

Design and Optimization of Wireless Information and Power Transfer Networks

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Statement of Originality

I hereby declare that except where specific reference is made to the work of others, the contents of this dissertation are original and have not been submitted in whole or in part for consideration for any other degree or qualification in this, or any other University. This dissertation is the result of my own work and includes nothing which is the outcome of work done in collaboration, except where indicated in the text.

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January 2018

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Abstract

Energy harvesting from the Radio Frequency (RF) has been identified as one of the solution to alleviate the power constraint in wireless device. RF can carry both information and power. In recent years, the studies of Wireless Information and Power Transfer (WIPT) has attracted significant interest among researchers. WIPT has appeared in multiple forms such as simultaneous wireless information and power transfer (SWIPT), wireless powered communication and wireless powered backscatter communication.

First, we analysed the trade-off between harvested energy and communication reliability in SWIPT in a K-user Orthogonal Frequency Division Multiplexing (OFDM) interference channel. When the transmitters are non-cooperative, it was found that the necessary condition for the optimal transmission strategy at high SNR is for each energy transmitter to transmit its signal on a single subcarrier. Based on one-subcarrier selection method at the energy transmitter, an iterative algorithm has been proposed to optimize the power allocation at all transmitters. Moreover, with a low-rate feedback from the information transmitters to the energy transmitters and with full cooperation among the transmitters respectively, the transmission strategy maximizing the trade-off region has been proposed.

Secondly, we analysed the coordinated waveform design in SWIPT. The waveform was designed by taking into account the the non-linearity of the rectifier. The trade-off between harvested energy and achievable rate was investigated. It was found that deterministic waveform has a better RF-DC efficiency and results in enlarged trade-off region compared to the modulated waveform. Accordingly, the superposition of modulated and deterministic waveform at the information transmitter was proposed. Another observation was that by accounting for the non-linearity of the rectifier, it increases the resource efficiency that leads to a larger trade-off region than those obtained by ignoring the non-linearity of the rectifier.

Finally, backscatter communication has emerged to be a promising technology for power limited devices. In this thesis, we analysed the trade-off between harvested energy and reliability of backscatter communication for wireless powered backscatter system. It was found that the performance of wireless powered backscatter is subject to a trade-off between harvested energy and SINR. The result highlights the importance of addressing the non-linearity of the diode for wireless powered backscatter system.

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Chapter 1

Introduction

Radio Frequency (RF) is a form of electromagnetic wave propagating through space and carrying electromagnetic energy [1], which for decades has been used in wireless information transfer (WIT) such as in TV, radio, cellular phone and Wi-Fi [2]. In the late 19th century, Nikola Tesla made a discovery that power can be transferred between two nodes without a physical connection [3]. Since then, several methods of wireless power transfer (WPT) have been introduced including inductive coupling, laser-based power transmission, near-field magnetic resonance and RF/microwave transmission. RF/microwave and laser are classified as far-field transfer which is suitable for long-range application.

While laser offers stronger power than RF/microwave and non-interference with existing communication systems, laser is least viable for practical purpose due to several significant weaknesses. Laser radiation can be hazardous and it requires direct line-of-sight from laser transmitter to receiver. Laser is also vulnerable to atmospheric absorption caused by dust and water vapor. On the other hand, RF has low safety and health issues [4] and can penetrate most transmission obstacle, making non-line-of-sight transmission possible as occurs in the current Wi-Fi network. With its lack of significant problem, RF is seen as the most promising method of WPT and WIT. In this thesis, our interest is on RF-based Wireless Information and Power Transfer (WIPT).

1.1 Motivation

WPT has emerged as the potential technology due to its benefits and advancement in the research studies. The benefits of WPT include free mobility with the elimination of power cord and charger and ability to deploy devices at inaccessible environment such as in rural areas and concrete structure [5]. WPT also offers longer battery life that reduces the scalability challenge of battery replacement and provides the chance of self-sustainability.

While extensive studies have been conducted on the device side to improve energy harvesting efficiency, the studies on the transmitter side has only been gaining interest recently [5].

Due to high propagation loss in wireless system, WPT has been primarily used for short-range wireless network with low power devices such as RFID/ backscatter [3, 6, 7] and wireless sensors [7, 8]. With the emergence of new wireless technologies such as small cell networks and massive MIMO, it significantly reduce the propagation loss by shortening communication distances and signal dispersion. Therefore, WPT is predicted to be an indispensable technology in the near future, primarily in the upcoming IoT technology where battery replacement is difficult for a large number of interconnected distributed sensors [6]. The IoT is seen to be as one of the next significant technology that connects real-world objects in a wide range of applications such as smart city, smart home, autonomous robots, vehicles and unmanned aerial vehicles (UAVs) [6, 9].

WIPT has appeared in multiple forms such as simultaneous wireless information and power transfer (SWIPT) where simultaneous transmission of power and information from one transmitter to receivers [11], wireless powered communication where the transmitter transmits power to the receiver which is then converted to DC power and reused for information transmission [12] and wireless powered backscatter communication where the transmitter transmits power/unmodulated signal to the backscatter which is then harvest the energy and backscatter the received signal to the reader [13]. The performance of WIPT relies on the ability to design efficient WIT and WPT blocks with limited wireless resources such as transmission power and bandwidth.

The challenge in WIT is to increase communication reliability. Several techniques have been designed to improve the performance of WIT, one of those is multi-carrier transmission. Multi-carrier transmission has been used in wireless system to reduce the effect of frequency selective fading. The robustness of multi-carrier transmission against frequency selective fading helps to improve communication performance [14].

On the other hand, the challenge in WPT is to increase the DC power level (i.e., RF-DC efficiency) at the output of the rectenna. A typical rectenna is made of a non-linear diode followed by a low-pass filter to extract the DC power out of the RF input signal [15]. Major efforts have been made to design efficient rectenna [11]. However, RF-DC efficiency is not only depend on the rectenna design, but it is also a function of the input waveform [16–22]. Interestingly, multi-carrier transmission has also been found to increase RF-DC efficiency in WPT [22].

Designing an efficient WIPT system is not as easy as overlaying a WIT and a WPT systems. WIPT requires redesigning system and network architectures, transmission strategy, resource allocation and transceiver circuit [7]. The main objectives of this research are to

design and optimize the wireless information and power transfer networks. It was shown in [23] that WIPT is subject to a trade-off between achievable rate and harvested energy. Motivated by the promising result of multi-carrier transmission in WPT and WIT networks [14, 22], we leverage multi-carrier excitation to the WIPT network. Moreover, in order to enhance the performance of WIPT, the design of multi-carrier transmission is formulated adaptive to the CSI knowledge available at the transmitter.

In [15], the paper has introduced the linear model of energy harvester. In the linear model, the DC power at the output of the rectenna is proportionate to the input power. The linear model has been used widely in SWIPT researches [24–42]. In this thesis, SWIPT in a K-user interference channel will be studied, the EH is based on the linear model introduced in [15].

Recently, an analytical EH model that accounts for the non-linearity of the rectifier is introduced in [17, 18]. It was observed that by accounting for the non-linearity of the rectifier, it increases RF-DC efficiency [17–20] and leads to a larger trade-off region than those obtained by ignoring the non-linearity of the rectifier [19]. This observation motivates us to study the coordinated waveform design for SWIPT based on non-linear EH model.

Backscatter communication has emerged to be a promising technology for power limited devices. In wireless powered backscatter communication, the device harvests energy from the upcoming RF signal and rely on backscattering modulation to modulate and reflect the signal for its communication with the RFID reader. The limitation of power at the backscatter leads to active research in wireless powered backscatter communication. However, existing studies in waveform optimization does not take into accounts for the non-linearity of the rectifier [13, 67]. This motivates us to study waveform design for wireless powered backscatter communication that accounts for the non-linearity of the rectifier, with multi-carrier excitation and adaptive to the CSI.

1.2 Contributions

The following are the contributions included in this thesis:

- To the best of author's knowledge, the trade-off between the achievable rate and harvested energy has not been studied for SWIPT based on linear EH model in OFDM IFC. First, it was assumed that the transmitters are non-cooperative and optimize their transmission strategy in a distributed manner. From the analysis, it was found that the necessary condition for the optimal transmission strategy at high SNR is for each energy transmitter to transmit its signal on a single subcarrier. Accordingly, an optimal one-subcarrier method has been proposed. Based on the proposed one-subcarrier method, an iterative algorithm has been designed to optimize the power

allocation at the transmitters. Moreover, with a low-rate feedback from the information transmitters, the new transmission strategy has been proposed to further enlarge the trade-off region. Finally, with full cooperation among the transmitters, a transmission strategy maximizing the trade-off between DoF at the information users and total harvested energy has also been proposed.

- To the best of author's knowledge, the coordinated waveform design for wireless information and power transfer accounting for the non-linearity of the rectifier has not been analysed yet. This thesis covers the design of coordinated waveform in a two-user system with one energy user and one information user (with their corresponding transmitter respectively denoted as energy transmitter and information transmitter). In this thesis, the design of multisine waveform with non-linearity of the rectifier model has been addressed. The trade-off between harvested energy and achievable rate for information transfer was investigated. The numerical simulation has shown that the deterministic waveform can enlarge the trade-off region over the modulated waveform. This observation motivates the new design of the superposition of modulated and deterministic waveform at the information transmitter. This new waveform design has a non-zero mean Gaussian distribution [19]. Numerical simulation in this thesis has shown that it can improve the trade-off region. This observation implies that the transmit waveform at the information transmitter should be designed differently from the capacity achieving in the conventional WIT that has a zero mean Gaussian distribution. Moreover, the numerical simulation has also shown that accounting for the non-linearity of the rectifier in the coordinated waveform design increases the resource efficiency that leads to a larger trade-off region than those obtained by ignoring the non-linearity of the rectifier.
- To the best of author's knowledge, the trade-off between harvested energy and the reliability of backscatter communication has not been studied for wireless powered backscatter system. In addition, existing waveform optimization does not take into account the non-linearity of the diode [13, 67]. In this thesis, the multisine transmit waveform at the reader was optimized so as to maximize the harvested energy subject to the transmit power and SINR constraints. From the numerical simulation, it was found that the performance of wireless powered backscatter was subject to a trade-off between harvested energy and SINR. The result also highlight the importance of addressing the non-linearity of the diode for wireless powered backscatter system.

1.3 Publication

The following papers have been published:

- Z. B. Zawawi, J. Park and B. Clerckx, "Simultaneous Wireless Information and Power Transfer in a Two-User OFDM Interference Channel," IEEE ISWCS 2015, Aug. 2015.
- B. Clerckx, Z. B. Zawawi and K. Huang "Wirelessly Powered Backscatter Communications: Waveform Design and SNR-Energy Tradeoff," IEEE Commun. Letters, vol. 21, no. 10, Oct. 2017.

The following paper has been submitted:

- Z. B. Zawawi, Y. Huang and B. Clerckx, "Multi-User Wirelessly Powered Backscatter Communications: Non-linearity, Waveform Design and SINR-Energy Tradeoff," 2018.

1.4 Structure of the Thesis

Chapter 2 of this thesis provides a review of literature on wireless information and power transfer, analytical energy harvesting model, the limitation of energy harvester and interference. Chapter 2 also provides an extensive review of literatures on SWIPT in OFDM and interference channel, waveform design in WPT and SWIPT and waveform optimization in backscattering system.

In Chapter 3, SWIPT in SISO K-user OFDM interference channel (IFC) with the knowledge of local channel state information (CSI) is investigated. There are three different scenarios namely K-ID user, K-EH user and multiple- EH user and multiple-ID user. The transmission strategy for each scenario is discussed. For a system with at least one information decoding (ID) user and one energy harvesting (EH) user, the necessary condition for an optimal transmission strategy that maximize the achievable rate-energy region is proposed. Accordingly, the transmission strategy that satisfy the necessary condition is designed and the achievable rate-energy region is identified. Moreover, the transmission strategy with additional low rate-feedback by the information transmitter is discussed. Chapter 3 also covered the transmission strategy with the knowledge of global CSI.

The coordinated waveform design for SWIPT is considered in Chapter 4. This chapter covers the superposition and non-superposition of deterministic and modulated waveforms. Perfect and no cancellation of interference signal at the information decoding user are considered. The trade-off between achievable rate and energy is analysed. Waveform

optimization is conducted maximizing the harvested energy subject to the transmit power and SINR constraints. The numerical simulation is performed for 1MHz and 10MHz bandwidths.

Chapter 5 studies waveform design for wireless powered backscatter. A point-to-point and multi-user backscatter are considered. The trade-off between harvested energy and backscatter communication is analysed. The optimization problem is formulated maximizing the harvested energy subject to the transmit power and Signal to Interference and Noise Ratio (SINR) constraints for all backscatters. Accordingly, the transmit waveform and receive combiner are optimized. The numerical simulation is then conducted.

Chapter 6 presents the conclusion of this research work and areas that can be further explored in the future.

1.5 Notations

The following notations are used throughout this thesis. A bold capital letter and a lower case letter represents a matrix and a vector, respectively. The notations $(\mathbf{A})^H$, $[\mathbf{A}]_i$, $(\mathbf{A})_i$, $\text{tr}(\mathbf{A})$, $\det(\mathbf{A})$ and $\|\mathbf{A}\|$ represent the conjugate transpose, the i -th column, the i -th row, the trace, the determinant and the 2-norm of a matrix $\mathbf{A}$, respectively. $|a|$ denotes the absolute value of a , a^* represents the conjugate of scalar a , $(a)^+ \triangleq \max(a, 0)$ and $\mathbf{A} \geq \mathbf{0}$ indicates that matrix $\mathbf{A}$ is a positive semi-definite matrix. The matrix with $\text{diag}(a_1, \dots, a_N)$ is a block diagonal matrix with elements a_n and $\mathbf{I}_N$ is $N \times N$ identity matrix. $\mathcal{A}\{.\}$ indicates the DC component of a signal.

Chapter 2

Literature Review

The literature review is divided into four sections as follows: an overview and analytical model of wireless information transfer (WIT) and wireless power transfer (WPT), SWIPT system, waveform design in WPT and SWIPT and backscattering system.

2.1 Wireless Information and Power Transfer

Radio Frequency (RF) is a form of electromagnetic wave propagating through space carrying electromagnetic energy [1]. RF has been widely used in wireless communication system for information transfer. On the other hand, RF has also been used as an energy source in wireless power transfer. In this section, an overview and analytical model of WIT and WPT are provided. Two types of waveform are considered which is modulated and deterministic multisine waveforms.

2.1.1 The Modulated Multisine Waveform

Since modulated waveform can carry both information and power, it can be used for WIT and WPT. In the wireless system, multi-carrier transmission has been used to reduce the effect of frequency selective fading. The robustness of multi-carrier transmission against frequency selective fading allows high data rate transmission [14]. The concept of multi-carrier transmission has been exploited in frequency division multiplexing (FDM) and Orthogonal Frequency Division Multiplexing (OFDM) systems as depicted in Fig. (2.1).

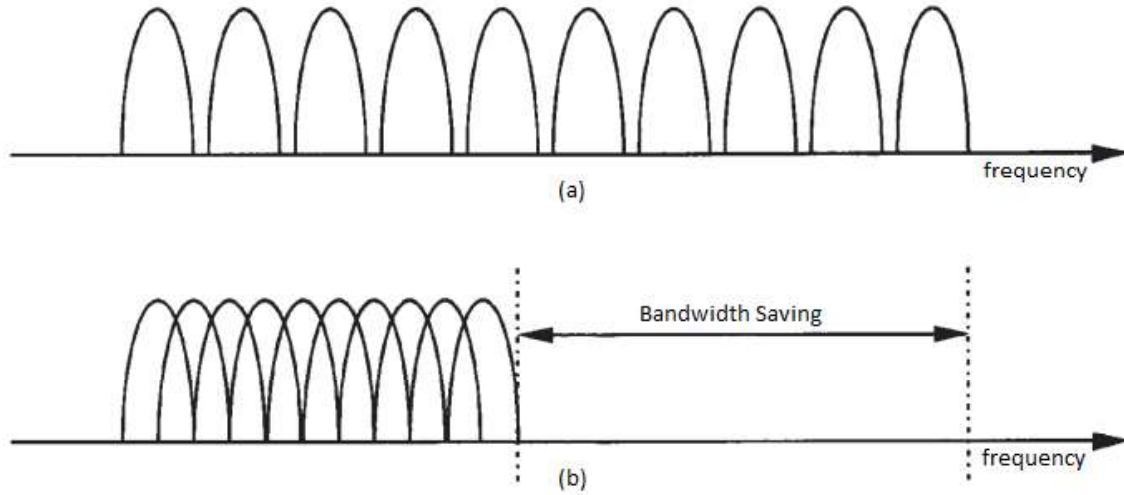


Fig. 2.1 Multicarrier modulation system (a) FDM and (b) OFDM [14].

Frequency Division Multiplexing (FDM)

The classical multi-carrier transmission has been developed in [44, 46] namely frequency division multiplexing (FDM). FDM has been used in TV system, telephone and radio transmission [2]. FDM divides the total bandwidth available into a non-overlapping frequency sub-channel (i.e., subcarrier). Each sub-channel can carry different data. For the non-overlapping sub-channel, each sub-channel are spaced far enough in the frequency domain as depicted in Fig. (2.1 (a)). Hence, each sub-channel can transfer information without interfering other sub-channel.

Orthogonal Frequency Division Multiplexing (OFDM)

In the wireless communication system, the transmitter can transfer information through multisine waveform based on OFDM system. OFDM system has been used in various wireless standards such as Wireless LAN (WLAN) and 3GPP Long Term Evolution (LTE)[63]. The frequency spacing Δ_w in OFDM system is designed such that the sub-channel are orthogonal and overlapped as depicted in Fig. (2.2). Accordingly, each sub-channel can transfer information without causing interference to other sub-channel.

As depicted in Fig. (2.1 (b)), due to the absence of guard interval between sub-channels in OFDM, OFDM can provide better spectral efficiency and high data rate over FDM [48]. In the wireless system, multi-path channel can caused inter-symbol interference (ISI). OFDM system can mitigate the inter-symbol interference by transmitting long symbol period such that ISI occurs at a small fraction of time [49]. Therefore, cyclic prefix (i.e., guard interval)

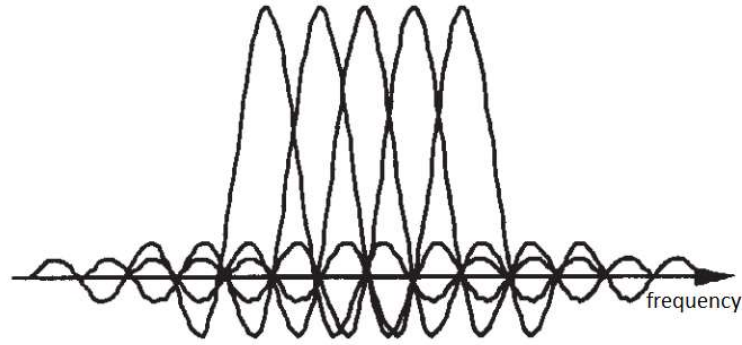


Fig. 2.2 Orthogonal sub-channel [14].

can be inserted between two consecutive symbols to eliminate the ISI. Cyclic prefix is a prefix of the symbol with a copy of the end. The cyclic prefix length should be larger than the maximum delay of multipath channel to eliminate ISI [50].

Moreover, cyclic prefix is used because it can provides simple frequency-domain equalization [51]. The convolution-multiplication property mention that the multiplication in the frequency domain corresponds to the cyclic convolution in the time domain [52]. The cyclic prefix converts the linear convolution of a multipath channel into circular convolution [52]. Therefore, simple equalization can be used to extract the transmitted data symbols [51].

