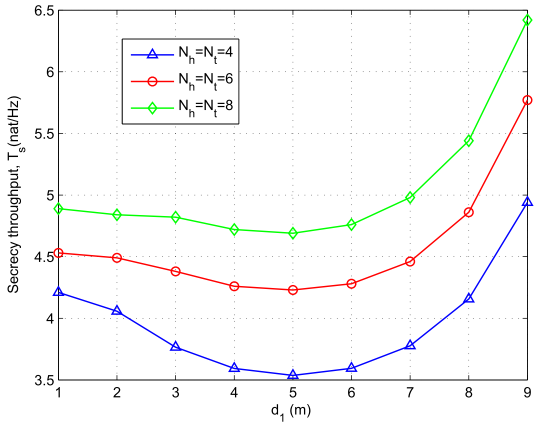

FIGURE 3. Secrecy throughput T_s versus the distance d_1 with perfect CSI.

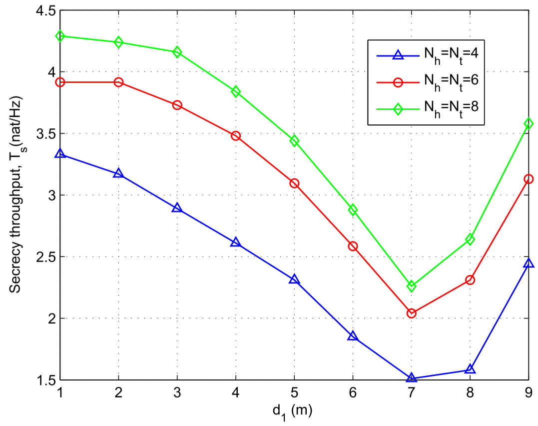

FIGURE 5. Secrecy throughput T_s versus the distance d_1 with imperfect CSI.

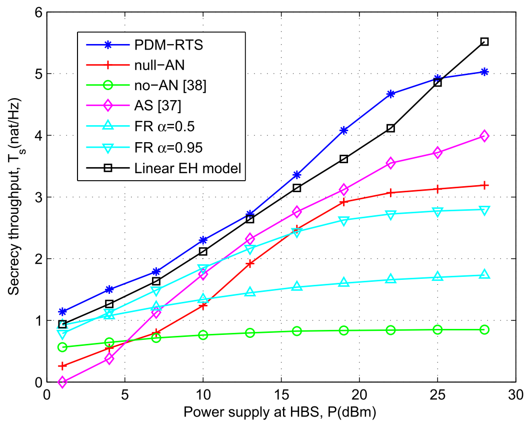

FIGURE 4. Secrecy throughput T_s versus the power supply at HBS P with imperfect CSI.

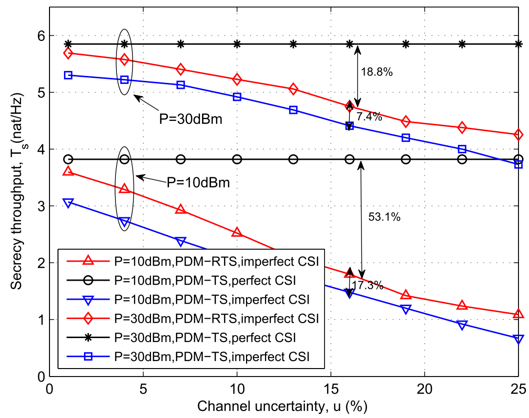

FIGURE 6. Secrecy throughput T_s versus channel uncertainty u.

the secrecy throughput. This is because that more transmit antennas provide more spatial degrees of freedom for beamforming design.

