ZF-Based Beamforming for Wireless Powered Cognitive Satellite-Terrestrial Networks

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Abstract—**In this paper, we propose a novel zero-forcing (ZF)-based beamforming (BF) scheme for a wireless powered cognitive satellite-terrestrial network (CSTN) operated in the millimeter wave band. Assuming that the satellite and base station are equipped with multiple antennas, we aim at maximizing the sum rate of the CSTN while satisfying the signal-to-interference-plus-noise-ratio requirements for both the information receivers (IRs) and earth stations, the energy harvesting requirements of the energy receivers (ERs), and the secrecy constraints at the ERs. Since the resulting optimization problem is mathematically intractable, we propose a novel multi-beam-based ZF BF scheme to generate beamforming vectors to serve the IRs and ERs. Specifically, the original nonconvex problem is decomposed into two independent subproblems. The first subproblem, which features beam orthogonality constraints, leads to closed form solutions for the beamforming vectors. The second subproblem, aiming at finding the optimal power allocation, is solved via the S-procedure. Finally, the effectiveness of the proposed scheme is demonstrated by simulation results.**

I. Introduction

With the increasing demand for mobile broadband services, the current cellular network infrastructure is faced with huge data traffic surges, and may not be able to support such a large number of devices and huge traffic demands anticipated in the coming years [1]. One of the reasons is the saturation of the spectrum bands currently allocated for wireless communications, i.e. within the range from 750 MHz to 6 GHz. Hence, the key to the evolution of future wireless networks lies in exploiting less used higher frequency spectrum within the mmWave band, which has been shown to be feasible through recent theoretical studies and experimental results [2].

One of the candidate bands for the deployment of mmWave cellular mobile networks is the portion of the spectrum ranging from 17 to 30 GHz, which has been allocated on a co-primary basis to fixed services (FS), fixed satellite services (FSS) and cellular network backhaul [3]-[6]. To efficiently utilize spectrum resources and address the scarcity issue, cognitive radio techniques have recently been applied to satellite-terrestrial networks, an architecture that is referred to as a cognitive satellite-terrestrial network (CSTN) [7], [8]. In this framework, the satellite (SAT), referred to as the primary network (PN), shares the radio frequency band with the terrestrial, referred to as the secondary network (SN), through dynamic spectrum access and beamforming (BF) technology [9], thereby significantly enhancing the spectrum utilization [10], [11].

Recently, the simultaneous wireless information and power transfer (SWIPT) techniques, which are capable of extracting electrical energy from information bearing RF signals, has emerged as a promising energy harvesting solution in power-constrained wireless networks [12]-[15]. Under the SWIPT framework, a base station (BS) can serve a set of distributed users, among which some intend to decode information from the received signal, referred to as information receivers (IRs), while others are interested in energy harvesting (EH), called energy receivers (ERs). It is well-known that security is a critical issue in SWIPT systems since ERs, which in general have better channel quality than IRs to guarantee efficient energy harvesting, can intercept more easily the confidential message sent to IRs [12].

The authors in [13] have investigated secure BF design through artificial noise in a SWIPT heterogeneous cellular network. The authors in [14] studied a multi-objective optimization problem for a secure SWIPT system with two conflicting objectives: total transmit power minimization and EH efficiency maximization; they then adopted a semi-define relaxation (SDR) approach to obtain an approximate solution. In [15], the authors investigated SWIPT for a secure multiuser multi-antenna scenario, and proposed a successive convex approximation (SCA) to obtain the optimal BF weight vectors. It is noted that the aforementioned works can indeed enhance security in SWIPT networks, but at the price of a heavy computational burden. Besides, the research on secure communication for wireless powered CSTN is still in its infancy. These considerations motivate our current work.