However, OFDM may suffer high Peak to Average Power Ration (PAPR) due to the overlapping of many subcarrier signals. High PAPR can reduce the power efficiency of RF amplifier [14].

Next, the model of OFDM system is discussed as follows. The input data is first mapped into modulation symbol by using some modulation constellation such as QAM and PSK. Then, the data is demultiplexed into N parallel streams $\tilde{x}_n$ by using a serial-to-parallel port. Here, $\tilde{x}_n = \tilde{s}_n e^{j\tilde{\phi}_n}$ is the information symbol transmitted on subcarrier n . A capacity achieving has a Gaussian input distribution such that $\tilde{x}_n$ is i.i.d circularly symmetric complex Gaussian distributed with zero mean [54]. Then, inverse FFT is performed to convert the frequency domain data to the OFDM symbol in time domain given by

$$\bar{x}(t) = \sum_n \tilde{x}_n e^{j2\pi nt/T}, \quad 0 \leq t < T \quad (2.1)$$

where T is the OFDM symbol duration given as $T = \frac{2\pi}{\Delta_w}$. T must be the reciprocal of Δ_w to maintain the orthogonality of the subcarrier.

In order to avoid the inter-symbol interference due to the multipath propagation, the cyclic prefix is inserted to the OFDM block. The signal after adding the cyclic prefix with



Fig. 2.3 Wireless system with EH/ID receiver.

duration T_g is given by

$$\tilde{x}(t) = \sum_n \tilde{x}_n e^{j2\pi nt/T}, \quad -T_g \leq t < T \quad (2.2)$$

This signal is finally up-converted to a carrier frequency and transmitted given by

$$x(t) = \Re\{\tilde{x}(t)e^{j\omega_0 t}\} = \Re\left\{\sum_{n=0}^{N-1} \tilde{x}_n e^{j\omega_n t}\right\} = \sum_{n=0}^{N-1} \tilde{s}_n \cos(\omega_n t + \tilde{\phi}_n). \quad (2.3)$$

The transmit signal for WIT is subject to the transmit power constraint (at the passband) given by $\mathcal{E}\{|x|^2\} = \frac{1}{2}\mathcal{E}\{\|\tilde{\mathbf{s}}\|_F^2\} \leq P$.

2.1.2 The Deterministic Multisine Waveform

The WPT waveform does not carry random information and is thus can be transmitted by using deterministic signal. Let us consider N multisine waveform, the transmit RF waveform at time t can be written as

$$x(t) = \sum_{n=0}^{N-1} s_n \cos(\omega_n t + \phi_n) = \Re\left\{\sum_{n=0}^{N-1} w_n e^{j\omega_n t}\right\}, \quad (2.4)$$

where $w_n = s_n e^{j\phi_n}$. ω_n is the frequency component given as $\omega_n = \omega_0 + n\Delta_w$ with $\Delta_w = 2\pi\Delta_f$ frequency spacing. Here, s_n and ϕ_n are the amplitude and the phase of n -th sinewave at frequency ω_n . The magnitudes and phases in Eq. (2.4) can be collected into vectors $\mathbf{s}$ and $\boldsymbol{\phi}$, respectively. The transmit waveform is subject to the transmit power constraint (at the passband) P , $\mathcal{E}\{|x|^2\} = \frac{1}{2}\|\mathbf{s}\|^2 \leq P$.

2.1.3 The Receive Signal

The transmitted signal of Eq. (2.1) and Eq. (2.4) propagate through the wireless channel to the receiver. After multipath propagation, the received signal for OFDM waveform can be

written as

$$y(t) = \Re\left\{\sum_{n=0}^{N-1} h_n \tilde{x}_n e^{j\omega_n t}\right\} = \sum_{n=0}^{N-1} \tilde{s}_n A_n \cos(\omega_n t + \tilde{\psi}_{j,n}) \quad (2.5)$$

and for WIT can be written as

$$y(t) = \Re\left\{\sum_{n=0}^{N-1} h_n w_n e^{j\omega_n t}\right\} = \sum_{n=0}^{N-1} s_n A_n \cos(\omega_n t + \psi_{j,n}). \quad (2.6)$$

Here, $h_n = A_n e^{j\tilde{\psi}_{j,n}} = \sum_{l=0}^{L-1} \alpha_l \cos(\omega_n \tau_l + \varepsilon_l)$ is the frequency response of the channel at frequency component n , $\psi_n = \tilde{\psi}_n + \phi_n$ and $\tilde{\psi}_n = \tilde{\psi}_n + \tilde{\phi}_n$. τ_l , ε_l and α_l are the delay, phase and amplitude of the l -th path, respectively. The antenna noise at n -th sinewave v_n is usually modelled as a circularly symmetric complex Gaussian noise with zero mean and variance σ_n^2 . The received signal for modulated OFDM waveform is given by

$$y(t) = \Re\left\{\sum_{n=0}^{N-1} (h_n \tilde{x}_n + v_n) e^{j\omega_n t}\right\} \quad (2.7)$$

and for deterministic waveform is given by

$$y(t) = \Re\left\{\sum_{n=0}^{N-1} (h_n w_n + v_n) e^{j\omega_n t}\right\}. \quad (2.8)$$

2.1.4 Information Receiver

At the information decoding receiver, we extract the information transmitted through the wireless channel from the received OFDM signal $y(t)$ in Eq. (2.7). The received OFDM signal $y(t)$ is first down-converted to form a baseband signal given by

$$y(t) = \sum_{n=0}^{N-1} (h_n \tilde{x}_n + v_n) e^{j2\pi n t/T}. \quad (2.9)$$

Then, the cyclic prefix is removed and FFT operation is performed on the time-domain signal to obtain the individual frequency components. Let assume that the cyclic prefix is long enough to avoid the ISI. The SISO received signal at n -th subcarrier is given as

$$y_n = h_n \tilde{x}_n + v_n, \quad (2.10)$$

where v_n is the zero mean complex additive white Gaussian (AWGN) noise. The achievable rate of point-to-point SISO is given by [54]

$$R = \sum_{n=1}^N \log_2 \left(1 + \frac{s_n^2 |h_n|^2}{\sigma_n^2} \right). \quad (2.11)$$

The deterministic multisine signal does not contain information such that when the transmitter transmits deterministic signal as in Eq. (2.4), $R = 0$.

2.1.5 Energy Receiver

At the energy harvesting receiver, the received RF signal is converted into direct current (DC) by using a rectenna. A rectenna is a combination of antenna and rectifier circuit. A typical rectifier circuit consists of a Schottky diode and a low-pass filter (LPF) [15]. The Schottky diode is usually used due to its high switching capacity that enables it to follow the high-frequency input signal [45].

The current $i_d(t)$ flowing through the diode is given by $i_d(t) = i_s(e^{\frac{v_d(t)}{nv_t}} - 1)$ where $v_d(t) = v_{in}(t) - v_{out}(t)$ is the voltage drop across the diode, i_s is the reverse bias saturation current, v_t is the thermal voltage and n is the ideality factor. By using the Taylor expansion of the exponential function around a fixed voltage drop $v_d = a$, the diode current can be written as

$$i_d(t) = \sum_{i=0}^{\infty} k_i (v_d(t) - a)^i, \quad (2.12)$$

where $k_0 = i_s(e^{\frac{a}{nv_t}} - 1)$, $k_i = i_s(\frac{e^{\frac{a}{nv_t}}}{i!(nv_t)^i})$ for $i = 1, \dots, \infty$.

Assuming a steady state response, the output voltage delivered to the load is a constant v_{out} that is the largest peaks of the input voltage $v_{in}(t)$. Since $\mathcal{E}\{v_{in}(t)\} = 0$ and $a = \mathcal{E}\{v_d(t)\} = \mathcal{E}\{v_{in}(t) - v_{out}\} = -v_{out}$, (2.12) can then be written as

$$i_d(t) = \sum_{i=0}^{\infty} k_i v_{in}(t)^i. \quad (2.13)$$

2.1.6 Linear Energy Harvesting Model

In this section, we discuss the linear model of energy harvester as introduced in [15]. The non-linear model of energy harvester will be discussed in Section (2.3.1).

Since $y(t)$ is practically small, the higher terms in Eq. (2.13) does not contribute much to the DC current. By ignoring the terms higher than 2 and because an input voltage is

proportional to $y(t)$ [15], optimizing Eq. (2.13) is equivalent to optimizing

$$i_d'(t) = \sum_{i=0}^2 k_i y(t)^i. \quad (2.14)$$

Since $\mathcal{E}\{y(t)\} = 0$ and k_0 is not a function of $y(t)$, the DC current at energy receiver can be written as

$$i_{DC}(t) = k_2 \mathcal{E}\{y(t)^2\}. \quad (2.15)$$

The harvested energy from the noise is typically small and thus assumed to be negligible. k_2 is the diode characteristic and from [15], for simplification k_2 is assumed to be 1. In addition, it is assumed that the converted energy to be stored in the battery is linearly proportional to (2.15) with a conversion efficiency $0 < \zeta \leq 1$ [15]. Accordingly, the harvested power (specifically, energy normalized by the baseband symbol period) can be written as

$$E = \zeta \mathcal{E}[y(t)^2] = \zeta \sum_n p_n h_n. \quad (2.16)$$

Without loss of generality, it is normally assumed that $\zeta = 1$.

From Eq. (2.16), the maximum harvested power for modulated OFDM and deterministic waveform is given by

$$E = 2P(\max_n \{|h_n|^2\}). \quad (2.17)$$

It is noted here that, the modulated and deterministic waveforms have the same RF-DC conversion efficiency. It is also noted that a single frequency component strategy maximizes the harvested energy for both modulated and deterministic waveforms.

2.1.7 Energy Harvester Circuit Limitation

Due to the energy harvester circuit limitation, the receiver is not able to decode information and harvest energy simultaneously [15]. As a result, various studies have investigated SWIPT with time switching [24–27, 33–35] and power splitting [15, 27] at the receiver as shown in Figure 2.4 and Figure 2.5. In time switching, the receiver either decode the information or harvest the energy at one time. In power splitting, the received signal is split into two streams such that one stream for information decoding and the other stream is for energy harvesting. Power splitting incurs higher receiver complexity over time switching and has additional optimization parameter of power splitting ratio [57].

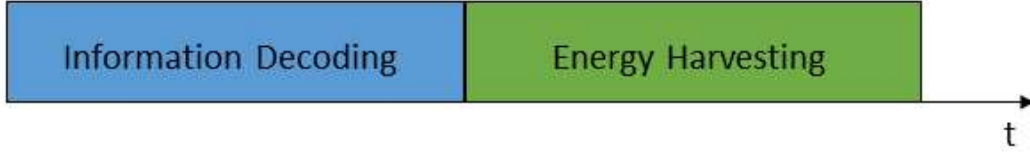


Fig. 2.4 Time Switching.

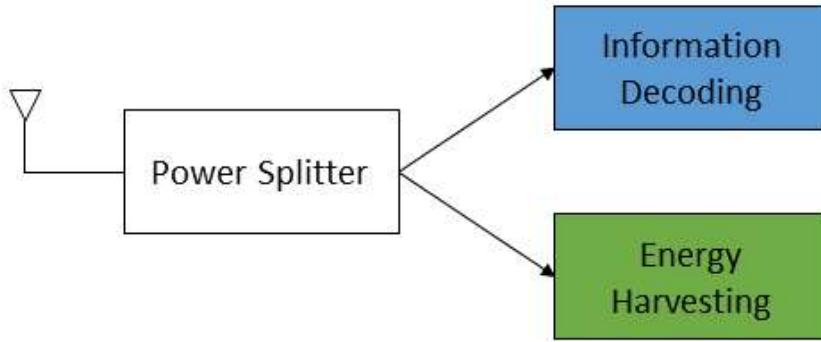


Fig. 2.5 Power Splitting.

2.2 Simultaneous Wireless Information and Power Transfer

SWIPT architecture consists of the fundamental block of WIT and WPT. SWIPT performance depends on the design of WIT and WPT blocks. Channel capacity is the limit of achievable rate for reliable communication. A capacity achieving has a Gaussian input distribution such that $\tilde{x}_n$ is i.i.d circularly symmetric complex Gaussian distributed with zero mean [54]. The challenge in WIT is to maximize the achievable rate.

On the other hand, the challenge in WPT is to increase the DC power level at the output of the rectifier. A typical rectifier consists of non-linear diode and low-pass filter to extract the DC power out of the RF input signal. Major effort has been made for efficient design of rectifier [11, 16]. However, the RF-DC efficiency is not only depend on the rectifier design, but it is also a function of the input waveform [16–18]. This has attracted the interest in the design and optimization of SWIPT networks.

The trade-off between achievable rate and harvested energy has been studied in [59] for the binary symmetric and additive white Gaussian noise channels. It was found that the performance of SWIPT is subject to a trade-off between maximizing the information rate and harvested energy. It was also observed in [23] for the frequency selective channel, the

power allocation based on the water-filling algorithm maximized the information rate, while harvested energy is maximized by transmitting at a single frequency. Similar observation has also been observed in [32] for the multiple-input multiple-output (MIMO) system. In MIMO system, the information rate is maximized by allocating the power based on the water-filling algorithm over several eigenvalues of the channel. On the other hand, the harvested energy is maximized by focusing the power into one direction corresponding to the maximum eigenvalue of the channel. Further researches have been exploited in Multiple-Input and Multiple-Output (MIMO) [32–35], relay [64, 65], multisine/OFDM [24–27] and IFC [33–39] systems.

Overwhelming majority of the SWIPT researches have used the linear EH model as discussed in Section (2.1.6). Some works on SWIPT based on the linear EH model and multisine/OFDM signal will be discussed next. The non-linear EH model will be considered later in Section (2.3).

2.2.1 SWIPT with Linear EH Model in OFDM System

In this section, we provide the state-of-the-art of SWIPT in OFDM system relying on the linear EH model. SWIPT with multisine/OFDM signal has been studied in [24–27].

[24] has studied the downlink and uplink transmission in a single and multi-user OFDM system. During downlink, the base station transmit RF waveform carrying the information signal. At the receiver, power splitting is performed to decode the information and harvest the energy. During uplink, the user and the base station transfer information and power, respectively. The transmitter is equipped with multiple antenna and the transmit beamforming vector is assumed to be given. The works focus on optimizing the transmit power allocation and power splitting ratio such that the achievable rate is maximized with given transmit power and minimum receive power.

In [25], SWIPT in a multiuser OFDM system is discussed. Each transmitter and receiver is equipped with a single antenna. Two transmission schemes, time division multiple access (TDMA) and orthogonal frequency division multiple access (OFDMA) are considered. Due to the EH circuit limitation, the receiver is not able to decode information and harvest energy simultaneously. Accordingly, the receiver performs time switching and power splitting for TDMA and OFDMA schemes, respectively. The time/power splitting ratio and power allocation across time or across frequency for TDMA and OFDMA, respectively are optimized such that the weighted sum-rate is maximized for given transmit power and energy constraint. Accordingly the achievable rate-energy region is obtained.

The work in [26] consider a single user OFDM system. The transmitter and receiver is equipped with a single antenna. The receiver performed power splitting to decode information

and harvest energy. The power allocation and power splitting ratio are optimized such that the achievable rate is maximized for given transmit power constraint and minimum received power. The interference power from other transmitter has been considered in the achievable rate. However, the harvested power from interference signal and the interference signal to the other user are ignored.

[27] consider a multiuser OFDM system. The transmitter and the receivers are equipped with a single antenna. The receiver employs power splitting to decode information and harvest received RF power. The resource allocation is optimized such that the energy efficiency is maximized with given transmit power, minimum received power and rate constraints. Here, the energy efficiency parameter is defined as the ratio of the achievable rate and power consumption of the system. Continuous and fixed power splitting scheme are considered in which the power splitter splits the received power with arbitrarily and discrete levels, respectively.

2.2.2 Interference channel

An Interference Channel (IFC) occurs when a system with multiple transmitters transmit their signal via a shared channel. Consequently, the intended received signal from the corresponding transmitter interferes with the unintended received signal from the other transmitters. The un-decodable interference signal can cause disruption and degrades the achievable rate of information transfer. In the presence of interference, the achievable rate at j -th receiver is given by [47]

$$R_j = \sum_{n=1}^N \log_2 \left(1 + \frac{p_{j,n} |h_{jj,n}|^2}{\sigma_n^2 + \sum_{k \neq j} p_{k,n} |h_{jk,n}|^2} \right) \quad (2.18)$$

where $p_{k,n}$ is the power allocation at the k -th transmitter and n -th subcarrier, $h_{jk,n}$ is the channel between k -th transmitter and j -th receiver and $p_{k,n} |h_{jk,n}|^2$ is the interference power from k -th transmitter at n -th subcarrier.

In contrast to WIT, interference signal increase the DC power at the EH receiver as it provides additional source of RF power for energy harvesting. In the presence of interference signal, the total harvested energy at j -th EH receiver is given by

$$E_j = \sum_{k=1}^K \sum_n p_{k,n} h_{jk,n}. \quad (2.19)$$

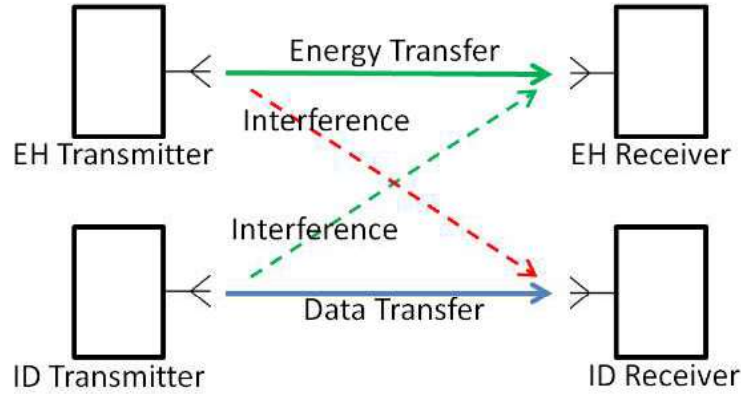


Fig. 2.6 Interference Channel.

2.2.3 SWIPT with Linear EH Model in Interference Channel

In this section, we provide the state-of-the-art of SWIPT in interference channel (IFC). SWIPT addressing IFC has been studied in [33–39]. The system model of these works have considered the linear EH model.

SWIPT in a two-user and K-user MIMO system have been considered in [33] and [34], respectively. In [33], the receiver either decode the incoming information or harvest the RF energy. There are three different scenarios in a two-user system, the two ID-users, two-EH users and one ID user and one EH user. The two ID users has been discussed in the literature of WIT. For a system with two EH users, the optimal transmission strategy maximizing the total harvested energy is obtained. For a system with one ID and one EH users, the optimal transmission strategy for WPT has been found to has at most rank-one. Accordingly, the transmission strategy with a rank-one energy beamforming has been proposed. The Maximum Energy Beamforming (MEB) and Minimum Leakage Beamforming (MLB) are the beamforming strategy maximizing the energy transfer to the EH user and minimizing the interference power to the ID user, respectively. Signal to Leakage Energy Ratio (SLER) is the beamforming strategy that simultaneously maximize the harvested energy to EH user and minimize the interference power to ID user. Accordingly, the rate-energy region was identified. For given energy beamforming strategy, an iterative algorithm was proposed to optimize the transmit covariance matrix for WIT and transmit power for WPT. The simulation result shows that SLER has a larger rate-energy region than the other two transmission strategies.