When the CSI of wiretap channel is imperfect, Fig. 4 verifies the secrecy performance of the proposed PDM-RTS, where $u_g = u_h = 0.1$ and $N_h = N_t = 4$. It can be found that the imperfect CSI degrades the secrecy performance. Different from the perfect CSI case, the PDM-RTS outperforms the null-AN scheme for high P, e. g., when $P \geq 19dBm$, the gap is larger than 1.1 nat/Hz. It implies that with imperfect CSI, the AN in null space of main channel obviously distinguishes from the optimal AN. The no-AN scheme shows worse performance which verifies the importance of AN for the imperfect CSI case. It should be noticed that the null-AN scheme is shown worse compared to the no-AN and FR schemes when P is low, because the sub-optimal AN signal leads to the reduction of beamforming power at transmitter. As P increases, the transmitter can harvest more power, hence the advantage of AN is enhanced, even it is sub-optimal. For the AS scheme, the performance gap caused by the sub-optimal diversity method is more obvious when the P is low. When the power supply for the transmitter is enough, the performance of AS scheme is much better.

Fig. 5 shows the secrecy throughput T_s versus the distance d_1 with imperfect CSI, where $P = 10dBm$ and $u_g = u_h = 0.1$. Similar to the perfect CSI case, the secrecy throughput T_s decreases first to the minimum and then increases as the distance d_1 grows. However, the turning point becomes larger compared with the perfect CSI case. This is because that the transmitter requires more harvested RF energy for the robust transmit scheme design with imperfect CSI, hence it is more difficult to overcome the decrease of harvested power in the PT. In addition, it can also be found that the imperfect CSI degrades secrecy performance compared with the results in Fig. 3.

Next, the influence of channel uncertainty is investigated in Fig. 6 with $u_g = u_h = u$, $d_1 = 5m$ and $N_h = N_t = 4$. For the deterministic model in (8), the larger channel uncertainty u extends feasible domain of CSI errors, which can enhance the intercepting capability of Eves for the worst case. Hence, the secrecy throughput T_s monotonically decreases as the channel uncertainty u grows using the PDM-RTS. The performance of the PDM-TS with perfect CSI is shown as a benchmark. It can be found that due to the CSI errors, when $u = 0.16$ the secrecy performance decreases by 53.1% and 18.8% for $P = 10dBm$ and $P = 30dBm$, respectively. How-

$$\max_{\mathbf{W},\mathbf{Q}_J} \log({\mathbf{g}_r}^H\mathbf{Q}_J\mathbf{g}_r+{\sigma_r}^2+{\mathbf{h}_r}^H\mathbf{W}\mathbf{h}_r)-\log a_r-Tr[\frac{\mathbf{g}_r{\mathbf{g}_r}^H}{a_r}(\mathbf{Q}_J-\mathbf{Q}_{J0})]-\phi \tag{37a}$$

$$s.t.\ \log b_k+Tr[\frac{\mathbf{g}_k{\mathbf{g}_k}^H}{b_k}(\mathbf{Q}_J-\mathbf{Q}_{J0})]+Tr[\frac{\mathbf{h}_k{\mathbf{h}_k}^H}{b_k}(\mathbf{W}-\mathbf{W}_0)]-\log({\mathbf{g}_k}^H\mathbf{Q}_J\mathbf{g}_k+{\sigma_k}^2)\le\phi, K\in\{1,\ldots,K\} \tag{37b}$$

$$Tr(\mathbf{W})\le\frac{(1-\alpha)P_{hst\_m}}{\alpha},\quad Tr(\mathbf{Q}_J)\le P \tag{37c}$$

$$\mathbf{W}\succeq\mathbf{0},\quad \mathbf{Q}_J\succeq\mathbf{0} \tag{37d}$$

ever, considering the CSI errors, the PDM-TS leads to performance degradation, e.g., the secrecy throughput decreases by 17.3% and 7.4% for $P = 10dBm$ and $P = 30dBm$ when $u = 0.16$ compared to the PDM-RTS.

VI. CONCLUSION

In this paper, we have studied the secure communication in the WPCNs based on the practical non-linear EH model. To guarantee the communication security, the secure two-phase communication protocol has been proposed. Based on this protocol, we designed the PDM-TS for the STM with perfect CSI and the PDM-RTS for the wSTM with imperfect CSI, respectively. Simulation results have been provided to demonstrate the effectiveness of the proposed transmit schemes. In addition, it was also shown that different from the linear EH model, the secrecy performance would reach a bottleneck for high power supply with the non-linear EH model, which indicates the existence of optimal power supply.