In this paper, we assume a scenario with a satellite that employs a steerable multi-beam antenna array and a BS that uses a uniform planar array (UPA) to obtain high gain with compact size, we investigate the sum rate maximization (SRM) problem for beamformer design in wireless powered CSTNs in the mmWave band while satisfying essential constraints, i.e.: signal-to-interference-plus-noise-ratio (SINR) requirements of the IRs and earth stations (ESs), EH requirements, and the secrecy constraints at the ERs. Unlike the work in [15], where the computationally complex SCA method is exploited to solve

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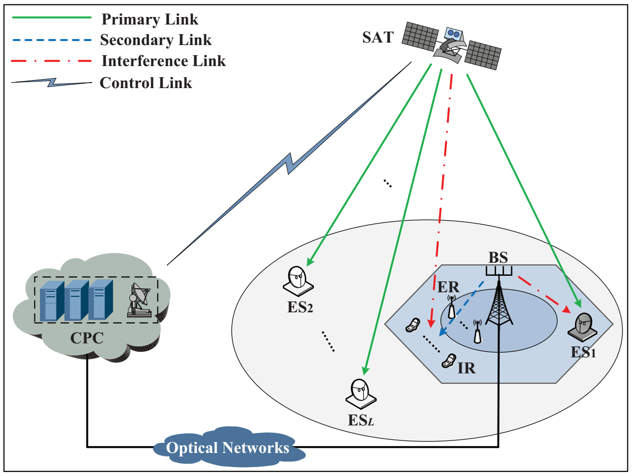

Fig. 1: System model of wireless powered CSTN.

the SRM problem, we propose a novel multi-beam-based zero-forcing (MB-ZF) scheme with low computational complexity. Specifically, the original nonconvex optimization problem is decomposed into two independent subproblems. The first subproblem, which features beam orthogonality constraints, leads to closed form solutions for the beamforming vectors at the SAT and BS. The second subproblem, aiming at finding the optimal power allocation, is solved via the S-procedure, which involves the application of second order cone programming (SOCP) and linear matrix inequality (LMI) optimizations. The effectiveness of the proposed scheme is demonstrated by simulations using realistic channel models for the satellite and terrestrial links.

II. System Model and Problem Formulation

As shown in Fig. 1, we consider the downlink of a CSTN in which the available network resources are allocated dynamically at a processing center through cloud computing to achieve optimal performance. Specifically, the geostationary orbit (GEO) satellite serves L ESs simultaneously through multicast communication. Meanwhile, the cellular network base station is located within the coverage area of the satellite transmissions. In this work, for simplicity, we only consider a single cell, but extension or our approach to multiple cells is possible. The BS is equipped with a large-scale (or massive) UPA to combat the large path loss and shadowing from radio propagation in the mmWave band. In addition, there are K IRs and M ERs ($K \geq M$) distributed in the cellular network, and the ERs are closer to the BS than the IRs for the purpose of efficient energy harvesting. Based on the locations of the cellular users, we assume that the ER is a potential eavesdropper to the closest IR, thus the IRs and ERs can be clustered into K pairs. For simplicity, all the CSTN users are equipped with single antenna, but generalizations are possible.

A. Channel Model

For satellite systems operating in high frequency bands, the propagation channel is impaired by various effects such as the

TABLE I: Description of the Parameters

Parameter	Definition
L / K / M	number of ESs / IRs and ERs
N_s / N_b	antenna number of SAT / BS
N_1 / N_2	number of array elements placed along the X / Z-axis
d_1 / d_2	inter-element spacing placed along the X / Z-axis
$\mathbf{h}_{ir,k}$ / $\mathbf{f}_{ir,k}$	channel vector between BS/SAT and the k-th IR
$\mathbf{h}_{er,m}$ / $\mathbf{f}_{er,m}$	channel vector between BS/SAT and the m-th ER
$\mathbf{h}_{es,l}$ / $\mathbf{f}_{es,l}$	channel vector between BS/SAT and the l-th ES
$\mathbf{w}_k$ / $\mathbf{v}$	BF weight vector toward the k-th IR / ES

path loss, rain attenuation and satellite antenna gain, which should be considered to properly model the satellite channel. According to the free space equation, the path loss can be expressed as

$$C_L \approx \left(\frac{\lambda}{4\pi}\right)^2 \frac{1}{d_h^2 + d_0^2} \tag{1}$$

where λ denotes the carrier wavelength, d_h the height of the GEO satellite and d_0 the distance from the beam center to the user.

Besides path loss, the radiowave propagation at high frequencies is dominantly affected by rain attenuation. The rain attenuation in dB, $\xi_{\text{dB}} = 20\log_{10}(\xi)$, commonly follows a lognormal distribution [16], namely, $\xi_{\text{dB}} \sim \mathcal{CN}\left(\mu, \sigma^2\right)$, where μ and σ, both in dB, represent the location and scale parameters, respectively. Thus, the rain attenuation fading vector $\mathbf{r} \in \mathbb{C}^{N_s \times 1}$ between the satellite and user can be written as

$$\mathbf{r} = \xi^{-\frac{1}{2}} e^{-j\boldsymbol{\alpha}} \tag{2}$$

where $\boldsymbol{\alpha} \in \mathbb{C}^{N_s \times 1}$ represents a phase vector whose components are uniformly distributed over $[0\ , 2\pi)$.

