A New Receiver Design: Simultaneous Wireless Power Transfer With Modulation Classification

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Abstract—This work proposes a new simultaneous wireless power transfer and modulation classification (SWPTMC) scheme, appropriate for internet of things (IoT) and military applications. The problem of SWPTMC is investigated for various modulation formats, i.e., quadrature phase-shift-keying (QPSK), 16-pulse amplitude modulation (16-PAM), π/4-QPSK, minimum shift keying (MSK), offset QPSK (OQPSK), and 16-quadrature amplitude modulation (16-QAM). We propose a new receiver architecture that incorporates conventional power splitting under a linear model with a certain level of sensitivity. The blind modulation classification algorithm is based on the higher-order cumulants and cyclic cumulants of the received signal. The cyclic cumulants use the non-zero cycle frequency position, while the higher-order cumulants use threshold values for classifying modulation formats. Monte Carlo simulations are carried out to validate the accuracy of the proposed SWPTMC scheme.

Index Terms—Cyclic cumulants, cumulants, energy harvesting, modulation classification, wireless power transfer.

I. INTRODUCTION

E NERGY-efficient transmission is one of the primary concerns in modern wireless networks, such as wireless sensor networks (WSN) and Internet of Things (IoT) due to the limited lifespan of fixed energy supplies. Wireless power transfer (WPT) through radio-frequency (RF) signals has recently emerged as a candidate technology for providing power to remotely located sensors and IoT devices [1].

The WPT concept is extended to the simultaneous wireless information and power transfer (SWIPT), which allows data and power to be transmitted through the same electromagnetic waveform. SWIPT has recently gained significant attention as an integrated approach for information decoding (ID) and energy harvesting (EH) [2], [3]. Practical SWIPT receivers have been proposed by using time-switching (TS) and power splitting (PS) schemes [4].

Classification of modulation formats is a key process of smart receivers to ensure correct demodulation. It plays a significant role in the military, intelligence, and civilian applications [5]–[9]. The blind modulation classification (MC) algorithms can be divided into two general categories, i.e., likelihood-based (LB) and feature-based (FB) algorithms. The LB algorithms require signal preprocessing tasks and suffer from higher computational complexity [5]. On the other hand, FB algorithms are easier to implement and have lower computational complexity [6]–[9]. The higher-order cumulants and moments-based MC algorithms employ threshold values to classify the modulation formats [6]–[8]. The cyclic cumulants-based methods discussed in [8], [9] are more robust and use non-zero cycle frequencies of the received signals to identify the modulation schemes.

In this paper, we propose a new receiver architecture which simultaneously harvests power and performs blind MC; this joint design introduces a new concept which is called simultaneous wireless power transfer and modulation classification (SWPTMC). We consider the linear harvesting model with a certain level of sensitivity. The proposed MC algorithm can be applied to a wide range of modulations, i.e., quadrature phase-shift-keying (QPSK), 16-pulse amplitude modulation (16-PAM), π/4-QPSK, minimum shift keying (MSK), offset QPSK (OQPSK), and 16-quadrature amplitude modulation (16-QAM) over Rayleigh fading channels. It is based on the combination of higher-order cumulants and second-order cyclic cumulants. The second-order cyclic cumulants use the non-zero cycle frequency as a feature to classify MSK and OQPSK modulation formats. The remaining modulation formats are identified by fourth and eighth-order cumulants. To the best of the authors’ knowledge, this is the first work in the literature focused on SWPTMC scheme.

II. SIGNAL MODEL

We consider a SWPTMC system consisting of a single transmit and receive antenna as shown in Fig. 1. The transmitter sends the modulated signal with an average transmission power of $P_{tx}$ in each time slot. The receiver divides the received signal into two components, i.e., one is used for the blind MC with PS ratio $\rho_1$ and the other one is for EH with power ratio $\rho_2$, where $\rho \in [0,1]$ is the PS factor. The transmitted continuous-time domain passband signal is given as

$$x(t) = \Re\left\{s(t)e^{j(2\pi f_c t + \phi)}\right\}, \quad (1)$$

where $s(t)$ is the carrier phase and $f_c$ is the carrier frequency. The in-phase ($I$) and quadrature ($Q$) components of the transmitted signal $s(t)$ are given as:

$$s_I(t) = \sum_{k=0}^{K-1} a_I(k)g(t-kT)$$

and

$$s_Q(t) = \sum_{k=0}^{K-1} a_Q(k)g(t-kT)$$
\[ s_Q(t) = \sum_{k=0}^{K-1} a_Q(k) g(t - kT), \] respectively, where \( a(k) = a_I(k) + ja_Q(k) \), \( K \) is the number of symbols, \( g(t) \) is the root cosine (RRC) pulse shape filter, and \( T \) is the symbol period. For QPSK, OQPSK, MSK: \( a_I(k), a_Q(k) \in \{\pm 1, \pm \sqrt{2}\} \); \( \pi/4\)-QPSK: \( a_I(k), a_Q(k) \in \{\pm 1, \pm \sqrt{2}/\sqrt{2}\} \); 16-PAM: \( a_I(k) \in \{\pm (2d - 1)/\sqrt{8}\}, a_Q(k) = 0 \), with \( d \in \{1, \ldots, 8\} \); and 16-QAM: \( a_I(k), a_Q(k) \in \{\pm 3/\sqrt{10}, \pm 1/\sqrt{10}\} \). The discrete-time received signal over additive white Gaussian noise (AWGN) and Rayleigh fading can be written as

\[ y[n] = \Re \left\{ \sum_{l=0}^{L-1} h[l] x[n-l] + v[n] \right\}, \quad (2) \]

where \( v[n] \) is the passband AWGN with zero mean and variance of \( \sigma_n^2 \) and \( h[l] = h[0], \ldots, h[L-1] \) is the Rayleigh fading coefficients of a multipath channel with \( L \) taps having zero mean and variance one. The above real signal \( y[n] \) can be expressed as complex signal, i.e., \( \hat{y}[n] = y[n] + j\hat{y}[n] \), where \( \hat{y}[n] \) is the Hilbert transform of \( y[n] \). At the receiver, we consider a linear EH model with a certain level of sensitivity \( \eta \) be expressed as complex signal, i.e., \( \hat{y}[n] = y[n] + j\hat{y}[n] \), where \( \hat{y}[n] \) is the Hilbert transform of \( y[n] \). The amount of harvested power can be expressed as [4]

\[ \mathcal{P}_{\text{eh}} = \frac{\eta (P_{\text{r}} - P_{\text{th}})}{P_{\text{r}}}, \quad P_{\text{r}} \geq P_{\text{th}}; \quad P_{\text{r}} < P_{\text{th}}, \quad (3) \]

where \( \eta \in [0, 1] \) is the RF-EH conversion efficiency and \( P_{\text{th}} \) is the RF-EH sensitivity level.

### III. PROPOSED RECEIVER ARCHITECTURE

The proposed separate receiver splits the received signal into two streams as shown in Fig. 1. One stream for the blind MC with PS ratio \( 1 - \rho \) and the other stream is for EH with PS ratio \( \rho \), described in detail below.

1) Modulation Classification: The received signal at MC circuit is \( y_m[n] = \sqrt{1 - \rho} \hat{y}[n] \). After down conversion, we can obtain the lowpass discrete signal as

\[ r_m[n] = \sqrt{1 - \rho} \sum_{l=0}^{L-1} h[l] x[n-l] + w[n], \quad (4) \]

where \( w[n] \) is the lowpass AWGN with zero mean and variance of \( \sigma_m^2 \). The classification performs in three stages: (a) at the first-stage, we use the second-order cyclic cumulants to identify OQPSK and MSK modulation formats; (b) at the second-stage, \( \pi/4\)-QPSK modulation is identified using the fourth-order cumulant; (c) at the last-stage, we employ eight-order cumulant to identify the remaining modulation formats.

The second-order cyclic cumulant of the baseband signal \( r_m[n] \) for different modulation formats are discussed below.

The Fourier series coefficient of the second-order time-varying correlation function, \( c_{[r_m,2,1]}(\tau) \), is known as cyclic cumulant, and at \( \tau = 0 \) lag can be expressed [9]

\[ C_{[r_m,2,1]}(\alpha; 0) \triangleq \frac{1}{N} \sum_{n=0}^{N-1} c_{[r_m,2,1]}(n; 0)e^{-j2\pi \alpha n}, \quad (5) \]

where \( \alpha \) is the cycle frequency and \( N \) is the received signal length. Once \( C_{[r_m,2,1]}(\alpha; 0) \) is determined, frequency estimation is given by [9]

\[ \hat{f}_b = \arg \max \left| C_{[r_m,2,1]}(\alpha; 0) \right|. \quad (6) \]

From (6), we obtain the second-order cycle frequency for MSK at \( \hat{f}_b = 2f_c \). For QPSK, \( \pi/4\)-QPSK, 16-QAM, and 16-PAM modulation formats, we get the same feature value, i.e., \( \hat{f}_b = f_c \) and there is no peak for OQPSK, \( \hat{f}_b = 0 \) [9], where \( f_c \) is the symbol rate. To classify the remaining modulation formats, we further examine their distinctive features in the second-stage. The fourth-order cumulant with zero conjugations can be expressed as [6]

\[ \hat{c}_{[r_m,4,0]} = \frac{1}{N} \sum_{n=0}^{N-1} r_m^4[n] - 3 \left( \frac{1}{N} \sum_{n=0}^{N-1} r_m^2[n] \right)^2. \quad (7) \]

