ROBUST HYBRID BEAMFORMING FOR SATELLITE-TERRESTRIAL INTEGRATED NETWORKS

Zhi Lin[1,2] *Min Lin*[3] *Benoit Champagne*[2] *Wei-Ping Zhu*[3,4] *Naofal Al-Dhahir*[5]

[1]College of Communications Engineering, Army Engineering University of PLA, China
[2]Department of Electrical and Computer Engineering, McGill University, Canada
[3]School of Communication and Information Engineering,
Nanjing University of Posts and Telecommunications, Nanjing, China
[4]Department of Electrical and Computer Engineering, Concordia University, Canada
[5]Department of Electrical and Computer Engineering, The University of Texas at Dallas, USA

ABSTRACT

In this paper, we propose a novel robust downlink beamforming (BF) design for satellite-terrestrial integrated networks. Under a realistic assumption that the angular information of eavesdroppers is not perfectly known, we establish an optimization framework for hybrid BF at the terrestrial base station and digital BF at the satellite to maximize the secrecy-energy efficiency of the system, while satisfying the quality-of-service constraints of both earth station and cellular user. Since the formulated optimization problem is mathematically intractable, we present an iterative algorithm based on the Charnes-Cooper approach to optimize the BF weight vectors. The effectiveness and superiority of the proposed robust hybrid BF scheme are validated via computer simulations.

Index Terms— Robust hybrid beamforming, satellite-terrestrial integrated networks, secrecy-energy efficiency

1. INTRODUCTION

Recently, the concept of satellite-terrestrial integrated networks (STIN) has received increased attention due to its advantages in providing both terrestrial and satellite network users with flexible access to new wireless communication services. [1]-[4]. STIN can maximize the utilization of wireless resources and make the transmissions within the satellite sub-networks and terrestrial cellular sub-networks more efficient. In this context, the authors in [5] provided a general framework of cooperative transmission in STIN, where the cases of unicast and multicast transmissions were discussed separately. By exploiting the *green* interference to enhance the security of satellite downlink wiretap channels, the authors of [6] proposed two zero-forcing based beamforming (BF) schemes to improve the system performance. This work was then extended to a more general case in [7], where a joint BF scheme was proposed to minimize the total transmit power of the satellite and base station under security constraints.

The secrecy-energy efficiency (SEE), expressed as the secrecy rate per Hertz per Joule, has become an important factor from both secrecy and ecological perspectives in addressing the demands of current communication networks while balancing the needs for data security and energy consumption [8]. Recently, large-scale antenna arrays have become a key element to combat the severe path loss anticipated for future wireless communication systems operating in the millimeter wave (mmWave) spectrum. Despite its theoretical advantages, implementing a massive antenna array with a large number of RF chains can be problematic, since it increases the computational complexity and power consumption, and lowers energy efficiency. To address these issues, hybrid beamforming has been proposed [9], [10], whereby the number of RF chains can be reduced through a cascade of digital and analog beamformers. The authors in [11] proposed a general optimization framework for hybrid analog-digital beamforming in STIN. However, the SEE optimization for hybrid BF design in STIN has not been addressed in the open literature so far. We first formulate a constrained optimization problem to maximize the SEE of the considered system while satisfying the signal-to-interference-plus-noise ratios (SINR) requirements of both the earth station and cellular user. Robustness is incorporated in the design by considering imperfect knowledge of the angles of departure for the downlink wiretap channels. We then propose a new iterative search algorithm based on the Charnes-Cooper approach to solve the optimization problem and obtain the desired hybrid BF weight vectors. Finally, we present simulation results to verify the effectiveness and advantages of the proposed hybrid BF scheme.

2. SYSTEM MODEL

As shown in Fig. 1, the satellite and cellular sub-networks share the same mmWave spectrum band, where the geostationary orbit (GEO) satellite serves an earth station (ES) in the presence of K eavesdroppers (Eves), while the BS serves

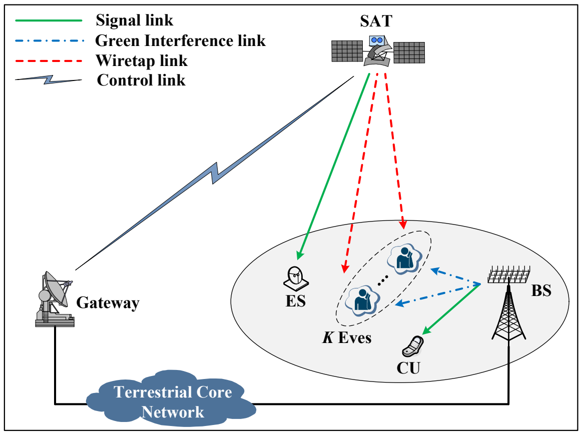

Fig. 1. System model of the considered STIN

a cellular user (CU). It is assumed that the Eves, but not the ES, are under coverage of the cellular sub-network, and therefore receive interference from the BS. In our model, all wireless channels are assumed to be frequency flat slow fading. For the Eves, in contrast to the ES and CU, only imperfect angles of departures (AoD) are available at the BS [12].