The multibeam antenna gain is determined by the satellite antenna pattern and the user position. Define ϕ_m as the angle between the m-th ES user and the m-th beam boresight with respect to the satellite, $\phi_m^{3\text{dB}}$ as the 3dB angle of the m-th beam, and $\mathbf{b} = [b_1, b_2, \cdots, b_{N_s}]^T$ the $N_s \times 1$ beam gain vector. Then the antenna gain from the m-th on-board beam to the user can be expressed as [17]

$$b_m = b_{\max}\left(\frac{J_1(u_m)}{2u_m} + 36\frac{J_3(u_m)}{u_m^3}\right)^2 \tag{3}$$

where $b_{\max}$ stands for the maximal satellite antenna gain, $u_m = 2.07123 \sin\phi_m / \sin\phi_m^{3\text{dB}}$, and $J_i(\cdot)$ represents the first-kind Bessel functions of order i.

Combining the above three factors, the satellite channel gain can be expressed as

$$\mathbf{f} = \sqrt{C_L G_r}\mathbf{r} \odot \mathbf{b}^{\frac{1}{2}} \tag{4}$$

where G_r denotes the ES off-boresight antenna gain [18].

For the terrestrial channel model, the BS employs a UPA of dimension $N_b = N_1 \times N_2$, with N_1 and N_2 representing the number of elements uniformly placed along the x-axis and the z-axis with inter-element spacing d_1 and d_2, respectively.

$\theta \in [0, \pi/2)$ denotes the elevation angle of arrival (AoA) and $\varphi \in [0, \pi)$ the azimuth AoA. The three-dimentional sparse channel model for a generic user can thus be expressed as [11]

$$\mathbf{h} = \sqrt{g(\theta_0, \varphi_0)}\rho_0 \mathbf{a}_h(\theta_0, \varphi_0) \otimes \mathbf{a}_v(\theta_0) + \sqrt{\frac{1}{N}} \sum_{n=1}^{N} \sqrt{g(\theta_n, \varphi_n)}\rho_n \mathbf{a}_h(\theta_n, \varphi_n) \otimes \mathbf{a}_v(\theta_n) \tag{5}$$

where $g(\theta, \varphi)$ represents the array element gain pattern, ρ_n the complex path gain, $\mathbf{a}_h(\theta, \varphi)$ and $\mathbf{a}_v(\theta)$ the azimuth and elevation array steering vectors (SVs), which are, respectively, given by

$$\mathbf{a}_h(\theta, \varphi) = \left[e^{-j\beta((N_1+1)/2)d_1 \sin\theta \cos\varphi}, \cdots, e^{+j\beta((N_1+1)/2)d_1 \sin\theta \cos\varphi}\right]^T, \tag{6a}$$

$$\mathbf{a}_v(\theta) = \left[e^{-j\beta((N_2+1)/2)d_2 \cos\theta}, \cdots, e^{+j\beta((N_2+1)/2)d_2 \cos\theta}\right]^T. \tag{6b}$$

B. Problem Formulation

Let $s(t)$ be the multicast signal sent by the satellite to the ESs, satisfying $E[|s(t)|^2] = 1$, and $x_k(t)$ the signal sent by the BS to the k-th IR with $E[|x_k(t)|^2] = 1$. Before transmission, the signals $s(t)$ and $x_k(t)$ are mapped with the beamformers $\mathbf{v} \in \mathbb{C}^{N_s \times 1}$ and $\mathbf{w}_k \in \mathbb{C}^{N_b \times 1}$, respectively. Then, the received signals at the l-th ES, the k-th IR and its potential eavesdropper can be expressed as

