Error Performance of Wireless Powered Cognitive Relay Networks with Interference Alignment

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Abstract

This paper studies a two-hop decode-and-forward underlay cognitive radio system with interference alignment technique. An energy-constrained relay node harvests the energy from the interference signals through a power-splitting (PS) relaying protocol. Firstly, the beamforming matrices design for the primary and secondary networks is demonstrated. Then, a bit error rate (BER) performance of the system under perfect and imperfect channel state information (CSI) scenarios for PS protocol is calculated. Finally, the impact of the CSI mismatch parameters on the BER performance is simulated.

Keywords

Decode-and-forward (DF) relaying, energy harvesting (EH), cognitive radio (CR), interference alignment (IA), wireless power transfer, bit error rate (BER).

I. INTRODUCTION

C OGNITIVE RADIO (CR) has attracted a significant attention by being an intelligent technology that can utilize radio spectrum and increase the spectral efficiency in wireless networks [1]. The main idea of CR is to provide secondary users (SUs), which are unlicensed nodes, with possibility to communicate in a licensed spectrum on the condition that primary users (PUs) should not receive harmful interference [1], [2]. There are three types of spectrum sharing CR paradigms, namely, interweave, underlay and overlay [1]. The interference mitigation at receivers can be managed by a promising technique named interference alignment (IA). IA provides free signaling dimensions by aligning the interference into one subspace [3], [4]. IA can be also applied in CR to cancel interference at primary and secondary receivers. A mitigation of the severe effect of interference at primary receivers allows secondary transmitters to increase their transmit power which consequently leads to an improvement of the performance of the secondary network (SN) [5]. In [6], degrees of freedom (DoFs) of the network were increased by implementing an IA technique in a multiple-input multiple-output (MIMO) relay CR.

Another promising technique, known as energy harvesting (EH), which harvests energy from ambient signals through time-switching (TS) and power-splitting (PS), was introduced in [7], [8]. IA and simultaneous wireless information and power transfer (SWIPT) in MIMO networks were jointly studied in [9], where an improvement of the performance of the network was analyzed through a dynamical selection of users as EH or information decoding (ID) terminals. The EHbased CR network was studied in [10], where authors developed a spectrum sensing policy in TS mode to guarantee EH possibility for SUs from primary signals. An identical system was studied in [11], where an optimal information and energy cooperation methods between primary and secondary networks was investigated. Finally, the work in [12] represented a resource allocation method in EH-based CR network with imperfect channel state information (CSI) cases.

In this paper, we study an underlay IA-based CR with an energy-restricted relay operating in PS mode. The performance of both the primary network (PN) and SN after interference mitigation is analyzed. In particular, a bit error rate (BER) performance of the proposed system model is calculated for PS relaying protocol under different imperfect CSI cases.

