Energy-Efficient Two-Way Relaying with Multiple Antennas

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***Abstract*—In this paper, we propose various kinds of energy-efficient two-way multi-antenna relaying with simultaneous wireless information and power transfer (SWIPT) and investigate their performance. Specifically, we first consider a two-way relay network where a pair of single-antenna end nodes communicate with each other through a multi-antenna relay node that is energy constrained. This relay node harvests energy from the two end nodes and use the harvested energy for forwarding their information. Three relaying schemes which support the considered network then build on the power splitting-based relaying protocol. The average bit error rates of these schemes are evaluated and compared by computer simulations considering several network parameters, including the number of relay antennas and the power splitting ratio. Such evaluation and comparison provide useful insights into the performance of SWIPT-based two-way multi-antenna relaying.**

Keywords—bit error rate; simultaneous wireless information and power transfer; two-way multi-antenna relaying.

I. INTRODUCTION

Recently, simultaneous wireless information and power transfer (SWIPT) has gained great interest due to its capability to deal with the energy scarcity in energy-constrained wireless networks [1-6]. In the seminal work [1], the fundamental trade-off between information and power transfer in different point-to-point wireless channels was studied. On the other hand, a pair of practical receiver designs for SWIPT, namely power splitting (PS) and time switching (TS), were firstly presented in [2]. Specifically, the PS-based receiver spits the received radio-frequency signal into two streams of different power for harvesting energy and decoding information, whereas the TS-based receiver switches over time between those two operations. The SWIPT has been adopted later in more complicated communication scenarios, including the broadband wireless system [3], the cellular network [4], the interference channel [5], and the relay channel [6]. This paper focuses on the last scenario.

Many works in the literature have been devoted to two-way multi-antenna relaying (without SWIPT) as this approach can not only extend communication range but also improve spectral efficiency. In a basic two-way multi-antenna relay network (see Fig. 1), an intermediate relay node equipped with multiple antennas is used to assist two end nodes in exchanging their information. Nevertheless, application of SWIPT to this kind of network is still in its infancy [7-9]. In [7], the SWIPT-based beamforming design for a multi-antenna relay was considered to maximize the sum rate of its two-way relay network. In [8], the authors presented a three-phase two-way relay network where an energy-constrained multi-antenna relay node harvests energy from a pair of single-antenna source nodes, and presented an optimal power allocation solution. In [9], an optimal joint source and relay beamforming scheme for two-way multi-antenna relay networks with SWIPT was proposed based on the principle of singular value decomposition.

Depending on the nature and complexity of relays, relaying schemes can be classified into two main categories: non-regenerative relaying and regenerative relaying. Non-regenerative relaying generally implies that the relays only amplify their received signals before retransmitting them. Then, it is often referred to as amplify-and-forward (AF) relaying in the literature. Note that all existing works on two-way multi-antenna relaying with SWIPT, i.e., [7-9], are non-regenerative. On the other hand, regenerative relaying requires the relays to change the waveforms and/or the data contents by performing some processing in the digital domain. An example is the decode-and-forward (DF) relay, which receives the data from its immediate predecessor, decodes, re-encodes, and finally retransmits it. To the best of our knowledge, regenerative relaying has not yet been considered in the SWIPT-based two-way multi-antenna relay networks.

In this paper, we consider a two-way relay network in which a pair of single-antenna end nodes communicate with each other through a multi-antenna relay node that is energy constrained. This relay node harvests energy from the two end nodes and use the harvested energy for forwarding their information. Based on a half-duplex relaying protocol, called power splitting-based relaying (PSR) [6], for separate energy harvesting and information processing at the relay node, three multiple-antenna relaying schemes, namely PS-AF, PS-DF, and PS-DF with space-time coding (PS-DF-STC), are designed for the considered network. In the DF-oriented design, network coding (NC) [10] is applied to the end nodes' information that is decoded at the relay node. Moreover, by having multiple antennas at the relay node, STC [11] is used in the PS-DF-STC scheme with the aim of achieving a better end-to-end decoding performance. Unlike the aforementioned works [7-9] which are devoted to analyzing the relevant sum-rate performance, this paper will investigate the average bit error rates (BERs) of the proposed relaying schemes as a function of the number of relay antennas and power splitting ratio.

