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Beamforming for Secure Wireless Information and Power Transfer in Terrestrial Networks Coexisting With Satellite Networks

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***Abstract*—This letter proposes a beamforming (BF) scheme to enhance wireless information and power transfer in terrestrial cellular networks coexisting with satellite networks. By assuming that the energy receivers are the potential eavesdroppers overhearing signals intended for information receivers (IRs), we first formulate a constrained optimization problem to maximize the minimal achievable secrecy rate of the IRs subject to the constraints of energy harvest requirement, interference threshold, and transmit power budget. Through exploiting the sequential convex approximation method, we convert the original problem into a linear one with a series of linear matrix inequality and second-order cone constraints. An iterative algorithm is then proposed to obtain the BF weight vectors. Finally, simulation results demonstrate the effectiveness and superiority of the proposed scheme.**

***Index Terms*—Secure communications, wireless information and power transfer, sequential convex approximation (SCA).**

I. INTRODUCTION

TO ACHIEVE gigabit-per-second data rates in the next generation mobile communications, one option is to leverage the available bandwidths at millimeter wave (mmWave) band. However, some portion of the mmWave spectrum has been allocated to satellite services. This form of the spectral coexistence of satellite network and terrestrial network, termed as cognitive satellite–terrestrial network (CSTN), is becoming a hot topic to tackle the spectrum congestion problem [1]–[8].

Wireless information and power transfer (WIPT) has been considered as a promising technology for small cell networks [9]–[13], where a base station (BS) serves a number of distributed users, such as information receivers (IRs) and energy receivers (ERs). However, the feature of energy harvesting give rise to security issues, because ERs can operate as potential eavesdroppers to intercept the private information sent to IRs [14]. In this context, considering that beamforming (BF) can both enhance the desired signal to intended users and suppress the interference to other users [15]–[17]; its application to WIPT systems to guarantee secure communications has been widely investigated (e.g., see [18] and [19], and the references there in). However, to the best of our knowledge, the secure communication of WIPT in CSTN is yet an open topic, and no result in this field has been reported so far. This observation motivates the work in this letter.

This letter studies secure communication for WIPT in terrestrial networks coexisting with satellite networks. We aim to maximize the minimal achievable secrecy rate (ASR) of multiple IRs subject to the constraints of energy harvest requirement, interference threshold of earth station (ES), and the transmit power budget of the BS. As the original problem is nonconvex, we employ the sequential convex approximation (SCA) method and propose an iterative BF algorithm to obtain the optimal solution, which is different from the conventional semidefinite relaxation (SDR) based methods [3], [14]. Simulation results confirm the effectiveness and superiority of our proposed BF scheme.

Notation: $(\cdot)^H$ stands for the Hermitian transpose of a matrix. $\|\cdot\|$ and $|\cdot|$ denote Euclidean norm and absolute value of a vector, and $\mathbf{x} \otimes \mathbf{y}$ denotes the Kronecker product of two vectors $\mathbf{x}$ and $\mathbf{y}$. $C^{M \times N}$ denotes the complex space of $M \times N$, $\log(\cdot)$ denotes the logarithmic function, and $\mathcal{CN}(\mu, \sigma^2)$ denotes the complex Gaussian distribution with mean μ and variance σ^2.

II. SYSTEM MODEL AND PROBLEM FORMULATION

We consider the forward link of a satellite network, where the satellite (SAT) generates K adjacent beams on the ground, and a single user per beam is served in each time slot through time division multiple access. Meanwhile, in the terrestrial network, the BS having multiantenna serves M IRs with the mth IR surrounded by K_m ERs. We assume that each of IRs and ERs is employed with a single antenna, and the channel

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state informations are available at BS, which is commonly used in most of related works [14], [19], [20]. However, different from the previous works, here the ES employs a parabolic antenna and the BS employs a uniform planar array (UPA) of dimension $N = N_1 \times N_2$ with N_1 and N_2 being the number of UPA uniformly placed along the horizontal axis and that along the vertical axis with interelement spacing d_1 and d_2, respectively. The two-dimensional (2-D) array steering vector (SV) of BS antenna can then be expressed as [21]

$$\mathbf{a}(\theta, \varphi) = \mathbf{a}_h(\theta, \varphi) \otimes \mathbf{a}_v(\theta) \tag{1}$$

