Optimization for DF Relaying Cognitive Radio Networks with Multiple Energy Access Points

Hong Xing[†], Xin Kang[§], Kai-Kit Wong[‡], and Arumugam Nallanathan[†]

[†] Department of Informatics, King's College London. E-mails: {hong.1.xing, arumugam.nallanathan}@kcl.ac.uk
[§]University of Electronic Science and Technology of China. E-mail:kangxin83@gmail.com
[‡] Department of Electronic and Electrical Engineering, University College London. E-mail: kai-kit.wong@ucl.ac.uk

***Abstract*—Cognitive radio (CR) has been advocated to improve the network spectrum efficiency for decades, and the cooperation between the primary and secondary systems has become a new paradigm to further improve the spectrum utilization. However, in practice, secondary transmitters (STs) are usually power constrained, which limits the application of cooperative cognitive radio networks (CCRN). In this paper, to tackle this, we consider a novel spectrum sharing CCRN powered by energy access points (EAPs) that can charge users wirelessly, in which a multi-antenna secondary user (SU) solely powered by its harvested energy seeks cooperation with a single-antenna primary user (PU) by serving as a deocde-and-forward (DF) relay. We investigate a payoff maximization problem from the SU's perspective, who gets paid by offering data relaying service for PU but has to pay for WEH, and obtain its optimal DF relay and WEH strategy. A greedy-based algorithm that can assign the ST to right EAPs is also proposed for the ease of implementation. The proposed scheme is shown to be effective by simulations with a negligible gap to the optimal solution.**

I. Introduction

With the rapid development of wireless services and applications, the demand for frequency resources has dramatically increased. How to accommodate these new wireless services and applications within the limited radio spectrum becomes a big challenge facing the modern society [1]. The compelling need to establish more flexible spectrum regulations motivates the advent of cognitive radio (CR) [2–4]. Cooperative cognitive radio networks (CCRN) further paves way to improve the spectrum efficiency of a CR system by advocating cooperation between the primary and secondary systems for their mutual benefits. Compared with classic CR approaches, i.e., *underlay*, *overlay*, and *interweave*, CCRN enables cooperative gains on top of CR in the sense that the secondary transmitter (ST) helps to improve the diversity of the primary transmission via relaying the primary user (PU)'s message while being allowed to access its bandwidth.

Although the conventional CCRN benefits from information-level cooperation, its implementation in real world might be compromised due to the power constraint of SUs, especially when the SUs are widespread low-power application nodes, such as energy constrained wireless sensors and small cell relays. With the advent of energy harvesting (EH) technologies, CCRN has now been envisioned to substantially improve the overall system spectrum efficiency by enabling both information-level and energy-level cooperation [5, 6]. Besides the well-known energy sources such as solar and wind, ambient radio signal has recently been regarded as a new viable source for wireless energy harvesting (WEH) [7–9]. Joint information and energy cooperation has thus been extended to new dimensions with the recent WEH advances [10–12]. In [10], the ST receives information from the primary transmitter (PT) and is also fed with energy by the PT using power splitting (PS) and/or time switching (TS) receiver, which brought more incentives for the ST to join in CCRN, and thus enlarges the achievable primary-secondary rate region.

In this paper, we consider a spectrum-sharing *decode-and-forward (DF)* relay CCRN powered by multiple multi-antenna wireless energy access points (EAPs), which improve the wireless power transfer (WPT) efficiency due to their array gain and dedicated energy beamforming (EB) design [13]. The ST is required to help with the primary transmission for the purpose of sharing the PU's spectrum with the power solely supplied by EAPs. The EAPs can be part of the infrastructure in future communication systems, e.g., road side units preferably in vehicle communications. The main contribution of this paper is summarized as follows:

- From an economic point of view, the payoff maximization problem for the SU that gains profit by assisting with the primary transmission but has to pay for the harvested energy from EAPs is investigated.
- The optimum strategy is obtained by jointly optimizing the ST's WEH strategies, multi-antenna beamforming, and PT's power allocation over two transmission phases of the DF relaying.
- A centralized greedy-based heuristic algorithm is developed to substantially reduce the integer programming (IP) induced complexity, and is also verified by simulations to well approach the optimal scheme.

Notation—We use the upper case boldface letters for matrices and lower case boldface letters for vectors. $(\cdot)^T$, $(\cdot)^H$, and $\mathsf{Tr}(\cdot)$ denote the transpose, conjugate transpose, and trace operations on matrices, respectively. $\|\cdot\|$ is the Euclidean norm of a vector. $\mathbb{E}[\cdot]$ stands for the statistical expectation of a random variable. $\boldsymbol{A} \succeq 0$ indicates that $\boldsymbol{A}$ is a positive semidefinite matrix and $\boldsymbol{I}$ denotes an identity matrix with appropriate size.

II. System Model

In this paper, we consider a WEH-enabled CCRN operating with DF relaying that consists of a primary transmitter-receiver pair, a secondary transmitter-receiver pair, and K EAPs. The primary transmitter and receiver are denoted by PT and PR, respectively. The secondary transmitter and receiver are denoted by ST and SR, respectively. PT and PR are equipped with one antenna each, while ST and SR are equipped with N and M antennas, respectively. The number of antennas at the kth EAP is denoted by N_k, $\forall k \in \mathcal{K}$, $\mathcal{K} = \{1, \cdots, K\}$.

In order to share the spectrum with the PUs, we assume that the ST is required to assist with the primary transmission via DF relaying, by which the ST receives the PT's signal in the first time slot, re-encodes it with its own message, and broadcasts the superimposed signal to the PR and the SR in the second time slot. In this paper, we assume that the ST is battery limited, and thus it resorts to WEH as its only means of power supply for the spectrum-sharing cooperative transmission. As illustrated by Fig. 1, a two-equal slot transmission protocol is assumed to be adopted.

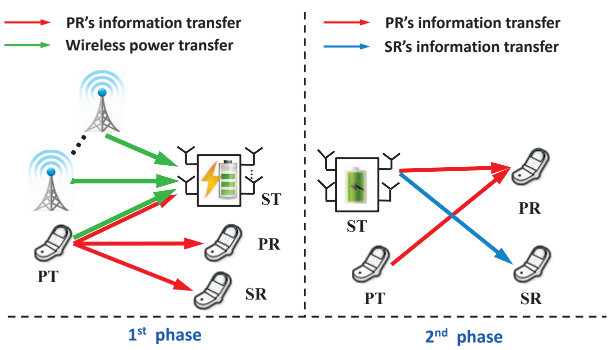

Fig. 1. Transmission protocol for a wireless powered CCRN.

