Two-Way Full-Duplex Spatial Modulation Systems With Wireless Powered AF Relaying

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***Abstract*—This letter proposes a dual-hop two-way full-duplex cooperative system, where spatial modulation is used at the sources and energy harvesting is employed at the wireless powered relay. We combine the physical-layer network coding and the fixed-gain amplify-and-forward relaying techniques for two-way transmission. For the simultaneous information and wireless power transfer, the power-splitting protocol is employed at the relay, which has no external power supply. A new unified tight upper-bound bit error rate (BER) expression is derived in a closed-form for both full-duplex and half-duplex transmission schemes. The accuracy of our analyses is verified by various Monte Carlo type computer simulations. The numerical results reveal that the full-duplex scheme can provide considerably better BER performance compared to the conventional half-duplex scheme as the quality of loop-interference cancellation improves and/or the spectral efficiency requirement increases.**

***Index Terms*—Energy harvesting, two-way full-duplex, spatial modulation, amplify-and-forward, power-splitting.**

I. Introduction

THE WIRELESS devices in the conventional communication systems use replaceable or rechargeable batteries, which induce a dependency on the limited battery lifetime and the necessity of an external power supply. Furthermore, the battery replacement operations can be either impossible or dangerous under some circumstances (e.g., military conditions, chemical environment, etc.). In order to prolong the fixed operation time of the batteries, simultaneous wireless information and power transfer (SWIPT) has gained remarkable attention, where the major purpose is to harvest the required energy from the RF signals used for the information transmission [1]. There are two common SWIPT protocols as power-splitting (PS) and time-switching (TS) [2]. The SWIPT technique is also very suitable for the cooperative networks in order to scavenge the energy at the relay node, which may not have an external power supply [3].

In order to enhance the capacity of the wireless systems, full-duplex (FD) communication is a promising technique by allowing the simultaneous transmission and reception over the same frequency band [4]. Contrary to the conventional half-duplex (HD) communication, a strong loop-interference (LI) occurs between the transmitter and receiver. However, the experimental results in [4] shows that FD communications can perform better when the quality of the LI cancellation increases. Another promising technique for next-generation wireless technologies is spatial-modulation (SM), which is proposed for multiple-input multiple-output (MIMO) architectures [5]. In SM, the information bits are carried by both the modulated signals and active transmit antenna indices to increase the spectral efficiency. Recently, an FD SM system is studied in [6] in terms of theory and applications (e.g., drone communications), where authors show that this system can reduce the complexity of LI cancellation in MIMO systems.

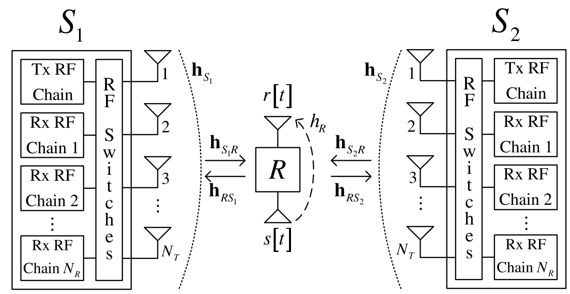

Fig. 1. The system model of FDSM-AF-PS.

In the literature, two-way FD cooperative networks utilizing physical-layer network coding (PLNC) and amplify-and-forward (AF) relaying have been widely investigated, e.g., [7] and [8], where two way transmission can be accomplished in a single time-slot. Recently, SWIPT technique has been also adopted to two-way FD cooperative networks for wireless-powered relays. In [9], the outage probability of two-way wireless powered relay networks has been analyzed for four different protocols. The numerical results show that FD relaying with PS protocol can provide higher throughput for the two-way networks. The cases of single and multiple relay selection for two-way FD relaying using the PS protocol have been investigated in [10], where the outage performance and capacity analyses are provided. Besides, the achievable rate of the SM technique with the TS protocol has been studied in [11].

To the best of authors' knowledge, this letter is the first attempt to apply the SM technique to two-way FD networks, where the relay utilizes fixed-gain AF relaying and the PS protocol for energy harvesting. It is assumed that the assisted relay has no external power supply and the required energy is obtained from the RF signals. So, in this letter, a new energy and bandwidth two-way FD relay system is proposed. We derive a closed-form upper-bound BER expression, which is also valid for the HD transmission scheme. The accuracy of our analyses is verified by comprehensive computer simulation results. We show that FD transmission can provide

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comprehensive better BER performance as the spectral efficiency and the quality of LI cancellations increases.