The multibeam satellite employs a fed reflector antenna array with N_s feeds uniformly positioned along a circle with radius d plus another feed at the center of the circle. By taking the effects of path loss, rain attenuation and satellite beam gains into account, the GEO downlink channel between the SAT and any user can be expressed as [13]

$$\mathbf{f} = \sqrt{C_L G_R / \xi}\, \mathbf{b}_g(\phi, \psi) \odot \mathbf{a}_c(\phi, \psi). \tag{1}$$

where C_L represents the path loss coefficient, G_R denotes the off-boresight antenna gain pattern, ξ is the rain attenuation coefficient, $\mathbf{b}_g(\phi, \psi)$ is the beam pattern vector of the satellite antenna feeds, and $\mathbf{a}_c(\phi, \psi)$ denotes the steering vector with $\phi \in [0, \pi/2)$ and $\psi \in [0, 2\pi)$ being the elevation and the azimuth angles, respectively.

In addition, we assume that the BS is equipped with a uniform planar array (UPA) of dimension $N_b = N_1 \times N_2$ to achieve high gain with compact size. Due to the highly directional and quasi-optical nature of mmWave transmissions, the terrestrial downlink channel can be modeled as the superposition of a predominant line-of-sight (LoS) component and a sparse set of single-bounce non-LoS (NLoS) components, which is adequate for urban environments. Hence, the terrestrial downlink channel can be expressed as [13]

$$\begin{aligned}\mathbf{h} = &\sqrt{g(\theta_0, \varphi_0)}\rho_0 \mathbf{a}_a(\theta_0, \varphi_0) \otimes \mathbf{a}_e(\theta_0, \varphi_0) \\ &+ \sqrt{\tfrac{1}{N}} \sum_{n=1}^{N} \sqrt{g(\theta_n, \varphi_n)}\rho_n \mathbf{a}_a(\theta_n, \varphi_n) \otimes \mathbf{a}_e(\theta_n, \varphi_n)\end{aligned} \tag{2}$$

where $g(\theta_0, \varphi_0)$ and $g(\theta_n, \varphi_n)$, $n = \{1, \cdots, N\}$ represent the directivity pattern of the LoS and the n-th NLoS links, ρ_0 and ρ_n denote the complex channel gains associated with the LoS path and the n-th NLoS link, $\mathbf{a}_a$ and $\mathbf{a}_e$ are the azimuth and elevation steering vectors of the UPA, respectively.

Let $s(t)$ denote the signal transmitted by the satellite to the ES. Prior to its transmission, the signal is mapped onto the BF weight vector $\mathbf{w} \in \mathbb{C}^{N_s \times 1}$ before transmission, which is intercepted by K Eves. Meanwhile, the BS sends signal $x(t)$ to the CU. Let $\mathbf{P} \in \mathbb{C}^{N_b \times N_r}$ denote analog precoder, consisting of phase shifters, and $\mathbf{v} \in \mathbb{C}^{N_r \times 1}$ denote the digital BF weight vector for signal $x(t)$. Hence, the received signals at the CU, ES, and k-th Eve are, respectively, expressed as

$$\begin{aligned} y_c(t) &= \mathbf{h}_c^H \mathbf{P}\mathbf{v} x(t) + \mathbf{f}_c^H \mathbf{w} s(t) + n_c(t), \\ y_s(t) &= \mathbf{f}_s^H \mathbf{w} s(t) + n_s(t), \\ y_k(t) &= \mathbf{f}_k^H \mathbf{w} s(t) + \mathbf{h}_k^H \mathbf{P}\mathbf{v} x(t) + n_k(t) \end{aligned} \tag{3}$$

where $\{n_c(t), n_s(t), n_k(t)\} \sim \mathcal{CN}(0, \sigma^2)$. According to (3), the SINR at the CU, ES, and k-th Eve are given by

$$\gamma_c = \frac{\left|\mathbf{h}_c^H \mathbf{P}\mathbf{v}\right|^2}{\left|\mathbf{f}_c^H \mathbf{w}\right|^2 + \sigma_c^2}, \gamma_s = \frac{\left|\mathbf{f}_s^H \mathbf{w}\right|^2}{\sigma_s^2}, \gamma_k = \frac{\left|\mathbf{f}_k^H \mathbf{w}\right|^2}{\left|\mathbf{h}_k^H \mathbf{P}\mathbf{v}\right|^2 + \sigma_k^2}. \tag{4}$$