A. The First Time Slot

Received signal at the PR. Let s denote PT's transmitted signal, and $\boldsymbol{x}_k s_k$ denote the kth EAP's energy signal where $\boldsymbol{x}_k \in \mathbb{C}^{N_k \times 1}$ is the beamforming vector for the kth EAP. For the convenience of exposition, we introduce an indicator function ρ_k for the kth EAP, $k \in \mathcal{K}$, which is defined as

$$\rho_k = \begin{cases} 1 & \text{the } k\text{th EAP is selected for WPT,} \\ 0 & \text{otherwise.} \end{cases} \tag{1}$$

Accordingly, the signal received at the PR can be expressed as

$$y_{PR}^{(1)} = h_{pp}\sqrt{\beta P_p} s + n_{PR}^{(1)}, \tag{2}$$

where h_{pp} denotes the complex channel from the PT to the PR; β is a power allocation factor that decides the amount of power used to transmit the primary information in the first time slot; P_p is the total power available to the PT for two time slots; and $n_{PR}^{(1)}$ denotes the circularly symmetric complex Gaussian (CSCG) additive noise at the PR, i.e., $\boldsymbol{n}_{PR}^{(1)} \sim \mathcal{CN}(0, \sigma_{PR}^2)$.

It is worth pointing out that part of the received signal from EAPs, i.e., $\sum_{k=1}^{K} \rho_k \boldsymbol{h}_{k,p}^H \boldsymbol{x}_k s_k$, where $\boldsymbol{h}_{k,p}$'s represent the complex channels from the kth EAP to the PR, can be perfectly cancelled by the PR assuming that the kth EAP transmits a constant signal s_k *a priori*, $\forall k \in \mathcal{K}$. This will consequently facilitate the decoding of the PT's signal s without causing interference.

Received signal at the SR. The received signal at the SR, with EAPs' energy signals perfectly cancelled, is given by

$$\boldsymbol{y}_{SR}^{(1)} = \boldsymbol{h}_{ps}\sqrt{\beta P_p} s + \boldsymbol{n}_{SR}^{(1)}, \tag{3}$$

where $\boldsymbol{h}_{ps}$ denotes the complex channels from the PT to the SR, and $\boldsymbol{n}_{SR}^{(1)}$ is the CSCG noise received at the SR, denoted by $\boldsymbol{n}_{SR}^{(1)} \sim \mathcal{CN}(0, \sigma_{SR}^2 \boldsymbol{I})$.

Received signal at the ST. In this paper, we assume that the ST employs a dynamic PS receiver for concurrent EH and information decoding (ID) from the same stream of received signal, where ϱ portion of the received power is used to feed the energy supply versus the remaining $1-\varrho$ portion reserved for ID. As a result, the signal received by the ST available for ID is given by

$$\boldsymbol{y}_{ST}^{(1)} = \sqrt{1-\varrho}(\boldsymbol{h}_{p,ST}\sqrt{\beta P_p} s + \boldsymbol{n}_a) + \boldsymbol{n}_c, \tag{4}$$

where $\boldsymbol{h}_{p,ST}$ denotes the complex channels from the PT to the ST; $\boldsymbol{n}_a$ denotes the antenna noise at the RF-band with each entry a CSCG variable of zero mean and a variance of $\sigma_{n_a}^2$; and $\boldsymbol{n}_c$ is the RF-band to baseband signal conversion noise, denoted by $\boldsymbol{n}_c \sim \mathcal{CN}(0, \sigma_{n_c}^2 \boldsymbol{I})$. Note that the simultaneously received energy signals $\sqrt{1-\varrho}\sum_{k=1}^{K} \rho_k \boldsymbol{H}_{k,ST} \boldsymbol{x}_k s_k$, where $\boldsymbol{H}_{k,ST} \in \mathbb{C}^{N \times N_k}$ represents the complex channels from the kth EAP to the ST, are assumed to have been cancelled by the ST for the same reason as that for the PR and the SR.

B. The Second Time Slot

Transmitted signal at the ST. In the second time slot, the ST extracts the PR's desired message and superimposes it with its own message as follows.

$$\boldsymbol{x}_{ST}^{(2)} = \boldsymbol{w}_p s + \boldsymbol{q}_s, \tag{5}$$

where $\boldsymbol{w}_p$ denotes the beamforming vector for the PR's desired signal s, while $\boldsymbol{q}_s$ is the transmit signal conveying the SR's information aimed for multiplexing MIMO transmission, the covariance matrix of which is $\mathbb{E}[\boldsymbol{q}_s \boldsymbol{q}_s^H] = \boldsymbol{Q}_s$. As mentioned before, the transmit power for the ST is solely supplied by its harvest power, i.e.,

$$\mathsf{Tr}(\boldsymbol{Q}_s) + \|\boldsymbol{w}_p\|^2 \leq \eta\varrho P_{\mathrm{EH}}(\beta), \tag{6}$$

where $P_{\mathrm{EH}}(\beta) = \sum_{k=1}^{K} \rho_k \|\boldsymbol{H}_{k,ST}\boldsymbol{x}_k\|^2 + \beta P_p \|\boldsymbol{h}_{p,ST}\|^2$, and η denotes the EH conversion efficiency.

Received signal at the PR. In the second time slot, PT uses the rest of its available power to transmit a duplicate copy of its previously transmitted signal given by $\sqrt{(1-\beta)P_p}s$. Thus, the received signal at the PR is given by

$$y_{PR}^{(2)} = h_{pp}\sqrt{(1-\beta)P_p} s + \boldsymbol{g}_{sp}^H \boldsymbol{x}_{ST}^{(2)} + n_{PR}^{(2)}, \tag{7}$$

where $\boldsymbol{g}_{sp}$ denotes the conjugate transpose of the complex channel from the ST to the PR, and $n_{PR}^{(2)}$ denotes the additive noise at the PR. Plugging (5) into (7), $y_{PR}^{(2)}$ can be rewritten in a more compact form as follows.

$$y_{PR}^{(2)} = \boldsymbol{g}_{sp}^H \boldsymbol{q}_s + \boldsymbol{h}'^H_{pp} \boldsymbol{w}'_p s + n_{PR}^{(2)}, \tag{8}$$

where $\boldsymbol{h}'_{pp} = [\boldsymbol{g}_{sp}^H,\ h_{pp}]^H$ is an equivalent channel carrying the primary information s, and $\boldsymbol{w}'_p = [\boldsymbol{w}_p^H,\ \sqrt{(1-\beta)P_p}\ ]^H$ is an equivalent beamforming vector.

Received signal at the SR. The received signal at the SR is given by

$$\boldsymbol{y}_{SR}^{(2)} = \boldsymbol{G}_{ss}\boldsymbol{q}_s + \left(\boldsymbol{G}_{ss}\boldsymbol{w}_p + \boldsymbol{h}_{ps}\sqrt{(1-\beta)P_p}\right) s + \boldsymbol{n}_{SR}^{(2)}, \tag{9}$$

where $\boldsymbol{G}_{ss} \in \mathbb{C}^{M\times N}$ denotes the MIMO channel between the ST and the SR, and $\boldsymbol{n}_{SR}^{(2)}$ is the receiving noise at the SR, denoted by $\boldsymbol{n}_{SR}^{(2)} \sim \mathcal{CN}(0, \sigma_{SR}^2 \boldsymbol{I})$.