II. SYSTEM MODEL

The proposed system model is composed of a PN with two pairs of PUs and a SN with three SUs. Each primary transmitter (T_i) transmits to its corresponding receiver by interfering with another primary receiver (R_i) and relay as shown in Fig. 1. The SN is composed of a source (S), a relay (R) and a destination (D) nodes. An energy constrained R operates in decode-andforward (DF) half-duplex mode by relaying the signal from S to D in two time periods. Ruses harvested energy from interference signals as its transmit power, while S and D supplied with stationary power sources. Also, it is assumed that D is located far from PN and does not receive any interference. All nodes of the network are assumed to have MIMO antennas. We also assume that all interference at R_j are canceled by IA technique, thus, S and R are not restricted by the level of transmit power. Another assumption is that the channels remain constant during a transmission block time T, but vary independently from one block to another. The definition of channel links between nodes can be denoted by the next. For channels of PN nodes, $\mathbf{H}_{j,i}^{[k]} \in \mathbb{C}^{N_j \times M_i}$, $\forall i, j \in \{1, 2\}$ denotes the channel between R_j and T_i , where superscript k indicates a certain time period when the data transmission occurs. N_j and M_i are the numbers of antennas at R_j and T_i , respectively. For channels of SN nodes, $\mathbf{H}_{R,S}$ and $\mathbf{H}_{D,R}$ denote the channel links related to the S-R and R-D transmissions while the inter-network channels are given by $\mathbf{H}_{j,R} \in \mathbb{C}^{N_j \times N_R}$, $\mathbf{H}_{j,S} \in \mathbb{C}^{N_j \times N_S}$ and $\mathbf{H}_{R,i} \in \mathbb{C}^{N_R \times M_i}$, where N_S , N_R and N_D denote the numbers of antennas at S, R and D, respectively. Each entry of any matrix **H** is assumed to be independent and identically distributed (i.i.d.) random variables according to $\mathcal{CN}(0,1)$, where $\mathcal{CN}(0,1)$ denotes the complex normal distribution with zero mean and unit variance. Also, note that each channel link is characterized by the corresponding distance and path-loss exponent denoted by $d_{m,n}$ and $\tau_{m,n}$, $\forall m \in \{1, 2, R, D\}$, $\forall n \in \mathcal{A} = \{1, 2, S, R\}$, respectively. We assume that each node is deployed with N antennas ($M_i = N_j = N_S = N_R = N_D = N$) and IA is exploited at R and R_i , accordingly. Therefore, each transmit node l with the power P_l employs a precoding matrix $\mathbf{V}_l \in \mathbb{C}^{(M_l \text{ or } N_l) \times f_l}$, with trace $\{\mathbf{V}_l \mathbf{V}_l^H\} = 1, \forall l \in \mathcal{A}$, where f_l is the number of the transmitted data streams. Then, each receive node, except D, employs the interference suppression matrix $\mathbf{U}_t \in \mathbb{C}^{N_t \times f_t}$, $\forall t \in \{1, 2, R\}$, where f_t is the number of data streams that needs to be decoded at the corresponding receiver. Thus, the received signal at R_j , in two time

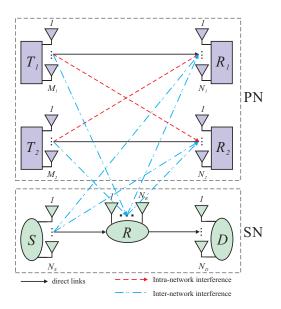


Fig. 1. IA and EH-based CRN with two PUs and one SU sharing the spectrum simultaneously.

periods, can be written as

$$\mathbf{y}_{j}^{[k]} = \underbrace{\sqrt{\frac{P_{j}}{d_{j,j}^{\tau_{j,j}}}} \mathbf{U}_{j}^{[k]H} \mathbf{H}_{j,j}^{[k]} \mathbf{V}_{j}^{[k]} \mathbf{s}_{j}}_{\text{desired signal}} + \underbrace{\mathbf{A}_{\text{interference from SN}}^{[k]} + \underbrace{\mathbf{A}_{j,i}^{[k]}}_{\text{interference from PN, } i \neq j} \mathbf{U}_{j}^{[k]H} \mathbf{H}_{j,i}^{[k]} \mathbf{V}_{i}^{[k]} \mathbf{s}_{i} + \tilde{\mathbf{n}}_{j}^{[k]}, \ k \in \{1, 2\}, \ (1)$$

where the effective noise term $\tilde{\mathbf{n}}_{j}^{[k]} = \mathbf{U}_{j}^{[k]H} \mathbf{n}_{j}^{[k]}$ is a zero-mean additive white Gaussian noise (AWGN) vector, with $\mathbb{E}{\{\tilde{\mathbf{n}}_{j}^{[k]}\tilde{\mathbf{n}}_{j}^{[k]H}\}} = \sigma_{\tilde{n}_{j}}^{2}\mathbf{I}$, where $\mathbb{E}{\{\cdot\}}$ denotes an expectation operator. Moreover, we have $\mathbb{E}{\{\mathbf{s}_{l}\mathbf{s}_{l}^{H}\}} = \mathbf{I}$, with $l \in \mathcal{A}$, since \mathbf{s}_{l} is assumed to be a vector consisting of symbols generated as i.i.d. Gaussian inputs. Finally, interference from SN to R_{j} can be determined as