Then, the achievable secrecy rate of the ES is given by

$$R_s = \left[\log_2(1 + \gamma_s) - \max_{k \in \{1, \cdots, K\}} \log_2(1 + \gamma_k)\right]^+ \tag{5}$$

where $[x] = \max(x, 0)$. The total power consumption of the considered system is modeled as

$$P_{tot} = \eta_1 \|\mathbf{w}\|^2 + \eta_2 \|\mathbf{v}\|^2 + P_S + P_B \tag{6}$$

where $\eta_1 > 1$ and $\eta_2 > 1$ are constants which account for the power amplifier inefficiency of the satellite and BS, respectively. P_S and P_B represent the fixed circuit power consumptions of the satellite and base station, respectively [14].

Considering the complex nature of information exchange, the high directionality of the mmWave channel and the limited feedback of the wiretap channels, the angular information based uncertainty model has been exploited in STIN [12]. Accordingly, we assume that the available cellular wiretap channel of the k-th Eves belongs to a given AoD uncertainty set Δ_k defined by $\theta_k \in \left[\theta_k^L, \theta_k^U\right]$ and $\varphi_k \in \left[\varphi_k^L, \varphi_k^U\right]$, this model is also applicable to the satellite wiretap channels. In this paper, we aim to maximize the SEE, while satisfying the SINR requirements of the ES and CU, analog precoder modula constraint and transmit power constraint. Mathematically, this optimization problem can be formulated as

$$\max_{\mathbf{w},\mathbf{v},\mathbf{P}} \min_{\Delta_k} R_s / P_{tot} \tag{7a}$$

$$\text{s.t. } \gamma_c \geq \Gamma_c, \tag{7b}$$

$$\gamma_s \geq \Gamma_s, \tag{7c}$$

$$\left|[\mathbf{P}]_{i,j}\right|^2 = 1/N_b,\ i = 1, \cdots, N_b,\ j = 1, \cdots, N_r, \tag{7d}$$

$$\|\mathbf{v}\|_F^2 \leq P_b,\ \|\mathbf{w}\|_F^2 \leq P_s \tag{7e}$$

where the fully-connected hybrid analog-digital architecture is considered in the above problem.

3. HYBRID BEAMFORMING SCHEME

Since the objective function (7a) is non-convex due to the imperfect wiretap channels with a continuum of possibilities, we exploit a discretization method to convert the wiretap channel boundary constraints into tractable forms. Specifically, we assume that the elevation and azimuth AoD angles for the wiretap channel of the k-th Eve can only take uniformly spaced values within their respective range $\theta_k \in \left[\theta_k^L, \theta_k^U\right]$ and $\varphi_k \in \left[\varphi_k^L, \varphi_k^U\right]$, as given by [15]

$$\begin{aligned} \theta_k^{(i)} &= \theta_k^L + i\Delta\theta,\ \ i = 0, \cdots, M_1, \\ \varphi_k^{(j)} &= \varphi_k^L + j\Delta\varphi,\ \ j = 0, \cdots, M_2 \end{aligned} \tag{8}$$

where $\Delta\theta = (\theta_k^U - \theta_k^L)/M_1$ and $\Delta\varphi = (\varphi_k^U - \varphi_k^L)/M_2$. The above formulation is also suitable for the AoD angles ϕ and ψ in the satellite downlink channel, but the details are omitted for brevity. Then, we define $\tilde{\mathbf{H}} = \sum_{i=0}^{M_1}\sum_{j=0}^{M_2} \mu_{i,j}\mathbf{H}^{(i,j)}$ and $\tilde{\mathbf{F}} = \sum_{i=0}^{M_1}\sum_{j=0}^{M_2} \mu_{i,j}\mathbf{F}^{(i,j)}$, where $\mathbf{H}^{(i,j)} = \mathbf{h}^{(i,j)}\left(\mathbf{h}^{(i,j)}\right)^H$, $\mathbf{F}^{(i,j)} = \mathbf{f}^{(i,j)}\left(\mathbf{f}^{(i,j)}\right)^H$, $\mu_{i,j} = \frac{1}{(M_1+1)(M_2+1)}$. By using these *averaged* channel matrices in problem (7) instead of the *imperfect* ones, the minimization over Δ_k can be removed. The validity of this approach has been demonstrated in [12].