The authors then leverage the results in [33] for SWIPT in a K-user MIMO system [34]. In a K-user system with at least one EH user and one ID user, the optimal transmissions strategy maximizing the rate-energy region has at most rank one. Accordingly, rank-one

beamforming strategies (MEB, MLB and SLER) as in [33] were applied. For given energy beamforming, the covariance matrix for WIT and transmit power for WPT were optimized iteratively to identify the achievable rate-energy region.

[35] has investigated SWIPT in a two-user MIMO system with reduced feedback by exploiting the Geodesic energy beamforming scheme. It has been observed that the Geodesic energy beamforming is an optimal transmissions strategy in a two-user MIMO system. By adding a rank-one constraint on the information transfer, the feedback overhead was further reduced by exploiting the Geodesic information beamforming. Accordingly, the extension of rank-one Geodesic information/energy beamforming strategies to the general K -user MIMO IFC was proposed. It is observed that the performance of Geodesic beamforming is close to the performance of SLER beamforming. Furthermore, the adaptive bit allocation strategy has been proposed to further reduce the feedback overhead.

[36] considers SWIPT in a multiuser MISO system. The ideal receiver has been considered where the receiver decodes the information and harvests the energy simultaneously. Due to the circuit limitation, the practical schemes have also been considered namely time-division mode switching (TDMS), time-division multiple access (TDMA) and power splitting. In TDMS scheme, the transmission time is divided into two time slots where all user harvest energy and decode information in the first and second time slot, respectively. In TDMA scheme with K user, the transmission time was divided into K time slots. At each time slot, one user perform information decoding and the other user perform energy harvesting. Because WPT was deterministic, the interference signal was considered to be decodable at the ID user. The resource allocation and time/power splitting were optimized maximizing the rate with given transmit power and energy constraints.

SWIPT in a two-user and K -user SISO IFC has been considered in [37]. It was assumed that each transmitter has a perfect knowledge of global CSI. First, the transmission strategy for a two-user system was discussed. Since WPT signal is deterministic, it was assumed that the interference signal can be estimated and canceled at the ID user. The WPT at the two transmitters was jointly optimize such as the energy beamforming was maximized. The optimal resource allocation for WPT and WIT were then obtained. In SWIPT with the general K -user, to avoid high complexity due to signal splitting and joint optimization of WPT at the transmitters, a suboptimal scheme was proposed. Two-user (one ID and one EH) are grouped together and in each group, the optimization for the two-user case is applied. Due to the high complexity in searching for the best grouping, a suboptimal grouping algorithm that leads to a weak inter-group interference and strong intra interference is proposed. Since the intra-group interference can be cancelled at the ID and provide an additional source for energy harvesting, this grouping strategy benefits both WIT and WPT. Additionally, because

the ID user is not able to decode inter-group interference, a weak inter-group interference is good for user grouping.

[38] considers SWIPT in a two-user MISO IFC with one EH user and one ID user. The energy and information transmitters are equipped with multiple antennas. The receiver performed time switching to decode information and harvest energy. Information transmitter considers two transmission strategies where the transmitter either transmits WIT signal or the superposition of WIT and WPT signals. For the second strategy, WPT signal was jointly optimized at the transmitters. The optimal beamforming was optimized maximizing the rate with given transmit power and energy harvesting constraints.

A two-user SISO IFC with energy harvesting transmitter has been considered in [39]. The transmitter transmits its signal subject to power availability in the battery. The power allocation was optimized maximizing the sum-rate subject to the transmit power, battery, data and energy causality constraints.

SWIPT in OFDM Interference Channel

By looking at the SWIPT literature, SWIPT in OFDM IFC has not been addressed yet. Therefore, we investigate SWIPT in an OFDM IFC. The details of this work is discussed in Chapter 3.

2.3 Waveform Design for Wireless Information and Power Transfer based on non-linear EH Model

Recently, an analytical EH model that accounts for the non-linearity of the diode has been introduced in [17, 18]. The waveform design and optimization based on the non-linear EH model have been studied in [17–20]. It has been shown to lead to a significant enhancement in RF to DC conversion efficiency and a larger achievable rate-energy region. The analytical non-linear EH model introduced in [17, 18] will be discussed next.

2.3.1 Non-linear Energy Harveting Model

In this section, we provide an analytical non-linear energy harveting model as introduced in [17, 18]. It is assumed that a loseless antenna model with voltage source $v_s(t)$ and antenna impedance $R_{ant} = 50\Omega$ is connected with an energy harvester with input impedance R_{in} as shown in Figure 2.7. The received signal at the antenna, $y(t)$ has an average power $P_{av} =$

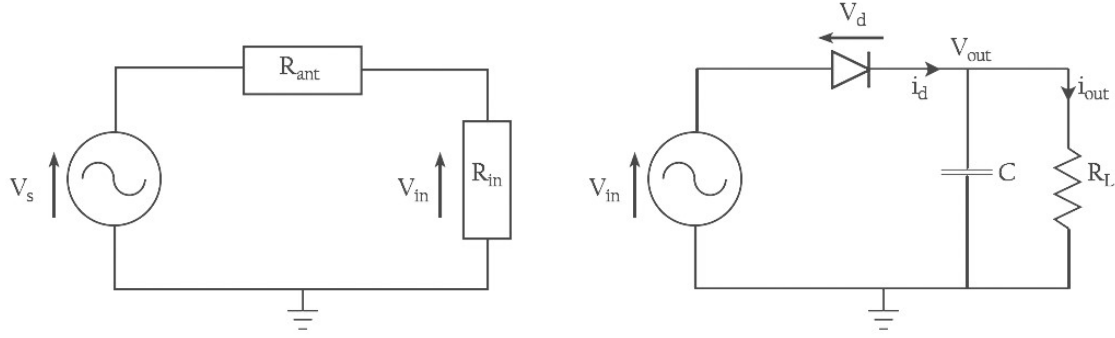


Fig. 2.7 Rectenna Circuit.

$\mathcal{E}\{|y(t)|^2\}$. Assuming a perfect impedance matching $R_{in} = R_{ant}$, the power is completely transferred to the energy harvester input impedance such that $P_{av} = \mathcal{E}\{|v_{in}(t)|^2\}/R_{in}$.

Accordingly, the output current in Eq. (2.13) can be written as

$$i_d(t) = \sum_{i=0}^{\infty} k_i R_{ant}^{i/2} \{y(t)^i\}. \quad (2.20)$$

Assuming an ideal rectifier, the current delivered to the load is constant and given as $i_{out} = \mathcal{A}\{i_d(t)\}$ where $\mathcal{A}\{\cdot\}$ indicates the DC component of a signal.

In order to make the optimization tractable, the Taylor expansion in Eq. (2.20) is truncated to the n_0^{th} order. The DC current is approximated as

$$i_{out}(t) = \mathcal{A}\{i_d(t)\} \approx \sum_{i=0}^{n_0} k_i R_{ant}^{i/2} \mathcal{A}\{y(t)^i\}. \quad (2.21)$$

By ignoring the terms higher than 4, the DC component in Eq. (2.21) can be written as $i_{out} = k_0 + z_{DC}$, where

$$z_{DC} = \sum_{i=1}^4 k_i R_{ant}^{i/2} \mathcal{A}\{y(t)^i\}. \quad (2.22)$$

It has been mentioned in [22] that the 4-th order truncation is the minimum order to sufficiently demonstrate the non-linearity of the rectifier process. Following [18], $k_i = \frac{i_s}{i!(nv_t)^i}$, $i_s = 5\mu A$, diode ideality factor $n = 1.05$ and $v_t = 25.86mV$. Here, k_0 is not a function of the received waveform such that optimizing i_{out} is equivalent to optimizing z_{DC} .

The received signal of deterministic and modulated waveforms are given in Eq. (2.5) and Eq. (2.6), respectively. Let us first consider the deterministic waveform. By averaging the

deterministic waveform over time, $\mathcal{A}\{y_{1,P}(t)^2\}$ and $\mathcal{A}\{y_{1,P}(t)^4\}$ are given by

$$\mathcal{A}\{y_{1,P}(t)^2\} = \frac{1}{2} \sum_{n=0}^{N-1} s_{1,1,n}^2 A_{11,n}^2 \quad (2.23)$$

and

$$\mathcal{A}\{y_{1,P}(t)^4\} = \frac{3}{8} \left[\sum_{\substack{n_0, n_1, n_2, n_3 \\ n_0 + n_1 = n_2 + n_3}} \left[\prod_{j=0}^3 s_{1,1,n_j} A_{11,n_j} \right] \cos(\psi_{11,n_0} + \psi_{11,n_1} - \psi_{11,n_2} - \psi_{11,n_3}) \right]. \quad (2.24)$$

The odd terms is neglected since $\mathcal{A}\{y_{1,P}(t)^i\} = 0$ for $i = 1, 3$.

The modulated waveform (i.e., OFDM) exhibits randomness due to the information symbol $\tilde{x}_I$. The modulated waveform can be seen as the multisine waveform for a fixed input symbols $\tilde{x}_I$. Hence, similar expression as eq. (1.22) and (1.23) can be written for $\mathcal{A}\{y_{1,I}(t)^2\}$ and $\mathcal{A}\{y_{1,I}(t)^4\}$, respectively. However, the information symbols change randomly every symbol duration T . For a given impulse response, the DC component is obtained by averaging over the distribution of the information symbols $\tilde{x}_I$ such that for $\mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^i\}\}$, $i = 2, 4$ is first extracted for given set of amplitudes and phases and then over the randomness of the information symbols $\tilde{x}_I$. The information symbol has a complex Gaussian distribution with $|\tilde{x}_{I,n}|^2$ is exponentially distributed such that $\mathcal{E}\{|\tilde{x}_{I,n}|^2\} = p_{2,n}$ and $\tilde{\phi}_{2,n}$ is uniformly distributed. From the moments of exponential distribution, we also have $\mathcal{E}\{|\tilde{x}_{I,n}|^4\} = 2p_{2,n}^2$. Therefore, by defining $s_{k,n} = \sqrt{p_{k,n}}$, $\mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^2\}\}$ and $\mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^4\}\}$ can be written as

$$\mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^2\}\} = \frac{1}{2} \mathcal{E}\left\{ \sum_{n=0}^{N-1} \tilde{s}_{2,2,n}^2 A_{12,n}^2 \right\} = \frac{1}{2} \sum_{n=0}^{N-1} \mathcal{E}\{\tilde{s}_{2,2,n}^2\} A_{12,n}^2 = \frac{1}{2} \sum_{n=0}^{N-1} s_{2,2,n}^2 A_{12,n}^2 \quad (2.25)$$

and

$$\begin{aligned} \mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^4\}\} &= \frac{3}{8} \mathcal{E}\left\{ \sum_{\substack{n_0, n_1, n_2, n_3 \\ n_0 + n_1 = n_2 + n_3}} \tilde{s}_{2,2,n_0} \tilde{s}_{2,2,n_1} \tilde{s}_{2,2,n_2} \tilde{s}_{2,2,n_3} A_{12,n_0} A_{12,n_1} A_{12,n_2} A_{12,n_3} \right. \\ &\quad \left. \cos(\tilde{\psi}_{12,n_0} + \tilde{\psi}_{12,n_1} - \tilde{\psi}_{12,n_2} - \tilde{\psi}_{12,n_3}) \right\} \\ &= \frac{3}{8} \left[\sum_n \mathcal{E}\{\tilde{s}_{2,2,n}^4\} A_{12,n}^4 + 2 \sum_{\substack{n_0, n_1, \\ n_0 \neq n_1}} \mathcal{E}\{\tilde{s}_{2,2,n_0}^2\} \mathcal{E}\{\tilde{s}_{2,2,n_1}^2\} A_{12,n_0}^2 A_{12,n_1}^2 \right] \\ &= \frac{3}{4} \sum_{n_0} \sum_{n_1} s_{2,2,n_0}^2 s_{2,2,n_1}^2 A_{12,n_0}^2 A_{12,n_1}^2. \end{aligned} \quad (2.26)$$



Fig. 2.8 Point to point WPT [17, 18].

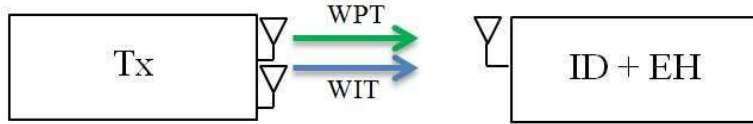


Fig. 2.9 Point to point SWIPT [19].

2.3.2 Waveform Design Literature

In this section, we provide the state-of-the-art of waveform design based on non-linear EH model for WPT and WIT.

A point-to-point WPT has been considered in [17, 18] as shown in Figure 2.8. The transmitter and receiver are equipped with multiple and single antenna, respectively. An analytical model of a rectenna adaptive to the CSI and multisine waveform has been introduced as discussed in Section (2.3.1). A waveform design maximizing the DC current subject to the transmit power constraint was discussed. It was found that the linear and non-linear models lead to different waveform design. The linear and non-linear models favor to allocate power over a single frequency component and multiple frequency components, respectively. The numerical and PSpice simulation has been performed. The results has shown that waveform design based on non-linear model provides a significant gain compared to the linear model and non-adaptive waveform design. The result has also shown that accounting for the non-linearity of the rectifier can enhances RF-to-DC conversion efficiency.

[19] has extended the waveform design problem to a point-to-point SWIPT as shown in Figure 2.9. The analytical model of the rectenna in [18] was leveraged and the performance of WPT and WIT in term of harvested energy were investigated. It was found that, due to the non-linearity of the rectifier and the randomness of information symbols, the OFDM waveform is less efficient than the deterministic WPT waveform. However, based on the linear model, the performance for both WPT and OFDM waveforms are similar. Accordingly, superposed SWIPT waveform is proposed to increase the efficiency of the rectenna. The superposed SWIPT waveform is optimized maximizing the rate-energy region for given transmit power. Power splitting is performed at the receiver to decode information and harvest energy.

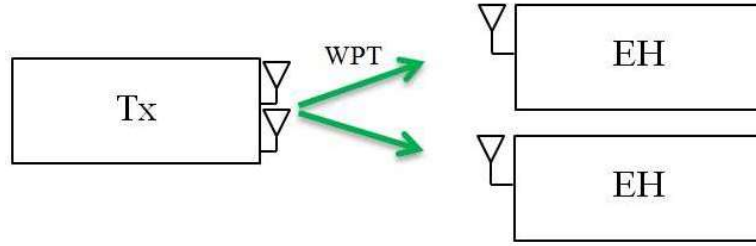


Fig. 2.10 WPT to multiuser EH [20].

[20] has leveraged the results in [18] and introduced an optimization methodology suitable for multi-antenna multisine multi-user WPT as shown in Figure 2.10. The transmitter and receiver are equipped with multiple and single antenna, respectively. In the presence of large antennas and sinewaves, optimization algorithm that exploits channel hardening was proposed. The waveform design was optimized such that the output DC is maximized for given transmit power. The result highlights the essential of accounting for the non-linearity of the rectifier in multi-user WPT.

2.3.3 Coordinated Waveform Design for SWIPT

In this thesis, the coordinated waveform design and optimization for SWIPT based on non-linear EH model was studied. The details of this work is discussed in Chapter 4.

2.4 Backscattering System

In the backscattering system as depicted in Figure 2.11, the continuous RF waveform is transmitted at the reader and propagate through the channel. The backscatter transfers information to the reader by modulating the RF received signal. Intuitively, waveform propagation through two medias with different impedance results in reflected wave [67, 68]. The modulation is performed by changing the antenna impedance in the presence of an incident signal. By modulating the antenna impedance, backscatter can control the reflected waveform such that it can be used for information transfer. The radiated waveform is reflected back to the reader and decoded to extract the information. Backscattering system benefits of low power consumption as the amplification and re-transmission are not performed at the backscatter.

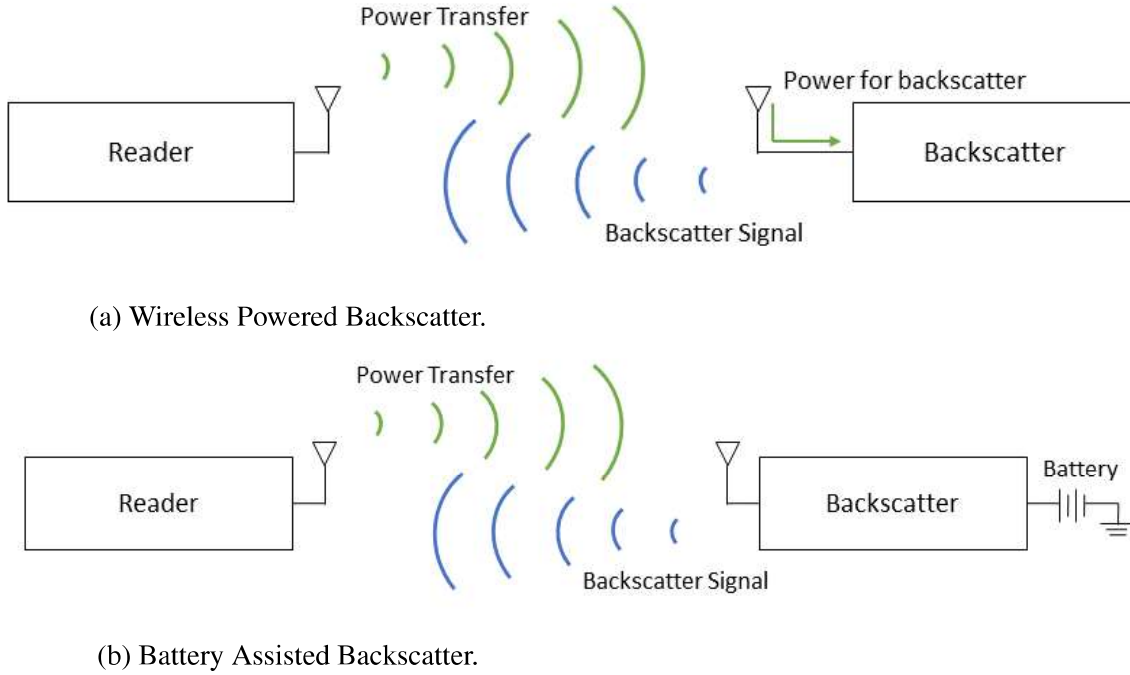


Fig. 2.11 Backscattering System [69].

2.4.1 Backscatter Type

Backscatter can be categorized into wireless and battery powered backscatters.

A wireless powered backscatter (passive backscatter) is entirely powered by the received signal from the transmitter. The backscatter operation is fully dependent on the wireless power transfer. Diode-based envelope detectors are used to harvest DC power at the tag and provides the power to other electronic parts. The backscatter transmission is performed by modulating the received RF signal. The limitation of power availability at the tag leads to an active research in the wireless powered backscatter.

A battery assisted backscatter (semi-passive backscatter) use battery power source to operate the circuit. Similar to the wireless powered backscatter, the transmission in battery assisted backscatter is performed by modulating the RF received signal.

2.4.2 Waveform Optimization in Backscattering System

In this section, we provide the state-of-the-art of waveform optimization in backscattering system.

[13] has studied WPT in a multiuser backscattering system. The reader and tag are equipped with multiple and single antennas, respectively. Energy beamforming was exploited to increase the total harvested energy at the tags. The reader are assumed to has no CSI

knowledge of the forward link. Therefore, the optimization at the reader was performed based on backscatter channel. The resource allocation was optimized such that the total harvested energy is maximized. For comparison, two baseline schemes, the optimization based on perfect forward channel and estimated forward channel have been considered. The result has shown that optimization based on backscatter channel suffer slight reduction as compared to the two baseline schemes.