II. SYSTEM MODEL

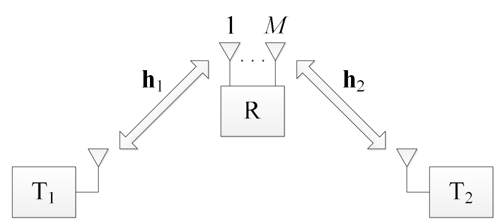

Fig. 1. System model.

Consider a two-way relay network as shown in Fig. 1, where end nodes T_1 and T_2, each of which is equipped with one antenna, exchange information through an energy-constrained intermediate relay node, R, possessing M antennas. This relay node will harvest energy from the two end nodes and use the harvested energy for forwarding their information. The relay node's antennas are spatially spaced in such a way that the received/transmitted signals undergo statistically independent fading. Throughout this paper, perfect timing and synchronization among T_1, T_2, and R are assumed, and binary phase shift keying (BPSK) modulation is used at T_1 and T_2. Let $CN\left(\mu,\sigma^2\right)$ denote a circularly symmetric complex Gaussian random variable with mean μ and variance σ^2, and $h_{1,m} \sim CN\left(0,d_1^{-v}\right)$ (or $h_{2,m} \sim CN\left(0,d_2^{-v}\right)$) denote the channel gain between the antenna of T_1 (or T_2) and the *m*-th antenna of R, where d_1 (or d_2) is the distance of the $\mathrm{T}_1-\mathrm{R}$ link (or the $\mathrm{T}_2-\mathrm{R}$ link), v is the path loss exponent, and $m=1,2,\ldots,M$. We presume that all the channels are static in an interval of $2N$, which denotes the total block time in which a certain block of information is exchanged between T_1 and T_2 (see Fig. 2(a)), and ignore the direct link between the end nodes owing to the larger distance compared with the $\mathrm{T}_1-\mathrm{R}$ and $\mathrm{T}_2-\mathrm{R}$ links.

III. PSR Protocol

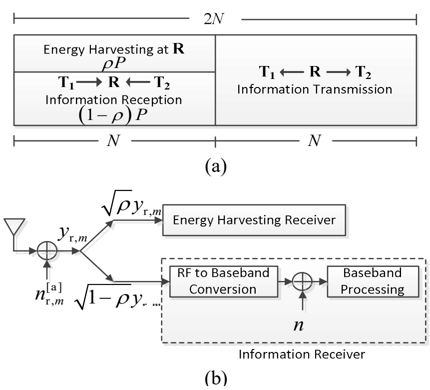

Fig. 2. (a) Key parameter in the PSR protocol for energy harvesting and information processing at the relay node; (b) Block diagram of the relay receiver (with a focus on its *m*-th antenna) in the PSR protocol.

Fig. 2 illustrates the key parameters in the PSR protocol for energy harvesting and information processing at the relay node R and the block diagram of the corresponding receiver. In Fig. 2(a), the first block time N is used for multiple access (MA) where the end nodes T_1 and T_2 transmit their signals simultaneously, and P is the total signal power. In the second block time N, the relay node processes this signal (according to the schemes that will be presented below) and broadcasts it. During the MA phase, the fraction of the received signal power ρP is used for energy harvesting, and the remaining received power $(1-\rho)P$ is used for information transmission, where $0<\rho<1$.

A. PS-DF Scheme

In the MA phase, the received radio-frequency (RF) signal at the relay node can be modeled as

$$\mathbf{y}_{\mathrm{r}} = \begin{bmatrix} y_{\mathrm{r},1} \\ y_{\mathrm{r},2} \\ \vdots \\ y_{\mathrm{r},M} \end{bmatrix} = \begin{bmatrix} \sqrt{P_1}h_{1,1}x_1 + \sqrt{P_2}h_{2,1}x_1 + n_{\mathrm{r},1}^{[\mathrm{a}]} \\ \sqrt{P_1}h_{1,2}x_1 + \sqrt{P_2}h_{2,2}x_2 + n_{\mathrm{r},2}^{[\mathrm{a}]} \\ \vdots \\ \sqrt{P_1}h_{1,M}x_1 + \sqrt{P_2}h_{2,M}x_2 + n_{\mathrm{r},M}^{[\mathrm{a}]} \end{bmatrix} = \begin{bmatrix} \sqrt{P_1}\mathbf{h}_1 & \sqrt{P_2}\mathbf{h}_2 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \mathbf{n}_{\mathrm{r}}^{[\mathrm{a}]} \quad (1)$$