where $\theta \in [0, \pi/2)$ and $\varphi \in [0, \pi)$ denote the vertical and horizontal angle-of-arrival, respectively. $\mathbf{a}_h(\theta, \varphi)$ and $\mathbf{a}_v(\theta)$ denote the $N_1 \times 1$ horizontal and $N_2 \times 1$ vertical SV, which are, respectively, given by

$$\mathbf{a}_h(\theta, \varphi) = \left[e^{-j\beta(n_1-(N_1-1)/2)d_1 \sin\theta \cos\varphi}\right]_{n_1 \in \Psi(N_1)} \tag{2a}$$

$$\mathbf{a}_v(\theta, \varphi) = \left[e^{-j\beta(n_2-(N_2-1)/2)d_2 \cos\theta}\right]_{n_2 \in \Psi(N_2)} \tag{2b}$$

with $\beta = 2\pi/\lambda$, λ being the carrier wavelength, and $\Psi(N) = \{n : n = 0, \ldots, N-1\}$. The geometry-based mmWave 3-D sparse channel between the BS and any user (ES, IR, ER) can be uniformly written as [21]

$$\mathbf{h} = \rho_0 \mathbf{a}(\theta_0, \varphi_0) + \sqrt{\frac{1}{L}} \sum_{l=1}^{L} \rho_l \mathbf{a}(\theta_l, \varphi_l) \tag{3}$$

where *L* denotes the total number of non-line-of-sight (NLoS) paths, and ρ_l denotes the complex path loss associated with the *l*th path. Here, the amplitudes of the NLoS components, namely $|\rho_l|^2, (l = 1, \ldots, L)$, are typically 5 to 10 dB weaker than that of LoS component $|\rho_0|^2$.

By supposing that the transmit signal from BS to the *m*th IR is $x_m(t)$ with normalized power, the received signals at the *m*th IR and the surrounding ERs are, respectively, given by

$$y_m(t) = \mathbf{h}_m^H \mathbf{w}_m x_m(t) + \sum_{i=1, i\neq m}^{M} \mathbf{h}_m^H \mathbf{w}_i x_i(t) + n_m(t) \tag{4a}$$

$$y_{m,k}(t) = \mathbf{g}_{m,k}^H \mathbf{w}_m x_m(t) + \sum_{i=1, i\neq m}^{M} \mathbf{g}_{m,k}^H \mathbf{w}_i x_k(t) + n_{m,k}(t) \tag{4b}$$

where $\mathbf{w}_m \in C^{N\times1}$ denotes the transmit BF weight vector at BS, $\mathbf{h}_m \in C^{N\times1}$ and $\mathbf{g}_{m,k} \in C^{N\times1}$ the channel vector between the BS and the *m*th IR and that between BS and the *k*th ER around the IR, respectively. Similar to [16], here we ignore the weak interference from SAT to IRs and ERs. In addition, the interference received at the ES from the BS can be expressed as

$$y_p(t) = \sqrt{G_p(\phi)} \sum_{m=1}^{M} \mathbf{h}_p^H \mathbf{w}_m x_m(t) + n_p(t) \tag{5}$$

where $\mathbf{h}_p \in C^{N\times1}$ denotes the channel vector between the BS and the ES. In (4) and (5), $n_i, i \in \{m, k, p\}$ are the additive white Gaussian noise with zero mean and variance of σ_i^2, satisfying $\sigma_i^2 = \kappa T_i B_i$, where κ denotes Boltzmann constant, T_i denotes the noise temperature, and B_i denotes the noise bandwidth. In (5), $G_p(\phi) = 10^{\hat{G}_p(\phi)/10}$, with $\hat{G}_p(\phi)$ being the off-boresight antenna gain pattern in dB, which is given by [22]

$$\tilde{G}_p(\phi) = \begin{cases} -2.5 \times 10^{-3}\left(\frac{D}{\lambda}\phi\right)^2, & 0^\circ < \phi < \phi_m \\ 2 + 15\log\frac{D}{\lambda}, & \phi_m \leq \phi < \phi_r \\ 32 - 25\log\phi, & \phi_r \leq \phi < 48^\circ \\ -10, & 48^\circ \leq \phi \leq 180^\circ \end{cases} \tag{6}$$

where $G_{\max}$ denotes the maximal gain of the ES antenna, *D* the antenna diameter, λ the wavelength, and ϕ the off-boresight angle. In (6), $\phi_m = \frac{20\lambda}{D}\sqrt{G_{\max} - \left(2 + 15\log\frac{D}{\lambda}\right)}$ and $\phi_r = 15.85\left(\frac{D}{\lambda}\right)^{-0.6}$ are in degrees.