APPENDIX A
PROOF OF PROPOSITION 2 AND 3

Proof of Proposition 2: We rewrite (19) for detail (see(37)), as shown at the top of this page, where $a_r={\mathbf{g}_r}^H\mathbf{Q}_{J0}\mathbf{g}_r+{\sigma_r}^2$ and $b_k={\mathbf{g}_k}^H\mathbf{Q}_{J0}\mathbf{g}_k+{\sigma_k}^2+{\mathbf{h}_k}^H\mathbf{W}_0\mathbf{h}_k$.

The Lagrange dual function of (37) is

$$\begin{aligned}L_1 &= \log({\mathbf{g}_r}^H\mathbf{Q}_J\mathbf{g}_r+{\sigma_r}^2+{\mathbf{h}_r}^H\mathbf{W}\mathbf{h}_r)-\log a_r\\&\quad -Tr[\frac{\mathbf{g}_r{\mathbf{g}_r}^H}{a_r}(\mathbf{Q}_J-\mathbf{Q}_{J0})]-\phi\\&\quad -\sum\nolimits_{k=1}^{K}\lambda_k\{-\log({\mathbf{g}_k}^H\mathbf{Q}_J\mathbf{g}_k+{\sigma_k}^2)+\log b_k\\&\quad +Tr[\frac{\mathbf{g}_k{\mathbf{g}_k}^H}{b_k}(\mathbf{Q}_J-\mathbf{Q}_{J0})]+Tr[\frac{\mathbf{h}_k{\mathbf{h}_k}^H}{b_k}(\mathbf{W}-\mathbf{W}_0)]-\phi\}\\&\quad -\nu_1[Tr(\mathbf{W})-\frac{(1-\alpha)P_{hst\_m}}{\alpha}]-\vartheta_1[Tr(\mathbf{Q}_J)-P]\\&\quad +Tr(\mathbf{A}_1\mathbf{W})+Tr(\mathbf{B}_1\mathbf{Q}_J)\end{aligned} \tag{38}$$

where $\lambda_k\ge0, k\in\{1,\ldots,K\}$, $\nu_1\ge0$, $\vartheta_1\ge0$, $\mathbf{A}_1\succeq\mathbf{0}$, $\mathbf{B}_1\succeq\mathbf{0}$ are the dual variables associated (37b), (37c) and (37d), respectively. The derivative w.r.t. $\mathbf{W}$ is

$$\frac{\partial L_1}{\partial\mathbf{W}}=\frac{\mathbf{h}_k{\mathbf{h}_k}^H}{{\mathbf{g}_r}^H\mathbf{Q}_J\mathbf{g}_r+{\sigma_r}^2+{\mathbf{h}_r}^H\mathbf{W}\mathbf{h}_r}-\sum\nolimits_{k=1}^{K}\lambda_k\frac{\mathbf{h}_k{\mathbf{h}_k}^H}{b_k}-\nu_1\mathbf{I}+\mathbf{A}_1 \tag{39}$$

The corresponding KKT conditions are

$$\frac{\partial L_1}{\partial\mathbf{W}^*}=\mathbf{0} \tag{40a}$$

$$\mathbf{A}_1\mathbf{W}^*=\mathbf{0} \tag{40b}$$

According to (40a), it can be obtained

$$\mathbf{A}_1=\sum_{k=1}^{K}\lambda_k\frac{\mathbf{h}_k{\mathbf{h}_k}^H}{b_k}+\nu_1\mathbf{I}-\frac{\mathbf{h}_k{\mathbf{h}_k}^H}{{\mathbf{g}_r}^H\mathbf{Q}_J\mathbf{g}_r+{\sigma_r}^2+{\mathbf{h}_r}^H\mathbf{W}^*\mathbf{h}_r} \tag{41}$$

Hence, we can get $rank(\mathbf{A}_1)\ge N_t-1$. According to (40b), we can get $rank(\mathbf{W}^*)=N_t-rank(\mathbf{A}_1)$, which implies $rank(\mathbf{W}^*)\le1$. Since $\mathbf{W}^*=\mathbf{0}$ is not a feasible solution to (19), we can conclude that $rank(\mathbf{W}^*)=1$. The proof is completed.