II. System Model

The system model of the proposed two-way FD SM system with AF relaying and PS (FDSM-AF-PS) consisting of two sources ($S_i,\ i \in \{1,2\}$) and one relay (R) is shown in Fig. 1. We assume that there is no direct link between the sources and they can only communicate via the relay. The sources convey their information by using the SM technique. For the FDSM-AF-PS system, all nodes operate in FD mode and two-way transmission is completed in a single time-slot. So, residual LI channels occur at both the relay and sources.

Each source is equipped with N_T antennas that can either operate as transmit or receive antenna by using the RF switches. Due to the SM transmission, only one antenna is utilized as the transmitter among N_T possible antennas in each symbol period, hence, it is required to place a single transmit RF chain at the source. The remaining $N_T - 1$ antennas can be used as receive antennas. Similar to [12], we place N_R receive RF chains ($1 \le N_R \le N_T - 1$) instead of using all remaining antennas in reception because it could be hard to employ that many receive RF chains for the large number of N_T (e.g., massive MIMO). N_R receive RF chains are arbitrarily connected to $N_T - 1$ available antennas via RF switches. Therefore, there are $N_R + 1$ RF chains at each source node. As seen from Fig. 1, there are one transmit and one receive antenna at the relay using the fixed-gain AF relaying and PS protocols. Although the average amount of the harvested and consumed energy at the relay are equal to each other, the instantaneous output energy can be higher or lower than the harvested energy due to the fixed-gain relaying. Hence, an additional battery is placed into the energy harvesting block, which is either charged or discharged according to the harvested energy and the required energy at the power amplifier. The received signal at the information receiver of the relay node is given by $r_{IR}[t] = \sqrt{\rho} r[t] + n_R[t]$, where

$$r[t] = \sqrt{P_S}\mathbf{h}_{S_1R}\mathbf{x}_1[t] + \sqrt{P_S}\mathbf{h}_{S_2R}\mathbf{x}_2[t] + h_R s[t], \tag{1}$$

$0 \le \rho \le 1$ is the PS ratio, P_S is the transmission power of the sources, $\mathbf{x}_i[t]$ is the $N_T \times 1$ unit-power signal vector transmitted from S_i, $s[t]$ is the transmitted signal from the relay, $\mathbf{h}_{S_iR}$ is the $1 \times N_T$ channel coefficient vector for $S_i - R$ link, h_R is the residual LI channel coefficient at the relay and $n_R[t]$ is the complex Gaussian noise. The entries of $\mathbf{h}_{S_iR}$ are i.i.d. according to $\mathcal{CN}(0, \Omega_i)$. Similarly, h_R and $n_R[t]$ are distributed as $\mathcal{CN}(0, \Omega_R)$ and $\mathcal{CN}(0, \sigma_R^2)$, respectively. Similar to [12] and [13], the average power of the residual LI channel is modeled as $\Omega_R = \beta P_R^{\lambda-1}$ where P_R is the average transmission power of the relay, β and λ are the constants representing the quality of LI cancellation ($0 \le \lambda \le 1$). According to the SM technique, the transmitted signal vector at S_i can be written as

$$\mathbf{x}_i[t] = [\ \cdots\ \ 0\ \ \underbrace{x_i^{q_i[t]}}_{k_i[t]^{th}\text{ position}}\ \ 0\ \ \cdots\ \ 0] \tag{2}$$

where $k_i[t]$ is the selected transmit antenna index at S_i according to the first $\log_2(N_T)$ bits and $x_i^{q_i[t]}$ denotes the $q_i[t]^{th}$ symbol from the M-PSK constellation according to following $\log_2(M)$ bits. Hence, $2\log_2(N_T M)$ bits are conveyed bidirectionally in the proposed systems in a single time-slot. By using (1) and (2), the received signal at the information receiver can be rewritten as $r_{IR}[t] = \sum_{i=1}^{2} \sqrt{P_S\rho} h_{S_iR}^{k_i[t]} x_i^{q_i[t]} + \sqrt{\rho} h_R s[t] + n_R[t]$, where $h_{S_iR}^{k_i[t]}$ is the $k_i[t]^{th}$ element of $\mathbf{h}_{S_iR}$. The transmitted signal from the relay is given by