III. Problem Formulation

We assume that the PR performs maximum ratio combining (MRC) on its received signal (c.f. (2) and (8)) from both time slots to jointly decode s, which yields its maximum achievable signal-to-interference-plus-noise ratio (SINR) given by

$$\mathrm{SINR}_{PR} = \frac{\beta P_p |h_{pp}|^2}{\sigma_{PR}^2} + \frac{|\boldsymbol{h}'^H_{pp}\boldsymbol{w}'_p|^2}{\boldsymbol{g}_{sp}^H \boldsymbol{Q}_s \boldsymbol{g}_{sp} + \sigma_{PR}^2}. \tag{10}$$

Then, taking the DF relay into account, the achievable rate for the PR, denoted by r_{PR}, is given by

$$r_{PR} = \min\Bigg\{\frac{1}{2}\log_2\left(1 + \frac{(1-\varrho)\beta P_p \|\boldsymbol{h}_{p,ST}\|^2}{(1-\varrho)\sigma_{n_a}^2 + \sigma_{n_c}^2}\right), \frac{1}{2}\log_2\left(1 + \frac{\beta P_p |h_{pp}|^2}{\sigma_{PR}^2} + \frac{|\boldsymbol{h}'^H_{pp}\boldsymbol{w}'_p|^2}{\boldsymbol{g}_{sp}^H \boldsymbol{Q}_s \boldsymbol{g}_{sp} + \sigma_{PR}^2}\right)\Bigg\}. \tag{11}$$

The achievable rate for the underlaying MIMO secondary transmission pair, denoted by r_{SR}, is given by

$$r_{SR} = \frac{1}{2}\log_2 \det\left(\boldsymbol{I} + \frac{\boldsymbol{G}_{ss}\boldsymbol{Q}_s\boldsymbol{G}_{ss}^H}{\sigma_{SR}^2}\right). \tag{12}$$

Note that we assume herein that the interference introduced by the PT's signal (c.f. (9)) is perfectly cancelled out due to the fact that the SR can perform successive interfere cancellation (SIC) to remove the PT's signal that is decoded in (3).

In this paper, we assume that the ST will be rewarded by the primary system in proportional to PR's achievable rate, i.e., $c_1 r_{PR}$ (c.f. (11)), where c_1 is a reward conversion factor that relates the PR's rate to the ST's revenue. As mentioned before, the EAPs as well as the PT charge $c_2\eta\varrho P_{\mathrm{EH}}(\beta)$ from the ST for the amount of harvested power, where c_2 is a cost conversion factor denoting the payment per unit power. As a result, the payoff for the ST is given by $c_1 r_{PR} - c_2\eta\varrho P_{\mathrm{EH}}(\beta)$.

Next, we investigate the payoff maximization problem for the ST as follows.

$$\begin{aligned}
(\mathrm{P1}):\ &\max_{\boldsymbol{w}_p,\boldsymbol{Q}_s,\varrho,\beta,\{\rho_k\}} \quad c_1 r_{PR} - c_2\eta\varrho P_{\mathrm{EH}}(\beta) && (13a)\\
&\mathtt{s.t.}\ r_{PR} \geq R_{PR}, && (13b)\\
&\quad r_{SR} \geq R_{SR}, && (13c)\\
&\quad \boldsymbol{w}'_p = [\boldsymbol{w}_p^H,\ \sqrt{(1-\beta)P_p}\ ]^H, && (13d)\\
&\quad \rho_k \in \{0,1\},\ \forall k, && (13e)\\
&\quad \mathsf{Tr}(\boldsymbol{Q}_s) + \|\boldsymbol{w}_p\|^2 \leq \eta\varrho P_{\mathrm{EH}}(\beta), && (13f)\\
&\quad 0\leq\varrho\leq 1,\ 0\leq\beta\leq 1,\ \boldsymbol{Q}_s \succeq \boldsymbol{0}, && (13g)
\end{aligned}$$

where R_{PR} and R_{SR} are QoS-required transmission rates for the PR and the SR, respectively.

IV. Proposed Solutions

To solve (P1), we first consider EAPs' optimal WPT for maximizing its own revenue. For the kth EAP, $\forall k \in \mathcal{K}$, if it is selected for transmission, its optimal transmission strategy is the well-known eigenmode EB [13], and can be explicitly expressed as $\boldsymbol{x}_k^* = \sqrt{P_0}\boldsymbol{v}_{\max}(\boldsymbol{H}_{k,ST})$,where P_0 denotes the transmit power of each EAP once connected; $\boldsymbol{v}_{\max}(\cdot)$ denotes the right singular vector that corresponds to the largest singular value of the given matrix. In addition, the (reduced) singluar value decomposition (SVD) of $\boldsymbol{H}_{k,ST}$ (assuming $N \leq N_k$) is expressed as $\boldsymbol{H}_{k,ST} = \boldsymbol{U}_{H,k}\boldsymbol{\Sigma}_{H,k}^{1/2}\boldsymbol{V}_k^H$, where $\boldsymbol{\Sigma}_{H,k} = diag(\lambda_{H,k,1},\ldots,\lambda_{H,k,N})$. When ρ_k's are given as $\bar{\rho}_k$'s, the maximum amount of WPT received by the ST is consequently given by $\bar{P}_{\mathrm{EH}}(\beta) = P_0\sum_{k=1}^K \bar{\rho}_k\lambda_{H,k,\max} + \beta P_p\|\boldsymbol{h}_{p,ST}\|^2$, where $\lambda_{H,k,\max} = \max\limits_n\{\lambda_{H,k,n}\}_{n=1}^N$. Thus, (P1) can be reduced to

$$\begin{aligned}
(\mathrm{P2}):\ &\max_{\boldsymbol{w}_p,\boldsymbol{Q}_s,\varrho,\beta} \quad c_1 r_{PR} - c_2\eta\varrho \bar{P}_{\mathrm{EH}}(\beta) && (14a)\\
&\mathtt{s.t.}\ (13b),\ (13c),\ (13d), && (14b)\\
&\quad \mathsf{Tr}(\boldsymbol{Q}_s) + \|\boldsymbol{w}_p\|^2 \leq \eta\varrho \bar{P}_{\mathrm{EH}}(\beta), && (14c)\\
&\quad 0\leq\varrho\leq 1,\ 0\leq\beta\leq 1,\ \boldsymbol{Q}_s \succeq \boldsymbol{0}. && (14d)
\end{aligned}$$

As a step stone to solving (P1), we first investigate the solution of (P2).