In spite of the above simplification, the objective function of (7a) remains intractable due to the fractional nature of the SEE metric. Then, by invoking the Charnes-Cooper approach and introducing auxiliary variables α and τ, it can be further transformed as

$$\max_{\mathbf{W},\mathbf{V},\mathbf{P}} \tau^{-1}\log_2\left(\frac{\sigma^2 + \mathrm{Tr}\left(\mathbf{F}_s\mathbf{W}\right)}{\alpha}\right) \tag{9a}$$

$$\text{s.t. } \eta_1\mathrm{Tr}\left(\mathbf{W}\right) + \eta_2\mathrm{Tr}\left(\mathbf{V}\right) + P_S + P_B = \tau, \tag{9b}$$

$$\frac{\mathrm{Tr}\left(\tilde{\mathbf{F}}_k\mathbf{W}\right)}{\mathrm{Tr}\left(\mathbf{P}^H\tilde{\mathbf{H}}_k\mathbf{P}\mathbf{V}\right) + \sigma^2} \le \alpha,\ \ \forall k, \tag{9c}$$

$$\mathrm{Tr}\left(\mathbf{P}^H\mathbf{H}_c\mathbf{P}\mathbf{V}\right) - \Gamma_c\left(\mathrm{Tr}\left(\mathbf{F}_c\mathbf{W}\right) + \sigma^2\right) \ge 0, \tag{9d}$$

$$\mathrm{Tr}\left(\mathbf{F}_s\mathbf{W}\right) - \Gamma_s\sigma^2 \ge 0, \tag{9e}$$

$$\left|[\mathbf{P}]_{i,j}\right|^2 = 1/N_b, i = 1, \cdots, N_b, j = 1, \cdots, N_r, \tag{9f}$$

$$\mathrm{Tr}\left(\mathbf{W}\right) \le P_s, \mathrm{Tr}\left(\mathbf{V}\right) \le P_b, \tag{9g}$$

$$\mathrm{rank}\left(\mathbf{W}\right) = 1, \mathrm{rank}\left(\mathbf{V}\right) = 1 \tag{9h}$$

where $\mathbf{W} = \mathbf{w}\mathbf{w}^H$, $\mathbf{V} = \mathbf{v}\mathbf{v}^H$.

We note that due to the coupling of the search variables $\mathbf{v}$ and $\mathbf{P}$, the optimization problem (9) is still non-convex. In order to solve this non-convex problem, we propose a separate iterative optimization scheme. Suppose that after the n-th iteration, we have obtained an analog precoder $\mathbf{P}^{(n)}$, then the optimization problem for the digital beamforming weight vector can be expressed as

$$\max_{\mathbf{W},\mathbf{V},\tau,\alpha} \log_2\left(\frac{\sigma^2 + \mathrm{Tr}\left(\mathbf{F}_s\mathbf{W}\right)}{\alpha}\right)\ \tau^{-1} \tag{10a}$$

$$\text{s.t. } \eta_1\mathrm{Tr}\left(\mathbf{W}\right) + \eta_2\mathrm{Tr}\left(\mathbf{V}\right) + P_S + P_B = \tau, \tag{10b}$$

$$\frac{\mathrm{Tr}\left(\tilde{\mathbf{F}}_k\mathbf{W}\right)}{\mathrm{Tr}\left(\mathbf{P}^{(n)H}\tilde{\mathbf{H}}_k\mathbf{P}^{(n)}\mathbf{V}\right) + \sigma^2} \le \alpha,\ \ \forall k, \tag{10c}$$

$$\mathrm{Tr}\left(\mathbf{P}^{(n)H}\mathbf{H}_c\mathbf{P}^{(n)}\mathbf{V}\right) - \Gamma_c\left(\mathrm{Tr}\left(\mathbf{F}_c\mathbf{W}\right) + \sigma^2\right) \ge 0, \tag{10d}$$