where $P_1=\zeta_1 P$ and $P_2=\zeta_2 P$ are the transmitted power from T_1 and T_2, respectively, $0<\zeta_1,\zeta_2<1$ are the power ratios of T_1 and T_2, respectively (i.e., $\zeta_1+\zeta_2=1$), x_1 and x_2 are the normalized information signals from T_1 and T_2, respectively (i.e., $\mathrm{E}\left[\left|x_1\right|^2\right]=\mathrm{E}\left[\left|x_2\right|^2\right]=1$), and $n_{\mathrm{r},m}^{[\mathrm{a}]} \sim CN\left(0,\sigma_{\mathrm{a}}^2\right)$ is the additive white Gaussian noise (AWGN) at the *m*-th antenna of R. The energy harvesting receiver in Fig. 2(b) rectifies the RF signal $\sqrt{\rho}y_{\mathrm{r},m}$ directly and gets the direct current to charge up the battery. Therefore, the harvested energy at the *m*-th antenna of the relay node during the MA phase is given by

$$E_m = \eta\rho N\left(\zeta_1 P d_1^{-v} + \zeta_2 P d_2^{-v} + \sigma_{\mathrm{a}}^2\right),\ m=1,2,\ldots,M \quad (2)$$

where $0<\eta\le 1$ is the energy conversion efficiency (which depends on the rectification process and the energy harvesting circuitry [6]). Meanwhile, the information receiver in Fig. 2(b) down-converts the RF signal $\sqrt{1-\rho}y_{\mathrm{r},m}$ to baseband and processes the baseband signal, where $n_{\mathrm{r},m}^{[\mathrm{c}]} \sim CN\left(0,\sigma_{\mathrm{c}}^2\right)$ is the AWGN due to RF-band-to-baseband signal conversion. After down conversion, the sampled baseband signal vector at the relay node is given by

$$\begin{aligned} \tilde{\mathbf{y}}_{\mathrm{r}} &= \sqrt{1-\rho}\mathbf{y}_{\mathrm{r}} + \mathbf{n}_{\mathrm{r}}^{[\mathrm{c}]} \\ &= \begin{bmatrix} \sqrt{(1-\rho)P_1}\mathbf{h}_1 & \sqrt{(1-\rho)P_2}\mathbf{h}_2 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \left(\sqrt{(1-\rho)}\mathbf{n}_{\mathrm{r}}^{[\mathrm{a}]} + \mathbf{n}_{\mathrm{r}}^{[\mathrm{c}]}\right) \\ &= \mathbf{\Psi}\begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \mathbf{n}_{\mathrm{r}}. \end{aligned} \quad (3)$$

Assuming that $\rho, P_1, P_2, \left\{h_{1,m}\right\}_{m=1}^{M}$, and $\left\{h_{2,m}\right\}_{m=1}^{M}$ are known at the relay node and applying zero-forcing (ZF) detection, estimates of x_1 and x_2, denoted by $\hat{x}_1$ and $\hat{x}_2$ respectively, are obtained as

$$\begin{bmatrix} \hat{x}_1 \\ \hat{x}_2 \end{bmatrix} = \left(\mathbf{\Psi}^{\mathrm{H}}\mathbf{\Psi}\right)^{-1}\mathbf{\Psi}\tilde{\mathbf{y}}_{\mathrm{r}}. \quad (4)$$

The relay node then performs NC of $\hat{x}_1$ and $\hat{x}_2$ at bit level to obtain the composite signal. Specifically, let $\hat{b}_i = \mathrm{demod}(\hat{x}_i)$ be the estimated information bit sequence corresponding to $\hat{x}_i$, where