WPT optimization in MIMO system has been considered in [67]. It was assumed that the reader has no CSI knowledge of forward link. Therefore, the optimization was performed based on backscatter channel so as to maximize the DC power from the backscattering signal at the reader. The result has shown that maximizing the backscattering receive power at the reader will also maximizing the received power at the backscatter.

[66] has considered a point to point WPT in backscattering system. The efficiency of several waveform were investigated to power up the backscatter. The result shows that the modulated signal (ASK, QPSK, BPSK and 16QAM) has a higher efficiency compared to the continuous waveform with the same transmit power. In addition, the result shows that the multisine waveform with equal power and phase-aligned over the frequency components increase the efficiency of RF-DC conversion.

[68] has investigated the performance of RF to DC conversion for several RFID tags. Power optimized waveform with short time window of large voltage peaks has been presented to improve the efficiency. The implementation of power optimized waveform at the RFID system was then discussed. Several power optimized waveform and the efficiency improvement at different RFID tags were discussed. It was observed that waveform with large peak power provides the highest RF to DC efficiency at the tag.

2.4.3 Waveform Design in Wireless Powered Backscattering System

In this thesis, waveform design and optimization in wireless powered backscattering system was studied. The details of this work is discussed in Chapter 5.

Chapter 3

Simultaneous Wireless Information and Power Transfer in a K-User Interference Channel

This chapter discusses SWIPT in SISO K-user OFDM IFC based on the linear EH model. SWIPT in MIMO IFC has been considered in [33, 34]. The results for SWIPT in MIMO IFC [33, 34] is leveraged to address the SWIPT in SISO K-user OFDM IFC. There are some similarities between a MIMO IFC and an OFDM IFC. Both MIMO and OFDM systems can provide high data rate and reliable communication by exploiting the spatial domain and frequency domain, respectively. However, a MIMO system needs to tackle the co-channel interference from all transmit antennas. In contrast, an OFDM transmission converts a frequency selective channel into a set of interference-free (i.e., parallel) sub-channels (denoted as subcarriers). The OFDM channel can therefore be modeled as a diagonal matrix. Unfortunately, an OFDM system have to deal with co-channel interference from other transmitters. In addition, a MIMO transmitter can beamform its signal by varying the transmit power and phase over different antennas, while an OFDM transmitter transmits its signal by allocating a proper transmit power at each subcarrier.

Interestingly, the result of a one-subcarrier transmission strategy of an OFDM system can be regarded as the rank one beamforming of the MIMO system in [33, 34]. Here, the signal at the energy transmitter in both OFDM and MIMO systems is transmitted onto a one-dimensional subspace by exploiting the frequency domain and spatial domain, respectively. However, the rank one beamforming of the MIMO system is in continuous space domain while the subcarrier transmission strategy in the OFDM system is in discrete frequency domain. It results in a simplistic optimal subcarrier selection for OFDM system compared to the MIMO Geodesic Beamforming in [35]. In addition, due to the discrete subcarrier in OFDM system,

the achievable rate-energy can be further enlarged region of a system with perfect knowledge of the local CSI with a low-rate feedback of the unutilized subcarrier indices by ID users.

3.1 System Model

This section introduce the OFDM system model for the SISO K-user IFC where information and energy transmitters transmit their OFDM/multisine signal to their corresponding (information decoding or energy harvesting) receivers. We will address the non-linear EH model in Chapter 4. Each OFDM symbol has N subcarriers and each transmitter and receiver has a single antenna. It is assumed that each transmitter has perfect knowledge of the local CSI, i.e. CSI of the links between itself and all receivers. The received signal at i -th receiver, $\mathbf{y}_i \in \mathbb{C}^{N \times 1}$ can be written as

$$\mathbf{y}_i = \sum_{k=1}^K \mathbf{H}_{ik} \mathbf{x}_k + \mathbf{n}_i, \quad (3.1)$$

where $\mathbf{n}_i \in \mathbb{C}^{N \times 1}$ is a complex white Gaussian noise vector with a covariance matrix $\sigma_n^2 \mathbf{I}_N$. For simplicity, it is assumed that $\sigma_n^2 = 1$. Moreover, it is assumed that the channel is slow fading, the transmitters are perfectly frequency synchronous and the cyclic prefix is long enough, such that the channel encompassing all subcarriers can be represented by a diagonal matrix structure as $\mathbf{H}_{ik} = \text{diag}(h_{ik,1}, \dots, h_{ik,N})$. Here, $h_{ik,n}$ is the channel coefficient for the n -th subcarrier between the k -th transmitter and the i -th receiver. The relative path loss of the cross link compared to the direct link is defined as $\delta_{ik} \in [0, 1]$. The vector $\mathbf{x}_k \in \mathbb{C}^{N \times 1}$ is the transmit signal at the k -th transmitter with a transmit power constraint (transmit power constraint at the baseband) P , such that

$$\mathcal{E} \left[\|\mathbf{x}_k\|^2 \right] \leq P \quad (3.2)$$

for $k = 1, 2, \dots, K$. In addition, $x_{k,n}$ is the transmit signal for the n -th subcarrier at the k -th transmitter and $p_{k,n} = \mathcal{E} \left[|x_{k,n}|^2 \right]$.

The time switching receiver is considered where the receiver switches its operation mode between ID and EH over time. When the receiver operates in information decoding (ID) mode, the achievable rate at the i -th ID receiver is given by

$$R_i = \log_2 \det(\mathbf{I}_N + \mathbf{H}_{ii}^H \mathbf{R}_{-i}^{-1} \mathbf{H}_{ii} \mathbf{P}_i) \quad (3.3)$$

where $\mathbf{P}_i = \text{diag}(p_{i,1}, \dots, p_{i,N})$ is the power allocation at the i -th transmitter and $\text{tr}(\mathbf{P}_i) \leq P$. $\mathbf{R}_{-i}$ is the covariance matrix of the noise plus interference at the i -th receiver and is given by

$$\mathbf{R}_{-i} = \mathbf{I}_N + \sum_{\substack{k=1 \\ k \neq i}}^K \mathbf{H}_{ik} \mathbf{P}_k \mathbf{H}_{ik}^H. \quad (3.4)$$

When the receiver operates in the energy harvesting (EH) mode, the total harvested energy at the i -th EH receiver (from Eq. (2.16)) is given by

$$E_i = \sum_{k=1}^K \text{tr}(\mathbf{H}_{ik} \mathbf{P}_k \mathbf{H}_{ik}^H + \mathbf{I}_N). \quad (3.5)$$

For simplicity, it is assumed that the noise power is negligible compared to the energy harvested from the transmitted signal such that

$$E_i \approx \sum_{k=1}^K \text{tr}(\mathbf{H}_{ik} \mathbf{P}_k \mathbf{H}_{ik}^H). \quad (3.6)$$

It is assumed that the interference signal from other transmitters is not decodable (If it is decodable, the interference term can simply be set to 0). Accordingly, it degrades the achievable rate performance at the ID users [8]. In contrast, the EH circuit can harvest energy from the interference signal. Therefore, the interference signal at the EH receiver is beneficial to increase the total harvested energy.

3.2 Transmission Strategy For The K-User OFDM IFC

This section discuss the transmission strategy for a K-user OFDM IFC with the local CSI at each transmitter. There are three different scenarios, namely K-ID user, K-EH user, and multiple-EH user and multiple-ID user. The main focus is on the multiple-ID user and multiple-EH user scenario. First, the transmission strategy for a K-ID user OFDM IFC and a K-EH user OFDM IFC is discussed. Then, the necessary condition for the optimal transmission strategy for a two-user OFDM IFC with one-EH user and one-ID user [70] is discussed. The necessary condition for the optimal transmission strategy for a K-user OFDM IFC with multiple-EH user and multiple-ID user is then obtained by leveraging the work in [34, 70]. Accordingly, the subcarrier selection strategies is proposed and the achievable rate-energy regions is identified. Furthermore, the achievable rate-energy region is further enlarged with some knowledge of the unutilized subcarrier indices by the information transmitters to the energy transmitters.

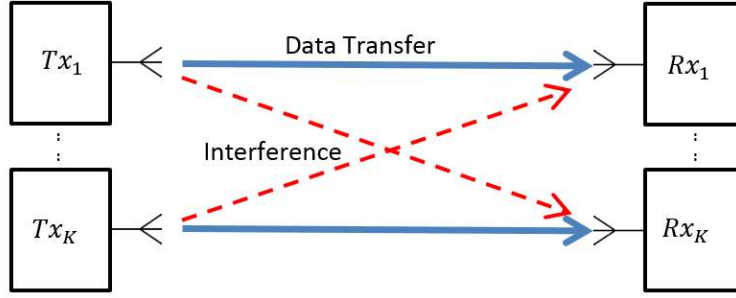


Fig. 3.1 K-ID User Scenario.

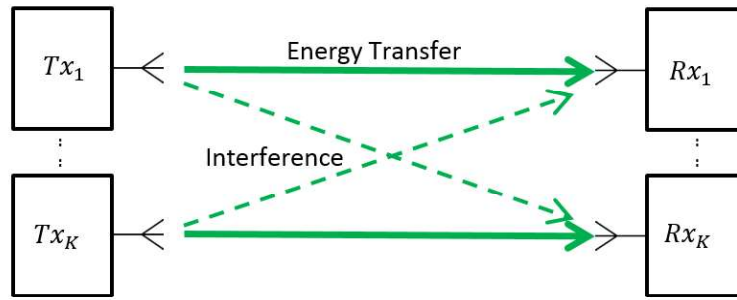


Fig. 3.2 K-EH User Scenario.

3.2.1 K-ID User Scenario

In the K-ID user scenario as depicted in Figure 3.1, the objective is to obtain the maximum achievable sum-rate subject to the transmit power constraint at each transmitter. Since the users decode data information, the total harvested energy is equal to zero. The problem in the K-ID user IFC has been considered in several studies. For example, non-cooperative game theory has been developed to maximize the achievable sum-rate with no CSI sharing among the transmitters [71, 72]. Each user iteratively updates their power allocation using waterfilling algorithm until they reach a Nash Equilibrium.

3.2.2 K-EH User Scenario

In the K-EH user scenario as depicted in Figure 3.2, the objective is to maximize the total harvested energy subject to the transmit power constraint at each transmitter. Since the users harvest RF energy, the achievable rate is equal to zero. The optimal transmission strategy in a K-EH user IFC is for the k-th EH transmitter to allocate its full transmit power to the subcarrier corresponding to the $\arg \max_{n=1,..,N} \sum_{j=1}^K |h_{jk,n}|^2$.

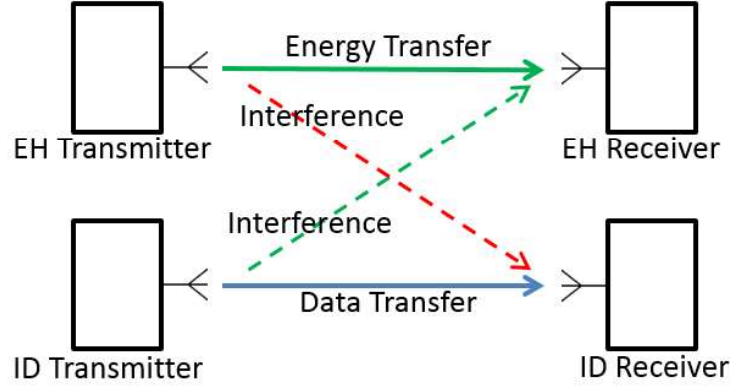


Fig. 3.3 One-EH User and One-ID User Scenario.

3.2.3 One EH-User and One ID-User

This section discuss SWIPT for a two-user with one-EH user and one-ID user (with their corresponding transmitter respectively denoted as EH transmitter and ID transmitter) as depicted in Figure 3.3. Without loss of generality, it is assumed that the first receiver is in EH mode and the second receiver is in ID mode. The achievable rate-energy region is then given by

$$C_{R-E}(P) \triangleq \left\{ R \leq \log_2 \det(\mathbf{I}_N + \mathbf{H}_{22}^H \mathbf{R}_{-2}^{-1} \mathbf{H}_{22} \mathbf{P}_2), \sum_{k=1}^2 \text{tr}(\mathbf{H}_{1k} \mathbf{P}_k \mathbf{H}_{1k}^H) \geq E, \right. \\ \left. \text{tr}(\mathbf{P}_k) \leq P, \mathbf{P}_k \geq \mathbf{0}, k = 1, 2 \right\}, \quad (3.7)$$

where $\mathbf{R}_{-2} = \mathbf{I}_N + \mathbf{H}_{21} \mathbf{P}_1 \mathbf{H}_{21}^H$.

Proposition 1: The optimal transmission strategy at high SNR for the EH transmitter in a two user OFDM IFC with one EH-user and one ID-user is to transmit its signal by allocating its transmit power on a single subcarrier.

Proof. First, let consider the boundary point $(\bar{R}, \bar{E})$ of the achievable rate-energy in Eq. (3.7), where $\text{tr}(\mathbf{H}_{12} \mathbf{P}_2 \mathbf{H}_{12}^H) \geq \bar{E}$. Then, for any given $\mathbf{P}_2$ on the boundary point, $\mathbf{P}_1 = \mathbf{0}$ is the optimal power allocation at the EH transmitter. That is, because the harvested energy from the ID transmitter is sufficient to meet the energy constraint $\bar{E}$, EH transmitter does not transmit any signal causing the interference to the ID receiver.

For $\text{tr}(\mathbf{H}_{12} \mathbf{P}_2 \mathbf{H}_{12}^H) < \bar{E}$, the EH transmitter should allocate its transmit power on a single subcarrier. By letting $\mathbf{A} = \mathbf{H}_{21} \mathbf{P}_1 \mathbf{H}_{21}^H$ and $\mathbf{B} = \mathbf{H}_{22} \mathbf{P}_2 \mathbf{H}_{22}^H$ such that $\mathbf{A} = \text{diag}(a_1, \dots, a_N)$ and

$\mathbf{B} = \text{diag}(b_1, \dots, b_N)$, the achievable rate at the boundary point $\bar{R}$ at high SNR is given by

$$\bar{R} = \log_2 \prod_{i=1}^N \left(1 + \frac{b_i}{1 + a_i} \right), \quad (3.8)$$

where b_i and a_i is the received power from the ID transmitter and the interference power from the EH transmitter at subcarrier i , respectively. First, let consider that the EH transmitter allocates its transmit power on m subcarriers such that $m < N$ and the ID transmitter allocates its transmit power on all available subcarriers (i.e., N subcarriers). Note that at high SNR (without interference), the ID transmitter allocates its transmit power on all available subcarriers (i.e., N subcarriers). For notation brevity, the subcarrier indices at both transmitters is sorted such that the EH transmitter allocates its transmit power to the first m subcarriers. Then the achievable rate can be written as

$$\bar{R} \approx \log_2 \left(\frac{\prod_{i=1}^m (1 + a_i + b_i) \prod_{j=m+1}^N (1 + b_j)}{\prod_{i=1}^m (1 + a_i)} \right). \quad (3.9)$$

At high SNR, because b_i for $n = 1, \dots, N$ are linearly proportional to P and for large transmit power, $\log(1 + P) \approx \log(P)$, Eq. (3.9) can be approximated as

$$\bar{R} \approx \log_2 \left(\prod_{i=1}^m \left(\frac{a_i}{1 + a_i} + \frac{b_i}{1 + a_i} \right) \prod_{j=m+1}^N b_j \right). \quad (3.10)$$

Because $b_i \propto P$ and $bi \gg a_i$, a_i is negligible with respect to b_i . Therefore, Eq. (3.10) can be written as

$$\bar{R} = \log_2 \left(\frac{\prod_{i=1}^m b_i}{\prod_{i=1}^m (1 + a_i)} \right). \quad (3.11)$$

Here, it is noted that $\prod_{i=1}^m (1 + a_i)$ in the denominator part of Eq. (3.11) has a minimum value when $m = 1$ (i.e., EH transmitter allocates its transmit power at one subcarrier) and a_i is small for given $\bar{E}$.

In addition, when the interference power from the EH transmitter is large for given $\bar{E}$, $\frac{a_i}{1 + a_i} \approx 1$. Accordingly, Eq. (3.10) can be written as

$$\begin{aligned} \bar{R} &\approx \log_2 \left(\prod_{i=1}^m \left(1 + \frac{b_i}{1 + a_i} \right) \prod_{j=m+1}^N b_j \right) \\ &= \log_2 \left(\prod_{j=m+1}^N b_j \right) + \log_2 \left(\prod_{i=1}^m \left(1 + \frac{b_i}{1 + a_i} \right) \right). \end{aligned} \quad (3.12)$$

Because $b_j \propto P$ for $j = 1, \dots, N$ and when a_i is large $\log_2(1 + \frac{b_i}{1+a_i})$ is small compared to $\log_2(b_j)$, then Eq. (3.12) is maximized when $m = 1$. $\square$

Remark 3.1: From the energy harvesting perspective, the EH transmitter should allocate its transmit power on a single subcarrier, that is the subcarrier corresponding to $\arg \max_{n=1, \dots, N} |h_{11,n}|^2$. From the information decoding perspective, the maximization of the achievable rate at high SNR is equivalent to the maximization of DOF. That is, from the DOF perspective, a larger number of subcarriers used to transmit power at the EH transmitter means that more subcarriers are interfered at the ID receiver. Therefore, for both information and energy transfer, EH transmitter should allocate its transmit power on a single subcarrier.

3.2.4 Multiple EH-User and Multiple ID-User

This section discuss the transmission strategy for the multiple EH-user and multiple ID-user as depicted in Figure 3.4. This discussion is also applicable for one EH/ID-user and multiple ID/EH-user scenarios. Without loss of generality, it is assumed that $\{Tx_k, Rx_k\}$ for $k = 1, \dots, K_1$ operate in EH mode and $k = K_1 + 1, \dots, K$ operate in ID mode. The achievable rate-energy region is given by

$$C_{R-E}(P) \triangleq \left\{ R \leq \sum_{k=K_1+1}^K \log_2 \det(\mathbf{I}_N + \mathbf{H}_{kk}^H \mathbf{R}_{-k}^{-1} \mathbf{H}_{kk} \mathbf{P}_k), \sum_{k=1}^K \sum_{i=1}^{K_1} \text{tr}(\mathbf{H}_{ik} \mathbf{P}_k \mathbf{H}_{ik}^H) \geq E, \right. \\ \left. \text{tr}(\mathbf{P}_k) \leq P, \mathbf{P}_k \geq \mathbf{0}, k = 1, \dots, K \right\}. \quad (3.13)$$

Power allocation $\mathbf{P}_k$ for $k = 1, \dots, K$ is coupled with power allocation from other transmitters which makes the problem above non-convex and therefore the optimal solution is infeasible in practice. Here, a necessary condition for the optimal transmission strategy for the EH transmitters with local CSI at high SNR is derived. Note that EH transmitters optimize their transmission strategy in a distributed manner.

Proposition 2: The optimal transmission strategy at high SNR for the EH transmitter in a K-user OFDM IFC with multiple EH-user and multiple ID-user is to transmit its signal by allocating its transmit power on a single subcarrier.