$$\begin{aligned} y_{es,l}(t) &= \mathbf{f}_{es,l}^H \mathbf{v} s(t) + \sum_{m=1}^{M} \mathbf{h}_{es,l}^H \mathbf{w}_k x_k(t) + n_{es,l}(t), \\ y_{ir,k}(t) &= \mathbf{h}_{ir,k}^H \mathbf{w}_k x_k(t) + \sum_{i=1, i\neq k}^{K} \mathbf{h}_{ir,k}^H \mathbf{w}_i x_i(t) + n_{ir,k}(t), \\ y_{er,k}(t) &= \mathbf{h}_{er,k}^H \mathbf{w}_k x_k(t) + \sum_{i=1, i\neq k}^{K} \mathbf{h}_{er,k}^H \mathbf{w}_i x_i(t) + n_{er,k}(t) \end{aligned} \tag{7}$$

where $n_{j,i}, i \in \{l, k\}, \forall l, k$ represents the additive Gaussian white noise (AWGN) with zero mean and variance $\sigma^2 = \kappa T B$, where $\kappa \approx 1.38 \times 10^{-23} J/K$ denotes Boltzmann constant, T the noise temperature, B the noise bandwidth, and the other parameters are described in Table I. It is noted that, since $K \geq M$, certain IR may not have a potential eavesdropper, but this does no affect subsequent developments. Then, the corresponding output SINRs for the received signal in (7) can be derived as

$$\begin{aligned} \gamma_{es,l}(t) &= \frac{\left|\mathbf{f}_{es,l}^H \mathbf{v}\right|^2}{\sum_{k=1}^{K} \left|\mathbf{h}_{es,l}^H \mathbf{w}_k\right|^2 + \sigma_{es,l}^2}, \\ \gamma_{ir,k}(t) &= \frac{\left|\mathbf{h}_{ir,k}^H \mathbf{w}_k\right|^2}{\sum_{i=1, i\neq k}^{K} \left|\mathbf{h}_{ir,k}^H \mathbf{w}_i\right|^2 + |\mathbf{f}_{ir,k}\mathbf{v}|^2 + \sigma_{ir,k}^2}, \\ \gamma_{er,k}(t) &= \frac{\left|\mathbf{h}_{er,k}^H \mathbf{w}_k\right|^2}{\sum_{i=1, i\neq k}^{K} \left|\mathbf{h}_{er,k}^H \mathbf{w}_i\right|^2 + |\mathbf{f}_{er,k}\mathbf{v}|^2 + \sigma_{er,k}^2}. \end{aligned} \tag{8}$$

Thus, the sum rate of the CSTN can be written as

$$R = \sum_{l=1}^{L} \log_2(1 + \gamma_{es,l}) + \sum_{k=1}^{K} \log_2(1 + \gamma_{ir,k}). \tag{9}$$

Each ER can harvest energy from the beams transmitted by the SAT and the BS. Let $0 < \xi < 1$ be the energy harvesting efficiency per a unit time. Then, the energy harvested by the m-th ER can be formulated as

$$E_m = \xi \left(\sum_{k=1}^{K} \left|\mathbf{h}_{er,m}^H \mathbf{w}_k\right|^2 + \left|\mathbf{f}_{er,m}^H \mathbf{v}\right|^2 + \sigma_{er,m}^2 \right). \tag{10}$$

The problem of interest is to maximize the sum rate of the CSTN by jointly designing the on-board and terrestrial BF weight vectors, subject to the SINR requirements at each IR and ES, the EH requirements and secrecy constraints at ERs, and the transmit power constraints at BS and SAT. Mathematically, the problem can be formulated as

$$\max_{\mathbf{w}_k, \mathbf{v}} \sum_{l=1}^{L} \log_2(1 + \gamma_{es,l}) + \sum_{k=1}^{K} \log_2(1 + \gamma_{ir,k}) \tag{11a}$$

$$\text{s.t. } \gamma_{es,l} \geq \Gamma_E, \quad \forall l, \tag{11b}$$

$$\gamma_{ir,k} \geq \Gamma_I, \quad \forall k, \tag{11c}$$

$$\gamma_{er,k} \leq \Gamma_S, \quad \forall k \tag{11d}$$

$$E_m \geq \Gamma_{th}, \quad \forall m, \tag{11e}$$

$$\|\mathbf{v}\|_F^2 \leq P_S, \sum_{k=1}^{K} \|\mathbf{w}_k\|_F^2 \leq P_B \tag{11f}$$

where Γ_E and Γ_I denote the SINR requirements for ES and IR, respectively, Γ_S and Γ_{th} the secrecy constraint and the EH requirement at ER, and P_S and P_B the transmit power budgets at SAT and BS, respectively. It is assumed that the channels are quasi-static and the channel state information is available at the processing center [17], [19].