$$\mathrm{Tr}\left(\mathbf{F}_s\mathbf{W}\right) - \Gamma_s\sigma^2 \ge 0, \tag{10e}$$

$$\mathrm{Tr}\left(\mathbf{W}\right) \le P_s, \mathrm{Tr}\left(\mathbf{V}\right) \le P_b, \tag{10f}$$

$$\mathrm{rank}\left(\mathbf{W}\right) = 1, \mathrm{rank}\left(\mathbf{V}\right) = 1 \tag{10g}$$

where $\mathbf{W} = \mathbf{w}\mathbf{w}^H$ and $\mathbf{V} = \mathbf{v}\mathbf{v}^H$. To reduce the computational complexity of (10), we propose s solution in two stages. Specifically, the outer problem can be written as

$$\max_{\tau} f\left(\tau\right)\ \tau^{-1} \tag{11}$$

where $f\left(\tau\right) = \log_2\left(\sigma^2 + \mathrm{Tr}\left(\mathbf{F}_s\mathbf{W}\right)/\alpha\right)$. Considering the monotonically increasing property of $\log_2\left(\cdot\right)$, the inner problem can be expressed as

$$\begin{aligned} &\max_{\mathbf{W},\mathbf{V},\alpha} \frac{\sigma^2 + \mathrm{Tr}\left(\mathbf{F}_s\mathbf{W}\right)}{\alpha} \\ &\text{s.t. } (10b) - (10g). \end{aligned} \tag{12}$$

Clearly, the outer problem can be solved by one-dimensional search on τ, while the inner problem can be solved by introducing $\beta = 1/\alpha$, $\bar{\mathbf{W}} = \mathbf{W}/\alpha$ and $\bar{\mathbf{V}} = \mathbf{V}/\alpha$. The inner optimization problem is then given by

$$\max_{\bar{\mathbf{W}},\bar{\mathbf{V}},\beta} \beta\sigma^2 + \mathrm{Tr}\left(\mathbf{F}_s\bar{\mathbf{W}}\right) \tag{13a}$$

$$\text{s.t. } \eta_1\mathrm{Tr}\left(\bar{\mathbf{W}}\right) + \eta_2\mathrm{Tr}\left(\bar{\mathbf{V}}\right) = \left(\tau - P_S - P_B\right)\beta, \tag{13b}$$

$$\beta\mathrm{Tr}\left(\tilde{\mathbf{F}}_k\bar{\mathbf{W}}\right) \le \mathrm{Tr}\left(\mathbf{P}^{(n)H}\tilde{\mathbf{H}}_k\mathbf{P}^{(n)}\bar{\mathbf{V}}\right) + \sigma^2,\ \forall k, \tag{13c}$$

$$\begin{aligned} &\mathrm{Tr}\left(\mathbf{P}^{(n)H}\mathbf{H}_c\mathbf{P}^{(n)}\bar{\mathbf{V}}\right) - \\ &\quad \Gamma_c\left(\mathrm{Tr}\left(\mathbf{F}_c\bar{\mathbf{W}}\right) + \sigma^2\right) \ge 0, \end{aligned} \tag{13d}$$

$$\mathrm{Tr}\left(\mathbf{F}_s\bar{\mathbf{W}}\right) - \Gamma_s\sigma^2\beta \ge 0, \tag{13e}$$

$$\mathrm{Tr}\left(\bar{\mathbf{W}}\right) \le P_s\beta,\ \ \mathrm{Tr}\left(\bar{\mathbf{V}}\right) \le P_b\beta, \tag{13f}$$

$$\mathrm{rank}\left(\bar{\mathbf{W}}\right) = 1, \mathrm{rank}\left(\bar{\mathbf{V}}\right) = 1. \tag{13g}$$

To handle the nonconvex constraint (13c), we introduce an auxiliary variable υ and obtain a second-order cone form as follows

$$\begin{aligned} &\tfrac{\beta + \mathrm{Tr}\left(\tilde{\mathbf{F}}_k\bar{\mathbf{W}}\right)}{2} \le \left\|\left[\tfrac{\beta - \mathrm{Tr}\left(\tilde{\mathbf{F}}_k\bar{\mathbf{W}}\right)}{2}, \upsilon\right]^T\right\|_2, \\ &\tfrac{c + \sigma^2 + 1}{2} \ge \left\|\left[\tfrac{c + \sigma^2 - 1}{2}, \upsilon\right]^T\right\|_2, \end{aligned} \tag{14}$$