$$s[t] = G r_{IR}[t-1] = G \sum_{u=0}^{\infty} \left(G\sqrt{\rho} h_R\right)^u q[t-u] \tag{3}$$

where $q[t] = \sum_{i=1}^{2} \sqrt{P_S\rho} h_{S_iR}^{k_i[t-1]} x_i^{q_i[t-1]} + n_R[t-1]$ and G is the amplification factor. In order to adjust the average transmission power as P_R, the fixed amplification factor has to be chosen as $G = \sqrt{\frac{P_R}{P_S\rho(\Omega_1+\Omega_2) + P_R\rho\Omega_R + \sigma_R^2}}$. Although the instantaneous transmission power of the relay varies according to time, by using (3), one can easily show that the average power consumption at the relay node is equal to $\mathrm{E}\{|s[t]|^2\} = P_R$. Hence, the average residual LI power in (1) is found as $\mathrm{E}\{|h_R s[t]|^2\} = \beta P_R^{\lambda}$. The average transmission power of the relay should be determined according to the harvested energy with the PS protocol. The energy harvesting operation is utilized by using $r_{EH}[t] = \sqrt{1-\rho}\, r[t]$. Then, the harvested energy during each symbol period is given by [9]

$$E_{EH}[t] = \varphi(1-\rho)\left(\sum_{i=1}^{2} P_S \left|h_{S_iR}^{k_i[t]}\right|^2 + \frac{E_{EH}[t-1]}{T}|h_R|^2\right)T \tag{4}$$

where $0 \le \varphi \le 1$ is the energy conversion efficiency and T is the symbol period. By assuming $\varphi(1-\rho)\Omega_R < 1$, the average harvested energy at the relay node can be obtained as $\mathrm{E}\{E_{EH}[t]\} = \frac{\varphi(1-\rho)P_S(\Omega_1+\Omega_2)T}{1-\varphi(1-\rho)\Omega_R}$. Afterwards, the average transmission power of the relay can be expressed as

$$P_R = \frac{\mathrm{E}\{E_{EH}[t]\}}{T} = \frac{\varphi(1-\rho)P_S(\Omega_1+\Omega_2)}{1-\varphi(1-\rho)\beta P_R^{\lambda-1}}. \tag{5}$$

To the best of our knowledge, unfortunately, it is not possible to solve (5) for P_R analytically due to the above intractable expression. However, it can be easily evaluated via common mathematical software. Then, the received signal at S_i is given by

$$\mathbf{y}_i[t] = \underbrace{\mathbf{h}_{RS_i} G\sqrt{P_S\rho} h_{S_jR}^{k_j[t-1]} x_j^{q_j[t-1]}}_{\text{Desired Signal}} + \underbrace{\mathbf{h}_{RS_i} G\sqrt{P_S\rho} h_{S_iR}^{k_i[t-1]} x_i^{q_i[t-1]}}_{\text{Removed by PLNC}} + \underbrace{\mathbf{h}_{RS_i} G\sqrt{\rho} h_R s[t-1] + \sqrt{P_S}\mathbf{h}_{S_i} x_i^{q_i[t-1]}}_{\text{Residual Loop-Interference}} + \underbrace{\mathbf{h}_{RS_i} G n_R[t-1] + \mathbf{n}_{S_i}[t]}_{\text{Noise}} \tag{6}$$

where $j \in \{1,2\}$, $j \ne i$, $\mathbf{h}_{RS_i}$ is the $N_R \times 1$ channel coefficient vector between the transmit antenna of the relay and the selected receive antennas of S_i, $\mathbf{h}_{S_i}$ is the $N_R \times 1$ residual LI channel coefficient vector between the active transmit and receive antennas at S_i and $\mathbf{n}_{S_i}[t]$ is the complex Gaussian noise vector. The entries of $\mathbf{h}_{RS_i}$, $\mathbf{h}_{S_i}$ and $\mathbf{n}_{S_i}[t]$ are respectively distributed according to $\mathcal{CN}(0, \Omega_i)$, $\mathcal{CN}(0, \Omega_S)$ and $\mathcal{CN}(0, \sigma_S^2)$. Similar to the relay, the average power of the residual LI channel at S_i is modeled as $\Omega_S = \beta P_S^{\lambda-1}$. As in [9], it is assumed that S_i has the knowledge of its own transmitted signal and the channel coefficients for PLNC. After applying PLNC, the resulting signal can be written as $\hat{\mathbf{y}}_i[t] = \mathbf{y}_i[t] - \mathbf{h}_{RS_i} G\sqrt{P_S\rho} h_{S_iR}^{k_i[t-1]} x_i^{q_i[t-1]}$. For given $\mathbf{h}_{S_jR}$ and $\mathbf{h}_{RS_i}$, the expected value and variance of $\hat{\mathbf{y}}_i[t]$ are given by