A. Feasibility

To investigate the feasibility issue of (P2), we need to investigate the feasible rate region that is supported by the system above. To study the feasible rate pairs, the following problem is formulated to find the maximum achievable R_{PR}.

$$\begin{aligned}
(\mathrm{P0\text{-}1}):\ &\max_{\boldsymbol{w}_p,\boldsymbol{Q}_s,\varrho,\beta} \quad r_{PR} && (15a)\\
&\mathtt{s.t.}\ (13d),\ (14c), && (15b)\\
&\quad 0\leq\varrho\leq 1,\ 0\leq\beta\leq 1,\ \boldsymbol{Q}_s \succeq \boldsymbol{0}. && (15c)
\end{aligned}$$

Denoting the optimum value for (P0-1) by R_{PR}^*, then given any $R_{PR} \in [0, R_{PR}^*]$, consider the maximum-R_{SR} problem as follows.

$$(\text{P0-2}): \max_{\boldsymbol{w}_p,\boldsymbol{Q}_s,\varrho,\beta} \quad r_{SR} \tag{16a}$$
$$\texttt{s.t.}\ (13b),\ (13d),\ (14c), \tag{16b}$$
$$0 \le \varrho \le 1,\ 0 \le \beta \le 1,\ \boldsymbol{Q}_s \succeq \boldsymbol{0}. \tag{16c}$$

We propose to attain the feasible rate region as follows. First, given any $\bar{\beta} \in [0,1]$, we exploit the following proposition to find the maximum R_{PR} to (P0-1), denoted by $\bar{R}^*_{PR}$; next, plugging any feasible R_{PR} into problem (P0-2), we obtain the optimum R_{SR} to (P0-2), denoted by $\bar{R}^*_{SR}(R_{PR})$; then we study the Pareto boundary $(\bar{R}_{PR}, \bar{R}^*_{SR}(R_{PR}))$'s for each $\bar{\beta}$; finally, we characterize the feasible rate region for (P2) by taking their union sweeping over $\bar{\beta} \in [0,1]$.

It is also worthy of noting that in the sequel we constrain ourself to the channels related to $\boldsymbol{h}_{p,ST}$ and h_{pp} satisfying the following condition $|h_{pp}|^2 \le \frac{\sigma^2_{PR}}{\sigma^2_{n_a}+\sigma^2_{n_c}}\|\boldsymbol{h}_{p,ST}\|^2$, since otherwise the direct transmission has already outperformed the upper-bound of the DF relaying, which reduces to a trivial case that is out of the main focus of this paper.

To solve (P0-1), we rewrite its objective function into a tractable form that is associated with the SINR of the DF relaying in two transmission, phases, respectively, as follows.

$$(\text{P0-1}'): \max_{\boldsymbol{w}_p,\boldsymbol{Q}_s,\varrho,\beta,t} \quad t \tag{17a}$$
$$\texttt{s.t.}\ \frac{(1-\varrho)\beta P_p\|\boldsymbol{h}_{p,ST}\|^2}{(1-\varrho)\sigma^2_{n_a}+\sigma^2_{n_c}} \ge t, \tag{17b}$$
$$\frac{\beta P_p|h_{pp}|^2}{\sigma^2_{PR}} + \frac{|\boldsymbol{h}'^H_{pp}\boldsymbol{w}'_p|^2}{\boldsymbol{g}^H_{sp}\boldsymbol{Q}_s\boldsymbol{g}_{sp}+\sigma^2_{PR}} \ge t, \tag{17c}$$
$$(13d),\ (14c), \tag{17d}$$
$$0 \le \varrho \le 1,\ 0 \le \beta \le 1,\ \boldsymbol{Q}_s \succeq \boldsymbol{0}. \tag{17e}$$

Proposition 4.1: For any $\bar{\beta} \in [0,1]$, the optimum t to $(\text{P0-1}')$ is uniquely determined by the following equation w.r.t. t.

$$\frac{\left(\sqrt{\eta\varrho(t)\bar{P}_{\rm EH}(\bar{\beta})}\,\|\boldsymbol{g}_{sp}\| + \sqrt{(1-\bar{\beta})P_p}|h_{pp}|\right)^2}{\sigma^2_{PR}} + \frac{\bar{\beta}P_p|h_{pp}|^2}{\sigma^2_{PR}} - t = 0, \tag{18}$$

where $\varrho(t)$ is given by

$$\varrho(t) = 1 - \frac{\sigma^2_{n_c}t}{\bar{\beta}P_p\|\boldsymbol{h}_{p,ST}\|^2 - \sigma^2_{n_a}t}. \tag{19}$$

Proof: A sketch of the proof is outlined herein. First we show that the constraints in (17b)-(17c) are always active when $(\text{P0-1}')$ admits its optimum value by contradiction. Hence, (17c) implies that t^* is achieved when there is no QoS requirement for the SR. Next, given $\boldsymbol{Q}^*_s = 0$, looking into a subproblem of maximizing $|\boldsymbol{h}'^H_{pp}\boldsymbol{w}'_p|^2$, $\boldsymbol{w}^*_p$ proves to be aligned to the same direction as $\boldsymbol{g}_{sp}$ given by $\boldsymbol{w}'^*_p = [\sqrt{\eta\varrho(t)P_{\rm EH}(\bar{\beta})}\frac{\boldsymbol{g}_{sp}}{\|\boldsymbol{g}_{sp}\|}\exp\{-j\measuredangle h_{pp}\}; \sqrt{(1-\bar{\beta})P_p}]$. Then, substituting $\boldsymbol{w}'^*_p$ for $\boldsymbol{w}'_p$, we yield (18), where $\varrho(t)$ is obtained from (17b) with the inequality active. ■

With t^* numerically solved, the maximum achievable R_{PR} is consequently given by $R^*_{PR} = \frac{1}{2}\log_2(1+t^*)$. Next, given any $R_{PR} \in [\frac{1}{2}\log_2(1+\frac{\bar{\beta}P_p|h_{pp}|^2}{\sigma^2_{PR}}), R^*_{PR}]$, we consider to solve (P0-2) with fixed $\bar{\beta}$. Since given $\bar{\beta}$, (17b) is shown to be convex w.r.t. ϱ, by introducing $\boldsymbol{W}'_p = \boldsymbol{w}'_p\boldsymbol{w}'^H_p$ and ignoring the rank constraint that $\boldsymbol{W}'_p$ is subject to, (P0-2) turns out to be a semidefinite relaxation (SDR) problem, which is convex and thus can be efficiently solved by existing convex optimization toolbox.

It is worthy of noting that whether or not $\boldsymbol{W}'^*_p$ is achievable by a beamforming vector $\boldsymbol{w}'_p$ lies in the tightness of the SDR used in (P0-2-sub), which turns out to be tight and the proof will be discussed in the next subsection.