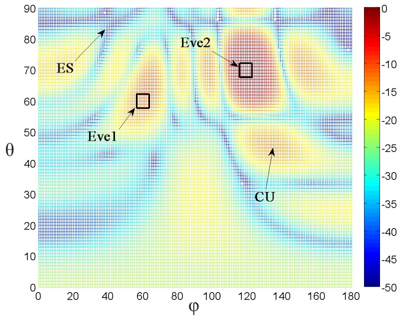

Fig. 2. 3D beampattern of $\mathbf{Pv}$

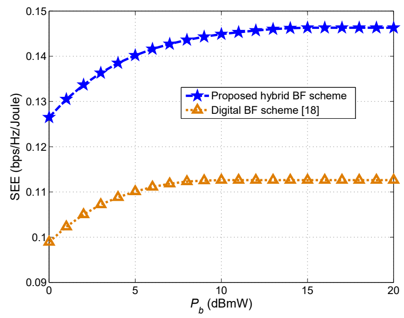

Fig. 3. SEE versus P_b

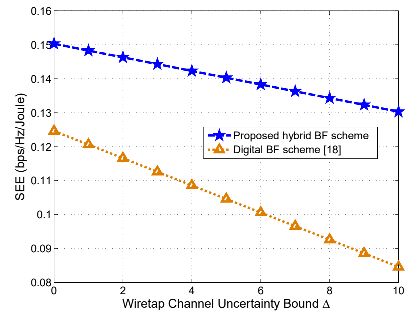

Fig. 4. SEE versus Δ

where $c = \mathrm{Tr}\left(\mathbf{P}^{(n)H}\tilde{\mathbf{H}}_k\mathbf{P}^{(n)}\bar{\mathbf{V}}\right)$. Thus, the problem (13) is convex except for the rank-1 constraints, which can be solved using semidefinite relaxation (SDR) and randomization as reported in [16].

Next, we focus on solving for the analog precoder $\mathbf{P}$. Once the solution of (13) $\{\mathbf{w}^{(n)}, \mathbf{v}^{(n)}\}$ is obtained, $\gamma_{s,l}$ and P_{tot} become constant with known $\{\mathbf{w}^{(n)}, \mathbf{v}^{(n)}\}$. The optimization problem of the analog precoder can then be written as follows

$$\begin{aligned} &\max_{\mathbf{P}} \min_{\Delta_k} \min_{k\in\{1,\cdots,K\}} \frac{\left|\mathbf{h}_k^H\mathbf{P}\mathbf{v}^{(n)}\right|^2+\sigma^2}{\left|\mathbf{f}_k^H\mathbf{w}^{(n)}\right|^2} \\ &\text{s.t. } \frac{\left|\mathbf{h}_c^H\mathbf{P}\mathbf{v}^{(n)}\right|^2}{\left|\mathbf{f}_c^H\mathbf{w}^{(n)}\right|+\sigma^2} \geq \Gamma_c, \\ &\quad \left|[\mathbf{P}]_{i,j}\right|^2 = \beta_{i,j}, i = 1,\cdots,N_b,\ j = 1,\cdots,N_r. \end{aligned} \tag{15}$$

It is noted that the constraints (9c) and (9e) can be directly removed because $\{\mathbf{w}^{(n)}, \mathbf{v}^{(n)}\}$ always satisfy these constraints. Then, (15) can be rewritten in a vector form that can be easily solved, namely

$$\begin{aligned} &\max_{\mathbf{p}} \min_{\Delta_k} \min_{k\in\{1,\cdots,K\}} \frac{\left|\mathbf{h}_k^H\hat{\mathbf{V}}^{(n)}\mathbf{p}\right|^2+\sigma^2}{\left|\mathbf{f}_k^H\mathbf{w}^{(n)}\right|^2} \\ &\text{s.t. } \frac{\left|\mathbf{h}_c^H\hat{\mathbf{V}}^{(n)}\mathbf{p}\right|^2}{\left|\mathbf{f}_c^H\mathbf{w}^{(n)}\right|^2+\sigma^2} \geq \Gamma_c, \\ &\quad \left|[\mathbf{p}]_q\right|^2 = \mathrm{vec}(\Phi)_q,\ q = 1,\cdots,N_bN_r \end{aligned} \tag{16}$$

where $\mathbf{p} = \mathrm{vec}\left(\mathbf{P}\right) \in \mathbb{C}^{N_bN_r\times 1}$, $\hat{\mathbf{V}}^{(n)} = \mathrm{block}-\mathrm{diag}(\mathbf{v}^{(n)T}, \cdots, \mathbf{v}^{(n)T}) \in \mathbb{C}^{N_b\times N_bN_r}$, $\Phi = \mathbf{1}_{N_b\times N_r}/N_b$.