We note that the constrained optimization problem (11) is intractable due to the non-convex objective function. To tackle this difficulty, in what follows, we will present a novel ZF-based BF scheme to efficiently obtain sub-optimal solutions.

III. A Novel ZF-based BF Solution

The sum rate maximization problem in (11) has been investigated in the open literature using the successive convex approximation approach [8], [15], which however leads to heavy computational burden. To more efficiently approach the problem (11) with satisfying performance, we present a MB-ZF scheme to obtain the near-optimal BF weight vectors. To this end, we exploit the fact that due to the large path loss and small antenna gain, the satellite signal received by the IRs and ERs can be ignored relative to the noise:

$$\left|\mathbf{f}_{ir,k}^H \mathbf{v}\right|^2 \ll \sigma_{ir,k}^2, \quad \left|\mathbf{f}_{er,k}^H \mathbf{v}\right|^2 \ll \sigma_{er,k}^2, \quad \forall k.$$

This assumption has already been utilized in previous work such as [11]. Meanwhile, in order to further simplify the SINR expression in (11) so that the original optimization problem is mathematically tractable, we impose ZF constraints on the unintended receivers to eliminate the mutual interference between IRs and ESs, and ensure that no private signals are wiretapped by a potential eavesdropper.

When the conventional ZF (CZF) BF scheme is adopted in (11), the BF weight vectors of the SAT and the BS should satisfy the following two optimization problems,

$$\begin{aligned} &\max_{\bar{\mathbf{v}}_l} \left|\mathbf{f}_{es,l}^H \bar{\mathbf{v}}_l\right|^2 \\ &\text{s.t.}\;\; \bar{\mathbf{v}}_l^H \bar{\mathbf{v}}_l = 1 \end{aligned} \tag{12}$$

and

$$\begin{aligned} &\max_{\bar{\mathbf{w}}_k} \left|\mathbf{h}_{ir,k}^H \bar{\mathbf{w}}_k\right|^2 \\ &\text{s.t.}\;\; \mathbf{h}_{ir,i}^H \bar{\mathbf{w}}_k = 0, \quad \forall i \neq k, \\ &\qquad \mathbf{h}_{er,k}^H \bar{\mathbf{w}}_k = 0, \\ &\qquad \mathbf{h}_{es,l}^H \bar{\mathbf{w}}_k = 0, \\ &\qquad \bar{\mathbf{w}}_k^H \bar{\mathbf{w}}_k = 1 \end{aligned} \tag{13}$$

where $\mathbf{v} = \sqrt{P_S} \sum_{l=1}^{L} \bar{\mathbf{v}}_l / \left\| \sum_{l=1}^{L} \bar{\mathbf{v}}_l \right\|$ and $\bar{\mathbf{w}}_k = \mathbf{w}_k / \sqrt{p_k}$ with p_k representing the power coefficient of $\mathbf{w}_k$. The solutions to (12) and (13) are given by

$$\bar{\mathbf{v}}_l = \mathbf{f}_{es,l} / \left\| \mathbf{f}_{es,l} \right\|, \tag{14a}$$

$$\bar{\mathbf{w}}_k = \frac{\left(\mathbf{I}_{N_b} - \mathbf{P}_k\right)\mathbf{h}_k}{\left\|\left(\mathbf{I}_{N_b} - \mathbf{P}_k\right)\mathbf{h}_k\right\|}. \tag{14b}$$

where $\mathbf{P}_k = \hat{\mathbf{H}}_k \left(\hat{\mathbf{H}}_k^H \hat{\mathbf{H}}_k\right)^{-1} \hat{\mathbf{H}}_k^H$ with $\hat{\mathbf{H}}_k = \left[\{\mathbf{h}_{ir,i}\}_{i\neq k}, \{\mathbf{h}_{er,k}\}, \mathbf{h}_{es,l}\right]$ $\forall k, l$.