B. Optimal Solution to (P1)

In this subsection, we study the optimal solution to (P1). For any set of given integer variables $\{\bar{\rho}_k\}$, assuming that (P2) proves to be feasible in accordance with Section IV-A, denote its optimum value as $f_1(\{\bar{\rho}_k\})$. The optimal solution to (P1) can thus be obtained by $\{\rho^*_k\} = \arg\max_{\{\bar{\rho}_k\}\in\boldsymbol{\Pi}} f_1(\{\bar{\rho}_k\})$, where $\boldsymbol{\Pi} = \{0,1\}^K$ is defined by length-K Cartesian product. Hence, we only focus on solving (P3) in the sequel.

We commence with recasting (P2) into a two-stage problem. First, with $\bar{\beta} \in [0,1]$ fixed, we solve the epigraph reformulation of (P2) as follows.

$$(\text{P2.1}): \max_{\boldsymbol{w}_p,\boldsymbol{Q}_s,\varrho,t} \frac{c_1}{2}\log_2(1+t) - c_2\eta\varrho\bar{P}_{\rm EH}(\bar{\beta}) \tag{20a}$$
$$\texttt{s.t.}\ \frac{(1-\varrho)\bar{\beta}P_p\|\boldsymbol{h}_{p,ST}\|^2}{(1-\varrho)\sigma^2_{n_a}+\sigma^2_{n_c}} \ge t, \tag{20b}$$
$$\frac{\bar{\beta}P_p|h_{pp}|^2}{\sigma^2_{PR}} + \frac{|\boldsymbol{h}'^H_{pp}\boldsymbol{w}'_p|^2}{\boldsymbol{g}^H_{sp}\boldsymbol{Q}_s\boldsymbol{g}_{sp}+\sigma^2_{PR}} \ge t, \tag{20c}$$
$$t \ge 2^{R_{PR}} - 1, \tag{20d}$$
$$r_{SR} \ge R_{SR}, \tag{20e}$$
$$\boldsymbol{w}'_p = [\boldsymbol{w}^H_p,\ \sqrt{(1-\bar{\beta})P_p}\ ]^H, \tag{20f}$$
$$\mathsf{Tr}(\boldsymbol{Q}_s) + \|\boldsymbol{w}_p\|^2 \le \eta\varrho\bar{P}_{\rm EH}(\bar{\beta}), \tag{20g}$$
$$0 \le \varrho \le 1. \tag{20h}$$

Denoting the optimum value of (P2.1) by $f_3(\bar{\beta})$, (P2) can be equivalently solved by $(\text{P2.2}): \max_{0\le\bar{\beta}\le1} f_3(\bar{\beta})$, which allows for a simple one-dimension search over $\bar{\beta}$. As a result, we aim for solving (P2.1) in the sequel.

It is observed from (P2.1) that the maximum payoff is always attained when $\mathsf{Tr}(\boldsymbol{Q}_s)$ takes on its minimum value such that (13c) holds. This can also be intuitively seen from the fact that the ST is rewarded only by the PR's achievable rate, and thus it attempts to fulfill its own user's QoS requirement, i.e., R_{SR}, with as little power as possible such that the remaining harvested power allocated for assisting with PR's transmission is larger, which may lead to a larger achievable rate for the PR. Hence, problem (P2.1) is readily decoupled into two subproblems without loss of optimality as follows. The first subproblem is given by $\max\limits_{\boldsymbol{Q}_s, r_{SR}\geq R_{SR}} \mathsf{Tr}(\boldsymbol{Q}_s)$. which is straightforward to be seen as a dual problem of the achievable rate maximization subject to the total transmit power for point-to-point MIMO channel. It admits the standard "water-filling" solution given by $\boldsymbol{Q}_s^* = \boldsymbol{V} diag(P_1,\ldots,P_r)\boldsymbol{V}^H$ [14], where $\boldsymbol{V} \in \mathbb{C}^{N\times r}$ is given by the (reduced) SVD of $\boldsymbol{G}_{ss} = \boldsymbol{U}_G \boldsymbol{\Sigma}_G^{1/2}\boldsymbol{V}_G^H$ with $\boldsymbol{\Sigma}_G = diag(\lambda_{G,1},\ldots,\lambda_{G,r})$, $r = \min\{M,N\}$, $p_i = (\frac{\nu}{\ln 2} - \frac{\sigma_{SR}^2}{\lambda_{G,i}})^+$, $i = 1,\ldots,r$, and ν is constant determining the "water-level" such that $\frac{1}{2}\sum_{i=1}^r \log_2(1+\frac{p_i\lambda_{G,i}}{\sigma_{SR}^2}) = R_{SR}$.

Next, plugging the $\boldsymbol{Q}_s^*$ into (P2.1), we continue with devising the technique of SDR such that (P2.1) can be reformulated as a convex problem (the second subproblem) as follows.

$$
\begin{aligned}
\text{(P2.1-SDR)}: &\max_{\boldsymbol{W}'_p,\varrho,t} \frac{c_1}{2}\log_2(1+t) - c_2\eta\varrho\bar{P}_{\mathrm{EH}}(\bar{\beta}) && \text{(21a)}\\
&\mathtt{s.t.}\ \frac{(1-\varrho)\bar{\beta}P_p\|\boldsymbol{h}_{p,ST}\|^2}{(1-\varrho)\sigma_{n_a}^2+\sigma_{n_c}^2} \geq t, && \text{(21b)}\\
&\frac{\bar{\beta}P_p|h_{pp}|^2}{\sigma_{PR}^2} + \frac{\mathsf{Tr}(\boldsymbol{H}_{pp}\boldsymbol{W}'_p)}{\boldsymbol{g}_{sp}^H\boldsymbol{Q}_s^*\boldsymbol{g}_{sp}+\sigma_{PR}^2} \geq t, && \text{(21c)}\\
&t \geq 2^{R_{PR}}-1, && \text{(21d)}\\
&\mathsf{Tr}(\bar{\boldsymbol{E}}\boldsymbol{W}'_p) \leq \eta\varrho\bar{P}_{\mathrm{EH}}(\bar{\beta}) - \mathsf{Tr}(\boldsymbol{Q}_s^*), && \text{(21e)}\\
&\mathsf{Tr}(\boldsymbol{U}_{N+1}\boldsymbol{W}'_p) = (1-\bar{\beta})P_p, && \text{(21f)}\\
&0\leq\varrho\leq 1,\ \boldsymbol{W}'_p \succeq \boldsymbol{0}, && \text{(21g)}
\end{aligned}
$$

where $\boldsymbol{H}_{pp} = \boldsymbol{h}'_{pp}\boldsymbol{h}'^H_{pp}$, $\bar{\boldsymbol{E}} = \boldsymbol{E}^H\boldsymbol{E}$, in which $\boldsymbol{E} = [\boldsymbol{I}_N\ \boldsymbol{0}]$, and $\boldsymbol{U}_{N+1} = diag(\boldsymbol{e}_{N+1})$, in which $\boldsymbol{e}_{N+1} \in \mathbb{R}^{(N+1)\times 1}$ denotes an element vector with the $(N+1)$th entry being 1 and all the other 0.