Based on (4), the output signal-to-interference-plus-noise-ratio of the *m*th IR and its surrounding ERs can be, respectively, expressed as

$$\gamma_m = \frac{\left|\mathbf{h}_m^H \mathbf{w}_m\right|^2}{\sum_{i=1, i\neq m}^{M} \left|\mathbf{h}_m^H \mathbf{w}_i\right|^2 + \sigma_m^2} \tag{7a}$$

$$\gamma_{m,k} = \frac{\left|\mathbf{g}_{m,k}^H \mathbf{w}_m\right|^2}{\sum_{i=1, i\neq m}^{M} \left|\mathbf{g}_{m,k}^H \mathbf{w}_i\right|^2 + \sigma_{m,k}^2} \forall k. \tag{7b}$$

The ASR of the *m*th IR is given by

$$R_m = \log_2 e \cdot \left[\log(1 + \gamma_m) - \max_{k \in \{1, \ldots, K_m\}} \log(1 + \gamma_{m,k})\right]^+ \tag{8}$$

where $[x]^+ = \max(x, 0)$. Due to the spatial filtering characteristics of BF technology, the achievable SR is always positive, and the superscript "+" can thus be removed. Now, we aim at designing the transmit BF weight vectors to maximize the minimal ASR, subject to the harvested energy constraint, the allowable interference, and the total transmit power budget, yielding the constrained optimization problem as

$$\max_{\{\mathbf{w}_m\}} \min_{m \in \{1, \ldots, M\}} R_m \tag{9a}$$

$$\text{s.t. } \xi\left(\sum_{i=1}^{M} \left|\mathbf{g}_{m,k}^H \mathbf{w}_i\right|^2 + \sigma_{m,k}^2\right) \geq \Gamma_{m,k} \forall m, k \tag{9b}$$

$$G_p(\phi) \sum_{m=1}^{M} \left|\mathbf{h}_p^H \mathbf{w}_m\right|^2 / \sigma_p^2 \leq \Gamma_p \tag{9c}$$

$$\sum_{m=1}^{M} \|\mathbf{w}_m\|_F^2 \leq P_{\text{tot}} \tag{9d}$$

where $\xi \in (0, 1]$ denotes the EH efficiency, $\Gamma_{m,k}$ denotes the prescribed EH threshold at $\text{ER}_{m,k}$, Γ_p denotes the allowable interference level of ES, P_{tot} denotes the power budget of BS.

Remark: We have formulated a constrained optimization problem to maximize the minimal ASR for WIPT system, where the satellite network shares the mmWave spectrum with a terrestrial network serving multiple IRs, each surrounded by many ERs. It is worth-noting that Liu *et al.* and Li *et al.* [14], [19] have focused only on a single IR situation without considering the interference at ES. Thus, we extend the previous works to a more complex and practical case in this paper.

III. PROPOSED BEAMFORMING SCHEME

In this section, we propose an SCA-based BF scheme to solve the complex problem (9). First of all, since the objective function of (9) is nonconvex and mathematically intractable, we introduce auxiliary variables t and $\{a_m, b_m\}$ along with (7) and (8) to rewrite it as

$$\max_{t,\{\mathbf{w}_m, a_m, b_m\}} t \tag{10a}$$

$$\text{s.t.} \log(1+a_m) - \log(1+b_m) \geq t \forall m \tag{10b}$$

$$\gamma_m \geq a_m \forall m \tag{10c}$$

$$\gamma_{m,k} \leq b_m \forall m, k \tag{10d}$$

$$(9b) - (9d). \tag{10e}$$

In order to solve (10) efficiently, we next adopt the SCA method to convert the nonconvex constraint (10b)–(10e) into LMI and SOC constraints. To this end, we introduce auxiliary variables $\{c_m, d_m, u_m, x_m\}$ into constraints (10c), leading to

$$\left|\mathbf{h}_m^H \mathbf{w}_m\right|^2 \geq x_m, \left|\mathbf{h}_m^H \mathbf{w}_i\right| \leq u_i, \forall m\, i \neq m \tag{11a}$$