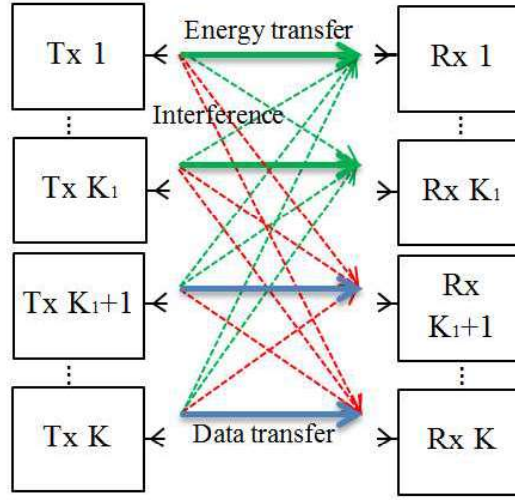


Fig. 3.4 K-user OFDM IFC with K_1 energy transmitters and $(K - K_1)$ information transmitters.

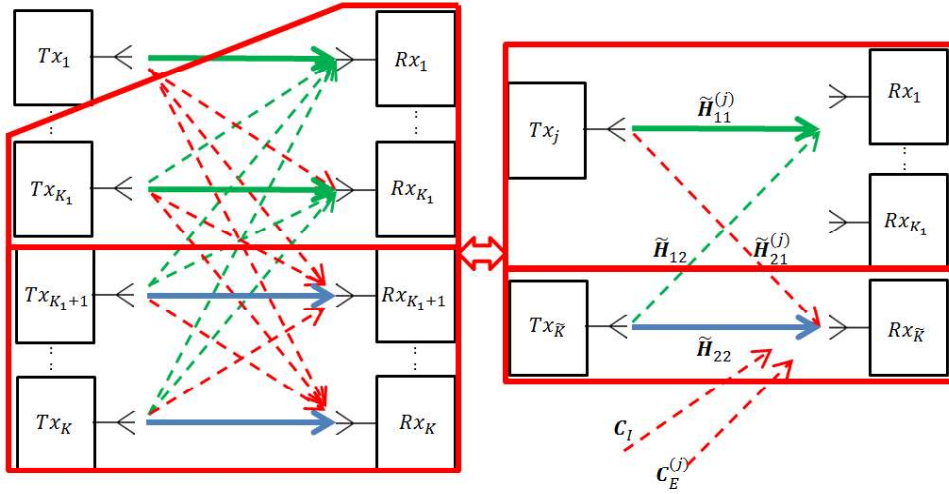


Fig. 3.5 Multiple ID and multiple EH in a K-user OFDM IFC.

Proof. Motivated by [34], the optimal transmission strategy for j -th EH transmitter in Eq. (3.13) is also the optimal solution for the boundary of

$$C_{R-E,j}(P) \triangleq \left\{ R \leq \log_2 \det \left(\mathbf{I}_{(K-K_1)N} + \tilde{\mathbf{H}}_{22}^H \left(\tilde{\mathbf{R}}_{-2}^{(j)} \right)^{-1} \tilde{\mathbf{H}}_{22} \tilde{\mathbf{P}}_2 \right), E \leq \text{tr}(\tilde{\mathbf{H}}_{11}^{(j)} \mathbf{P}_j (\tilde{\mathbf{H}}_{11}^{(j)})^H) + \text{tr}(\tilde{\mathbf{H}}_{12} \tilde{\mathbf{P}}_2 \tilde{\mathbf{H}}_{12}^H), \text{tr}(\mathbf{P}_k) \leq P, \mathbf{P}_k \geq \mathbf{0}, k = j, K_1 + 1, \dots, K \right\}, \quad (3.14)$$

for $j = 1, \dots, K_1$, where

$$\tilde{\mathbf{H}}_{22} = \text{diag} \{ \mathbf{H}_{(K_1+1)(K_1+1)}, \dots, \dots \mathbf{H}_{KK} \}, \quad (3.15)$$

$$\tilde{\mathbf{H}}_{12} = [\tilde{\mathbf{H}}_{12}^{(K_1+1)}, \dots, \dots \tilde{\mathbf{H}}_{12}^{(K)}] = \begin{bmatrix} \mathbf{H}_{1K_1+1} & \dots & \mathbf{H}_{1K} \\ \vdots & \ddots & \vdots \\ \mathbf{H}_{K_1K_1+1} & \dots & \mathbf{H}_{K_1K} \end{bmatrix}, \quad (3.16)$$

$$\tilde{\mathbf{H}}_{21}^{(j)} = \begin{bmatrix} \mathbf{H}_{(K_1+1)j} \\ \vdots \\ \mathbf{H}_{Kj} \end{bmatrix}, \quad \tilde{\mathbf{H}}_{11}^{(j)} = \begin{bmatrix} \mathbf{H}_{1j} \\ \vdots \\ \mathbf{H}_{K_1j} \end{bmatrix}, \quad (3.17)$$

$$\tilde{\mathbf{P}}_2 = \text{diag} \{ \mathbf{P}_{K_1+1}, \dots, \mathbf{P}_K \}, \quad (3.18)$$

and

$$\tilde{\mathbf{R}}_{-2}^{(j)} = \mathbf{I}_{(K-K_1)N} + \mathbf{C}_I + \mathbf{C}_E^{(j)} + \tilde{\mathbf{H}}_{21}^{(j)} \mathbf{P}_j (\tilde{\mathbf{H}}_{21}^{(j)})^H. \quad (3.19)$$

$\mathbf{C}_I = \text{diag} \{ \sum_{\substack{k=K_1+1 \\ k \neq K_1+1}}^K \mathbf{H}_{(K_1+1)k} \mathbf{P}_k \mathbf{H}_{(K_1+1)k}^H, \dots, \sum_{\substack{k=K_1+1 \\ k \neq K}}^K \mathbf{H}_{Kk} \mathbf{P}_k \mathbf{H}_{Kk}^H \}$ is the covariance matrix

of the interference from other ID transmitters and $\mathbf{C}_E^{(j)} = \sum_{k=1, k \neq j}^{K_1} \tilde{\mathbf{H}}_{21}^{(k)} \mathbf{P}_k (\tilde{\mathbf{H}}_{21}^{(k)})^H$ is the covariance matrix of the interference from all EH transmitters except for j -th EH transmitter.

The energy at the boundary point of Eq. (3.13) can be written as

$$\sum_{k=1}^K \sum_{i=1}^{K_1} \text{tr} (\mathbf{H}_{ik} \mathbf{P}_k \mathbf{H}_{ik}^H) = \text{tr} (\tilde{\mathbf{H}}_{11}^{(j)} \mathbf{P}_j (\tilde{\mathbf{H}}_{11}^{(j)})^H) + \sum_{k=1, k \neq j}^{K_1} \text{tr} (\tilde{\mathbf{H}}_{11}^{(k)} \mathbf{P}_k \tilde{\mathbf{H}}_{11}^{(k)}) + \text{tr} (\tilde{\mathbf{H}}_{12} \tilde{\mathbf{P}}_2 \tilde{\mathbf{H}}_{12}^H), \quad (3.20)$$

where the harvested energy from other EH transmitters $\sum_{k=1, k \neq j}^{K_1} \text{tr} (\tilde{\mathbf{H}}_{11}^{(k)} \mathbf{P}_k (\tilde{\mathbf{H}}_{11}^{(k)})^H)$, shifts the energy boundary of Eq. (3.14) along the energy-axis. It can be found that the optimal transmission strategy for j -th EH transmitter at the boundary point of Eq. (3.13) is also the optimal transmission strategy at the boundary point of Eq. (3.14).

Here, Eq. (3.14) can be seen as the achievable rate-energy region for the two-user OFDM IFC with the effective channel of $\tilde{\mathbf{H}}_{11}^{(j)}, \tilde{\mathbf{H}}_{12}, \tilde{\mathbf{H}}_{21}^{(j)}$ and $\tilde{\mathbf{H}}_{22}$ as depicted in Figure 3.5. But, in contrast to the achievable rate-energy region for the two-user OFDM IFC, the effective channel in Eq. (3.14) is subject to the covariance matrix $\mathbf{C}_I$ and $\mathbf{C}_E^{(j)}$ which are both unknown at j -th EH transmitter. Therefore from Section (3.2.3), the optimal transmission strategy for each EH transmitter is to allocate its transmit power on a single-subcarrier with a proper power allocation, which is a generalized version of the two-user OFDM IFC. $\square$

Remark 3.2: From the energy harvesting perspective, each EH transmitter (j-th EH transmitter) should allocate its transmit power on a single subcarrier corresponding to $\arg \max_{n=1,\dots,N} |\tilde{h}_{11,n}^{(j)}|^2$, where $(\tilde{\mathbf{H}}_{ik}^{(j)})^H \tilde{\mathbf{H}}_{ik}^{(j)} = \text{diag}(|\tilde{h}_{ik,1}^{(j)}|^2, \dots, |\tilde{h}_{ik,N}^{(j)}|^2)$. Here, $|\tilde{h}_{ik,n}^{(j)}|^2 = [\mathbf{H}_{ik}^{(j)}]_n^H [\mathbf{H}_{ik}^{(j)}]_n$. From the information decoding perspective, the sum-rate at high SNR is enhanced if the DOF at the ID users is maximized (i.e., the interference subspace at the ID receivers is minimized). Accordingly, each EH transmitter should allocate its transmit power on a single subcarrier, so that the number of subcarriers at the ID receivers subject to the interference from the EH transmitters is minimized. Therefore, for both information and energy transfer perspectives, each EH transmitter should allocate its transmit power on a single subcarrier.

3.2.5 Achievable Rate-Energy Region

In this section, following the necessary condition of a single subcarrier transmission strategy at the EH transmitters, a single subcarrier selection strategies at the EH transmitters is discussed. Then, the optimization problem for the achievable rate-energy region is formulated. Because the optimization problem is non-convex, an iterative algorithm that optimize the power allocation at the EH transmitters and ID transmitters simultaneously is proposed.

From an energy harvesting perspective, maximum harvested energy can be achieved by allocating the transmit power at an EH transmitter to the subcarrier corresponding to the maximum channel gain (MaxCG) of the concatenated channel matrix from that EH transmitter to all EH receivers. From an information decoding perspective, with the aim of minimizing the total leakage power to all ID receivers, an EH transmitter should allocate its transmit power to the subcarrier corresponding to the minimum channel gain (MinCG) of the concatenated channel matrix from that EH transmitter to all ID receivers. Motivated by [33, 34], the subcarrier index can be efficiently chosen such that the Signal to Leakage and Energy Ratio (SLER) is maximized, where the SLER at j-th EH transmitter is defined as $SLER_n^{(j)} =$

$$\frac{|\tilde{h}_{11,n}^{(j)}|^2}{|\tilde{h}_{21,n}^{(j)}|^2 + \beta \left(\max \left(\frac{\bar{E}}{\bar{K}_1 P} - |\tilde{h}_{11,n}^{(j)}|^2, 0 \right) \right)}, \quad (3.21)$$

where β is a fixed regularization parameter. Hence, the subcarrier index $\bar{n}^{(j)}$ for j -th EH transmitter is chosen as

$$\begin{aligned} \arg \max_{n=1, \dots, N} |\tilde{h}_{11,n}^{(j)}|^2 & \quad \text{for MaxCG} \\ \arg \min_{n=1, \dots, N} |\tilde{h}_{21,n}^{(j)}|^2 & \quad \text{for MinCG} \\ \arg \max_{n=1, \dots, N} SLE R_n^{(j)} & \quad \text{for SLER.} \end{aligned} \quad (3.22)$$

With the knowledge of local CSI at each transmitter, an optimal one-subcarrier transmission strategy is proposed namely Signal to Leakage Ratio with Energy Constraint (SLREC). Let $\bar{E}^{(j)}$ be the required energy to be harvested from j -th EH transmitter at the boundary point of Eq. (3.13) such that

$$\sum_{k=1}^{K_1} \text{tr} \left(\mathbf{H}_{kj} \mathbf{P}_j \mathbf{H}_{kj}^H \right) = \bar{E}^{(j)}. \quad (3.23)$$

Here $\bar{E}^{(j)} > 0$, j -th EH transmitter does not transmit any signal if $\bar{E}^{(j)} = 0$. Then, we have the following proposition.

Proposition 3: The optimal $\bar{n}^{(j)}$ for a single subcarrier transmission strategy at j -th EH transmitter with local CSI is given by

$$\bar{n}^{(j)} = \arg \max_{n \in \mathbf{N}^{(j)}} \frac{|\tilde{h}_{11,n}^{(j)}|^2}{|\tilde{h}_{21,n}^{(j)}|^2}, \quad (3.24)$$

where $\mathbf{N}^{(j)}$ is a set of subcarrier indices at j -th EH transmitter satisfying the energy constraint in Eq. (3.23). Here, if $\mathbf{N}^{(j)} = \{\}$, then no feasible solution can satisfy the energy constraint in Eq. (3.23).

Proof. For a positive definite matrix $\mathbf{S}$ and a positive semi-definite matrix $\mathbf{X}_i$, the maximization of

$$g(\mathbf{X}_i) = \sum_{i=1}^L \log_2 \det(\mathbf{I}_N + \mathbf{S}_i(\mathbf{I}_N + \mathbf{X}_i)^{-1}) \quad (3.25)$$

with respect to $\mathbf{X}_i$ ($i = 1, \dots, L$) is equivalent to the minimization of $\sum_i \det(\mathbf{I}_N + \mathbf{X}_i)$ with respect to $\mathbf{X}_i$ ([35], Lemma 1). Therefore, at the boundary of the achievable rate-energy region of Eq. (3.13),

$$\sum_{k=K_1+1}^K \det(\mathbf{R}_{-k}) = \sum_{k=K_1+1}^K \det \left(\mathbf{I}_N + \sum_{\substack{i \neq k \\ i \neq j}}^K \mathbf{H}_{ki} \mathbf{P}_i \mathbf{H}_{ki}^H + \mathbf{H}_{kj} \mathbf{P}_j \mathbf{H}_{kj}^H \right) \quad (3.26)$$

is minimized. Given the transmission on a single subcarrier n' at j -th EH transmitter, Eq. (3.26) can be written as

$$\sum_{k=K_1+1}^K \det(\mathbf{R}_{-k}) = \sum_{k=K_1+1}^K \left(\left(1 + \sum_{\substack{i \neq k \\ i \neq j}}^K p_{i,n'} |h_{ki,n'}|^2 + p_{j,n'} |h_{kj,n'}|^2 \right) \prod_{n \neq n'}^N \left(1 + \sum_{\substack{i \neq k \\ i \neq j}}^K p_{i,n} |h_{ki,n}|^2 \right) \right). \quad (3.27)$$

Because $\prod_{n \neq n'}^N \left(1 + \sum_{\substack{i \neq k \\ i \neq j}}^K p_{i,n} |h_{ki,n}|^2 \right)$ term in Eq. (3.27) is independent of $\mathbf{P}_j$ (each EH transmitter optimize their transmission strategy in a distributed manner) and unknown to j -th EH transmitter, the minimization of Eq. (3.27) with respect to the interference power from j -th EH transmitter is given by

$$\bar{n}^{(j)} = \arg \min_{n \in \mathbf{N}^{(j)}} \sum_{k=K_1+1}^K \left(1 + \sum_{\substack{i \neq k \\ i \neq j}}^K p_{i,n} |h_{ki,n}|^2 + p_{j,n} |h_{kj,n}|^2 \right). \quad (3.28)$$

Here, the interference power from other transmitter except from j -th EH transmitter $\sum_{\substack{i \neq k \\ i \neq j}}^K p_{i,n} |h_{ki,n}|^2$ for $n = 1, \dots, N$ is unknown to j -th EH transmitter. Therefore, the optimal subcarrier index $\bar{n}^{(j)}$ at j -th EH transmitter is chosen such that Eq. (3.27) is minimized with respect to the interference power from j -th EH transmitter given as

$$\bar{n}^{(j)} = \arg \min_{n \in \mathbf{N}^{(j)}} \sum_{k=K_1+1}^K p_{j,n} |h_{kj,n}|^2. \quad (3.29)$$

In addition, condition Eq. (3.23) is satisfied if $\sum_{k=1}^{K_1} p_{j,\bar{n}^{(j)}} |h_{kj,\bar{n}^{(j)}}|^2 = \bar{E}^{(j)}$. Hence, $p_{j,\bar{n}^{(j)}} = \bar{E}^{(j)} / (\sum_{k=1}^{K_1} |h_{kj,\bar{n}^{(j)}}|^2) = \bar{E}^{(j)} / |\tilde{h}_{11,\bar{n}^{(j)}}^{(j)}|^2$. Accordingly, Eq. (3.29) can be written as

$$\bar{n}^{(j)} = \arg \min_{n \in \mathbf{N}^{(j)}} \bar{E}^{(j)} \frac{|\tilde{h}_{21,n}^{(j)}|^2}{|\tilde{h}_{11,n}^{(j)}|^2}. \quad (3.30)$$

That is from Eq. (3.30), the optimal one subcarrier selection strategy at j -th EH transmitter can be written as in Eq. (3.24). $\square$

Based on Proposition 3, the subcarrier index $\bar{n}^{(j)}$ at j -th EH transmitter for SLREC is chosen as follows:

1. Possible subcarriers at j -th EH transmitter are chosen such that $P |\tilde{h}_{11,n}^{(j)}|^2 \geq \frac{\bar{E} - \bar{E}_{12}}{K_1}$, where $\bar{E}_{12}$ is the total harvested energy from the ID transmitters. In order to find the

possible subcarriers at j-th EH transmitter for the system with distributed transmission strategy, a uniform energy requirement from the EH transmitters is applied where $\bar{E} - \bar{E}_{12}$ is divided equally among K_1 EH transmitters.

2. Among the possible subcarriers at j-th EH transmitter, $\bar{n}^{(j)}$ is the subcarrier index with the maximum ratio of $\frac{|\tilde{h}_{11,n}^{(j)}|^2}{|\tilde{h}_{12,n}^{(j)}|^2}$.

Remark 3.3: If the interference signal can be cancelled at the ID receivers, then the interference term can be set to 0 (i.e., $\mathbf{R}_{-j} = \mathbf{I}_N$). To minimize the required energy from the ID transmitters, EH transmitters should allocate their transmit power at the MaxCG for any given energy constraint in Eq. (3.13). Hence, more transmit power can be allocated at the ID transmitters to maximize the achievable sum-rate.

Remark 3.4: From Eq. (3.27), the optimal subcarrier selection at j-th EH transmitter should depend on the interference created by other EH transmitters and ID transmitters. Then, if the centralized optimization is available, the transmitters could choose the subcarrier selection more adequately by accounting for other transmitters transmission strategies.

Based on the proposed one-subcarrier selection strategies at the EH transmitter (MaxCG, MinCG, SLER and SLREC), the optimization problem for the achievable rate-energy region in Eq. (3.13) is formulated and accordingly, an algorithm that iteratively optimize the power allocation at the ID transmitters and EH transmitters is proposed. The EH transmitter can choose to transmit its signal at a subcarrier corresponding to MaxCG, MinCG, SLER or SLREC. In this paper, it is assumed that each EH transmitter choose the same subcarrier selection strategy.

The optimization problem for the achievable rate-energy region in Eq. (3.13) can be written as

$$\begin{aligned}
 (P3.1) \quad & \text{maximize} \quad \sum_{k=K_1+1}^K \log_2 \det(\mathbf{I}_N + \mathbf{H}_{kk} \mathbf{R}_{-k}^{-1} \mathbf{H}_{kk}^H \mathbf{P}_k) \\
 & \text{subject to} \quad \sum_{k=1}^K \sum_{i=1}^{K_1} \text{tr}(\mathbf{H}_{ik} \mathbf{P}_k \mathbf{H}_{ik}^H) \geq \bar{E} \\
 & \quad \text{tr}(\mathbf{P}_k) \leq P \\
 & \quad \mathbf{P}_k \geq \mathbf{0} \text{ for } k = 1, \dots, K.
 \end{aligned} \tag{3.31}$$

Note that the optimization can also be formulated as maximizing the total harvested energy subject to the rate constraint.