Thus, problem (16) can be further rewritten as

$$\begin{aligned} &\max_{\hat{\mathbf{P}}}\ t \\ &\text{s.t. } \mathrm{Tr}\left(\hat{\mathbf{V}}^{(n)H}\tilde{\mathbf{H}}_k\hat{\mathbf{V}}^{(n)}\hat{\mathbf{P}}\right) + \sigma^2 \geq \\ &\qquad\qquad t\mathrm{Tr}\left(\tilde{\mathbf{F}}_k\mathbf{W}^{(n)}\right),\ \forall k, \\ &\quad \mathrm{Tr}\left(\hat{\mathbf{V}}^{(n)H}\tilde{\mathbf{H}}_c\hat{\mathbf{V}}^{(n)}\hat{\mathbf{P}}\right) \geq \\ &\qquad\qquad \Gamma_c\left(\mathrm{Tr}\left(\mathbf{F}_c\mathbf{W}^{(n)}\right) + \sigma^2\right), \\ &\quad \mathrm{diag}\left[\hat{\mathbf{P}}\right]_q = \left[\mathbf{q}\mathbf{q}^H\right]_q,\ q = 1,\cdots,N_bN_r, \\ &\quad \mathrm{rank}\left(\hat{\mathbf{P}}\right) = 1 \end{aligned} \tag{17}$$

where $\hat{\mathbf{P}} = \mathbf{p}\mathbf{p}^H$, $\mathbf{q} = \mathrm{vec}\left(\Phi\right)$. Thus, problem (17) can be solved using the SDR and randomization method.

4. NUMERICAL RESULTS

We consider a STIN scenario where an ES is intercepted by $K = 2$ Eves. The SINR thresholds of the IRs and ES are set as $\Gamma_c = \Gamma_s = -10$ dB. The power amplifier inefficiency of the satellite and BS as $\eta_1 = \eta_2 = 1/0.39$ [17]. The antennas number of the satellite and BS are $N_s = 7$ and $N_b = 8 \times 8$, with $N_r = 4$ RF chains. Unless otherwise indicated the transmit power budget or the satellite and BS are set as $P_b = 30$ dBmW and $P_s = 48$ dBmW, while the nominal value for the channel AoD uncertainty region is $\Delta = 4°$. In addition, the digital BF scheme in [18] is adopted as a benchmark.

Fig. 2 depicts the three-dimensional (3D) beampattern of the BS beamformer $\mathbf{Pv}$, where the two main lobes of the beampattern point to the Eves with a level of at least -10 dB within the uncertainty region. The received SINR of the intended CU also meets the required thresholds, while a null is generated with -45 dB at the ES. Fig. 4 depicts the SEE versus the BS transmit power budget P_b. Clearly, the SEE of the proposed hybrid BF scheme outperforms that of the digital BF scheme due to the lower power consumption of RF chains. Fig. 5 plots the SEE versus wiretap channel error bound, which shows the robustness of the proposed hybrid BF scheme to changes in the size of the AoD uncertainty region.

5. CONCLUSION

In this paper, we have proposed a hybrid BF scheme to achieve SEE maximization in STIN. To solve the original non-convex problem, we first used a discretization method to transform the constraints on the imperfect channel AoD into solvable ones. Then, an iterative BF algorithm based on the Charnes-Cooper method was conceived to solve the problem and obtain the digital and analog BF weight vectors. Finally, numerical results were given to demonstrate the superiority and effectiveness of the proposed hybrid BF scheme in comparison with an existing method.

6. REFERENCES

[1] M. Jia, X. Gu, Q. Guo, W. Xiang, and N. Zhang, "Broadband hybrid satellite-terrestrial communication systems based on cognitive radio toward 5G," *IEEE Wireless Commun.*, vol. 23, no. 6, pp. 96-106, Dec. 2016.

[2] B. Li, Z. Fei, Z. Chu, F. Zhou, K.-K. Wong, and Pei Xiao, "Robust chance-constrained secure transmission for cognitive satellite-terrestrial networks," *IEEE Trans. Veh. Technol.*, vol. 67, no. 5, pp. 4208-4219, May 2018.