$$x_m \geq c_m^2, \sum_{i=1, i\neq m}^{M} u_i^2 + \sigma_m^2 \leq d_m \forall m \tag{11b}$$

$$c_m^2/d_m \geq a_m \forall m. \tag{11c}$$

To deal with the nonconvex constraint (11a), by denoting $w_m^{(n)}$ as the value of w_m at the *n*th iteration and employing the first-order Taylor series expansion around $w_m^{(n)}$, the left-hand side of the first inequality in (11a) can be replaced by a lower bound as

$$\left|\mathbf{h}_m^H \mathbf{w}_m\right|^2 \geq \mathbf{h}_m^H \mathbf{W}_m \mathbf{h}_m \tag{12}$$

where $\mathbf{W}_m = \mathbf{w}_m^{(n)} \mathbf{w}_m^H + \mathbf{w}_m \left(\mathbf{w}_m^{(n)}\right)^H - \mathbf{w}_m^{(n)} \left(\mathbf{w}_m^{(n)}\right)^H$. Thus, (11a) can be transformed to a linear constraint as

$$\mathbf{h}_m^H \mathbf{W}_m \mathbf{h}_m \geq x_m, \left|\mathbf{h}_m^H \mathbf{w}_i\right| \leq u_i \forall m, i \neq m. \tag{13}$$

Then, (11b) can be directly transformed to an SOC form as

$$\begin{aligned} \frac{x_m+1}{2} &\geq \left\| \left[\frac{x_m-1}{2}, c_m\right]^T \right\|_2 \quad \forall m \\ \frac{d_m+1}{2} &\geq \left\| \left[\frac{d_m-1}{2}, \{u_i\}, \sigma_m\right]^T \right\|_2 \quad \forall m, i \neq m. \end{aligned} \tag{14}$$

Also, by employing the first-order Taylor series expansion around $c_m^{(n)}$ and $d_m^{(n)}$, (11c) can be reformulated as

$$2\left(c_m^{(n)}/d_m^{(n)}\right) c_m - \left(c_m^{(n)}/d_m^{(n)}\right)^2 d_m \geq a_m. \tag{15}$$

As for the nonconvex constraint (10d), by introducing variables $\{f_{m,k}, v_{m,k}, y_{m,k}\}$, it can be rewritten as

$$\left|\mathbf{g}_{m,k}^H \mathbf{w}_m\right| \leq y_{m,k}, \left|\mathbf{g}_{m,k}^H \mathbf{w}_i\right|^2 \geq v_{m,i} \forall m, k, i \neq m \tag{16a}$$

$$\left\| \left[\frac{b_m - f_{m,k}}{2}, y_{m,k}\right]^T \right\|_2 \leq \frac{b_m + f_{m,k}}{2} \forall m, k \tag{16b}$$

$$f_{m,k} \leq \sum_{i=1, i\neq m}^{M} v_{m,i} + \sigma_{m,k}^2 \forall m, k. \tag{16c}$$

With a manner similar to obtaining (13), constraint (9b) and (16a) can be approximated as

$$\xi \left(\sum_{i=1, i\neq m}^{M} \mathbf{g}_{m,k}^H \mathbf{W}_i \mathbf{g}_{m,k} + \sigma_{m,k}^2 \right) \geq \Gamma_{m,k} \forall m, k \tag{17a}$$

$$\left|\mathbf{g}_{m,k}^H \mathbf{w}_m\right| \leq y_{m,k}, \mathbf{g}_{m,k}^H \mathbf{W}_i \mathbf{g}_{m,k} \geq v_{m,i} \forall m, k, i \neq m. \tag{17b}$$

Besides, by introducing variables $\{p_m\}$, the constraint (9c) can be converted into SOC form as

$$\left|\mathbf{h}_p^H \mathbf{w}_m\right| \leq p_m \quad \forall m \tag{18a}$$

$$\left\| [p_1, \cdots, p_M]^T \right\|_2 \leq \sqrt{\Gamma_p \sigma_p^2 / G_p(\phi)}. \tag{18b}$$