From the objective function of Eq. (3.31), it can be found that the gradient of the achievable rate with respect to the power allocation at the EH transmitters is negative. Therefore, an iterative algorithm that optimizes the power allocation at the EH transmitters

is developed based on the steepest descent method. Given a matrix $\mathbf{A}(x)$, $\frac{d}{dx} \ln \det \mathbf{A}(x) = \text{tr} \left(\mathbf{A}(x)^{-1} \frac{d\mathbf{A}(x)}{dx} \right)$ [76]. By letting the power vector $\mathbf{P}$ such that

$$\mathbf{P} = [p_{1,\bar{n}^{(1)}}, \dots, p_{K_1,\bar{n}^{(K_1)}}]^T \in \mathbb{R}^{K_1 \times 1}, \quad (3.32)$$

where $p_{j,\bar{n}^{(j)}}$ is the power for subcarrier $\bar{n}^{(j)}$ (either MaxCG, MinCG, SLER or SLREC subcarrier selection strategy) at j-th EH transmitter, the gradient of the achievable rate in Eq. (3.31) with respect to $\mathbf{P}$ is given by

$$\nabla J(\mathbf{P}, \mathbf{P}_{K_1+1}, \dots, \mathbf{P}_K) \in \mathbb{R}^{K_1 \times 1} = [\nabla J_1, \dots, \nabla J_{K_1}]^T = \begin{bmatrix} \frac{1}{\ln(2)} \text{tr} \left(\sum_{i=K_1+1}^K ((\mathbf{D}_m)^{-1} - (\mathbf{D}_k)^{-1}) \Omega_{i1} \right) \\ \vdots \\ \frac{1}{\ln(2)} \text{tr} \left(\sum_{i=K_1+1}^K ((\mathbf{D}_m)^{-1} - (\mathbf{D}_k)^{-1}) \Omega_{iK_1} \right) \end{bmatrix}, \quad (3.33)$$

where $\mathbf{D}_m = \mathbf{R}_{-i} + \mathbf{H}_{ii} \mathbf{P}_i \mathbf{H}_{ii}^H$, $\mathbf{D}_k = \mathbf{R}_{-i}$ and $\Omega_{ij} = [\mathbf{H}_{ij}]_{\bar{n}^{(j)}} [\mathbf{H}_{ij}]_{\bar{n}^{(j)}}^H$.

Given $\mathbf{P}$ and total harvested energy from the EH transmitters $\bar{E}_{11} = \sum_{k=1}^{K_1} \sum_{i=1}^{K_1} \text{tr}(\mathbf{H}_{ik} \mathbf{P}_k \mathbf{H}_{ik}^H)$, the transmission strategy at the ID transmitters are then optimized. The transmission strategy at the ID transmitters can be solved by using Lagrangian method. The Lagrangian of the optimization problem (P 3.1) is given by

$$\begin{aligned} \mathcal{L}(\mathbf{P}_j, \lambda, \mu_j, j = K_1 + 1, \dots, K) = & \sum_{j=K_1+1}^K \log_2 \det(\mathbf{I}_N + \mathbf{H}_{jj} \mathbf{R}_{-j}^{-1} \mathbf{H}_{jj}^H \mathbf{P}_j) - \mu_j (\text{tr}(\mathbf{P}_j) - P) \\ & + \lambda \left(\sum_{j=K_1+1}^K \sum_{k=1}^{K_1} \text{tr}(\mathbf{H}_{kj} \mathbf{P}_j \mathbf{H}_{kj}^H) - (\bar{E} - \bar{E}_{11}) \right), \end{aligned} \quad (3.34)$$

where λ and μ_j are the Lagrangian dual variables for the transmit power constraint at j-th ID transmitter and the energy constraint, respectively. The dual function is given by

$$g(\lambda, \mu_j, j = K_1 + 1, \dots, K) = \max_{\mathbf{P}_j \geq 0} \mathcal{L}(\mathbf{P}_j, \lambda, \mu_j) \quad (3.35)$$

with the dual problem is then defined as $\min_{\lambda, \mu_j} g(\lambda, \mu_j)$ subject to $\lambda \geq 0$ and $\mu_j \geq 0$ for $j = K_1 + 1, \dots, K$. The minimization of $g(\lambda, \mu_j)$ over λ and μ_j can be solved using the subgradient method, where the subgradient of $g(\lambda, \mu_{K_1+1}, \dots, \mu_K)$ is given by $(\bar{E}_{12} - (\bar{E} -$

Algorithm 1 Optimization Algorithm for the ID transmitters

```

1: Initialize:  $\lambda$  and  $\mu_j$ .
2: repeat
3:   repeat
4:     for  $j = K_1 + 1, \dots, K$  do
5:       Update  $\mathbf{R}_{-j} = \mathbf{I}_N + \bar{\mathbf{R}}_{-j} + \sum_{k=K_1+1, k \neq j}^K \mathbf{H}_{jk} \mathbf{P}_k \mathbf{H}_{jk}^H$ .
6:       Compute  $\mathbf{P}_j$  as in Eq. (3.37) for given  $\lambda$ ,  $\mu_j$  and  $\mathbf{R}_{-j}$ .
7:     end for
8:   until  $\mathbf{P}_j$  for  $j = K_1 + 1, \dots, K$  converges.
9:   Update  $\lambda$  and  $\mu_j$  for given  $\mathbf{P}_j$  and  $\bar{E}_{11}$  using the subgradient method.
10: until  $\lambda$  and  $\mu_j$  converges.

```

$\bar{E}_{11}), P - \text{tr}(\mathbf{P}_{K_1+1}), \dots, P - \text{tr}(\mathbf{P}_K))$ [71]. For given λ and μ_j , Eq. (3.34) can be written as

$$\begin{aligned}
\max_{\mathbf{P}_j \geq 0} \mathcal{L}(\mathbf{P}_j, \lambda, \mu_j, j = K_1 + 1, \dots, K) &= \sum_{j=K_1+1}^K \log_2 \det(\mathbf{I}_N + \mathbf{H}_{jj} \mathbf{R}_{-j}^{-1} \mathbf{H}_{jj}^H \mathbf{P}_j) - \mu_j \sum_{j=K_1+1}^K \text{tr}(\mathbf{P}_j) + \\
&\quad \lambda \left(\sum_{j=K_1+1}^K \sum_{k=1}^{K_1} \text{tr}(\mathbf{H}_{kj} \mathbf{P}_j \mathbf{H}_{kj}^H) \right) \\
&= \sum_{j=K_1+1}^K (\log_2 \det(\mathbf{I}_N + \mathbf{H}_{jj} \mathbf{R}_{-j}^{-1} \mathbf{H}_{jj}^H \mathbf{P}_j) - \text{tr}(\mathbf{A}_j \mathbf{P}_j)),
\end{aligned} \tag{3.36}$$

where $\mathbf{A}_j = \mu_j \mathbf{I}_N - \lambda (\tilde{\mathbf{H}}_{12}^{(j)})^H \tilde{\mathbf{H}}_{12}^{(j)}$ subject to $\mathbf{A}_j \geq \mathbf{0}$. Note that Eq. (3.36) can be seen as a conventional rate maximization in the OFDM IFC subject to the power constraint. Accordingly, the power allocation at the ID transmitters are optimized by using the iterative water-filling algorithm [71, 72]. At each iteration, j -th ID transmitter updates its power allocation as

$$\mathbf{P}_j = \left(\gamma - \mathbf{R}_{-j} (\mathbf{H}_{jj} \mathbf{H}_{jj}^H)^{-1} \right)^+, \tag{3.37}$$

where $\gamma = (\ln 2 (\mathbf{A}_j))^{-1}$. Here, we can optimize the power allocation at the ID transmitters using the KKT method. However, an additional information of interference pricing is required. The iterative algorithm to optimize the power allocation at the ID transmitters is given in Algorithm 1.

Let $\omega = [|\tilde{h}_{11, \bar{n}(1)}^{(1)}|^2, \dots, |\tilde{h}_{11, \bar{n}(K_1)}^{(K_1)}|^2]$. The iterative algorithm to identify the achievable rate-energy (R-E) region is detailed in Algorithm 2. The step size Δ is given by a fixed value in

Algorithm 2 Identification of the R-E region

- 1: **Initialize:** $\mathbf{P}^{(0)}$ in (28) such that $\mathbf{P}^{(0)} = [P, \dots, P]^T$, $\bar{E}_{11}^{(0)} = \omega \mathbf{P}^{(0)}$ and $\bar{\mathbf{R}}_{-j}^{(0)} = \sum_{k=1}^{K_1} (\mathbf{P}^{(0)})_k [\mathbf{H}_{jk}]_{\bar{n}^{(k)}} [\mathbf{H}_{jk}]_{\bar{n}^{(k)}}^H$.
 - 2: **for** $i = I_{\max}$ **do**
 - 3: Solve optimization problem (P 3.1) for $\mathbf{P}_{(K_1+1)}^{(i)}, \dots, \mathbf{P}_K^{(i)}$ as in Algorithm 1 subject to $\bar{E}_{11}^{(i)}$ and $\bar{\mathbf{R}}_{-j}^{(i)}$.
 - 4: **if** $\bar{E}_{11}^{(i)} + \sum_{j=K_1+1}^K \sum_{k=1}^{K_1} \text{tr}(\mathbf{H}_{kj} \mathbf{P}_j^{(i)} \mathbf{H}_{kj}^H) > \bar{E}$ **then** update ∇J as in Eq. (3.33)
 - 5: **end if**.
 - 6: **if** $\nabla J_k = 0$ **then** set $\nabla J_k = \alpha$, where α is a fixed negative parameter.
 - 7: **end if**
 - 8: Update $\mathbf{P}^{(i+1)}$ such that $\mathbf{P}^{(i+1)} = \min(\max(\mathbf{P}^{(i)} + \Delta \nabla J, 0), P)$, where Δ is the step size.
 - 9: Update $\bar{E}_{11}^{(i+1)}$ and $\bar{\mathbf{R}}_{-j}^{(i+1)}$ with $\mathbf{P}^{(i+1)}$ similar to Step 1).
 - 10: **end for**
 - 11: The achievable rate-energy is determined as $(R, E) = (\sum_{j=K_1+1}^K \log_2 \det(\mathbf{I}_N + \mathbf{H}_{jj}(\mathbf{R}_{-j}^{(I_{\max}+1)})^{-1} \mathbf{H}_{jj}^H \mathbf{P}_j^{(I_{\max}+1)}), \bar{E}_{11}^{(I_{\max}+1)} + \sum_{j=K_1+1}^K \sum_{k=1}^{K_1} \text{tr}(\mathbf{H}_{kj} \mathbf{P}_j^{(I_{\max}+1)} \mathbf{H}_{kj}^H))$.
-

$[0, \Delta_{\max}]$, where the maximum allowable step size $\Delta_{\max}$ is given by

$$\Delta_{\max} = \frac{\bar{E} - \sum_{j=1}^K \sum_{k=1}^{K_1} \text{tr}(\mathbf{H}_{kj} \mathbf{P}_j^{(i)} \mathbf{H}_{kj}^H) - \omega \mathbf{P}^{(i)}}{\omega \nabla J(\mathbf{P}^{(i)}, \mathbf{P}_{K_1+1}^{(i)}, \dots, \mathbf{P}_K^{(i)})}. \quad (3.38)$$

In Step (8) of Algorithm 2, when the total harvested energy is larger than the energy constraint $\bar{E}$, EH transmitters reduce their transmit power so that the interference to the ID receivers is minimized. But, when the ID transmitters do not allocate their transmit power at the subcarrier chosen by k-th EH transmitter, then $\nabla J_k = 0$. Accordingly, k-th EH transmitter will decrease its transmit power at subcarrier $\bar{n}^{(k)}$ by setting $\nabla J_k = \alpha$, where α is a negative parameter.

Note that Algorithm 2 always has negative gradient and positive step size such that the power allocation at the EH transmitters is monotonically decreasing. In addition, given the power allocation at the EH transmitters, the power allocation of the iterative waterfilling in Algorithm 1 always converges to the Nash Equilibrium[71, 72]. Therefore, it can be concluded that Algorithm 2 always converges.

Remark 3.5: Depending on the *water level*, ID transmitters do not allocate their transmit power to the subcarrier with poor channel gain to noise plus interference ratio. If the subcarrier chosen at k-th EH transmitter is the subcarrier unutilized at the ID transmitters, then $\nabla J_k = 0$. Here, the interference from k-th EH transmitter does not degrade the achievable

sum-rate performance. Therefore, if the EH transmitter knows the subcarrier unutilized by the ID transmitters, then the EH transmitter can transfer energy without affecting the achievable sum-rate performance.

3.3 Information Sharing of Unutilized Subcarrier indices by ID Users

From Remark 3.5, with the cooperation among the ID transmitters to inform the EH transmitters of the indices of the subcarriers unutilized by the ID transmitters, the achievable R-E region can be further enlarged. Let $\mathbf{Z}$ be the set of unutilized subcarrier indices at all ID-users, i.e., $\mathbf{Z} = \{n_u | p_{k,n_u} = 0 \text{ for all } k = K_1 + 1, \dots, K\}$. Then, the achievable sum-rate R in Eq. (3.13) can be written as:

$$R = \sum_{n \notin \mathbf{Z}}^N \sum_{k=K_1+1}^K \log_2 \left(1 + p_{k,n} |h_{kk,n}|^2 \left(1 + \sum_{j \neq k}^K p_{j,n} |h_{kj,n}|^2 \right)^{-1} \right). \quad (3.39)$$

That is, the achievable sum-rate is not degraded by the interference signal on the subcarrier unutilized by the ID transceivers. Accordingly, with the information of unutilized subcarrier indices from the ID transmitters to the EH transmitters, EH transmitters can allocate their transmit power to the subcarrier unutilized at the ID transceivers. The one-subcarrier and two-subcarriers transmission strategies associated with the unutilized subcarrier at the ID transceivers are proposed.

The subcarrier index $\bar{n}^{(j)}$ for SLREC with information sharing of the unutilized subcarrier indices and one-subcarrier transmission strategy is chosen as: According to the energy constraint in Eq. (3.13), possible subcarriers at j -th EH transmitter are chosen such that $P |\tilde{h}_{11,n}^{(j)}|^2 \geq \frac{\bar{E} - \bar{E}_{12}}{K_1}$ is satisfied. Among them,

1. If there are unutilized subcarriers at the ID users, choose $\bar{n}^{(j)}$ as the subcarrier index with the largest channel gain of $|\tilde{h}_{11,n}^{(j)}|^2$ among them.
2. Else, choose $\bar{n}$ as the subcarrier index with the maximum ratio of $\frac{|\tilde{h}_{11,n}^{(j)}|^2}{|\tilde{h}_{21,n}^{(j)}|^2}$ (i. e., SLREC subcarrier selection method).

Then, the iterative algorithm to identify the achievable rate-energy region for SLREC with information sharing of unutilized subcarrier indices and one-subcarrier transmission strategy is given in Algorithm 2.

An upper-bound on the achievable R-E region with the non-cooperative strategy and a single subcarrier selection strategy is obtained by conducting an Exhaustive Search over all possible subcarriers. Specifically, the Exhaustive Search algorithm performs a full search for the optimal single-subcarrier selection and power allocation for the EH transmitters, in which the ID transmission strategy (i.e. ID transmitter's power allocation) is known at the energy transmitters. For a given energy constraint, the achievable rate is evaluated from Algorithm 1 for each subcarriers at the EH transmitters. Finally, the EH transmitters allocate its transmit power at the subcarrier with the highest achievable rate. Note that Exhaustive Search algorithm requires transmitter cooperation with global CSIT knowledge (CSI of the links between all transmitters and all receivers).

Allocating the EH power on a single subcarrier unutilized at the ID transmitters may not satisfy the required harvested energy. Therefore, a two-subcarrier transmission strategy (subcarrier indices $\bar{n}^{(j)}$ and $\bar{n}'^{(j)}$ at j-th EH transmitter) is proposed where the EH power is only partially allocated to the subcarrier unutilized by the ID-users. The subcarrier index $\bar{n}^{(j)}$ is chosen based on the SLREC subcarrier selection method and subcarrier index $\bar{n}'^{(j)}$ is the subcarrier index with the largest channel gain of $|\tilde{h}_{11,n}^{(j)}|^2$ among the unutilized subcarriers at the ID-users. Here, it is defined the power vector for the EH transmitters at subcarrier $\bar{n}^{(j)}$ as

$$\mathbf{P} = [p_{1,\bar{n}^{(1)}}, \dots, p_{K_1,\bar{n}^{(K_1)}}]^T \in \mathbb{R}^{K_1 \times 1} \quad (3.40)$$

and the power vector for the EH transmitters at subcarrier $\bar{n}'^{(j)}$ as

$$\mathbf{P}' = [p_{1,\bar{n}'^{(1)}}, \dots, p_{K_1,\bar{n}'^{(K_1)}}]^T \in \mathbb{R}^{K_1 \times 1}. \quad (3.41)$$

Let $\omega' = [|\tilde{h}_{11,\bar{n}^{(1)}}^{(1)}|^2, \dots, |\tilde{h}_{11,\bar{n}^{(K_1)}}^{(K_1)}|^2]$. The iterative algorithm to identify the achievable rate-energy region for SLREC with information sharing of the unutilized subcarrier indices and two-subcarriers transmission strategy is given in Algorithm 3.

The step size Δ' is given by a fixed value in $[0, \Delta'_{max}]$, where Δ'_{max} is given by

$$\Delta'_{max} = \frac{\bar{E} - \sum_{j=1}^K \sum_{k=1}^{K_1} \text{tr}(\mathbf{H}_{kj} \mathbf{P}_j^{(i)} \mathbf{H}_{kj}^H) - [\alpha, \dots, \alpha] \mathbf{P}'^{(i)}}{\omega' [\alpha, \dots, \alpha]^T}. \quad (3.42)$$

In Step (7), if the total harvested energy is larger than $\bar{E}$, the EH transmitters reduce their transmit power at subcarrier $\bar{n}$, so that the interference to the ID receivers is minimized. Then, the remaining power is allocated to subcarrier $\bar{n}'$.