$$\mathbf{A}^{[k]} = \begin{cases} \sqrt{\frac{P_S}{d_{j,S}^{\tau_{j,S}}}} \mathbf{U}_j^{[k]H} \mathbf{H}_{j,S} \mathbf{V}_S \mathbf{s}_S, & \text{if } k = 1, \\ \sqrt{\frac{P_R}{d_{j,R}^{\tau_{j,R}}}} \mathbf{U}_j^{[k]H} \mathbf{H}_{j,R} \mathbf{V}_R \mathbf{s}_R, & \text{if } k = 2. \end{cases}$$
(2)

The received signal at R, within transmission period of S-R, can be written as

$$\mathbf{y}_{R} = \underbrace{\sqrt{\frac{P_{S}}{d_{R,S}^{\tau_{R,S}}}} \mathbf{U}_{R}^{H} \mathbf{H}_{R,S} \mathbf{V}_{S} \mathbf{s}_{S}}_{\text{desired signal}} + \underbrace{\sqrt{\frac{P_{i}}{d_{R,i}^{\tau_{R,i}}}} \sum_{i=1}^{2} \mathbf{U}_{R}^{H} \mathbf{H}_{R,i} \mathbf{V}_{i}^{[1]} \mathbf{s}_{i}}_{\text{interference from PN}} + \tilde{\mathbf{n}}_{R},$$
(3)

where $\tilde{\mathbf{n}}_R = \mathbf{U}_R^H \mathbf{n}_R$ is the effective noise after interference suppression beamforming at the

relay.

Then, R decodes and forwards the desired signal s_S to D within R - D transmission period. Thus, D obtains the following signal

$$\mathbf{y}_D = \sqrt{\frac{P_R}{d_{D,R}^{\tau_{D,R}}}} \mathbf{H}_{D,R} \mathbf{V}_R \mathbf{s}_R + \mathbf{n}_D,$$
(4)

where \mathbf{n}_D is the AWGN vector, with $\mathbb{E}\{\mathbf{n}_D\mathbf{n}_D^H\} = \sigma_D^2\mathbf{I}$.

The interference in receive nodes can be assumed to be completely canceled if the following conditions are satisfied for R_j as [13], [14]

$$\mathbf{U}_{j}^{[k]H}\mathbf{H}_{j,i}^{[k]}\mathbf{V}_{i}^{[k]} = \mathbf{0}, \ \forall i, j \in \{1, 2\}, \ \forall i \neq j,$$
(5a)

$$\mathbf{U}_{j}^{[k]H}\mathbf{J}^{[k]} = \mathbf{0}, \text{ where } \mathbf{J}^{[k]} = \begin{cases} \mathbf{H}_{j,S}\mathbf{V}_{S}, \text{ if } k = 1, \\ \mathbf{H}_{j,R}\mathbf{V}_{R}, \text{ if } k = 2, \end{cases}$$
(5b)

$$\operatorname{rank}\left(\mathbf{U}_{j}^{[k]H}\mathbf{H}_{j,j}^{[k]}\mathbf{V}_{j}^{[k]}\right) = f_{j}, \ \forall j \in \{1,2\},$$
(5c)

and for R as

$$\mathbf{U}_{R}^{H}\mathbf{H}_{R,i}\mathbf{V}_{i}^{[1]} = \mathbf{0}, \ \forall i \in \{1,2\},$$
(6a)

$$\operatorname{rank}\left(\mathbf{U}_{R}^{H}\mathbf{H}_{R,S}\mathbf{V}_{S}\right) = f_{S}.$$
(6b)