$$\begin{aligned}\mathrm{E}\left\{\hat{\mathbf{y}}_i[t] \middle| \mathbf{h}_{S_jR}, \mathbf{h}_{RS_i}\right\} &= \mathbf{h}_{RS_i} G\sqrt{P_S\rho} h_{S_jR}^{k_j[t-1]} x_j^{q_j[t-1]} \\ \mathrm{Var}\left\{\hat{\mathbf{y}}_i[t] \middle| \mathbf{h}_{S_jR}, \mathbf{h}_{RS_i}\right\} &= \left\|\mathbf{h}_{RS_i}\right\|^2 G^2 \left\{\rho\Omega_R P_R + \sigma_R^2\right\} \\ &\quad + P_S\Omega_S + \sigma_S^2\end{aligned} \tag{7}$$

By omitting the time index t, the optimal ML detection rule for FDSM-AF-PS system is obtained as $[\hat{k}_j, \hat{q}_j] = \arg\min_{\tilde{k}_j, \tilde{q}_j} \|\hat{\mathbf{y}}_i - \mathbf{h}_{RS_i} G\sqrt{P_S\rho} h_{S_jR}^{\tilde{k}_j} x_j^{\tilde{q}_j}\|^2$, where $\hat{k}_j$ and $\hat{q}_j$ are the determined transmit antenna and M-PSK symbol index.

III. BER Analyses

A theoretical BER expression is derived for the proposed system throughout this section by using the pairwise error probability (PEP) approach. For given $\mathbf{h}_{S_jR}$ and $\mathbf{h}_{RS_i}$, the conditional PEP (CPEP) expression is given by

$$P\left(\mathbf{x}_j \to \hat{\mathbf{x}}_j \middle| \mathbf{h}_{S_jR}, \mathbf{h}_{RS_i}\right) = P\left\{ \left\| \hat{\mathbf{y}}_i - \mathbf{h}_{RS_i} G\sqrt{P_S\rho} h_{S_jR}^{k_j} x_j^{q_j} \right\|^2 \geq \left\| \hat{\mathbf{y}}_i - \mathbf{h}_{RS_i} G\sqrt{P_S\rho} h_{S_jR}^{\hat{k}_j} x_j^{\hat{q}_j} \right\|^2 \right\} = P(\xi_i \geq 0) \quad (8)$$

where $\hat{\mathbf{x}}_j$ is the decided symbol vector when $\mathbf{x}_j$ is transmitted and ξ_i is a Gaussian distributed random variable. By using (7), the expected value and variance of ξ_i for given $\mathbf{h}_{S_jR}$ and $\mathbf{h}_{RS_i}$ can be obtained as

$$\mathrm{E}\left[\xi_i | \mathbf{h}_{S_jR}, \mathbf{h}_{RS_i}\right] = -G\sqrt{P_S\rho}\left\|\mathbf{h}_{RS_i}\right\|^2 \left| h_{S_jR}^{k_j} x_j^{q_j} - h_{S_jR}^{\hat{k}_j} x_j^{\hat{q}_j} \right|^2$$

$$\mathrm{Var}\left[\xi_i | \mathbf{h}_{S_jR}, \mathbf{h}_{RS_i}\right] = 2\left\|\mathbf{h}_{RS_i}\right\|^2 \left| h_{S_jR}^{k_j} x_j^{q_j} - h_{S_jR}^{\hat{k}_j} x_j^{\hat{q}_j} \right|^2 \times \left(\frac{P_R\left\|\mathbf{h}_{RS_i}\right\|^2}{D} + P_S\Omega_S + \sigma_S^2 \right) \quad (9)$$

where $D = \frac{P_R}{G^2(\rho P_R \Omega_R + \sigma_R^2)}$. From (9), (8) can be rewritten as

