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 built 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.

II. SYSTEM MODEL

![System model](image1)

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 \( \mu \) and variance \( \sigma^2 \), and \( h_{m,n} \sim CN\left(0, d_{m,n}^{-1}\right) \) (or \( h_{m,n} \sim CN\left(0, d_{m,n}^{-1}\right) \)) denote the channel gain between the antenna of \( T_m \) (or \( T_2 \)) and the \( m \)-th antenna of \( R \), where \( d_{m,n} \) (or \( d_{m,n} \)) is the distance between the \( T_m - R \) link (or the \( T_2 - R \) link), \( \psi \) 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 \( T_1 - R \) and \( T_2 - R \) links.

III. PSR PROTOCOL

Fig. 2 illustrates the key parameters 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.

\[ y_i = \begin{bmatrix} y_{i,1} \\ y_{i,2} \\ \vdots \\ y_{i,M} \end{bmatrix} = \begin{bmatrix} \sqrt{P_{1} h_{1,1} x_1 + \sqrt{P_{1} h_{2,1} x_1 + n_{1,1}^{[1]}}} \\ \sqrt{P_{2} h_{1,2} x_1 + \sqrt{P_{2} h_{2,2} x_2 + n_{2,1}^{[1]}}} \\ \vdots \\ \sqrt{P_{M} h_{1,M} x_1 + \sqrt{P_{M} h_{2,M} x_2 + n_{M,1}^{[1]}}} \end{bmatrix} \]

where \( P_i = \zeta_i P \) 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., \( E\left[|x_1|^2\right] = E\left[|x_2|^2\right] = 1 \)), and \( n_{1,1}^{[1]} \sim CN\left(0, \sigma_1^2\right) \) 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 \( \sqrt{\rho y_{i,n}} \) 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

\[ E_m = \eta \rho N \left(\zeta_i P d_{1,m}^{-\psi} + \zeta_2 P d_{2,m}^{-\psi} + \sigma_1^2\right), \quad m = 1, 2, \ldots, M \]

where \( 0 < \eta \leq 1 \) 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 \( \sqrt{\rho y_{i,n}} \) to baseband and processes the baseband signal, where \( n_{1,1}^{[1]} \sim CN\left(0, \sigma_1^2\right) \) 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

\[ y_i = \begin{bmatrix} y_{i,1} \\ y_{i,2} \end{bmatrix} = \begin{bmatrix} \sqrt{(1-\rho) P_{1} h_{1,1} x_1 + \sqrt{(1-\rho) P_{1} h_{2,1} x_1 + n_{1,1}^{[1]}}} \\ \sqrt{(1-\rho) P_{2} h_{1,2} x_2 + \sqrt{(1-\rho) P_{2} h_{2,2} x_2 + n_{2,1}^{[1]}}} \end{bmatrix} \]

Assuming that \( \rho, P_1, P_2, \{h_{1,m}\}_{m=1}^M, \) and \( \{h_{2,m}\}_{m=1}^M \) 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

\[ \begin{bmatrix} \hat{x}_1 \\ \hat{x}_2 \end{bmatrix} = \begin{bmatrix} \Psi \Psi^\dagger \end{bmatrix} \begin{bmatrix} y_i \end{bmatrix} \]

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

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is obtained as \( y = \text{demod}(\cdot) \) is the demodulation function. The composite signal is given by \( x_i = \text{mod}(\hat{b}_i \oplus \hat{b}_j) \) where 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

\[
P_i = \sum_{n=1}^{M} E_{n,i} = \eta \rho M \left( \zeta_i \bar{P} d_{i}^{-1} + \zeta_j \bar{P} d_{j}^{-1} + \sigma^2_i \right)
\]

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

\[
y_i = \sqrt{P} \sum_{n=1}^{N} h_{i,n} x_n + n_i^{[1]} + n_i^{[2]} \tag{6}
\]

where \( n_i^{[1]} \sim \text{CN} \left( 0, \sigma_i^2 \right) \) and \( n_i^{[2]} \sim \text{CN} \left( 0, \sigma_i^2 \right) \) are the AWGN due to the antenna and that due to RF-band-to-baseband signal conversion, respectively. Assuming that \( \{h_{\alpha,i}\}_{\alpha=1}^{4} \) is known at \( T_i \), an estimate of \( x_i \) is obtained as

\[
\hat{x}_i = \frac{y_i}{\sum_{n=1}^{N} h_{i,n}} = \frac{\hat{y}_i}{\sum_{n=1}^{N} h_{i,n}} + n_i^{[1]} + n_i^{[2]} \tag{7}
\]