Proof of Proposition 3: We firstly transfer (30) into

$$\begin{aligned}\Xi_k &= \begin{bmatrix}\mathbf{I}_{N_s} & \mathbf{0}_{(N_h+1)\times N_s}\end{bmatrix}^H\mathbf{W}\begin{bmatrix}\mathbf{I}_{N_s} & \mathbf{0}_{(N_h+1)\times N_s}\end{bmatrix}\\&\quad +\begin{bmatrix}\mathbf{I}_{N_s} & \mathbf{0}_{(N_h+1)\times N_s}\end{bmatrix}^H\mathbf{W}\begin{bmatrix}\mathbf{0}_{N_s\times(N_h+1)} & \hat{\mathbf{h}}_k\end{bmatrix}\\&\quad +\begin{bmatrix}\mathbf{0}_{N_s\times(N_h+1)} & \hat{\mathbf{h}}_k\end{bmatrix}^H\mathbf{W}\begin{bmatrix}\mathbf{I}_{N_s} & \mathbf{0}_{N_s\times(N_h+1)}\end{bmatrix}\\&\quad +\begin{bmatrix}\mathbf{0}_{N_s\times(N_s+N_h)} & \hat{\mathbf{h}}_k\end{bmatrix}^H\mathbf{W}\begin{bmatrix}\mathbf{0}_{N_s\times(N_s+N_h)} & \hat{\mathbf{h}}_k\end{bmatrix}\\&\quad +\begin{bmatrix}\mathbf{0}_{(N_h+1)\times N_s} & \mathbf{I}_{N_h+1}\end{bmatrix}^H\Theta_k\begin{bmatrix}\mathbf{0}_{(N_h+1)\times N_s} & \mathbf{I}_{N_h+1}\end{bmatrix}\succeq\mathbf{0}\end{aligned} \tag{42}$$

with $\Theta_k=\begin{bmatrix}\mathbf{Q}_J & \mathbf{Q}_J\mathbf{g}_k\\ \hat{\mathbf{g}}_k\mathbf{Q}_J & \hat{\mathbf{g}}_k\mathbf{Q}_J\mathbf{g}_k\end{bmatrix}$. Combined with (42), the Lagrange dual function of (32) can be written as

$$\begin{aligned}L_2 &= \log({\mathbf{g}_r}^H\mathbf{Q}_J\mathbf{g}_r+{\sigma_r}^2+{\mathbf{h}_r}^H\mathbf{W}\mathbf{h}_r)-\log a_r\\&\quad -Tr[\frac{\mathbf{g}_r{\mathbf{g}_r}^H}{a_r}(\mathbf{Q}_J-\mathbf{Q}_{J0})]-\varphi-\sum\nolimits_{k=1}^{K}Tr(\mathbf{C}_k\Xi_k)\\&\quad -\nu_2[Tr(\mathbf{W})-\frac{(1-\alpha)P_{hst\_m}}{\alpha}]-\vartheta_2[Tr(\mathbf{Q}_J)-P]\\&\quad +Tr(\mathbf{A}_2\mathbf{W})+Tr(\mathbf{B}_2\mathbf{Q}_J)\end{aligned} \tag{43}$$

where $\mathbf{C}_k\succeq\mathbf{0}, k\in\{1,\ldots,K\}$, $\nu_2\ge0$, $\vartheta_2\ge0$, $\mathbf{A}_2\succeq\mathbf{0}$, $\mathbf{B}_2\succeq\mathbf{0}$ are the corresponding dual variables. Besides,

$$\frac{\partial L_2}{\partial\mathbf{W}}=\frac{\mathbf{h}_k{\mathbf{h}_k}^H}{{\mathbf{g}_r}^H\mathbf{Q}_J\mathbf{g}_r+{\sigma_r}^2+{\mathbf{h}_r}^H\mathbf{W}\mathbf{h}_r}-\sum\nolimits_{k=1}^{K}\Psi_k-\nu_2\mathbf{I}+\mathbf{A}_2 \tag{44}$$