$i = 1, 2$ and $\mathrm{demod}(\cdot)$ is the demodulation function. The composite signal is given by $x_{\mathrm{r}} = \mathrm{mod}(\hat{b}_1 \oplus \hat{b}_2)$ where $\mathrm{mod}(\cdot)$ and $\oplus$ denote the modulation function and the bit-wise XOR operator, respectively. As in [6], we assume that the processing power required by the transmit/receive circuitry at the relay node is negligible as compared to the power used for transmitting the composite signal in the broadcast (BC) phase. From (2), the latter power is given by

$$P_{\mathrm{r}} = \frac{\sum_{m=1}^{M} E_m}{N} = \eta\rho M\left(\zeta_1 P d_1^{-v} + \zeta_2 P d_2^{-v} + \sigma_{\mathrm{a}}^2\right) \qquad (5)$$

and the sampled received (baseband) signal at the end node T_i $(i = 1, 2)$ in the BC phase can be expressed as

$$y_i = \sqrt{P_{\mathrm{r}}} \sum_{m=1}^{M} h_{i,m} x_{\mathrm{r}} + n_i^{[\mathrm{a}]} + n_i^{[\mathrm{c}]} \qquad (6)$$

where $n_i^{[\mathrm{a}]} \sim CN\left(0, \sigma_{\mathrm{a}}^2\right)$ and $n_i^{[\mathrm{c}]} \sim CN\left(0, \sigma_{\mathrm{c}}^2\right)$ are the AWGN due to the antenna and that due to RF-band-to-baseband signal conversion, respectively. Assuming that $\left\{h_{i,m}\right\}_{m=1}^{M}$ is known at T_i, an estimate of x_{r} is obtained as

$$\hat{x}_{\mathrm{r}} = \frac{y_i}{\sum_{m=1}^{M} h_{i,m}} = \sqrt{P_{\mathrm{r}}} x_{\mathrm{r}} + \frac{n_i^{[\mathrm{a}]} + n_i^{[\mathrm{c}]}}{\sum_{m=1}^{M} h_{i,m}}. \qquad (7)$$

At the end node T_i, the intended signal x_j $(j = 1, 2;\ j \neq i)$ can be finally recovered by performing bit-level network decoding of $\hat{x}_{\mathrm{r}}$ with its own signal x_i.

B. PS-DF-STC Scheme

For the MA phase, the description of the signal transmissions from the end nodes T_1 and T_2 to the relay node R can be done as in the PS-DF scheme, i.e., (1)-(3). The aforementioned ZF estimation and bit-level NC also follow. However, instead of transmitting the same composite bit sequence $b_{\mathrm{r}} := \hat{b}_1 \oplus \hat{b}_2$ simultaneously via M antennas in the BC phase, the relay node performs space-time block coding [11] for this sequence, as outlined in [12]. Specifically, let $\mathbf{B}$ be the space-time block-coded composite bit matrix whose dimension is $M \times L$, where L is the block length of the corresponding space-time block code. If N consecutive composite bits, i.e., $b_{\mathrm{r}}[k], b_{\mathrm{r}}[k+1], ..., b_{\mathrm{r}}[k+N-1]$, are transmitted with this matrix, then the code rate is N/L. In this paper, we concentrate on the space-time block-coded composite bit matrices with the full code rate, i.e., $L = N$. Such matrices for two, three, and four antennas are shown in Table I. As a result, the sampled received (baseband) signal at the end node T_i $(i = 1, 2)$ in the BC phase can be expressed as

$$\begin{aligned} &\left[y_i[k] \quad y_i[k+1] \quad \cdots \quad y_i[k+N-1]\right] \\ &= \sqrt{P_{\mathrm{r}}}\left[x_{\mathrm{r}}[k] \quad x_{\mathrm{r}}[k+1] \quad \cdots \quad x_{\mathrm{r}}[k+N-1]\right]\mathbf{H} \\ &\quad + \left[n_i[k] \quad n_i[k+1] \quad \cdots \quad n_i[k+N-1]\right] \end{aligned} \qquad (8)$$

where $\mathbf{H}$ is exemplified in Table I, and $n_i \sim CN\left(0, \sigma_{\mathrm{a}}^2 + \sigma_{\mathrm{c}}^2\right)$

TABLE I. EXAMPLES OF $\mathbf{B}$ AND $\mathbf{H}$ FOR SPACE-TIME CODING.