$$P\left(\mathbf{x}_j \to \hat{\mathbf{x}}_j \middle| \mathbf{h}_{S_jR}, \mathbf{h}_{RS_i}\right) = Q\left(\sqrt{ \frac{\gamma_{S_jR}^{\mathbf{x}_j \to \hat{\mathbf{x}}_j} \gamma_{RS_i}}{\gamma_{RS_i} + D} } \right) \quad (10)$$

where $\gamma_{S_jR}^{\mathbf{x}_j \to \hat{\mathbf{x}}_j} = \frac{P_S |h_{S_jR}^{k_j} x_j^{q_j} - h_{S_jR}^{\hat{k}_j} x_j^{\hat{q}_j}|^2}{2(P_R\Omega_R + \sigma_R^2/\rho)}$ and $\gamma_{RS_i} = \frac{P_R\|\mathbf{h}_{RS_i}\|^2}{P_S\Omega_S + \sigma_S^2}$. One can easily show that $\gamma_{S_jR}^{\mathbf{x}_j \to \hat{\mathbf{x}}_j}$ and γ_{RS_i} are respectively exponentially and chi-squared distributed random variables. Thus, their probability density function (PDF) expressions are given as $f_{\gamma_{S_jR}^{\mathbf{x}_j \to \hat{\mathbf{x}}_j}}(x) = \frac{e^{-x/\bar{\gamma}_{S_jR}}}{\bar{\gamma}_{S_jR}}$ and $f_{\gamma_{RS_i}}(x) = \frac{x^{N_R - 1} e^{-x/\bar{\gamma}_{RS_i}}}{\Gamma(N_R)\bar{\gamma}_{RS_i}^{N_R}}$, where $\bar{\gamma}_{S_jR} = \frac{\sigma_{q_j}^2 P_S \Omega_j}{2(P_R\Omega_R + \sigma_R^2/\rho)}$, $\bar{\gamma}_{RS_i} = \frac{P_R\Omega_i}{P_S\Omega_S + \sigma_S^2}$ and

$$\sigma_{q_j}^2 = \begin{cases} \left| x_j^{q_j} - x_j^{\hat{q}_j} \right|^2, & k_j = \hat{k}_j \\ 2, & k_j \neq \hat{k}_j. \end{cases} \quad (11)$$

By defining $\kappa_i = \frac{\gamma_{S_jR}^{\mathbf{x}_j \to \hat{\mathbf{x}}_j} \gamma_{RS_i}}{\gamma_{RS_i} + D}$ and integrating the CPEP over the PDF of κ_i, the PEP can be written as

$$P\left(\mathbf{x}_j \to \hat{\mathbf{x}}_j\right) = \int_{y=0}^{\infty} Q\left(\sqrt{y}\right) f_{\kappa_i}(y) dy = \int_{y=0}^{\infty} \frac{e^{-y/2} F_{\kappa_i}(y)}{2\sqrt{2\pi y}} dy \quad (12)$$

where $F_{\kappa_i}(\cdot)$ is the cumulative distribution function (CDF) of κ_i. By using [14, eq. (3.471.9)], the CDF of κ_i is obtained as $F_{\kappa_i}(y) = 1 - \frac{2(y\mu_i)^{N_R/2}}{\Gamma(N_R) e^{y/\bar{\gamma}_{S_jR}}} K_{N_R}(2\sqrt{y\mu_i})$, where $\mu_i = \frac{D}{\bar{\gamma}_{S_jR}\bar{\gamma}_{RS_i}}$ and $K_{N_R}(\cdot)$ is the N_R^{th} order modified Bessel function of second kind [14, eq. (8.407.1)]. By substituting the CDF of κ_i into (12) and using [14, eq. (6.643.3)], the PEP expression of the proposed system can be derived as

$$P\left(\mathbf{x}_j \to \hat{\mathbf{x}}_j\right) = \frac{1}{2} - \frac{e^{\frac{\omega_i}{2}} \omega_i^{N_R} \Gamma\left(N_R + \frac{1}{2}\right)}{2\sqrt{2\mu_i}\Gamma(N_R)} W_{-\frac{N_R}{2}, \frac{N_R}{2}}\left(\omega_i^2\right) \quad (13)$$

where $\omega_i = \sqrt{\mu_i/(0.5 + 1/\bar{\gamma}_{S_jR})}$ and $W_{\cdot,\cdot}(\cdot)$ is the Whittaker function [14, eq. (9.220.4)]. Finally, a BER expression for the FDSM-AF-PS system at S_i can be obtained in closed-form by using the union-bound technique and given by

$$P_{b,i} \leq \sum_{k_j=1}^{N_T} \sum_{q_j=1}^{M} \sum_{\hat{k}_j=1}^{N_T} \sum_{\hat{q}_j=1}^{M} \frac{N\left(\mathbf{x}_j \to \hat{\mathbf{x}}_j\right) P\left(\mathbf{x}_j \to \hat{\mathbf{x}}_j\right)}{\log_2(N_T M) N_T M} \quad (14)$$

where $N(\mathbf{x}_j \to \hat{\mathbf{x}}_j)$ represents the number of bit errors associated with the corresponding pairwise error event.