with

$$\begin{aligned}\Psi_k &= \left[\mathbf{I}_{N_s} \;\; \mathbf{0}_{(N_h+1)\times N_s}\right]^H \mathbf{C}_k \left[\mathbf{I}_{N_s} \;\; \mathbf{0}_{(N_h+1)\times N_s}\right] \\ &+ \left[\mathbf{I}_{N_s} \;\; \mathbf{0}_{(N_h+1)\times N_s}\right]^H \mathbf{C}_k \left[\mathbf{0}_{N_s\times(N_h+1)} \;\; \hat{\mathbf{h}}_k\right] \\ &+ \left[\mathbf{0}_{N_s\times(N_h+1)} \;\; \hat{\mathbf{h}}_k\right]^H \mathbf{C}_k \left[\mathbf{I}_{N_s} \;\; \mathbf{0}_{N_s\times(N_h+1)}\right] \\ &+ \left[\mathbf{0}_{N_s\times(N_s+N_h)} \;\; \hat{\mathbf{h}}_k\right]^H \mathbf{C}_k \left[\mathbf{0}_{N_s\times(N_s+N_h)} \;\; \hat{\mathbf{h}}_k\right]\end{aligned}$$

Based on the KKT conditions, we get

$$\mathbf{A}_2 = -\frac{\mathbf{h}_k \mathbf{h}_k{}^H}{\mathbf{g}_r{}^H \mathbf{Q}_J \mathbf{g}_r + \sigma_r{}^2 + \mathbf{h}_r{}^H \mathbf{W}^* \mathbf{h}_r} + \sum_{k=1}^{K} \Psi_k + \nu_2 \mathbf{I} \tag{45}$$

Since $\Psi_k \succeq \mathbf{0}$, it can be obtained $rank(\mathbf{A}_2) \geq N_t - 1$. Using the similar method in the proof of Proposition 2, it can be verified that $rank(\mathbf{W}^*) = 1$. Proposition 3 is proved.

APPENDIX B
PROOF OF LEMMA 2

For the function

$$f_k(q) = 1 + C_r P_{hst} q - C_r q - e^{tq}(1 + C_k P_{hst} q - C_k P_{hst}) \tag{46}$$

The first derivative can be expressed as

$$\begin{aligned}\frac{df_k(q)}{dq} &= C_r P_{hst} - e^{tq} t(1 + C_k P_{hst} q - C_k P_{hst}) - e^{tq} C_k P_{hst} \\ &= -t C_k P_{hst} q e^{tq} + (t C_k P_{hst} - t - C_k P_{hst}) e^{tq} + C_r P_{hst}\end{aligned} \tag{47}$$

The second derivative is calculated as

$$\begin{aligned}&\frac{d^2 f_k(q)}{dq^2} \\ &= -t C_k P_{hst} e^{tq} - t^2 C_k P_{hst} q e^{tq} + (t C_k P_{hst} - t - C_k P_{hst}) t e^{tq} \\ &= (t C_k P_{hst} - t C_k P_{hst} q - t - 2 C_k P_{hst}) t e^{tq}\end{aligned} \tag{48}$$

Duo to $q > 1, t > 0, C_k > 0, P_{hst} > 0$, it can be obtained that

$$\frac{d^2 f_k(q)}{dq^2} < 0, \quad q > 1 \tag{49}$$

Hence, Lemma 2 is proved.

APPENDIX C
PROOF OF PROPOSITION 4

It can be found in the $\tilde{C}_k = \frac{(\hat{\mathbf{h}}_k + \delta_{hk})^H \bar{\mathbf{W}}_R (\hat{\mathbf{h}}_k + \delta_{hk})}{(\hat{\mathbf{g}}_k + \delta_{gk})^H \bar{\mathbf{Q}}_{JR} (\hat{\mathbf{g}}_k + \delta_{gk}) + \sigma_k{}^2}$, the molecule and denominator are uncoupled. Hence, for calculating the maximum value of $\tilde{C}_k$, we can derive the maximum value of molecule and the minimum value of denominator, respectively.