M	N	$\mathbf{B}$	$\mathbf{H}$
2	2	$\begin{bmatrix} b_{\mathrm{r}}[k] & -b_{\mathrm{r}}[k+1] \\ b_{\mathrm{r}}[k+1] & b_{\mathrm{r}}[k] \end{bmatrix}$	$\begin{bmatrix} h_{i,1} & h_{i,2} \\ h_{i,2} & -h_{i,1} \end{bmatrix}$
3	4	$\begin{bmatrix} b_{\mathrm{r}}[k] & -b_{\mathrm{r}}[k+1] & -b_{\mathrm{r}}[k+2] & -b_{\mathrm{r}}[k+3] \\ b_{\mathrm{r}}[k+1] & b_{\mathrm{r}}[k] & b_{\mathrm{r}}[k+3] & -b_{\mathrm{r}}[k+2] \\ b_{\mathrm{r}}[k+2] & -b_{\mathrm{r}}[k+3] & b_{\mathrm{r}}[k] & b_{\mathrm{r}}[k+1] \end{bmatrix}$	$\begin{bmatrix} h_{i,1} & h_{i,2} & h_{i,3} & 0 \\ h_{i,2} & -h_{i,1} & 0 & h_{i,3} \\ h_{i,3} & 0 & -h_{i,1} & -h_{i,2} \\ 0 & -h_{i,3} & h_{i,2} & -h_{i,1} \end{bmatrix}$
4	4	$\begin{bmatrix} b_{\mathrm{r}}[k] & -b_{\mathrm{r}}[k+1] & -b_{\mathrm{r}}[k+2] & -b_{\mathrm{r}}[k+3] \\ b_{\mathrm{r}}[k+1] & b_{\mathrm{r}}[k] & b_{\mathrm{r}}[k+3] & -b_{\mathrm{r}}[k+2] \\ b_{\mathrm{r}}[k+2] & -b_{\mathrm{r}}[k+3] & b_{\mathrm{r}}[k] & b_{\mathrm{r}}[k+1] \\ b_{\mathrm{r}}[k+3] & b_{\mathrm{r}}[k+2] & -b_{\mathrm{r}}[k+1] & b_{\mathrm{r}}[k] \end{bmatrix}$	$\begin{bmatrix} h_{i,1} & h_{i,2} & h_{i,3} & h_{i,4} \\ h_{i,2} & -h_{i,1} & -h_{i,4} & h_{i,3} \\ h_{i,3} & h_{i,4} & -h_{i,1} & -h_{i,2} \\ h_{i,4} & -h_{i,3} & h_{i,2} & -h_{i,1} \end{bmatrix}$

includes the antenna and signal-conversion AWGNs at the corresponding time instants. Following [12] and assuming that $\{h_{i,m}\}_{m=1}^{M}$ are known at T_i, an estimate of $\left[x_{\mathrm{r}}[k] \quad x_{\mathrm{r}}[k+1] \quad \cdots \quad x_{\mathrm{r}}[k+N-1]\right]$ can be obtained as

$$\begin{aligned} &\left[\hat{x}_{\mathrm{r}}[k] \quad \hat{x}_{\mathrm{r}}[k+1] \quad \cdots \quad \hat{x}_{\mathrm{r}}[k+N-1]\right] \\ &= \mathrm{Re}\left[y_i[k] \quad y_i[k+1] \quad \cdots \quad y_i[k+N-1]\right]\mathbf{H}^{\mathrm{H}} \end{aligned} \qquad (9)$$

where $(\cdot)^{\mathrm{H}}$ denotes the Hermitian of a matrix. At the end node T_i, the intended signal x_j $(j = 1, 2;\ j \neq i)$ can be finally recovered by performing bit-level network decoding of $\hat{x}_{\mathrm{r}}$ with its own signal x_i.