The main drawback of the CZF-BF algorithm is that only one directional beam pattern pointing to the intended IR is generated, and thus the received signal power at the ERs for energy harvesting can not be guaranteed, possibly rendering the original problem (11) infeasible. In this context, we propose a novel multi-beam-based ZF BF scheme to generate two beamforming vectors based on further orthogonality constraints, to serve ER and IR from different pair. It can be written as

$$\begin{aligned} \mathbf{w}_k &= \sqrt{p_k}\,\bar{\mathbf{w}}_k \\ &= \sqrt{p_k}\left(\sqrt{\lambda_{k,1}}\mathbf{w}_{k,1} + \sqrt{\lambda_{k,2}}\mathbf{w}_{k,2}\right), \end{aligned} \tag{15}$$

where $\lambda_{k,1}$ and $\lambda_{k,2}$, satisfying $\lambda_{k,1} + \lambda_{k,2} = 1$, are normalizing coefficients used to guarantee $\|\bar{\mathbf{w}}_k\|_F = 1$. The beamformer $\mathbf{w}_{k,1}$ represents the beam directed towards the intended IR, which is obtained from (13) and (14b) through replacing $\bar{\mathbf{w}}_k$ with $\mathbf{w}_{k,1}$, while the beamformer $\mathbf{w}_{k,2}$ represents the beam towards the i-th ER $i = k+1$ modulo K, which is determined by

$$\begin{aligned} &\max_{\mathbf{w}_{k,2}} \left|\mathbf{h}_{er,i}^H \mathbf{w}_{k,2}\right|^2 \\ &\text{s.t.}\;\; \mathbf{h}_{ir,k}^H \mathbf{w}_{k,2} = 0, \\ &\qquad \mathbf{h}_{er,k}^H \mathbf{w}_{k,2} = 0, \\ &\qquad \mathbf{h}_{es,l}^H \mathbf{w}_{k,2} = 0, \qquad \forall l, \\ &\qquad \mathbf{w}_{k,2}^H \mathbf{w}_{k,2} = 1. \end{aligned} \tag{16}$$

The proposed MB-ZF beamforming is illustrated in Fig. 2 for the case $M = K = 3$.

Similar to (14b), $\mathbf{w}_{k,2}$ can be obtained as

$$\mathbf{w}_{k,2} = \frac{\left(\mathbf{I}_{N_b} - \mathbf{P}_i\right)\mathbf{h}_{i,e}}{\left\|\left(\mathbf{I}_{N_b} - \mathbf{P}_i\right)\mathbf{h}_{i,e}\right\|}, \tag{17}$$

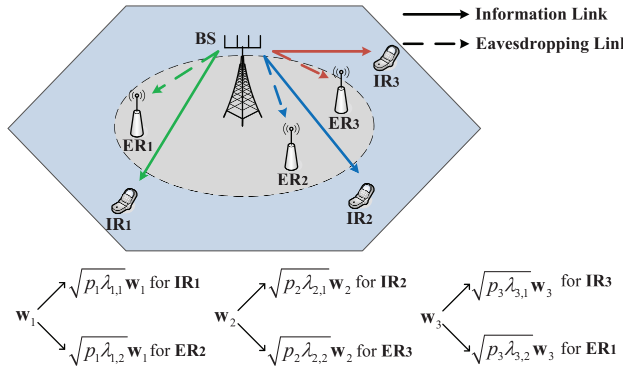

Fig. 2: Illustration of MB-ZF scheme for the case of $M = K = 3$.

where $\mathbf{P}_i = \hat{\mathbf{G}}_i \left(\hat{\mathbf{G}}_i^H \hat{\mathbf{G}}_i\right)^{-1} \hat{\mathbf{G}}_i^H$ with $\hat{\mathbf{G}}_i = \left[\mathbf{h}_k, \mathbf{h}_{k,e}, \mathbf{h}_l\right], \forall k, l$.

Substituting (14) and (17) into (11) and letting $\mathbf{W}_{k,1} = \mathbf{w}_{k,1}\mathbf{w}_{k,1}^H$, $\mathbf{W}_{k,2} = \mathbf{w}_{k,2}\mathbf{w}_{k,2}^H$, the original optimization problem can be transformed into the following problem

$$\max_{\{p_k, \lambda_{k,1}\}} \sum_{k=1}^{K} \log_2\left(1 + a_k p_k \lambda_{k,1}\right) \tag{18a}$$

$$\begin{aligned} \text{s.t.}\quad &\xi\left(p_k\left(1-\lambda_{k,1}\right)\mathrm{Tr}\left(\mathbf{H}_{er,i}\mathbf{W}_{k,2}\right)\right. \\ &\qquad\qquad \left.+\sigma_{er,k}^2\right) \geq \Gamma_{th},\ \forall k, \end{aligned} \tag{18b}$$