As stated in Section IV-A, the upper-bound solution of $\boldsymbol{W}'_p$ by relaxing its rank constraint can be achieved if and only if $\mathrm{rank}(\boldsymbol{W}'_p) = 1$, which is guaranteed by the following proposition.

Proposition 4.2: The optimal solution to (P3.1-2-SDR) always satisfies $\mathrm{rank}(\boldsymbol{W}'^*_p) \leq 1$.

Proof: Please refer to Appendix A. ■

C. Proposed Solution to (P1)

In this subsection, we study a suboptimal solution to (P1). It is known from Section IV-B that the optimal solution to (P1) requires exhaustive search over $\boldsymbol{\Pi}$, which induces computational complexity up to $\mathcal{O}(2^K)$, and is thus quite prohibitive in practical system. Hence, we propose to reduce this integer programming (IP)-induced complexity to $\mathcal{O}(\mathcal{K})$ by designing a bi-direction greedy-based algorithm as shown in Table I.

The main thrust of this scheme is two-fold. On one hand, we aim to be "greedy" in terms of the ST's revenue by incrementally decreasing its value from an all-on state ($\{\bar{\rho}_k = 1\}$), i.e., turning off each time one EAP that corresponds to presently the least $\lambda_{H,k,\max}$, until the payoff begins to decrease. On the other hand, we aim to be "greedy" in terms of saving cost by incrementally increasing P_{EAP}^* from an all-off state ($\{\bar{\rho}_k = 0\}$) vice versa. At last, the ST chooses $\{\bar{\rho}_k\}$ from the above two options that yields a larger payoff.

TABLE I
A GREEDY-BASED ALGORITHM FOR SOLVING (P1)

Require: (R_{PU}, R_{SU})
sort $diag(\boldsymbol{\Sigma}_{H,k})$ s.t. $\lambda_{H,n_1,\max} \geq \lambda_{H,n_2,\max} \geq \ldots \lambda_{H,n_K,\max}$
$ii \leftarrow K$, $\bar{\rho}_{n_k}^{(ii)} \leftarrow 1$, $k = 1,\ldots,K$
solve (P0) given $\{\bar{\rho}_{n_k}^{(ii)}\}$, and obtain $f_1(\{\bar{\rho}_{n_k}^{(ii)}\})$
if $R_{PU} > R_{PU}^*$ OR $R_{SU} > R_{SU}^*$ **then**
  return ('Infeasible')
else
  repeat
    $ii \leftarrow ii-1$, $\bar{\rho}_{n_{ii+1}}^{(ii)} \leftarrow 0$, and update $\{\bar{\rho}_{n_k}^{(ii)}\}$, $f_1(\{\bar{\rho}_{n_k}^{(ii)}\})$
  until $f_1(\{\bar{\rho}_{n_k}^{(ii)}\}) < f_1(\{\bar{\rho}_{n_k}^{(ii+1)}\})$ OR $ii = 0$
end if
return $ii^* = \arg\max_{ii\leq\bar{ii}\leq K} f_1(\{\bar{\rho}_{n_k}^{(\bar{ii})}\})$
sort $diag(\boldsymbol{\Sigma}_{H,k})$ s.t. $\lambda_{H,n'_1,\max} \leq \lambda_{H,n'_2,\max} \leq \ldots \lambda_{H,n'_K,\max}$
$jj \leftarrow 0$, $\bar{\rho}_{n'_k}^{(jj)} \leftarrow 0$, $k = 1,\ldots,K$
solve (P0) given $\{\bar{\rho}_{n'_k}^{(jj)}\}$, and obtain $f_1(\{\bar{\rho}_{n'_k}^{(jj)}\})$
repeat
  $jj \leftarrow jj+1$, $\bar{\rho}_{n'_{jj}}^{(jj)} \leftarrow 1$, and update $\{\bar{\rho}_{n'_k}^{(jj)}\}$, $f_1(\{\bar{\rho}_{n'_k}^{(jj)}\})$
until $f_1(\{\bar{\rho}_{n'_k}^{(jj)}\}) < f_1(\{\bar{\rho}_{n'_k}^{(jj-1)}\})$ OR $jj = K$
return $jj^* = \arg\max_{1\leq\bar{jj}\leq jj} f_1(\{\bar{\rho}_{n'_k}^{(\bar{jj})}\})$
Ensure: $f_1^{(\mathrm{grd})} = \max\{f_1(\{\bar{\rho}_{n_k}^{(ii^*)}\}), f_1(\{\bar{\rho}_{n'_k}^{(jj^*)}\})\}$,
$\{\rho_k^{(\mathrm{grd})}\} = \arg\max\{f_1(\{\bar{\rho}_{n_k}^{(ii^*)}\}), f_1(\{\bar{\rho}_{n'_k}^{(jj^*)}\})\}$

V. NUMERICAL RESULTS

In this section, we provide numerical results to validate the proposed complexity reduced joint EAP-selection and DF beamforming scheme for the considered wireless powered spectrum-sharing CRN, referred to as *proposed*. Denoted by *optimal*, the optimal solutions to problem (P1), in spite of their exponential complexity, serve as achievable performance upper-bound.

Taking the position of the ST as the origin of a polar coordinate, the PT, PR and SR are assumed to be located at $(-5,0)$, $(5,\frac{\pi}{6})$ and $(5,\frac{11\pi}{6})$, respectively, with the default unit for distance m. We also assume that K EAPs are uniformly located within a circle of radius $R_{\max}$m centred at the ST. The channel models consist of both large-scale and small-scale fading, the former of which is given by a simple path loss model with a path loss exponent of 2.5 in addition to 30dB free space attenuation, and the latter of which is multi-path fading following $i.i.d.$ complex Gaussian distribution with zero mean and unit variance. Unless otherwise specified, the simulation parameters are set as follows: $R_{\max} = 8$m, $P_p = 30$dBm, $P_0 = 40$dBm, $K = 5$, $M = 3$, $N = 3$, $\eta = 0.5$, $\sigma_{n_a}^2 = -170$dBm, $\sigma_{n_c}^2 = -130$dBm, $\sigma_{PR}^2 = \sigma_{SR}^2 = \sigma_{n_a}^2 + \sigma_{n_c}^2$, $\eta = 0.5$, $c_1 = 1$, and $c_2 = 100$. N_k's are all set to be 10.