At the end node \( T_i \), the intended signal \( y_j \) \( (j = 1, 2; j \neq i) \) can be finally recovered by performing bit-level network decoding of \( \hat{x}_i \) 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 \( T_i \) and \( T_j \) 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 \( \tilde{b}_j := \tilde{b}_i \oplus \tilde{b}_j \) 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 \( B \) 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., \( b_j[k], b_j[k+1], \ldots, b_j[k+N-1] \), 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 matrix 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

\[
y_i = \sqrt{P} \left[ x_i[k] x_i[k+1] \cdots x_i[k+N-1] \right] + n_i^{[1]} + n_i^{[2]} \tag{8}
\]

where \( H \) is exemplified in Table I, and \( n_i \sim \text{CN} \left( 0, \sigma_i^2 + \sigma_j^2 \right) \)

\[
\text{TABLE I. EXAMPLES OF B AND H FOR SPACE-TIME CODING.}
\]

<table>
<thead>
<tr>
<th>M</th>
<th>N</th>
<th>B</th>
<th>H</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>2</td>
<td>[ h_1^{[1]} ] [ h_2^{[1]} ]</td>
<td>[ h_1^{[1]} ] [ h_2^{[1]} ]</td>
</tr>
<tr>
<td>3</td>
<td>4</td>
<td>[ h_1^{[1]} ] [ h_2^{[1]} ] [ h_3^{[1]} ]</td>
<td>[ h_1^{[1]} ] [ h_2^{[1]} ] [ h_3^{[1]} ]</td>
</tr>
<tr>
<td>4</td>
<td>4</td>
<td>[ h_1^{[1]} ] [ h_2^{[1]} ] [ h_3^{[1]} ] [ h_4^{[1]} ]</td>
<td>[ h_1^{[1]} ] [ h_2^{[1]} ] [ h_3^{[1]} ] [ h_4^{[1]} ]</td>
</tr>
</tbody>
</table>

includes the antenna and signal-conversion AWGNs at the corresponding time instants. Following [12] and assuming that \( \{h_{\alpha,i}\}_{\alpha=1}^{4} \) are known at \( T_i \), an estimate of \( \tilde{x}_i[k] = x_i[k+1] \cdots x_i[k+N-1] \) can be obtained as

\[
\tilde{x}_i[k] = \frac{y_i[k]}{\text{Re} \left[ \text{det}(y_i) \right]} \left[ x_i[k+1] \cdots x_i[k+N-1] \right]{\text{H}}^H
\]

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

**C. PS-AF Scheme**

For the MA phase, the description of the signal transmissions from the end nodes \( T_i \) and \( T_j \) 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

\[
z_i = \frac{\tilde{y}_i}{\|\tilde{y}_i\|} \tag{10}
\]

where \( \|\tilde{y}_i\| = \sqrt{1-\rho^2} \left( \| \text{det}(y_i) \| + \| \text{det}(H) \| + M \sigma_i^2 \right) + M \sigma_j^2 \)

and the sampled received (baseband) signal at the end node \( T_j \) \( (i = 1, 2) \) is given by

\[
y_j = \sqrt{P} \left[ h_i^{[1]} \tilde{y}_i + n_i^{[1]} + n_i^{[2]} \right] \tag{11}
\]

where \( (\cdot)^T \) denotes the transpose of a matrix, and \( n_i^{[1]} \) and \( n_i^{[2]} \) are defined below (6). Assuming that \( \rho, P, \{h_{\alpha,i}\}_{\alpha=1}^{4}, \{h_{\alpha,i}\}_{\alpha=4}^{4}, \) and \( \|\tilde{y}_i\| \) are known at \( T_i \), an estimate of the intended signal \( x_j \) \( (j = 1, 2; j \neq i) \) can be obtained as

\[
\hat{x}_j = \frac{y_j - \sqrt{1-\rho^2} \text{det}(H) \tilde{y}_i}{\sqrt{1-\rho^2} \|\tilde{y}_i\|} \tag{12}
\]
IV. SIMULATION RESULTS AND CONCLUDING REMARKS

![Fig. 3. BER versus $\sigma^2_i$ ($\sigma^2_i = 0.01$ and $\rho = 0.5$).](image)

![Fig. 4. BER versus $\sigma^2_i$ ($\sigma^2_i = 0.01$ and $\rho = 0.5$).](image)

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

Fig. 4. BER versus $\sigma^2_i$ ($\sigma^2_i = 0.01$ 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 $T_2$, $\xi_1, \xi_2 = 0.5$, 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 $\sigma^2_i$ for different numbers of relay antennas $M$ (with fixed conversion noise variance $\sigma^2_c = 0.01$) in Fig. 3 and versus conversion noise variance $\sigma^2_c$ for different values of $M$ (with fixed antenna noise variance $\sigma^2_i = 0.01$) 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 $\rho$ ($\sigma^2_i = 0.01$ and $\sigma^2_c = 0.01$).](image)

Fig. 5. BER versus $\rho$ ($\sigma^2_i = 0.01$ and $\sigma^2_c = 0.01$).

It would be interesting to study the effect of the power splitting ratio $\rho$ on the BER performance. To this end, we show in Fig. 5 the BER as a function of $\rho$ 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 $\rho$ which minimizes the BER of the PS-DF scheme depends mainly on the number of relay antennas. For example, the optimal $\rho$ 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 $\rho$ 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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