$$p_k\lambda_{k,1}\mathrm{Tr}\left(\mathbf{H}_{ir,k}\mathbf{W}_{k,1}\right) - \Gamma_I \sigma_{ir,k}^2 \geq 0,\ \forall k, \tag{18c}$$

$$\sum_{k=1}^{K} p_k \leq P_B,\ 0 < \lambda_{k,1} < 1 \tag{18d}$$

where $a_k = \frac{\mathrm{Tr}(\mathbf{H}_{ir,k}\mathbf{W}_{k,1})}{\sigma_{ir,k}^2}$. Introducing the auxiliary variables $\{x_k\}$, we can obtain

$$\max_{\{p_k, \lambda_{k,1}, x_k\}} \log_2 e \sum_{k=1}^{K} 2x_k \tag{19a}$$

$$\text{s.t.}\quad p_k\lambda_{k,1}\mathrm{Tr}\left(\mathbf{H}_{ir,k}\mathbf{W}_{k,1}\right) + \sigma_{ir,k}^2 \geq e^{2x_k},\ \forall k, \tag{19b}$$

$$(18b) - (18d).$$

To tackle the non-convex constraint (19b), we introduce the auxiliary variables $\{y_k\}$, and ignore the noise term σ_k^2, which does not change the inequality and has little influence on the optimization problem, thereby obtaining

$$b_k\left(p_k + \lambda_{k,1}\right) \geq \left\|\left[b_k\left(p_k - \lambda_{k,1}\right), y_k\right]^T\right\|_2, \tag{20a}$$

$$y_k \geq e^{x_k}, \tag{20b}$$

where $b_k = \frac{\mathrm{Tr}(\mathbf{H}_{ir,k}\mathbf{W}_{k,1})}{2}$. Similar to [8], (20b) can be transformed into a series of SOC. In addition, by using S-procedure, the constraints (18b) and (18c) can be, respectively,

TABLE II: Main Simulation Parameters

Parameter	Value
Orbit	GEO
Carrier frequency	18 GHz
Maximal beam gain	$b_{\max} = 52$ dB
3dB angle	$\phi_{3\text{dB}} = 0.4°$
BS inter-element spacing	$d_1 = d_2 = \lambda/2$
Rain fading	μ=-3.125, σ=1.591
Satellite signal bandwidth	B = 500 KHz
Noise temperature	T = 300 K

converted into the semi-definite constraints

$$\begin{bmatrix} p_k \mathrm{Tr}\left(\mathbf{H}_{er,i}\mathbf{W}_{k,2}\right) & \sqrt{\Gamma_{th} - \xi\sigma_{er,k}^2} \\ \sqrt{\Gamma_{th} - \xi\sigma_{er,k}^2} & \xi\left(1 - \lambda_{k,1}\right) \end{bmatrix} \succeq 0, \quad \text{(21a)}$$

$$\begin{bmatrix} p_k \mathrm{Tr}\left(\mathbf{H}_{ir,k}\mathbf{W}_{k,1}\right) & \sqrt{\Gamma_I}\sigma_{ir,k} \\ \sqrt{\Gamma_I}\sigma_{ir,k} & \lambda_{k,1} \end{bmatrix} \succeq 0. \quad \text{(21b)}$$

Consequently, the optimization problem (19) can be expressed as the convex problem

$$\begin{aligned} &\max_{\{p_k,\lambda_{k,1},x_k,y_k\}} \sum_{k=1}^{K} x_k \\ &\text{s.t.} \quad (20)\,, (21)\,, (18d) \end{aligned} \quad (22)$$

which can be efficiently solved to obtain the power allocation coefficients by standard convex optimization tools.

Remark 1: It can be observed from (15) that if $\lambda_{k,2} = 0$, the proposed MB-ZF scheme reduces to the CZF BF method. However, compared with the CZF BF scheme, our proposed MB-ZF approach can control and utilize the interference from other IRs to satisfy the EH requirement of the ERs without sacrificing the system secrecy performance, making the solution of the optimization problem more reliable.