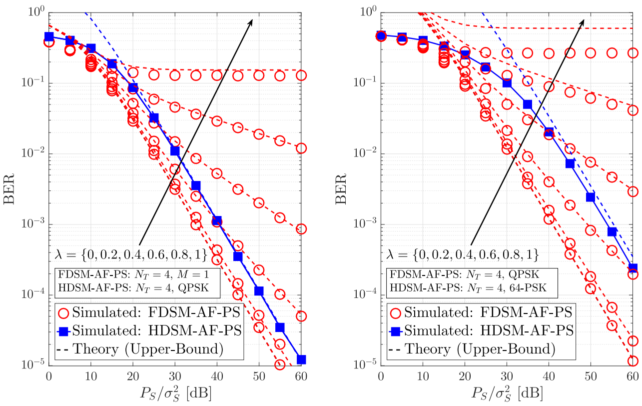
Remark 1: Two-way transmission can be completed also in two time-slots by the two-way HD SM system with AF relaying and PS (HDSM-AF-PS), where all nodes operate in HD mode and the residual LI channels do not exist. Please note that the resulting HDSM-AF-PS system is also novel and when $\beta = 0$, the BER expression given in (14) is valid for this system.

Remark 2: A total of $2\log_2(N_T M)$ bits are exchanged in either one or two time-slots when the FD or HD transmission scheme is adopted, respectively. Hence, the spectral efficiency expressions of FDSM-AF-PS and HDSM-AF-PS systems can be respectively given as [12] $\eta_{FD} = 2\log_2(N_T M)$ and $\eta_{HD} = \log_2(N_T M)$. Above expressions implies that when the HD transmission scheme is used, either M-PSK constellation size or N_T should be increased in order to obtain the same spectral efficiency.

IV. Numerical Results

In this section, theoretical and Monte Carlo simulation results for the BER performance of the FDSM-AF-PS and HDSM-AF-PS systems are represented for various scenarios. The BER curves are plotted according to S_1 (i.e., $P_{b,1}$). It is important to note that the total number of available RF chains are the same for the FD and HD transmission schemes to make a fair comparison. During the numerical analyses, we assume that $\Omega_1 = \Omega_2 = 1$, $\sigma_R^2 = \sigma_S^2$, $T = 1$ and the energy conversion operation handled without any loss (i.e., $\varphi = 1$) as in [3].

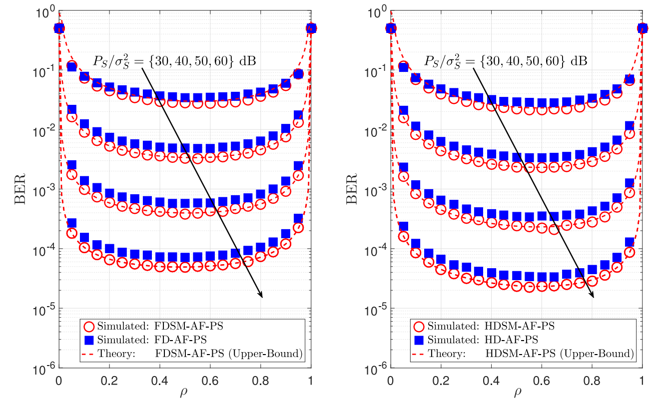
The BER performance comparison of the proposed systems versus P_S/σ_S^2 is plotted in Fig. 2, where $\eta \in \{4, 8\}$ [bits/s/Hz], $N_T = 4$, $N_R = 3$ and $\rho = 0.5$. The LI cancellation parameters for the FDSM-AF-PS system are chosen as $\beta = 0.1$ and $\lambda = \{0, 0.2, 0.4, 0.6, 0.8, 1\}$ similar to [13]. For the FDSM-AF-PS system, a constant-parameter carrier (e.g., $+1$ in baseband) and QPSK signals are utilized, on the other hand, QPSK and 64-PSK signals are used to ensure the same spectral efficiency in the HDSM-AF-PS system. It is shown that the computer simulation results for the FD and HD cases are both tightly upper-bounded by the theoretical results obtained from (14), which proves the accuracy of our analyses. The first important observation is that FDSM-AF-PS can provide better performance with respect to HDSM-AF-PS as long as the quality of LI cancellation increases. For instance, 4.5 dB and 3.4 dB performance improvement can be achieved for $\lambda = 0$