[3] Z. Lin, M. Lin, J. Ouyang, W.-P. Zhu, and S. Chatzinotas, "Beamforming for secure wireless information and power transfer in terrestrial networks coexisting with satellite networks," *IEEE Signal Process. Lett.*, vol. 25, no. 8, pp. 1166-1170, Aug. 2018.

[4] J. Du, C. Jiang, H. Zhang, X. Wang, Y. Ren, and M. Debbah, "Secure satellite-terrestrial transmission over incumbent terrestrial networks via cooperative beamforming," *IEEE J. Sel. Areas Commun.*, vol. 36, no. 7, pp. 1367-1382, Jul. 2018.

[5] X. Zhu, C. Jiang, L. Kuang, N. Ge, S. Guo, and J. Lu, "Cooperative transmission in integrated terrestrial-satellite networks," *IEEE Network*, vol. 36, no. 7, pp. 1367-1382, Jul. 2018.

[6] K. An, M. Lin, J. Ouyang, and W.-P. Zhu, "Secure transmission in cognitive satellite terrestrial networks," *IEEE J. Sel. Areas Commun.*, vol. 34, no. 11, pp. 3025-3037, Nov. 2016.

[7] M. Lin, Z. Lin, W.-P. Zhu, and J.-B. Wang, "Joint beamforming for secure communication in cognitive satellite terrestrial networks," *IEEE J. Sel. Areas Commun.*, vol. 36, no. 5, pp. 1017-1029, May 2018.

[8] J. Ouyang, M. Lin, Y. Zou, W.-P. Zhu, and D. Massicotte, "Secrecy energy efficiency maximization in cognitive radio networks," *IEEE Access*, vol. 5, pp. 2641-2650, Feb. 2017.

[9] I. Ahmed *et al.*, "A survey on hybrid beamforming techniques in 5G: architecture and system model perspectives," *IEEE Commun. Surveys Tut.*, vol. 20, no. 4, pp. 3060-3097, 4th Quart., 2018.

[10] A. F. Molisch *et al.*, "Hybrid beamforming for massive MIMO: a survey," *IEEE Commun. Mag.*, vol. 55, no. 9, pp. 134-141, Sep. 2017.

[11] M. A. Vaquez, L. Blanco, X. Artiga and A. I. P.-Neira, "Hybrid analog-digital transmit beamforming for spectrum sharing satellite-terrestrial systems," *IEEE 17th International Workshop on SPAWC*, pp. 1-5, July 2016.

[12] Z. Lin, M. Lin, J.-B. Wang, Y. Huang, and W.-P. Zhu, "Robust secure beamforming for 5G cellular networks coexisting with satellite networks," *IEEE J. Sel. Areas Commun.*, vol. 36, no. 4, pp. 932-945, Apr. 2018.

[13] Z. Lin, M. Lin, J. Wang, T. De Cola, and J. Wang, "Joint beamforming and power allocation for satellite-terrestrial integrated networks with non-orthogonal multiple access," *IEEE J. Sel. Topics Signal Process.*, vol. 13, no. 3, pp. 657-670, Jun. 2019.

[14] S. He, C. Qi, Y. Wu, and Y. Huang, "Energy-efficient transceiver design for hybrid sub-array architecture MIMO systems," *IEEE Access*, vol. 4, pp. 9895-9905, 2016.

[15] Z. Lin, M. Lin, Y. Huang, T. De Cola, and W.-P. Zhu, "Robust multi-objective beamforming for integrated satellite and high altitude platform network with imperfect channel state information," *IEEE Trans. Signal Process.*, vol. 67, no. 24, pp. 6384-6386, Dec. 2019.

[16] J. Huang and A. L. Swindlehurst, "Robust secure transmission in MISO channels based on worst-case optimization," *IEEE Trans. Signal Process.*, vol. 60, no. 4, pp. 1696-1707, Apr. 2012.

[17] X. Gao, L. Dai, and A. M. Sayeed, "Low RF-complexity technologies to enable millimeter-wave MIMO with large antenna array for 5G wireless communications," *IEEE Commun. Mag.*, vol. 56, no. 4, pp. 211-217, Apr. 2018.

[18] Y. Jiang, Y. Zou, J. Ouyang, and Jia Zhu, "Secrecy energy efficiency optimization for artificial noise aided physical-layer security in OFDM-based cognitive radio networks," *IEEE Trans. Veh. Technol.*, vol. 67, no. 12, pp. 11858-11872, Dec. 2018.