From (7), we can easily differentiate \( \pi/4\)-QPSK and the rest of modulations as shown in Table I.

| \( \pi/4\)-QPSK | 16-PAM | QPSK | 16-QAM |
|----------------|--------|------|--------|
| 0.00            | 0.68   | 1.00 | 0.68   |

At the final-stage, to identify the remaining modulation formats, we use eight-order cumulant with four conjugations, which can be expressed as [7]

\[ \hat{c}_{[r_m,8,4]} = \frac{1}{N} \sum_{n=0}^{N-1} \left[ r_m[n]^4 - 16 \hat{c}_{[r_m,6,3]}(c_{[r_m,2,1]} - \hat{c}_{[r_m,4,0]})^2 \right. \]

\[-18 \hat{c}_{[r_m,4,2]} - 72 \hat{c}_{[r_m,4,2]} \hat{c}_{[r_m,2,1]}^2 - 24 \hat{c}_{[r_m,2,1]}, \quad (8)\]

where \( \hat{c}_{[r_m,p,q]} = \sum_{\text{items}} \left( \sum_{\text{items}} \cdots \right), \) i.e., \( \hat{c}_{[r_m,4,2]} = \sum_{\text{items}} r_m[n]^4 \), \( \hat{c}_{[r_m,6,3]} = \sum_{\text{items}} r_m[n]^6 \), \( \hat{c}_{[r_m,8,4]} = \sum_{\text{items}} r_m[n]^8 \), and \( \hat{c}_{[r_m,4,2]} = \sum_{\text{items}} r_m[n]^4 r_m[n]^2 \).

From (8), we can easily differentiate 16-PAM, QPSK, and 16-QAM modulation formats as shown in Table II.

| \( \pi/4\)-QPSK | 16-PAM | QPSK | 16-QAM |
|----------------|--------|------|--------|
| -142.00        | -34.00 | -13.98 | -13.98 |

2) Energy Harvesting: The received signal at the EH circuit is \( y_e[n] = \sqrt{\eta} \hat{y}[n] + w_e[n] \), where \( w_e[n] \) is the rectenna circuit AWGN noise with zero mean and variance of \( \sigma_e^2 \). From the received signal \( y_e[n] \), we can find the power for the i-th constellation point as \( P_{s_i} \approx \rho P_{s_i} \), assuming that the power harvested from the passband and circuit noise is negligible, where \( \sigma \) is the fading power gain and \( P_{s_i} \) is the transmit power for the i-th constellation point. By using (3) and averaging at the fading, we get the average harvested power as

\[ \mathcal{P}_{\text{eh}} = \frac{\eta \rho}{M} \sum_{i=1}^{M} P_{s_i} e^{-P_{s_i}/\rho P_{s_i}}, \quad (9) \]
where $M$ is the constellation size of the modulation formats.

Figs. 2 and 3 show the average harvested power ($P_{\text{Eh}}$) and the percentage of correct classification ($P_{cc}$) for the modulation formats considered, as a function of $\rho$ over a Rayleigh fading channel. It is known that a modulation scheme with higher peak-to-average power ratio (PAPR), increases the harvested power [4]. The PAPR of PSK based modulations, 16-QAM, and 16-PAM modulation formats are given as $\psi_{\text{PSK}} = 1$, $\psi_{\text{16-QAM}} = 1.8$, and $\psi_{\text{16-PAM}} = 2.64$, respectively. Hence, the average harvested power follows: $P_{\text{Eh}}^{\text{16-PAM}} > P_{\text{Eh}}^{\text{16-QAM}} > P_{\text{Eh}}^{\text{PSK}}$, as shown in Fig. 2. The classification performance trade-off has been observed in Fig. 3 for $0.5 < \rho \leq 1$. Thus, for $\rho \in [0, 0.5]$ we can simultaneously harvest power with slightly affecting the classifier performance.

V. CONCLUSION

A SWPTMC receiver architecture has been proposed and implemented for different modulation formats over Rayleigh fading channels. The EH circuit uses a linear model with certain level sensitivity and the MC circuit uses cumulants and cyclic cumulants at the baseband level to classify QPSK, 16-PAM, MSK, OQPSK, 16-QAM, and $\pi/4$-QPSK modulation formats. The results highlight that for $0 < \rho \leq 0.5$ we can harvest power without affecting the classifier performance.

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