Now, we turn our attention to constraint (10b). By introducing auxiliary variables $\{q_m\}$ and using Taylor series expansion, the constraint (10b) is recast to

$$1 + a_m \geq e^{q_m} \quad \forall m \tag{19a}$$

$$q_m - \log\left(1 + b_m^{(n)}\right) - \frac{b_m - b_m^{(n)}}{\left(1 + b_m^{(n)}\right)} \geq t \quad \forall m. \tag{19b}$$

To further reduce the computational complexity caused by the generalized nonlinear convex program of (19a), we approximate it by a series of SOC forms as [23]

$$\begin{aligned} &1 + z_{m,1} \geq \left\| \left[1 - z_{m,1}, 2 + q_m/2^{L-1}\right]^T \right\|_2 \\ &1 + z_{m,2} \geq \left\| \left[1 - z_{m,2}, 5/3 + q_m/2^{L}\right] \right\|_2 \\ &1 + z_{m,3} \geq \left\| [1 - z_{m,3}, 2z_{m,1}] \right\|_2 \\ &z_{m,4} \geq 19/72 + z_{m,2} + z_{m,3}/24 \\ &1 + z_{m,l} \geq \left\| [1 - z_{m,l}, 2z_{m,l-1}] \right\|_2, l = 5, \ldots, L+3 \\ &1 + z_{m,L+4} \geq \left\| [1 - z_{m,L+4}, 2z_{m,L+3}] \right\|_2 \\ &1 + a_m \geq z_{m,L+4} \end{aligned} \tag{20}$$

where $\mathbf{z}_m = [z_{m,1}, \ldots, z_{m,L+4}]^T \forall m$, are the introduced variables, and the accuracy of (20) would increase as *L* increases.

Finally, the (*n* + 1)th approximation of problem (10) can be expressed as

$$\begin{aligned} &\max_{\substack{t, \{\mathbf{w}_m, a_m, b_m, c_m, d_m, f_{m,k}, \\ u_m, v_{m,k}, p_m, q_m, x_m, y_{m,k}, \mathbf{z}_m\}}} \quad t \\ &\text{s.t.} \quad (9\text{d}), (13)\text{–}(15), (16\text{b}), (16\text{c}), \\ &(17), (18), (19\text{b}), (20) \end{aligned} \tag{21}$$

which can be efficiently solved by a software package, such as CVX. The feasibility and convergence as the iteration number *n* increases have been proved in [24]. As a result, the proposed BF scheme is summarized in Algorithm 1.

IV. NUMERICAL RESULTS

In this section, we provide simulation results to illustrate the superiority of the proposed BF scheme (PBF) by comparing with the SDR BF scheme [25], partial zero-forcing (PZF) [26]

TABLE I
MAIN SIMULATION PARAMETERS

Parameter	**Value**
Carrier frequency	$f = 18$ GHz
Number of BS antennas	$N = 4 \times 4$
Number of users and Eves	$M = 2, M_K = 2$
Maximal gain of the ES antenna	$G_{\max} = 38$ dB
Noise bandwidth	$B_i = 5$ MHz
Noise temperature	$T_i = 300$ K

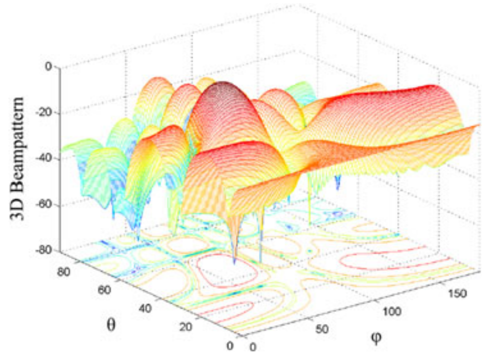

Fig. 1. 3-D beampattern of $\mathbf{w}_1$.

Algorithm 1: The Proposed Secure BF Scheme.