Maximizing the molecule can be expressed

$$\max_{\delta_{hk}} \delta_{hk}{}^H \bar{\mathbf{W}}_R \delta_{hk} + 2Re\{\hat{\mathbf{h}}_k^H \bar{\mathbf{W}}_R \delta_{hk}\} + \hat{\mathbf{h}}_k^H \bar{\mathbf{W}}_R \hat{\mathbf{h}}_k \tag{50a}$$

$$s.t. \; \delta_{hk}{}^H \delta_{hk} \leq \varepsilon_{hk} \tag{50b}$$

(50) is an inhomogeneous quadratic problem, which is actually NP-hard. We exploit the SDR to obtain the optimal solution.

Setting $\mathbf{l}_k \triangleq [\delta_{hk}{}^H, t]^H$ with $t = 1$, (50) can be re-expressed as

$$\max_{\delta_{hk}} \mathbf{l}_k{}^H \begin{bmatrix} \bar{\mathbf{W}}_R & \bar{\mathbf{W}}_R \hat{\mathbf{h}}_k \\ \hat{\mathbf{h}}_k^H \bar{\mathbf{W}}_R & \hat{\mathbf{h}}_k^H \bar{\mathbf{W}}_R \hat{\mathbf{h}}_k \end{bmatrix} \mathbf{l}_k \tag{51a}$$

$$s.t. \; \delta_{hk}{}^H \delta_{hk} \leq \varepsilon_{hk} \tag{51b}$$

$$t = 1 \tag{51c}$$

With $\mathbf{L}_{hk} = \mathbf{l}_k \mathbf{l}_k{}^H$, $\mathbf{Q}_{1hk} = \begin{bmatrix} \bar{\mathbf{W}}_R & \bar{\mathbf{W}}_R \hat{\mathbf{h}}_k \\ \hat{\mathbf{h}}_k^H \bar{\mathbf{W}}_R & \hat{\mathbf{h}}_k^H \bar{\mathbf{W}}_R \hat{\mathbf{h}}_k \end{bmatrix}$, $\mathbf{Q}_{2h} = diag(\mathbf{I}_{N_t}, 0)$ and $\mathbf{Q}_{3h} = diag(\mathbf{0}_{N_t}, 1)$, it can be obtained that

$$\mathbf{l}_k{}^H \begin{bmatrix} \bar{\mathbf{W}}_R & \bar{\mathbf{W}}_R \hat{\mathbf{h}}_k \\ \hat{\mathbf{h}}_k^H \bar{\mathbf{W}}_R & \hat{\mathbf{h}}_k^H \bar{\mathbf{W}}_R \hat{\mathbf{h}}_k \end{bmatrix} \mathbf{l}_k = Tr(\mathbf{L}_{hk} \mathbf{Q}_{1hk}) \tag{52a}$$

$$\delta_{hk}{}^H \delta_{hk} \leq \varepsilon_{hk} \Leftrightarrow Tr(\mathbf{L}_{gk} \mathbf{Q}_{2g}) \leq \varepsilon_{gk} \tag{52b}$$

$$t = 1 \Leftrightarrow Tr(\mathbf{L}_{hk} \mathbf{Q}_{3h}) = 1 \tag{52c}$$

Since $\mathbf{L}_{hk} \succeq \mathbf{0}$, then, (51) can be transferred into (34). It can be proved that the optimal solution of (34) satisfies $rank(\mathbf{L}_{hk}{}^*) = 1$, hence the SDR solution of (34) is tight for (51).

Similarly, it can be derived that the problem of minimizing the denominator is equal to (35). Then, proof of Proposition 4 is complete.

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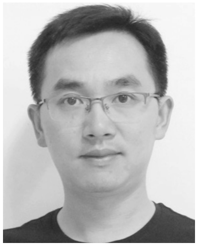

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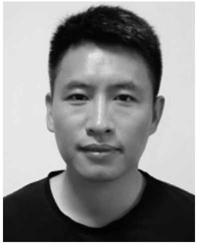

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