C. PS-AF Scheme

For the MA phase, the description of the signal transmissions from the end nodes T_1 and T_2 to the relay node R can be done as in the PS-DF scheme, i.e., (1)-(3). In the BC phase, the relay node amplifies and forwards the information signal as

$$\mathbf{z}_{\mathrm{r}} = \frac{\tilde{\mathbf{y}}_{\mathrm{r}}}{\left\|\tilde{\mathbf{y}}_{\mathrm{r}}\right\|} \qquad (10)$$

where $\left\|\tilde{\mathbf{y}}_{\mathrm{r}}\right\| = \sqrt{(1-\rho)\left(P_1\left\|\mathbf{h}_1\right\|^2 + P_2\left\|\mathbf{h}_2\right\|^2 + M\sigma_{\mathrm{a}}^2\right) + M\sigma_{\mathrm{c}}^2}$ and the sampled received (baseband) signal at the end node T_i $(i = 1, 2)$ is given by

$$\begin{aligned} y_i &= \sqrt{P_{\mathrm{r}}}\mathbf{h}_i^{\mathrm{T}} \frac{\tilde{\mathbf{y}}_{\mathrm{r}}}{\left\|\tilde{\mathbf{y}}_{\mathrm{r}}\right\|} + n_i^{[\mathrm{a}]} + n_i^{[\mathrm{c}]} \\ &= \frac{\sqrt{(1-\rho) P_{\mathrm{r}} P_1}}{\left\|\tilde{\mathbf{y}}_{\mathrm{r}}\right\|}\mathbf{h}_i^{\mathrm{T}}\mathbf{h}_1 x_1 + \frac{\sqrt{(1-\rho) P_{\mathrm{r}} P_2}}{\left\|\tilde{\mathbf{y}}_{\mathrm{r}}\right\|}\mathbf{h}_i^{\mathrm{T}}\mathbf{h}_2 x_2 \\ &\quad + \frac{\sqrt{(1-\rho) P_r}}{\left\|\tilde{\mathbf{y}}_{\mathrm{r}}\right\|}\mathbf{h}_i^{\mathrm{T}}\mathbf{n}_{\mathrm{r}}^{[\mathrm{c}]} + n_i^{[\mathrm{a}]} + n_i^{[\mathrm{c}]} \end{aligned} \qquad (11)$$

where $(\cdot)^{\mathrm{T}}$ denotes the transpose of a matrix, and $n_i^{[\mathrm{a}]}$ and $n_i^{[\mathrm{c}]}$ are defined below (6). Assuming that $\rho, P_{\mathrm{r}}, \{h_{1,m}\}_{m=1}^{M}, \{h_{2,m}\}_{m=1}^{M}$, and $\left\|\tilde{\mathbf{y}}_{\mathrm{r}}\right\|$ are known at T_i, an estimate of the intended signal $x_j\,(j = 1, 2;\ j \neq i)$ can be obtained as

$$\hat{x}_j = \frac{y_i - \sqrt{(1-\rho) P_{\mathrm{r}} P_i}\mathbf{h}_i^{\mathrm{T}}\mathbf{h}_i x_i \Big/ \left\|\tilde{\mathbf{y}}_{\mathrm{r}}\right\|}{\sqrt{(1-\rho) P_{\mathrm{r}}}\mathbf{h}_i^{\mathrm{T}}\mathbf{h}_j \Big/ \left\|\tilde{\mathbf{y}}_{\mathrm{r}}\right\|}. \qquad (12)$$

IV. SIMULATION RESULTS AND CONCLUDING REMARKS

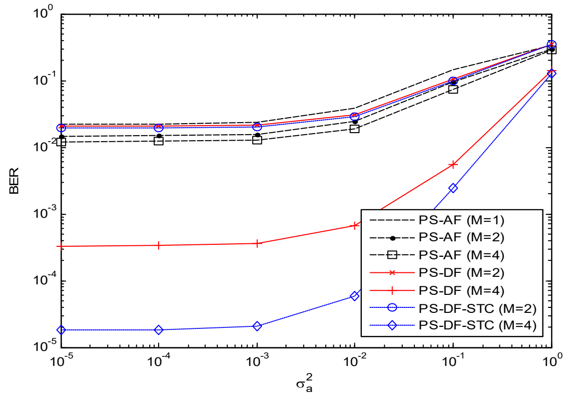

Fig. 3. BER versus σ_a^2 ($\sigma_c^2 = 0.01$ and $\rho = 0.5$).