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Input: $\{\mathbf{h}_m, \mathbf{h}_p, \mathbf{g}_{m,k}, \Gamma_{m,k}, \Gamma_p, P_{\text{tot}}\}$.
Set the tolerance of accuracy $\varepsilon$ and the iteration number $n = 0$;
Randomly generate $\mathbf{w}_m^{(0)}, b_m^{(0)}, c_m^{(0)}, d_m^{(0)}$;
repeat
    $n := n + 1$;
    Solve the problem (21) through SCA approach;
    Set $\mathbf{w}_m^{(n)} := \mathbf{w}_m, b_m^{(n)} = b_m, c_m^{(n)} = c_m, d_m^{(n)} = d_m$;
    $\eta := \left\|\mathbf{w}_m^{(n)} - \mathbf{w}_m^{(n-1)}\right\|_F + \left|b_m^{(n)} - b_m^{(n-1)}\right| + \left|c_m^{(n)} - c_m^{(n-1)}\right| + \left|d_m^{(n-1)} - d_m^{(n-1)}\right|$;
until $\eta \leq \varepsilon$ or $n = N_{\max}$ with $N_{\max}$ being the maximal number of iterations
Output: Optimal beamforming vectors $\{\mathbf{w}_m\}$.
```

that nulls the signal to surrounding ERs, and the maximum ratio transmission (MRT) toward the intended user. We set the parameter $L = 4$, the energy conversion efficiency as $\xi = 0.6$, the EH threshold as $\Gamma_{m,k} = -15$ dBmW [14], the interference threshold of ES as $\Gamma_p = -10$ dB [19], the distance between BS and ERs, and IRs as 20 and 40 m, respectively. Other parameters are listed in Table I.

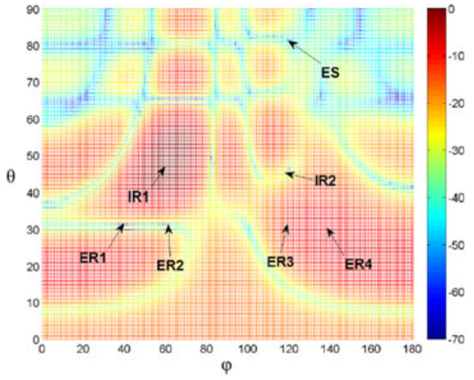

Fig. 2. Beampattern of $\mathbf{w}_1$ from vertical version.

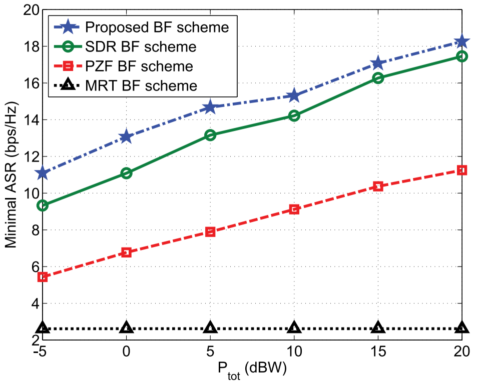

Fig. 3. Minimal ASR versus transmit power budget P_{tot}.

By letting the transmit power as $P_{\text{tot}} = 10$ dBW, Figs. 1 and 2 show the normalized 3-D beampattern and corresponding vertical view of BF weight vector $\mathbf{w}_1$, respectively. It is observed that the maximal direction of beampattern points to IR_1, while two nulls each with -50 dB toward the direction to two ERs surrounding IR_1 and two nulls each with -40 dB toward the direction to IR_2 and ES are generated. Furthermore, it is found that -10 dB sidelobes are generated, which point to two ERs surrounding IR_2, meaning that the proposed BF scheme can efficiently achieve EH and suppress the interception process of IR_2.

Fig. 3 depicts the minimal ASR versus the transmit power budget of the BS for different BF schemes. As we see, the minimal ASR of PBF, SDR, and PZF would increase as the power budget P_{tot} increases, meanwhile, the performance of MRT scheme remains a certain value due to the fixed MRT weight vectors. Besides, the minimal ASR of the proposed BF scheme is enhanced by about 1.8, 6.7, and 11.2 bps/Hz in comparison with SDR, PZF, and MRT schemes when the transmit power is 5 dBW, indicating the superiority of the proposed BF scheme. The reason is that the scheme can efficiently suppress the interference between IRs and the private signal leakage to ERs, while meeting the EH requirements at ERs.

V. CONCLUSION

In this letter, a secure BF scheme for WIPT in a terrestrial network coexisting with satellite network has been developed to maximize the minimal ASR subject to the EH requirement, interference at ES, and the total power of the BS. Specifically, we have adopted the SCA method to transform the original nonconvex problem into a problem easier to be solved with LMI and SOC constraints, and then obtained the BF weight vectors through an iterative algorithm. Simulation results have confirmed the effectiveness and superiority of the proposed BF scheme.

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