Algorithm 3 R-E region for two-subcarriers transmission strategy

-
- 1: **Initialize:** $\mathbf{P}^{(0)} = [P, \dots, P]^T$, $\mathbf{P}'^{(0)} = [0, \dots, 0]^T$, $\bar{E}_{11}^{(0)} = \omega \mathbf{P}^{(0)} + \omega' \mathbf{P}'^{(0)}$ and $\bar{\mathbf{R}}_{-j}^{(0)} = \sum_{k=1}^{K_1} (\mathbf{P}^{(0)})_k [\mathbf{H}_{jk}]_{\bar{n}(k)} [\mathbf{H}_{jk}]_{\bar{n}(k)}^H + \sum_{k=1}^{K_1} (\mathbf{P}'^{(0)})_k [\mathbf{H}_{jk}]_{\bar{n}'(k)} [\mathbf{H}_{jk}]_{\bar{n}'(k)}^H$.
 - 2: **for** $i = I_{\max}$ **do**
 - 3: **if** $\bar{E}_{11}^{(i)} + \sum_{j=K_1+1}^K \sum_{k=1}^{K_1} \text{tr}(\mathbf{H}_{kj} \mathbf{P}_j^{(i)} \mathbf{H}_{kj}^H) > \bar{E}$ and $\mathbf{P}^{(i)} \neq \mathbf{0}$ **then** update ∇J as in Eq. (3.33)
 - 4: **end if.**
 - 5: **if** $\nabla J_k = 0$ **then** set $\nabla J_k = \alpha$.
 - 6: **end if**
 - 7: Update $\mathbf{P}^{(i+1)}$ such that $\mathbf{P}^{(i+1)} = \min(\max(\mathbf{P}^{(i)} + \Delta \nabla J, 0), P)$ and $\mathbf{P}'^{(i+1)} = P - \mathbf{P}^{(i+1)}$.
 - 8: Update $\bar{E}_{11}^{(i+1)}$ and $\bar{\mathbf{R}}_{-j}^{(i+1)}$ similar to Step 1.
 - 9: **end for**
 - 10: The achievable rate-energy is determined as $(R, E) = (\sum_{j=K_1+1}^K \log_2 \det(\mathbf{I}_N + \mathbf{H}_{jj} (\mathbf{R}_{-j}^{(I_{\max}+1)})^{-1} \mathbf{H}_{jj}^H \mathbf{P}_j^{(I_{\max}+1)}), \bar{E}_{11}^{(I_{\max}+1)} + \sum_{j=K_1+1}^K \sum_{k=1}^{K_1} \text{tr}(\mathbf{H}_{kj} \mathbf{P}_j^{(I_{\max}+1)} \mathbf{H}_{kj}^H))$.
-

Note that the power allocation at the EH transmitters at subcarrier index $\bar{n}'$ in Algorithm 3 does not degrade the achievable rate in the objective function of optimization problem (P 3.1). Therefore, since Algorithm 2 converges, we can conclude that Algorithm 3 also converges.

3.4 Extension: Transmission Strategy with global CSI

The capacity of the Gaussian IFC at high SNR is dominated by the DoF of the ID transceivers [73, 74]. Here, the DoF of the multiple ID-user is a pre-log factor of the capacity at high SNR. That is, at high SNR, the achievable sum-rate scales with the total achievable DoF η , such that $R \propto \eta \log(\text{SNR})$. With the knowledge of the global CSI at the transmitters, the trade-off region of DoF and harvested energy is identified. Accordingly, the power allocation that maximizes the DoF at the ID users for given energy constraint is discussed. Let consider $K_1 = K - 1$ (i.e., one ID and multiple EH transceiver pairs), but it can be directly extended to general K_1 and K . It is noted that here interference alignment can achieve larger total DoF in an OFDM IFC. However, interference alignment would require transmitter precoding and interference suppression which is out of scope of this thesis. The optimization problem for the achievable DoF-Energy region is given by

$$\begin{aligned}
 (P3.2) \quad & \text{maximize} \quad \text{DoF} \\
 & \text{subject to} \quad \sum_{j=1}^{K_1} \sum_{k=1}^K \text{tr} \left(\mathbf{H}_{jk} \mathbf{P}_k \mathbf{H}_{jk}^H \right) \geq \bar{E} \\
 & \quad \text{tr}(\mathbf{P}_k) \leq P \\
 & \quad \mathbf{P}_k \geq \mathbf{0} \text{ for } k = 1, 2, \dots, K,
 \end{aligned} \tag{3.43}$$

where $\bar{E}$ is the required total harvested energy. The total harvested energy in optimization problem (P 3.2) can be written as

$$\begin{aligned}
 E &= \sum_{k=1}^{K_1} \text{tr} \left(\tilde{\mathbf{H}}_{11}^{(k)} \mathbf{P}_k (\tilde{\mathbf{H}}_{11}^{(k)})^H \right) + \text{tr} \left(\tilde{\mathbf{H}}_{12}^{(K)} \mathbf{P}_K (\tilde{\mathbf{H}}_{12}^{(K)})^H \right) \\
 &= \bar{E}_{11} + \bar{E}_{12}.
 \end{aligned} \tag{3.44}$$

From optimization problem (P 3.2), we first discuss the power allocation at the EH transmitters. Note that the DoF can be regarded as the number of interference-free subcarriers available at the ID transceiver because $R \propto \sum_{n \in \mathbf{Z}_{free}} \log(P)$ at high SNR, where $\mathbf{Z}_{free}$ is an interference-free subcarrier index set. To maximize the number of *interference-free* subcarriers available at the ID transceiver, denoted as $\tilde{N}$, the subcarriers where each EH transmitter allocates its transmit power should be overlapped with other EH transmitters as much as possible (i.e., the subcarriers used by energy transmitters should be aligned). However, because subcarrier indices having the maximum channel gain are different over EH transmitters, the subcarrier alignment into a small number of subcarriers may not satisfy the harvesting energy constraint. Accordingly, to find the maximum $\tilde{N}$ satisfying the harvesting energy constraint, the exhaustive search method is utilized. Note that the total harvested energy for given $\tilde{N}$ is maximized when each EH transmitter allocate its maximum power on a single subcarrier, where the subcarrier chosen is the subcarrier with the maximum channel gain among the interference subcarriers at the ID transceiver.

From Eq. (3.44), the required harvested energy from the EH transmitters (i.e., $\bar{E}_{11}$) is minimized when $\bar{E}_{12}$ is maximized. In addition, the DoF at the ID user is maximized when the ID transmitter allocates its transmit power at all interference-free subcarriers (i.e., $\mathbf{Z}_{free}$). However, maximum harvested energy by the ID transmitter is given by $\bar{E}_{12,max} = P |\tilde{h}_{12,n'}^{(K)}|^2$, where $n' = \arg \max_{n=1,\dots,N} |\tilde{h}_{12,n}^{(k)}|^2$. Here, $\bar{E}_{12}$ and the DoF at the ID user can be maximized simultaneously when $p_{K,n'} \rightarrow P$ and $p_{K,n} \rightarrow 0$ for $n \neq n'$.

The algorithm to identify the upper bound of DoF-energy region is given in Algorithm 4

Algorithm 4 Algorithm for DoF-Energy Region

- 1: Find the maximum $\tilde{N}$ satisfying the energy constraint by using the exhaustive search method.

Here, EH transmitters allocate maximum power on the subcarrier with the maximum channel gain among the interference subcarrier at the ID transceiver. The power allocation at the ID transmitter is given as $p_{K,n'} = P$ and $p_{K,n} = 0$ for $n \neq n'$.

- 2: Update DoF= $\min(\tilde{N}, 0)$ and E as in (3.44).

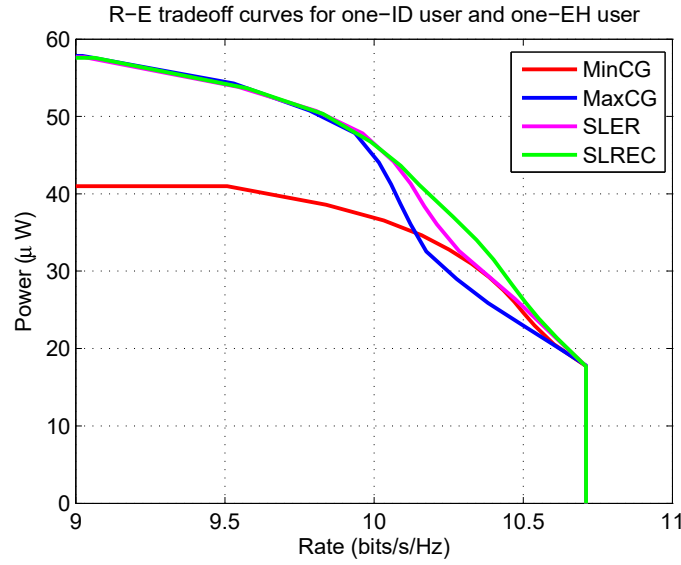


Fig. 3.6 R-E tradeoff for MaxCG, MinCG, SLER and SLREC for one ID-user and one EH-user.

3.5 Simulation Results

This section presents the simulation results. It is assumed that the path loss is $10^{-3/2}$ (with power path loss exponent and distance is 3 and 10m, respectively) such that the channel $\mathbf{H}_{ik} = 10^{-3/2} \tilde{\mathbf{H}}_{ik}$. $\tilde{\mathbf{H}}_{ik} = \text{diag}(\mathbf{F}[\mathbf{h}_{ik} \mathbf{0}_{N-L}]^T)$, where $\mathbf{h}_{ik} \in \mathbb{C}^{1 \times L}$ is the multipath channel of length $L = 3$, the frequency selective fading channel is modeled using three-tap exponentially distributed power profile and $\mathbf{F} \in \mathbb{C}^{N \times N}$ is the normalized Fourier matrix. Here, $\delta_{i,k} = 1$ and $\delta_{i,k} = 0.8$ for direct link ($i = k$) and cross link ($i \neq k$), respectively. In addition, the maximum transmit power, P is set to $50mW$, the noise power at each subcarrier is set to $1\mu W$ and the number of subcarriers N is set to 8.

Figure 3.6, Figure 3.7 and Figure 3.8 show the R-E tradeoff for a two user and K-user with $K_1 = [4, 3]$ and $K = 5$. The EH transmitters allocate their transmit power on a single subcarrier based on MaxCG, MinCG, SLER and SLREC subcarrier selection methods as described in section (3.2.5) and the ID transmitter(s) optimizes its transmit power as in

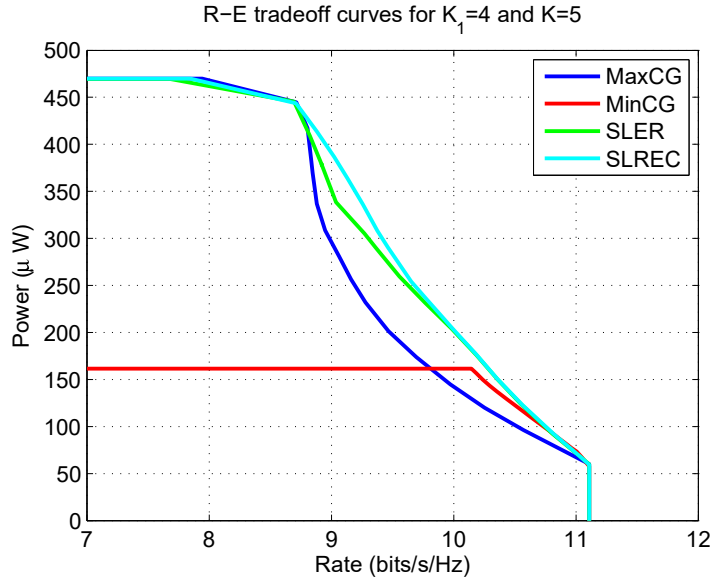


Fig. 3.7 R-E tradeoff for MaxCG, MinCG, SLER and SLREC when $K_1 = 4$ and $K = 5$.

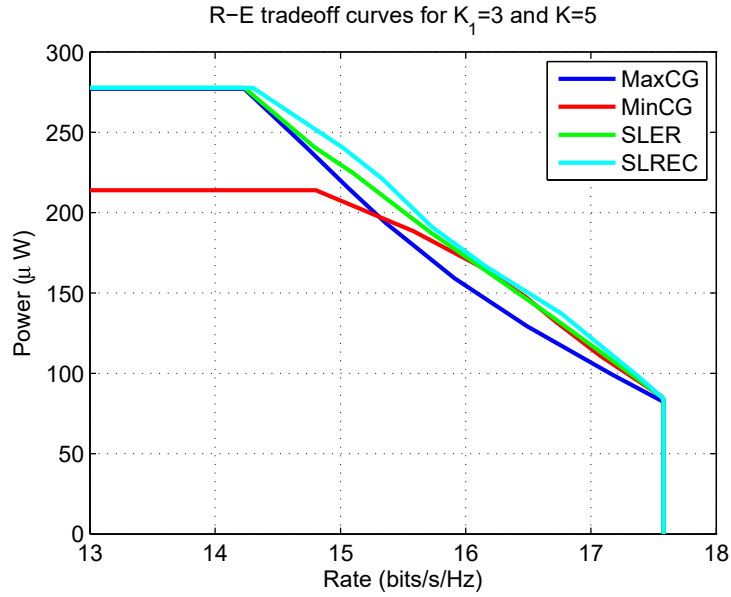


Fig. 3.8 R-E tradeoff for MaxCG, MinCG, SLER and SLREC when $K_1 = 3$ and $K = 5$.

Algorithm 1. The achievable rate-energy region is determined based on Algorithm 2. The simulation result shows that the MinCG increases the achievable rate at the ID receiver(s) while the MaxCG achieves a larger harvested energy at the EH receivers. The R-E region for SLER covers both MaxCG and MinCG R-E regions. In addition, it can be found that the proposed SLREC subcarrier selection strategy achieves a larger R-E region than SLER. Here, SLREC minimizes the interference gain at the ID receiver(s) while energy receivers harvest

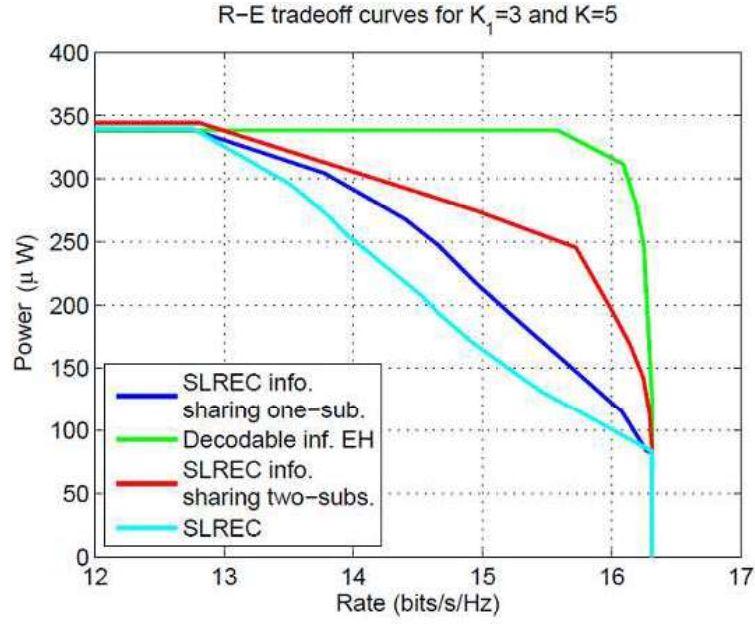


Fig. 3.9 R-E tradeoff for SLREC with and without information sharing of unutilized subcarrier indices by ID users and for Decodable interference from EH transmitters when $K_1 = 3$ and $K = 5$.

sufficient energy to meet the energy constraint. From Figure 3.7 and Figure 3.8, it can be found that as the number of EH-users increased, the total harvested energy increased. From Figure 3.6, Figure 3.7 and Figure 3.8, in the region where the harvested energy is less than around $[18, 60, 90]\mu\text{W}$ respectively, the achievable rate is unchanged. That is, because the harvested energy from the ID transmitter(s) is sufficient to meet the energy constraint $\bar{E}$, EH transmitters do not have to transmit their signal that caused interference to the ID-user(s).

Figure 3.9 shows the R-E tradeoff of a K -user OFDM IFC with a low-rate feedback of the unutilized subcarrier indices by ID-users (local CSI with the information of the unutilized subcarrier indices by ID-users) as discussed in section III-E. It is observed that with the low-rate feedback of unutilized subcarrier indices by ID-users, the achievable R-E region is enlarged. Here, because the interference power at the unutilized subcarriers at ID-users do not degrade the achievable sum-rate, EH transmitters can transfer energy without compromising the performance of ID-users. It is observed that by allocating the transmit power on two-subcarriers, where one of them is the unutilized subcarrier at the ID-users, the achievable rate-energy is enlarged. That is, if the total harvested energy is sufficient, EH transmitters reduce their transmit power on the chosen subcarrier based on SLREC. By allocating the remaining transmit power on the chosen unutilized subcarrier at the ID-users, we can further reduce the interference level at ID-users. In contrast, when the harvested energy cannot be satisfied by the unutilized subcarriers at the ID transceiver, one-subcarrier selection strategy

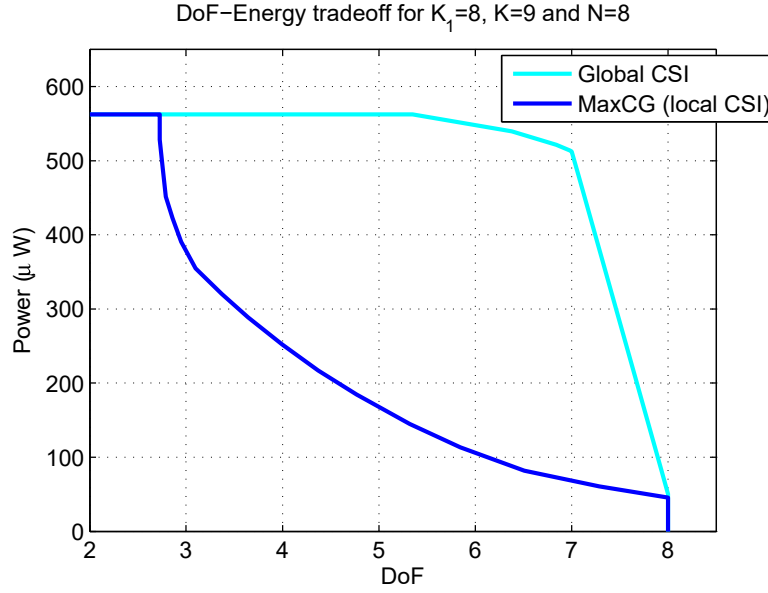


Fig. 3.10 DoF-Energy tradeoff curves for the multiple-EH users with Global CSI and local CSI (MaxCG) when $K_1 = 8$ and $K = 9$.

caused larger interference to the ID receiver. From Figure 3.9, we can find that when the interference signal from the EH transmitters is decodable at the ID receivers, the achievable sum-rate at the ID-users increased. That is, because the interference signal from the EH transmitters can be removed at the ID-users, interference signal from the EH transmitters does not degrade the achievable sum-rate performance. However, when the energy constraint is large such that the harvested energy from the EH transmitters is not sufficient to meet the energy constraint, ID transmitters have to divert their power allocation to satisfy the energy constraint that degrades their achievable sum-rate.

Figure 3.10 shows the trade-off between DoF and harvested energy for a system with $K_1 = 8$ and $K = 9$. It can be found that the achievable DoF at the ID receiver decreases as the target energy increases. As expected, the multiple EH-users with the knowledge of the global CSI at the transmitters has a larger achievable DoF-energy region than local CSI with MaxCG performed at the EH transmitters. Note that in term of DoF and when the transmitters only have the knowledge of local CSI, the MaxCG is the optimal subcarrier strategy at the EH transmitters as it provides the maximum harvested energy for given number of subcarrier interfered by the energy transfer. Here, because the transmitters have the knowledge of the global CSI, EH transmitters and ID transmitter can cooperatively allocate their transmit power so that the achievable DoF at the ID user is maximized for given energy constraint. In the region when the harvested energy is less than around $50\mu\text{W}$, ID-user can achieve maximum DoF. Here, because the harvested energy from the ID transmitter is sufficient

to meet the energy constraint $\bar{E}$, EH transmitters do not transmit any signal. Hence, all subcarriers at the ID-user are available for data transfer without EH transmitters interfering the operation.