Fig. 2 depicts the Pareto boundaries of the feasible rate regions by setting $\bar{\rho}_k = 1$, $\forall k \in \mathcal{K}$. The feasible rate regions are accordingly given by $\{(r_{PR}, r_{SR}) \mid r_{PR} \leq R_{PR}, r_{SR} \leq R_{SR}\}$. It is observed from Fig. 2 that the same R_{SR} can be supported by a large range of R_{PR}. It is also seen that given

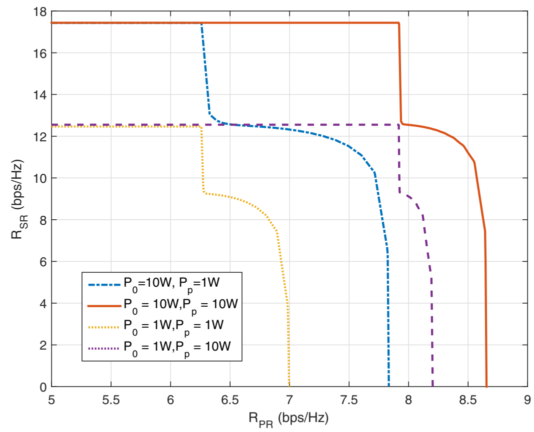

Fig. 2. The feasible rate region of (R_{PR}, R_{SR}), $K = 10$, $L = 10$, $M = N = 3$ under different P_0 and P_p.

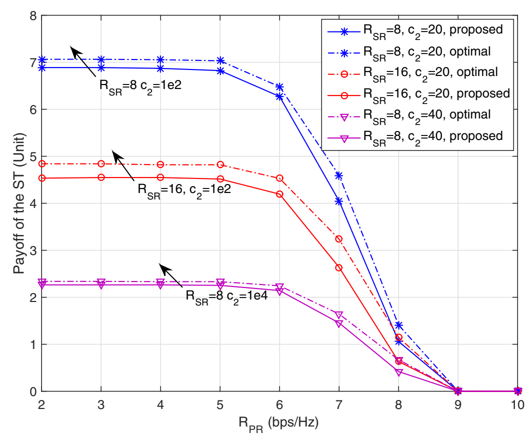

Fig. 3. The average payoff of the ST by different schemes vs R_{PR}, $K = 5$.

the same P_p, both R^*_{PR}'s and R^*_{SR}'s response positively with the increase in P_0, which implies that both PR and SR can benefit from the WPT phase.

Fig. 3 shows the average ST's payoff versus the PR's transmission rate requirement. As expected intuitively, the average payoff of the ST remains stable for small and mild value of R_{PR}, and is seen to first climb down with higher QoS requirement on PR's achievable rate and then to arrive at zero (infeasible) when R_{PR} increases to 9bps/Hz. Given the same cost conversion factor c_2, a larger transmission rate required by the SR also results in an obvious decrease in the ST's payoff. Furthermore, the proposed schemes do not exhibit much performance loss from the optimal solutions. In particular, when c_2 increases by 20dB, the optimal scheme outperforms the proposed one with negligible gap. This can be explained as follows. When c_2 is large, the ST attempts to connect to as fewer EAPs as possible, as long as the required R_{PR} and R_{SR} are satisfied. Accordingly, the optimum selection scheme tends to activate only one EAP, which can be easily reached by one of our greedy method's options.

VI. Conclusion

This paper considers a novel WEH-enabled CCRN operating with DF relaying completely powered by the PT and multiple EAPs. Assuming that direct links are active, a multi-antenna SU pair assists a single-antenna PU pair in DF relaying, and in return is allowed to superimpose its own signal with the PU's. A joint optimization of EAP-selection scheme, WEH-enabled information reception as well as the DF relay beamforming for the ST, and the power allocations for the PT has been investigated to maximize the ST's payoff, subject to QoS requirements of PR and SR, and the harvested power of the ST. We optimally solve this IP-mixed non-convex optimization problem using the technique of SDR which is proved to be tight. We also propose a low complexity suboptimal algorithm based on greedy EAP-selection, the effectiveness of which is validated by numerical results.

Appendix A

First, we present the Lagrangian of (P2.1-SDR) as follows.

$$\begin{aligned}
&\mathcal{L}(\boldsymbol{W}'_p, \varrho, t) = \frac{c_1}{2}\log_2(1+t) - c_2\eta\varrho\bar{P}_{\text{EH}}(\bar{\beta}) \\
&+\alpha\left(\frac{\bar{\beta}P_p\|h_{p,ST}\|^2}{\sigma^2_{n_a}} - \frac{\sigma^2_{n_c}/\sigma^2_{n_a}\bar{\beta}P_p\|h_{p,ST}\|^2}{(1-\varrho)\sigma^2_{n_a}+\sigma^2_{n_c}} - t\right) \\
&+\gamma\left(\frac{\bar{\beta}P_p|h_{pp}|^2}{\sigma^2_{PR}} + \frac{\mathsf{Tr}(\boldsymbol{H}'_{pp}\boldsymbol{W}'_p)}{\boldsymbol{g}^H_{sp}\boldsymbol{Q}^*_s\boldsymbol{g}_{sp}+\sigma^2_{PR}} - t\right) \\
&+\lambda(t - 2^{R_{PR}}+1) - \mu\left(\mathsf{Tr}(\bar{\boldsymbol{E}}\boldsymbol{W}'_p) - \eta\varrho\bar{P}_{\text{EH}}(\bar{\beta}) + \mathsf{Tr}(\boldsymbol{Q}^*_s)\right) \\
&+\nu\left(\mathsf{Tr}(\boldsymbol{U}_{N+1}\boldsymbol{W}'_p) - (1-\bar{\beta})P_p\right) + \mathsf{Tr}(\boldsymbol{V}\boldsymbol{W}'_p), \qquad (22)
\end{aligned}$$

where α and γ are dual variables associated with SINR constraints in, respectively; λ, μ and ν are those associated with; finally, $\boldsymbol{V}$ is the dual variable for semidefinite $\boldsymbol{W}'_p$.

In accordance with Karush-Kuhn-Tucker (KKT) conditions [15], from (22), we can derive that

$$\frac{\gamma\boldsymbol{H}'_{pp}}{\boldsymbol{g}^H_{sp}\boldsymbol{Q}^*_s\boldsymbol{g}_{sp}+\sigma^2_{PR}} - \mu\bar{\boldsymbol{E}} + \nu\boldsymbol{U}_{N+1} + \boldsymbol{V} = \boldsymbol{0}, \qquad (23a)$$

$$\boldsymbol{V}\boldsymbol{W}'_p = \boldsymbol{0}. \qquad (23b)$$

Post-multiply (23a) with $\boldsymbol{W}'_p$, combing with (23b), it follows that

$$\left(\mu\bar{\boldsymbol{E}} - \nu\boldsymbol{U}_{N+1}\right)\boldsymbol{W}'_p = \frac{\gamma\boldsymbol{H}'_{pp}}{\boldsymbol{g}^H_{sp}\boldsymbol{Q}^*_s\boldsymbol{g}_{sp}+\sigma^2_{PR}}\boldsymbol{W}'_p. \qquad (24)$$