A. Beamforming Design

If the space of the desired signal is linearly independent from that of the interference signal, then the desired signal can be easily decoded from the received one. Hence, the design of precoding matrices should be in such a way that interference in all receivers need to span to one another. Thus, in the first time period, the interference at R_1 , R_2 and R can be spanned as span $\left(\mathbf{H}_{1,2}^{[1]}\mathbf{V}_2^{[1]}\right) = \operatorname{span}\left(\mathbf{H}_{1,S}\mathbf{V}_S\right)$, span $\left(\mathbf{H}_{2,1}^{[1]}\mathbf{V}_1^{[1]}\right) = \operatorname{span}\left(\mathbf{H}_{2,S}\mathbf{V}_S\right)$ and span $\left(\mathbf{H}_{R,1}\mathbf{V}_1^{[1]}\right) =$ span $\left(\mathbf{H}_{R,2}\mathbf{V}_2^{[1]}\right)$, respectively, where span(\mathbf{X}) is the vector space spanned by the column vectors of \mathbf{X} . After spanning all interference, the precoding matrices $\mathbf{V}_1^{[1]}$, $\mathbf{V}_2^{[1]}$ and \mathbf{V}_S can be obtained as [15]

$$\mathbf{V}_{2}^{[1]} = (\mathbf{H}_{R,2})^{-1} \mathbf{H}_{R,1} \mathbf{V}_{1}^{[1]},$$
(7a)

$$\mathbf{V}_{S} = (\mathbf{H}_{2,S})^{-1} \mathbf{H}_{2,1}^{[1]} \mathbf{V}_{1}^{[1]},$$
(7b)

where $\mathbf{V}_{1}^{[1]}$ is derived using $\mathbf{V}_{1}^{[1]} = \operatorname{eig}(\mathbf{Z})$, with $\mathbf{Z} = (\mathbf{H}_{R,1})^{-1} \mathbf{H}_{R,2} \left(\mathbf{H}_{1,2}^{[1]}\right)^{-1} \mathbf{H}_{1,S}(\mathbf{H}_{2,S})^{-1} \mathbf{H}_{2,1}^{[1]}$ and $\operatorname{eig}(\mathbf{X})$ are the eigenvectors of \mathbf{X} .

Interference suppression matrices $\mathbf{U}_{j}^{[k]}$ during two time slots, need to be orthogonalized to the interference at R_{j} to meet conditions in (5). Similarly, \mathbf{U}_{R} needs to be orthogonalized to the interference at R in S-R transmission period. Derivations of those matrices can be written as

$$\mathbf{U}_{j}^{[k]} = \operatorname{null}\left(\left[\mathbf{H}_{j,i}^{[k]}\mathbf{V}_{i}^{[k]}\right]^{H}\right), \ j \neq i,$$
(8a)

$$\mathbf{U}_{R} = \operatorname{null}\left(\left[\mathbf{H}_{R,1}\mathbf{V}_{1}^{[1]}\right]^{H}\right).$$
(8b)

In the 2^{nd} time period, S stays silent, while R establishes its own communication. The design of precoding and interference suppression matrices for this time period can be done by following the same step in (7)-(8).

B. Imperfect CSI

The assumption of perfect CSI in wireless networks is highly idealistic due to channel estimation error. Thus, the following model can be deployed for an imperfect CSI estimation [4]

$$\hat{\mathbf{H}} = \mathbf{H} + \mathbf{E},\tag{9}$$

where $\hat{\mathbf{H}}$ is the observed mismatched channel, $\mathbf{H} \sim \mathcal{CN}(0, \mathbf{I})$ represents the real channel matrix and \mathbf{E} is the error matrix which represents an inaccuracy degree in the estimated CSI. It is also assumed that \mathbf{E} is independent of \mathbf{H} . Considering the signal-to-noise ratio (SNR), θ , \mathbf{E} is described as

$$\mathbf{E} \sim \mathcal{CN}(0, \lambda \mathbf{I}) \text{ with } \lambda = \psi \theta^{-\kappa}, \tag{10}$$

where λ is an error variance, $\kappa \ge 0$ and $\psi > 0$ determine various CSI scenarios. Finally, the real channel matrix, conditioning on $\hat{\mathbf{H}}$, [16], can be described as

$$\mathbf{H} = \frac{1}{1+\lambda}\hat{\mathbf{H}} + \tilde{\mathbf{H}},\tag{11}$$

where $\tilde{\mathbf{H}} \sim \mathcal{CN}(0, \frac{\lambda}{1+\lambda}\mathbf{I})$ is independent of $\hat{\mathbf{H}}$.