IV. Numerical Results

This section presents the results of numerical simulations to characterize the performance of the proposed suboptimal BF scheme. We set $K = M = 3$, i.e. the BS serces three IRs and ERs, respectively. The energy conversion efficiency is set as $\xi = 0.6$, the EH threshold as $\Gamma_{th} = -20$ dBm [20], the SINR threshold of the IRs and ESs as $\Gamma_I = \Gamma_E = -10$ dB [12], the secrecy constraint at ERs as $\Gamma_S = -30$ dB [20], and the other parameters are listed in Table II. In the simulations, the SCA-BF scheme [15] and the maximum ratio transmission (MRT) scheme are adopted as benchmarks.

Fig. 3 shows the system sum rate comparison of the three methods versus the transmit power budget of the SAT and BS, where the transmit power budget of the SAT and BS is assumed as P, and the BS is deployed with a $N_b = 12\times 12 = 144$ UPA. As we can see, the performance of the proposed suboptimal BF algorithm is very close to that of the SCA-BF scheme, while the computational burden of the latter is about N_b^2 times than that of our proposed BF scheme. It can also be seen that the proposed BF scheme outperforms the MRT-BF scheme with 11.3 bps/Hz while $P = 10$ dBW, verifying that the intra-cell interference can be exploited to enhance the secure communication, while the interference is also utilized to meet the EH requirements at the ERs.

Fig. 4 illustrates the system sum rate versus the type of UPA employed at the BS. It can be seen that the sum rate increases with the size of the array, especially when the array increases from 4×4 to 8×8. However, further increases in the array size only achieves a small performance, and the sum rate gradually reach a constant value. When a 4×4 UPA is deployed, the proposed BF scheme performs better than the MRT-BF scheme by 15 bps/Hz. It is observed that the sum rate of MRT-BF scheme is still lower than that of our proposed scheme even with large size of the array, since the MRT scheme does not take the intra-cell interference into consideration and the multi-user interference significantly suppresses the performance.

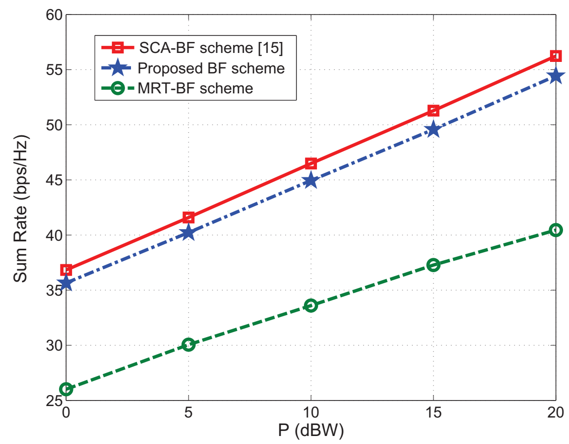

Fig. 3: Sum rate versus transmit power.

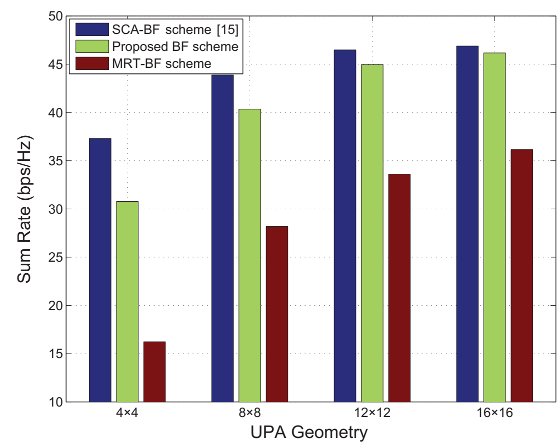

Fig. 4: Sum rate versus UPA Geometry.

V. Conclusion

In this paper, a novel ZF-based BF scheme for wireless powered CSTN was investigated. Based on a practical satellite and terrestrial model in high frequency bands, we first presented a constrained optimization problem to maximize the sum rate of a CSTN subject to the SINR requirements at the IRs and

ESs, the secrecy constraint for potential eavesdroppers, the EH requirement at the ERs and the transmit power budget at the SAT and BS. To address the resulting intractable optimization problem, we proposed a MB-ZF beamforming scheme to decompose the original nonconvex problem into two independent subproblems. The first subproblem, which features beam orthogonality constraints, leads to closed form solutions for the beamforming vectors. The second subproblem, aiming at finding the optimal power allocation, is solved via the S-procedure. Finally, our simulation results verified the effectiveness of the proposed novel BF scheme.

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