(a) $\eta = 4$ [bits/s/Hz]

(b) $\eta = 8$ [bits/s/Hz]

Fig. 2. BER performance of FDSM-AF-PS with $\beta = 0.1$ and HDSM-AF-PS versus P_S/σ_S^2, where $\eta \in \{4, 8\}$ [bits/s/Hz], $N_T = 4$, $N_R = 3$ and $\rho = 0.5$.



(a) FD transmssion for $\eta = 10$ [bits/s/Hz]

(b) HD transmssion for $\eta = 5$ [bits/s/Hz]

Fig. 3. BER performance comparison for FD transmission with $\beta = 0.1$, $\lambda = 0.2$, $\eta = 10$ [bits/s/Hz] and HD transmission with $\eta = 5$ [bits/s/Hz] versus ρ, where $N_T = 16$ and $N_R = 3$.

and $\lambda = 0.2$, respectively, for $\eta = 4$ [bits/s/Hz] at a BER value of 10^{-3}. Furthermore, the performance improvement of the FD transmission increases for the case of higher spectral efficiency. To illustrate, the performance improvements rise to 12.9 dB and 11.6 dB for $\lambda = 0$ and $\lambda = 0.2$, respectively, for $\eta = 8$ [bits/s/Hz] at a BER value of 10^{-3}. However, for the higher values of λ (e.g., poor LI cancellation), it is convenient to employ HDSM-AF-PS instead of the FD transmission scheme.

In Fig. 3, the BER performance of the FDSM-AF-PS and HDSM-AF-PS systems with BPSK is demonstrated versus ρ, where $P_S/\sigma_S^2 \in \{30, 40, 50, 60\}$ dB, $N_T = 16$ and $N_R = 3$. Therefore, the spectral efficiency of the FD and HD transmission schemes are adjusted to $\eta = 10$ [bits/s/Hz] and $\eta = 5$ [bits/s/Hz], respectively. In order to represent the advantages of the SM technique, we compare the proposed schemes with FD and HD systems employing AF relaying and PS (FD-AF-PS and HD-AF-PS), where all information bits are conveyed via M-PSK signals. In the reference systems, one transmit and three receive antennas are placed for the same number of the RF chains. Additionally, 32-PSK signals are employed to ensure the same spectral efficiency. Here, we assume that $\beta = 0.1$ and $\lambda = 0.2$ for both FDSM-AF-PS and FD-AF-PS in accordance with the experimental results in [4]. As seen from the simulation results, both of the proposed systems using the SM technique can outperform FD-AF-PS and HD-AF-PS. Moreover, we observe that the performance improvement by means of SM technique increases for the higher spectral efficiency. In terms of PS ratio for FDSM-AF-PS, the optimum PS ratio providing the lowest BER value equals $\rho \in \{0.55, 0.53, 0.51, 0.48\}$ when $P_S/\sigma_S^2 \in \{30, 40, 50, 60\}$ dB, respectively. In contrast to the FD transmission scheme, the optimum PS ratio is constant as $\rho = 0.59$ and independent from P_S/σ_S^2 when HDSM-AF-PS is utilized in this scenario.

V. Conclusion

In this letter, we have proposed FDSM-AF-PS and HDSM-AF-PS systems, using SM and PLNC techniques at the sources and employing fixed-gain AF relaying at the relay. There is no external power supply at the relay, so the required energy is obtained from the RF signals via the PS protocol. By using the PEP, a new unified closed-form upper-bound BER expression has been derived for both schemes. The accuracy of our analyses has been verified by computer simulation results. It has been shown that FDSM-AF-PS can perform considerably better than its HD counterpart as the quality of LI cancellation and/or the spectral efficiency increases. Our future work may focus on the optimization of system parameters to enhance the performance.

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