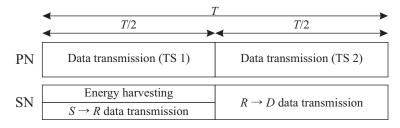


Fig. 2. Time frame structure of PSR.

III. POWER-SPLITTING RELAYING

The PSR for SWIPT is shown in Fig. 2, where the total time is split into two equal portions, one for the S-R and the rest for R-D data transmissions [?]. Within the 1st time fraction, an energy portion of ρ , with $0 < \rho < 1$, at R is allocated for EH, while the remaining power of $(1 - \rho)$ is conveyed to data transmission purpose.

Hence, the R obtains the following signal for EH

$$\mathbf{y}_{R}^{EH} = \sqrt{\frac{\rho P_{S}}{d_{R,S}^{\tau_{R,S}}}} \mathbf{H}_{R,S} \mathbf{V}_{S} \mathbf{s}_{S} + \sum_{i=1}^{2} \sqrt{\frac{\rho P_{i}}{d_{R,i}^{\tau_{R,i}}}} \mathbf{H}_{R,i} \mathbf{V}_{i}^{[1]} \mathbf{s}_{i} + \sqrt{\rho} \mathbf{n}_{R}.$$
 (12)

The power harvested from the noise is insignificant which can be neglected. Thus, the instantaneous harvested energy at R can be derived from (12) as [7]

$$P_{R} = \eta \rho \left(\frac{P_{S}}{d_{R,S}^{\tau_{R,S}}} \left\| \left| \mathbf{H}_{R,S} \mathbf{V}_{S} \right| \right|^{2} + \sum_{i=1}^{2} \frac{P_{i}}{d_{R,i}^{\tau_{R,i}}} \left\| \left| \mathbf{H}_{R,i} \mathbf{V}_{i}^{[1]} \right| \right|^{2} \right),$$
(13)

where $|| \cdot ||$ denotes the Euclidean norm. Then, by using (11), the received information signal with power $(1 - \rho)$ at R can be derived as (14) at the top of the next page.

The corresponding signal-to-interference-noise ratio (SINR) for R from (14) is derived by

$$\mathbf{y}_{R}^{IT} = \sqrt{1-\rho} \mathbf{U}_{R}^{H} \left(\sqrt{\frac{P_{S}}{d_{R,S}^{\tau_{R,S}}}} \left(\frac{1}{1+\lambda} \hat{\mathbf{H}}_{R,S} + \tilde{\mathbf{H}}_{R,S} \right) \mathbf{V}_{S} \mathbf{s}_{S} + \sum_{i=1}^{2} \sqrt{\frac{P_{i}}{d_{R,i}^{\tau_{R,i}}}} \left(\frac{1}{1+\lambda} \hat{\mathbf{H}}_{R,i} + \tilde{\mathbf{H}}_{R,i} \right) \mathbf{V}_{i}^{[1]} \mathbf{s}_{i} + \mathbf{n}_{R} \right)$$
(14)

following

$$\gamma_{R} = \frac{\frac{P_{S}(1-\rho)}{d_{R,S}^{\tau_{R,S}}(1+\lambda)^{2}} ||\mathbf{U}_{R}^{H}\hat{\mathbf{H}}_{R,S}\mathbf{V}_{S}||^{2}}{\frac{P_{S}(1-\rho)}{d_{R,S}^{\tau_{R,S}}} ||\mathbf{U}_{R}^{H}\tilde{\mathbf{H}}_{R,S}\mathbf{V}_{S}||^{2} + I_{PN} + \sigma_{\tilde{n}_{R}}^{2}},$$
(15)

where $I_{PN} = \frac{P_i(1-\rho)}{d_{R,i}^{TR,i}} \sum_{i=1}^2 ||\mathbf{U}_R^H \tilde{\mathbf{H}}_{R,i} \mathbf{V}_i^{[1]}||^2$ defines the interference from primary transmitters.