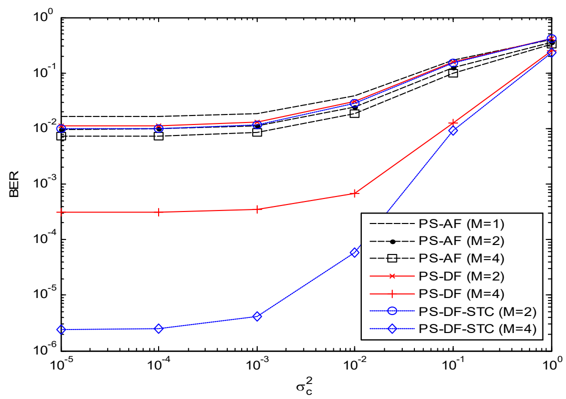

Fig. 4. BER versus σ_c^2 ($\sigma_a^2 = 0.01$ and $\rho = 0.5$).

In this section, we evaluate the performance of the proposed multiple-antenna relaying schemes (i.e., PS-DF, PS-DF-STC, and PS-AF) in terms of average BER of the end nodes T_1 and T_2. Suppose that T_1 and T_2 are separated by a distance of 2 m, and the relay R is located halfway between them. Unless stated otherwise, we set the total signal power, $P = 1$ W, the power ratios of T_1 and T_2, $\zeta_1, \zeta_2 = 0.5$, the path loss exponent, $\nu = 2.7$, the power energy conversion efficiency, $\eta = 1$, and the power splitting ratio, $\rho = 0.5$.

The BERs of the PS-DF, PS-DF-STC, and PS-AF schemes are plotted versus antenna noise variance σ_a^2 for different numbers of relay antennas M (with fixed conversion noise variance $\sigma_c^2 = 0.01$) in Fig. 3 and versus conversion noise variance σ_c^2 for different values of M (with fixed antenna noise variance $\sigma_a^2 = 0.01$) in Fig. 4. To make the BER curves readable in all these figures, the results at $M = 3$ are excluded. It is clear that increasing the number of relay antennas generally improves the BER performance. From Figs. 3 and 4, we can see that the BERs of the PS-DF, PS-DF-STC, and PS-AF schemes are comparable when $M = 2$, and their difference becomes significant when $M = 4$. In the latter case, the PS-DF-STC scheme performs best while the PS-AF scheme does worst.

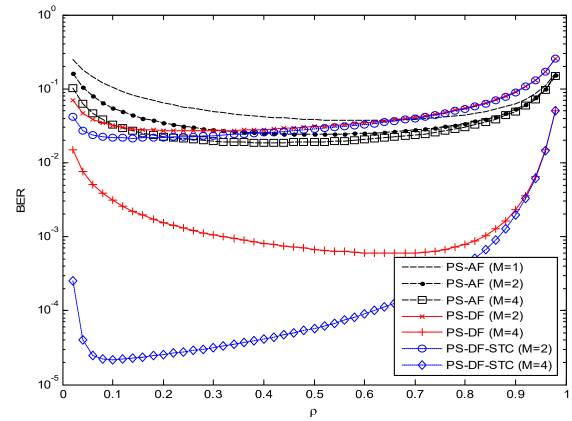

Fig. 5. BER versus ρ ($\sigma_a^2 = 0.01$ and $\sigma_c^2 = 0.01$).

It would be interesting to study the effect of the power splitting ratio ρ on the BER performance. To this end, we show in Fig. 5 the BER as a function of ρ for the PS-DF, PS-DF-STC, and PS-AF schemes. From this figure, we observe that in general, equal power allocation, i.e., $\rho = 0.5$, is a good strategy for the PS-AF scheme. However, the optimal ρ which minimizes the BER of the PS-DF scheme depends mainly on the number of relay antennas. For example, the optimal ρ for the two-antenna PS-DF scheme is approximately 0.26 while that for the four-antenna PS-DF scheme is around 0.65 (See Fig. 5). In addition, the optimal ρ for the two-antenna PS-DF-STC scheme is nearly the same as that for the four-antenna PS-DF-STC scheme.

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