First, we show that $\mu > 0$ by contradiction. Assuming $\mu = 0$, in line with (23a), by denoting $\boldsymbol{g}^H_{sp}\boldsymbol{Q}^*_s\boldsymbol{g}_{sp}+\sigma^2_{PR}$ by c, we arrive at $\frac{\gamma}{c}\boldsymbol{H}'_{pp} + \nu\boldsymbol{U}_{N+1} = -\boldsymbol{V}$, where the LHS is positive semidefinite, and the RHS is negative semidefinite as a result of $\boldsymbol{V} \succeq \boldsymbol{0}$. Thus, it holds if and only if $\gamma = 0$ and $\nu = 0$. Consequently, the KKT condition in terms of $\mathcal{L}(\boldsymbol{W}'_p, \varrho, t)$'s derivative w.r.t. ϱ is given by

$$-c_2\eta\bar{P}_{\text{EH}}(\bar{\beta}) - \alpha\frac{\sigma^2_{n_c}/\sigma^2_{n_a}\bar{\beta}P_p\|h_{p,ST}\|^2}{\left((1-\varrho)+\sigma^2_{n_c}/\sigma^2_{n_a}\right)^2} = 0, \qquad (25)$$

which causes contradiction. Henc $\mu \neq 0$ is proved.

Next, based upon $\mu > 0$, we show that $\mathrm{rank}(\boldsymbol{W}_p'^{*}) \leq 1$ in the following two cases. In the first case, $\nu > 0$. $\mu\bar{\boldsymbol{E}} - \nu\boldsymbol{U}_{N+1}$ is thus a full-rank diagonal matrix, and it follows from (24) that

$$\boldsymbol{W}_p' = \left(\mu\bar{\boldsymbol{E}} - \nu\boldsymbol{U}_{N+1}\right)^{-1} \frac{\gamma\boldsymbol{H}_{pp}'}{\boldsymbol{g}_{sp}^H\boldsymbol{Q}_s^*\boldsymbol{g}_{sp} + \sigma_{PR}^2}\boldsymbol{W}_p', \tag{26}$$

which implies that $\mathrm{rank}(\boldsymbol{W}_p') \leq \mathrm{rank}(\boldsymbol{H}_{pp}') = 1$. In the second case, $\nu = 0$, from which two subcases are developed. In the first subcase, $\gamma = 0$. It thus suggests that $\mu\bar{\boldsymbol{E}}\boldsymbol{W}_p' = \boldsymbol{0}$ (c.f. (24)), from which it is easily seen that $\mathrm{rank}(\boldsymbol{W}_p') = 1$, since $\mathrm{rank}(\bar{\boldsymbol{E}}) = N$. In the second subcase, $\gamma > 0$. We show that in the sequel this subcase is impossible by contradiction. In accordance with (23a), if $\nu = 0$, $\boldsymbol{V} = \mu\bar{\boldsymbol{E}} - \frac{\gamma}{c}\boldsymbol{H}_{pp}'$, which yields $[\boldsymbol{V}]_{N+1,N+1} = -\frac{\gamma}{c}|h_{pp}|^2 < 0$ that contradicts to $\boldsymbol{V} \succeq \boldsymbol{0}$. Hence when $\nu = 0$, it follows that $\gamma = 0$ and thus $\boldsymbol{W}_p' = \boldsymbol{0}$, the rank of which satisfies $\mathrm{rank}(\boldsymbol{W}_p') \leq 1$.

References

[1] , "Spectrum policy task force," Federal Communications Commission, ET Docket No. 02-135, Tech. Rep., Nov. 2002.

[2] J. Mitola and G. Q. Maguire, "Cognitive radio: Makeing software radios more personal," *IEEE Pers. Commun.*, vol. 6, no. 6, pp. 13–18, Aug. 1999.

[3] S. Haykin, "Cognitive radio: brain-empowered wireless communications," *IEEE J. Sel. Areas Commun.*, vol. 23, no. 2, pp. 201–220, Feb. 2005.

[4] N. Devroye, P. Mitran, and V. Tarokh, "Achievable rates in cognitive radio channels," *IEEE Trans. Inf. Theory*, vol. 52, no. 5, pp. 1813–1827, May 2006.

[5] Y. Guo, J. Xu, L. Duan, and R. Zhang, "Joint energy and spectrum cooperation for cellular communication systems," *IEEE Trans. Commun.*, vol. 62, no. 10, pp. 3678–3691, Oct. 2014.

[6] J. Xu, Y. Guo, and R. Zhang, "CoMP meets energy harvesting: A new communication and energy cooperation paradigm," in *Proc. IEEE Global Communications Conference (GLOBECOM'13)*, Atlanta, GA, USA, Dec. 2013, pp. 2508–2513.

[7] L. R. Varshney, "Transporting information and energy simultaneously," in *Proc. IEEE International Symposium on Information Theory (ISIT'08)*, Toronto, ON, CA, July 2008, pp. 1612–1616.

[8] K. Huang and V. K. N. Lau, "Enabling wireless power transfer in cellular networks: architecture, modeling and deployment," *IEEE Trans. Wireless Commun.*, vol. 13, no. 99, pp. 1–11, Jan. 2014.

[9] X. Kang, C. K. Ho, and S. Sun, "Full-duplex wireless-powered communication network with energy causality," *IEEE Trans. Wireless Commun.*, vol. 14, no. 10, pp. 5539–5551, Oct. 2015.

[10] G. Zheng, Z. Ho, E. A. Jorswieck, and B. Ottersten, "Information and energy cooperation in cognitive radio networks," *IEEE Trans. Signal Process.*, vol. 62, no. 9, pp. 2290–2303, May 2014.

[11] L. Mohjazi, M. Dianati, G. K. Karagiannidis, S. Muhaidat, and M. Al-Qutayri, "RF-powered cognitive radio networks: technical challenges and limitations," *IEEE Commun. Mag.*, vol. 53, no. 4, pp. 94–100, April 2015.

[12] V. Rakovic, D. Denkovski, Z. Hadzi-Velkov, and L. Gavrilovska, "Optimal time sharing in underlay cognitive radio systems with RF energy harvesting," in *Proc. IEEE International Conference on Communications (ICC'15)*, London, UK, Jun. 2015, pp. 7689–7694.

[13] R. Zhang and C. K. Ho, "MIMO broadcasting for simultaneous wireless information and power transfer," *IEEE Trans. Wireless Commun.*, vol. 12, no. 5, pp. 1989–2001, May 2013.

[14] T. M. Cover and J. A. Thomas, *Elements of Information Theory*, 2nd ed. John Wiley & Sons, 2006.

[15] S. Boyd and L. Vandenberghe, *Convex Optimization*. Cambridge Univ. Press, 2004.