Then, the received signal at D can be written as

$$\mathbf{y}_{D} = \sqrt{\frac{P_{R}}{d_{D,R}^{\tau_{D,R}}}} \left(\frac{1}{1+\lambda}\hat{\mathbf{H}}_{D,R} + \tilde{\mathbf{H}}_{D,R}\right) \mathbf{V}_{R}\mathbf{s}_{R} + \mathbf{n}_{D}.$$
 (16)

and SINR from (16) can be derived as

$$\gamma_D = \frac{\frac{P_R}{d_{D,R}^{\tau_{D,R}}(1+\lambda)^2} ||\hat{\mathbf{H}}_{D,R} \mathbf{V}_R||^2}{\frac{P_R}{d_{D,R}^{\tau_{D,R}}} ||\tilde{\mathbf{H}}_{D,R} \mathbf{V}_R||^2 + \sigma_D^2},$$
(17)

where σ_D^2 is the noise power.

Also, the received SINR for R_j is shown as

$$\gamma_{j}^{[k]} = \frac{\frac{P_{j}}{d_{j,j}^{\tilde{\tau}_{j,j}}(1+\lambda)^{2}} ||\mathbf{U}_{j}^{[k]H}\hat{\mathbf{H}}_{j,j}^{[k]}\mathbf{V}_{j}^{[k]}||^{2}}{B^{[k]} + C^{[k]} + \sigma_{\tilde{n}_{j}}^{2}},$$
(18)

where the intra-network interference of the PN due to the CSI mismatch is given by $B^{[k]} = \frac{P_j}{d_{j,j}^{\tau_{j,j}}} ||\mathbf{U}_j^{[k]H} \tilde{\mathbf{H}}_{j,j}^{[k]} \mathbf{V}_j^{[k]}||^2 + \frac{P_i}{d_{j,i}^{\tau_{j,i}}} ||\mathbf{U}_j^{[k]H} \tilde{\mathbf{H}}_{j,i}^{[k]} \mathbf{V}_i^{[k]}||_{i\neq j}^2$ while the inter-network interference from the SN is expressed by

$$C^{[k]} = \begin{cases} \frac{P_S}{d_{j,S}^{\tau_{j,S}}} ||\mathbf{U}_j^{[k]H} \tilde{\mathbf{H}}_{j,S} \mathbf{V}_S||^2, & \text{if } k = 1, \\ \frac{P_R}{d_{j,R}^{\tau_{j,R}}} ||\mathbf{U}_j^{[k]H} \tilde{\mathbf{H}}_{j,R} \mathbf{V}_R||^2, & \text{if } k = 2. \end{cases}$$
(19)

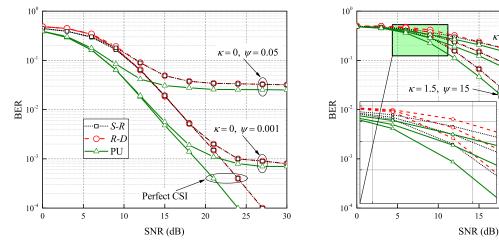
The BER of symbol \mathbf{s}_m for binary phase shift keying (BPSK) can be derived as [17]

$$BER_m = Q(\sqrt{\gamma_m}), \ m \in \{1, 2, R, D\},$$
 (20)

where $Q(x) = 1/\sqrt{2\pi} \int_x^\infty \exp(-t^2/2) dt$ is the Gaussian-Q function.

IV. SIMULATION RESULTS

This section presents the simulation results for the proposed system model in Rayleigh fading channels with BPSK modulation. The system parameters are as follows: $d_{m,n} = 3$ and $\tau_{m,n} =$ 2.7 $\forall m \in \{1, 2, R, D\}, \forall n \in \{1, 2, S, R\}$ and equal transmit power at T_i and S. The calculated optimal values of $\rho = 0.75$ with $\eta = 0.8$ in [18] is considered. Furthermore, the following values of (κ, ψ) such as (0, 0.001), (0, 0.05), (0.75, 10), (1, 10) and (1.5, 15) are used to investigate the impact of CSI mismatch.



(a) BER performance for perfect CSI and SNR-independent CSI mismatch ((0, 0.001) and (0, 0.005)).

(b) BER performance for SNR-dependent CSI mismatch ((0.75, 10), (1, 10) and (1.5, 15)).

15

 $\kappa = 0.75, \ \psi = 10$ 10

 $\cdots \Box \cdots R - D$ O - S - R

20

·PU

25

30

Fig. 3. BER vs. SNR of the primary user, the relay and the destination node operating in the PSR protocol under different CSI scenarios.

Fig. 3 shows how imperfect CSI parameters impact on the BER performance of the PU and SUs. For the case of SNR-independent CSI mismatch (when $\kappa = 0$, Fig. 3a), the BER degrades as ψ increases because the channel error variance does not depend on the SNR and the BER curves saturate after some SNR values, e.g. at 15 dB and 21 dB for 0.05 and 0.001, respectively. Furthermore, it is worth to note that the BER performance is not affected by ψ in the low SNR region, i.e. ψ starts playing a role at 3 dB and 6 dB for the BER of PU and SUs, respectively. This can be explained by the fact that small values of ψ do not increase much the error rate of the received signal at low SNR.

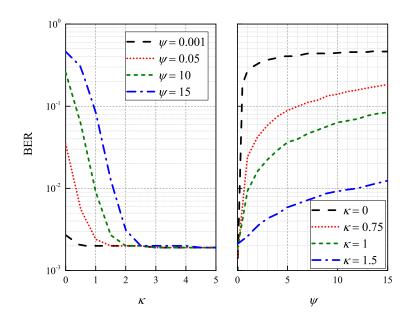


Fig. 4. BER vs. the CSI mismatch parameters κ and ψ of SU at 20 dB.

In general, BER performance of PU outperforms those of R and D because of the power portion $1 - \rho$ devoted for data transmission at R. The less the value of the allocated power for data transmission is, the higher the probability of the incorrect data detection becomes. To compare the performance of R and D at low SNR, R performs better than D because R transmits its data with a certain number of errors which in turn affects the error rate of D. However, the BER performances of SUs match after 10 dB due to the fact that R harvests more energy at high SNR and D consequently receives a strong signal to detect. When $\kappa \neq 0$, the channel error variance becomes SNR-dependent (see Fig. 3b), which implies no saturation of the BER performance. An increase of κ leads to the BER improvement. At 30 dB, the BER performance of SUs obtains 0.0353, 0.0079 and 0.0005 for (0.75, 10), (1, 10) and (1.5, 15), respectively.

A more deeper analysis on the impact of κ and ψ on the BER performance can be obtained from Fig. 4, where the BER performance of SUs is built up as a function of different values of κ and ψ at 20 dB. It can be noticed that the BER performance improves as κ increases, while an increase of ψ results in the BER degradation. In the first subfigure, the BER curves for different values of ψ approach 0.0019 at certain κ values. Meanwhile, in the second subfigure, the BER performance for different κ degrades as ψ increases. It is observed that small values of κ correspond to more abrupt BER degradation, and vise versa.

V. CONCLUSION

In this paper, we analyzed the BER performance of EH-based DF CRN with PS relaying protocols and embedded IA technique. The five special scenarios of the imperfect CSI given by (0, 0.001), (0, 0.05), (0.75, 10), (1, 10) and (1.5, 15) were studied to analyze the impact of the CSI quality on the BER performance of PU and SUs. The presented results with $\rho = 0.75$ showed that the BER of PU outperforms those of SUs in perfect and imperfect SCI cases. Moreover, the BER curve degraded as ψ increased while rise of κ leaded to the BER improvement.

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