3.6 Summary

This chapter has discussed SWIPT in SISO K-user OFDM IFC. First, it is assumed that each transmitter has perfect knowledge of the local CSI of the link between itself and all receivers, such that the energy transmitters optimize their transmission strategy in a distributed manner. We have found that, the necessary condition for the optimal transmission strategy at high SNR is for each EH transmitter to allocate its transmit power on a single subcarrier. Accordingly, the single subcarrier selection strategies are proposed. The subcarrier selection maximizing harvested energy and minimizing interference power at the ID receivers are proposed namely MaxCG and MinCG, respectively. Motivated by [33, 34] and signal to leakage and noise ratio (SLNR) beamforming [77], signal to leakage and energy ratio (SLER) subcarrier selection is proposed. Moreover, an optimal single subcarrier for a distributed transmission strategy namely signal to leakage ratio with energy constraint (SLREC) have been proposed. It is noted that if the interference signal can be decodable at the ID receiver, the optimal subcarrier selection strategy is MaxCG. Based on the proposed single subcarrier selection method, an iterative algorithm is proposed.

It is noted that, depending on the water level, the information transmitter may not allocate their transmit power to the subcarrier with poor channel gain to noise plus interference ratio. If the subcarrier chosen at the EH transmitter is the subcarrier unutilized by the information transmitter, then the power transfer does not degrade the achievable sum-rate. Motivated by this observation, with low-rate feedback from the information transmitter of their unutilized subcarrier to the energy transmitter, the achievable rate-energy region is enlarged. Accordingly, a single subcarrier transmission strategy and two-subcarriers transmission strategy (one of them is the unutilized subcarrier at the ID users) are discussed. It can be found that when EH transmitters allocate some of their transmit power on the unutilized subcarrier at the ID users, higher R-E performance is achieved than a single subcarrier transmission strategy.

Finally, with the knowledge of the global CSI of the link between all transmitters and all receivers at the transmitters, the power allocation strategy maximizing the DoF at the information users for given energy constraint is proposed.

In this chapter, the waveform optimization is based on the linear EH model. This will be revisited in the next chapter where the non-linear EH model will be used.

Chapter 4

Coordinated Waveform Design for Wireless Information and Power Transfer

In the previous chapter, waveform was optimized based on the linear EH model. Recently, [17–20] have introduced an analytical EH model that accounts for the non-linearity of the rectifier. It was observed that by accounting for the non-linearity of the rectifier, it increases RF-DC efficiency and leads to a larger trade-off region than those obtained by ignoring the non-linearity of the rectifier. Moreover, by considering non-linear EH model, a different waveform was observed compared to the linear model.

Motivated by the promising results for the waveform design based on the non-linear EH model in [17–20], we go one step further and consider waveform design across two coordinated transmitters serving two users, as occurring in a two-user interference channel. The waveform design is based on the non-linear EH model introduced in [17, 18]. There are three different scenarios for a two-user interference channel, namely two ID users, two EH users and one ID user and one EH user. The two ID user scenario is not associated with WPT and has been considered in several researches such as [71, 72]. From [18, 20], we can easily leverage the multi-rectenna WPT waveform design and tackle the problem of waveform design in the two EH user scenario.

Here, we expand this line of research and investigate coordinated waveform design for SWIPT in an interference channel with one EH user and one ID user (with their corresponding transmitter respectively denoted as EH transmitter and ID transmitter). Each transmitter and receiver have a single antenna. It is assumed that each transmitter has a perfect knowledge of the global CSI. The EH transmitter transmits deterministic multisine WPT waveform and the ID transmitter transmits OFDM waveform as shown in Figure 4.1. This system is

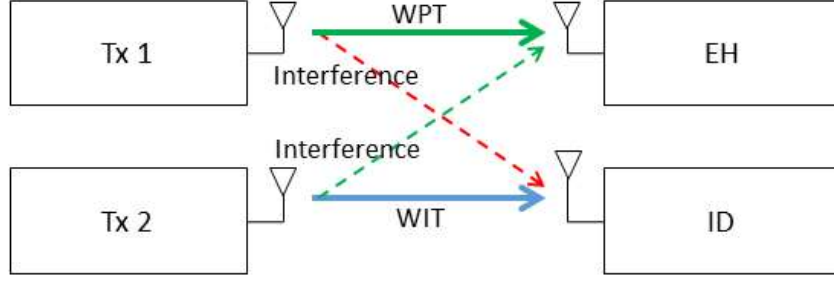


Fig. 4.1 Non-superposed Information and Power Transfer (NIPT).

named Non-superposed Information and Power Transfer (NIPT). NIPT is the distributed version of SWIPT point to point in [12], but it does not require the high hardware complexity power splitter. Here, the coordinated WPT and WIT waveforms as a function of the channel between each transmitter and receivers are designed.

4.1 Non-superposed Information and Power Transfer (NIPT)

In this section, a system model for NIPT as shown in Figure 4.1 is discussed. The first transmitter transmits wireless power using a deterministic multisine waveform and the second transmitter transmits information using a modulated waveform (e.g. OFDM). Each transmitter transmits N deterministic or modulated waveform (i.e., N subcarriers) at the same frequencies (i.e., $\omega_n = \omega_o + n\Delta_\omega$ with frequency spacing $\Delta_\omega = 2\pi\Delta_f$ and $n = 1, \dots, N$) and each transmitter and receiver have a single antenna. It is assumed that each transmitter has a perfect knowledge of the global CSI.

4.1.1 Transmit Signal

Let assume that the first and the second user is an EH and ID user, respectively. The first transmitter transmits deterministic multisine waveform given by

$$x_1(t) = \sum_{n=0}^{N-1} s_{1,n} \cos(\omega_n t + \phi_{1,n}) = \Re\left\{ \sum_{n=0}^{N-1} w_{1,n} e^{j\omega_n t} \right\}, \quad (4.1)$$

where $w_{1,n} = s_{1,n} e^{j\phi_{1,n}}$. Here, $s_{1,n}$ and $\phi_{1,n}$ are the amplitude and the phase of n -th sinewave at frequency ω_n . The magnitudes and phases in Eq. (4.1) can be collected into vectors $\mathbf{s}_1$ and $\boldsymbol{\phi}_1$. The transmit waveform at the first transmitter is subject to the transmit power constraint (at the passband) $\mathcal{E}\{|x_1|^2\} = \frac{1}{2} \|\mathbf{s}_1\|_F^2 \leq P$.

At the second transmitter, the ID transmitter transmits modulated waveform based on OFDM signal given by

$$x_2(t) = \sum_{n=0}^{N-1} \tilde{s}_{2,n} \cos(\omega_n t + \tilde{\phi}_{2,n}) = \Re\left\{ \sum_{n=0}^{N-1} \tilde{x}_{I,n} e^{j\omega_n t} \right\}. \quad (4.2)$$

Here, $\tilde{x}_{I,n} = \tilde{s}_{2,n} e^{j\tilde{\phi}_{2,n}}$ is the information symbol transmitted on subcarrier n . A capacity achieving Gaussian input distribution is considered such that $\tilde{x}_{I,n}$ is i.i.d circularly symmetric complex Gaussian distributed. The magnitudes and phases in Eq. (4.2) can be collected into vector $\tilde{\mathbf{s}}_2$ and $\tilde{\boldsymbol{\phi}}_2$. The transmit signal at the second transmitter is subject to the transmit power constraint $\mathcal{E}\{|x_2|^2\} = \frac{1}{2} \mathcal{E}\{\|\tilde{\mathbf{s}}_2\|_F^2\} \leq P$.

4.1.2 Receive Signal

Let $y_{j,P}(t)$ and $y_{j,I}(t)$ be the received deterministic and modulated waveforms, respectively at j -th receiver. After multipath propagation, the received signal at the j -th receiver can be written as $y_j(t) = y_{j,P}(t) + y_{j,I}(t)$, where from Eq. (4.1) and Eq. (4.2)

$$\begin{aligned} y_j(t) &= \sum_{n=0}^{N-1} s_{1,n} A_{j1,n} \cos(\omega_n t + \psi_{j1,n}) + \sum_{n=0}^{N-1} \tilde{s}_{2,n} A_{j2,n} \cos(\omega_n t + \tilde{\psi}_{j2,n}) \\ &= \Re\left\{ \sum_{n=0}^{N-1} (h_{j1,n} w_{1,n} + h_{j2,n} \tilde{x}_{I,n}) e^{j\omega_n t} \right\} \end{aligned} \quad (4.3)$$

Here, $h_{jk,n} = A_{jk,n} e^{j\tilde{\psi}_{jk,n}} = \sum_{l=0}^{L-1} \alpha_{jk,l} \cos(\omega_n \tau_{jk,l} + \varepsilon_{jk,l})$ is the frequency response of the channel between k -th transmitter and j -th receiver at subcarrier n , $\psi_{jk,n} = \tilde{\psi}_{jk,n} + \phi_{k,n}$ and $\tilde{\psi}_{j2,n} = \tilde{\psi}_{jk,n} + \tilde{\phi}_{2,n}$. $\tau_{jk,l}$, $\varepsilon_{jk,l}$ and $\alpha_{jk,l}$ are the delay, phase and amplitude of the l -th path from k -th transmitter to j -th receiver, respectively.

4.1.3 Energy Harvester

A Non-Linear Model

The EH model has been introduced in Section (2.3.1). By ignoring the terms higher than 4, the DC component in Eq. (2.21) can be written as $i_{out} = k_0 + z_{DC}$, where

$$z_{DC} = \sum_{i=1}^4 k_i R_{ant}^{i/2} \mathcal{A}\{y(t)^i\}. \quad (4.4)$$

Here, $k_i = \frac{i_s}{i!(nv_i)^i}$, $i_s = 5\mu A$, diode ideality factor $n = 1.05$ and $v_t = 25.86mV$. Nevertheless, this approach is general and work for any order. It has been mentioned in [22] that the 4-th order truncation is the minimum order to sufficiently demonstrate the non-linearity of the rectifier process. Here, k_0 is not a function of the received waveform such that optimizing i_{out} is equivalent to optimizing z_{DC} .

The receive signal at first receiver $y_1(t)$ consists of deterministic waveform $y_{1,p}(t)$ and modulated waveform $y_{1,I}(t)$ given in Eq. (4.3). Let us first consider the deterministic waveform. By averaging the deterministic waveform over time

$$\mathcal{A}\{y_{1,p}(t)^2\} = \frac{1}{2} \sum_{n=0}^{N-1} s_{1,1,n}^2 A_{11,n}^2 \quad (4.5)$$

and

$$\mathcal{A}\{y_{1,p}(t)^4\} = \frac{3}{8} \left[\sum_{\substack{n_0, n_1, n_2, n_3 \\ n_0 + n_1 = n_2 + n_3}} \left[\prod_{j=0}^3 s_{1,1,n_j} A_{11,n_j} \right] \cos(\psi_{11,n_0} + \psi_{11,n_1} - \psi_{11,n_2} - \psi_{11,n_3}) \right]. \quad (4.6)$$

The odd terms is neglected since $\mathcal{A}\{y_{1,p}(t)^i\} = 0$ for $i = 1, 3$.

The modulated waveform (i.e., OFDM) exhibits randomness due to the information symbol $\tilde{x}_I$. The modulated waveform can be seen as the multisine waveform for a fixed input symbols $\tilde{x}_I$. Hence, similar expression as Eq. (4.5) and Eq. (4.6) can be written for $\mathcal{A}\{y_{1,I}(t)^2\}$ and $\mathcal{A}\{y_{1,I}(t)^4\}$. However, the information symbol change randomly every symbol duration T . For a given impulse response, the DC component is obtained by averaging over the distribution of the information symbols $\tilde{x}_I$ such that for $\mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^i\}\}$, $i = 2, 4$ is first extracted for given set of amplitudes and phases and then over the randomness of the information symbols $\tilde{x}_I$. The information symbol has a complex Gaussian distribution with $|\tilde{x}_{I,n}|^2$ is exponentially distributed such that $\mathcal{E}\{|\tilde{x}_{I,n}|^2\} = p_{2,n}$ and $\tilde{\phi}_{2,n}$ is uniformly distributed. From the moments of exponential distribution, $\mathcal{E}\{|\tilde{x}_{I,n}|^4\} = 2p_{2,n}^2$. Therefore, by defining $s_{k,n} = \sqrt{p_{k,n}}$,

$$\mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^2\}\} = \frac{1}{2} \mathcal{E}\left\{ \sum_{n=0}^{N-1} \tilde{s}_{2,2,n}^2 A_{12,n}^2 \right\} = \frac{1}{2} \sum_{n=0}^{N-1} \mathcal{E}\{\tilde{s}_{2,2,n}^2\} A_{12,n}^2 = \frac{1}{2} \sum_{n=0}^{N-1} s_{2,2,n}^2 A_{12,n}^2 \quad (4.7)$$

and

$$\begin{aligned}
\mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^4\}\} &= \frac{3}{8} \mathcal{E}\left\{ \sum_{\substack{n_0, n_1, n_2, n_3 \\ n_0 + n_1 = n_2 + n_3}} \tilde{s}_{2,2,n_0} \tilde{s}_{2,2,n_1} \tilde{s}_{2,2,n_2} \tilde{s}_{2,2,n_3} A_{12,n_0} A_{12,n_1} A_{12,n_2} A_{12,n_3} \right. \\
&\quad \left. \cos(\tilde{\psi}_{12,n_0} + \tilde{\psi}_{12,n_1} - \tilde{\psi}_{12,n_2} - \tilde{\psi}_{12,n_3}) \right\} \\
&= \frac{3}{8} \left[\sum_n \mathcal{E}\{\tilde{s}_{2,2,n}^4\} A_{12,n}^4 + 2 \sum_{\substack{n_0, n_1, \\ n_0 \neq n_1}} \mathcal{E}\{\tilde{s}_{2,2,n_0}^2\} \mathcal{E}\{\tilde{s}_{2,2,n_1}^2\} A_{12,n_0}^2 A_{12,n_1}^2 \right] \quad (4.8) \\
&= \frac{3}{4} \sum_{n_0} \sum_{n_1} s_{2,2,n_0}^2 s_{2,2,n_1}^2 A_{12,n_0}^2 A_{12,n_1}^2.
\end{aligned}$$

Then, Eq. (4.4) at the first receiver can be written as

$$\begin{aligned}
z_{DC} &= k_2 R_{ant} \mathcal{A}\{y_{1,P}(t)^2\} + k_4 R_{ant}^2 \mathcal{A}\{y_{1,P}(t)^4\} + k_2 R_{ant} \mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^2\}\} + k_4 R_{ant}^2 \\
&\quad \mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^4\}\} + 6k_4 R_{ant}^2 \mathcal{A}\{y_{1,P}(t)^2\} \mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^2\}\}. \quad (4.9)
\end{aligned}$$

The odd terms is neglected since $\mathcal{E}\{y_{1,I}(t)^i\} = 0$ and $\mathcal{A}\{y_{1,P}(t)^i\} = 0$ for $i = 1, 3$.

A Linear Model

The linear EH model has been introduced in Section (2.1.5). This linear model has been extensively used in SWIPT literature [24–42]. The harvested energy based on the linear model at the first receiver is given by

$$E = \mathcal{E}\{y_1(t)^2\}. \quad (4.10)$$

It is noted that the linear EH model only address the second order term in Eq. (2.21) and Eq. (4.4).

4.1.4 Information Receiver

Perfect Cancellation

The WPT waveform is deterministic such that it can be cancelled at the ID. For perfect cancellation (PC) of the deterministic signal at the second receiver, the baseband received signal at subcarrier n is given as

$$y_{2,n} = h_{22,n} \tilde{x}_{I,n} + v_{2,n}, \quad (4.11)$$

where $v_{2,n}$ is the complex white Gaussian noise at the second receiver and subcarrier n with variance $\sigma_{2,n}^2$. The achievable rate at the second receiver can be written as

$$R(\mathbf{s}_2) = \sum_{n=0}^{N-1} \log_2 \left(1 + \frac{s_{2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2} \right), \quad (4.12)$$

where $\mathbf{s}_k \in \mathbb{R}^{N \times 1}$ for $k = 1, 2$ and $s_{2,n}^2 |h_{22,n}|^2$ is the received power of the baseband signal from the second transmitter and n -th subcarrier for WIT.

No-Cancellation

For no-cancellation (NC) of WPT signal at the ID, the baseband received signal is written as

$$y_{2,n} = h_{22,n} \tilde{x}_{I,n} + h_{21,n} w_{1,n} + v_{2,n}, \quad (4.13)$$

The achievable rate at the second receiver is given as

$$R(\mathbf{s}_2) = \sum_{n=0}^{N-1} \log_2 \left(1 + \frac{s_{2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2 + s_{1,n}^2 |h_{21,n}|^2} \right), \quad (4.14)$$

where $s_{1,n}^2 |h_{21,n}|^2$ is the interference power from the first transmitter.

4.2 Coordinated Waveform Optimization for NIPT

This section discusses the optimization of coordinated waveform for NIPT with and without interference cancellation at the ID. The SWIPT waveform is optimized so as to maximize the rate-energy region of the system.

4.2.1 Perfect Cancellation

The achievable rate-energy (R - z_{DC}) region for PC and non-linear EH model is given as

$$C_{\bar{R}-\bar{z}_{DC}}(P) \triangleq \{ \bar{R}, \bar{z}_{DC} : \bar{R} \leq \sum_{n=1}^N \log_2 \left(1 + \frac{s_{2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2} \right), \quad (4.15)$$

$$\bar{z}_{DC} \leq z_{DC}(\mathbf{s}_1, \mathbf{s}_2, \phi_1), \frac{1}{2} \|\mathbf{s}_1\|_F^2 \leq P, \frac{1}{2} \|\mathbf{s}_2\|_F^2 \leq P \}$$

where z_{DC} is given in Eq. (4.9). In order to maximize z_{DC} , it should be guaranteed that $\psi_{11,n_0} + \psi_{11,n_1} - \psi_{11,n_2} - \psi_{11,n_3}$ in Eq. (4.6) is equal to zero [18]. Note that $\psi_{11,n} = \bar{\psi}_{11,n} +$

$\phi_{1,n}$. Therefore, the DC current is maximized when the phase of WPT waveform is matched to the phase of the channel given as $\phi_{1,n}^* = -\bar{\psi}_{11,n}$.

Given the optimal phase for WPT ϕ_1^* , Eq. (4.15) can be written as

$$\begin{aligned}
 (P\ 4.1) \quad & \max_{\mathbf{s}_1, \mathbf{s}_2} \quad z_{DC}(\mathbf{s}_1, \mathbf{s}_2, \phi_1^*) \\
 \text{s. t.} \quad & \sum_{n=0}^{N-1} \log_2 \left(1 + \frac{s_{2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2} \right) \geq \bar{R} \\
 & \frac{1}{2} \|\mathbf{s}_1\|^2 \leq P, \\
 & \frac{1}{2} \|\mathbf{s}_2\|^2 \leq P.
 \end{aligned} \tag{4.16}$$

The posynomial maximization problem in Eq. (4.16) can be solved by using a Reverse GP approach [18, 78]. Throughout this chapter, we use GP as a general approach to solve the optimization problem, unless stated otherwise.

By introducing t_0 , the objective function is changed into minimizing $1/t_0$ subject to $z_{DC} \geq t_0$. Then, the monomial approximation is applied to the lower bound inequality constraint. Let $u_i(x)$ is the monomial terms in the lower bound inequality constraint such that $f(x) = \sum_i u_i(x)$, the lower bound inequality of a posynomial can be approximated with an upper bound inequality $\prod_i (\frac{u_i(x)}{\alpha_i})^{-\alpha_i}$. Here, α_i as $\alpha_i = u_i(x)/f(x)$ is choosen with $\sum_i \alpha_i = 1$ and $\alpha_i \geq 0$. An iterative algorithm is used to tighten the upper bound approximation such that at each iteration, the standard GP is solved for an updated a_i until convergence [18, 78]. Accordingly, the standard GP problem in Eq. (4.16) is written as

$$\begin{aligned}
 (P\ 4.2) \quad & \min_{\mathbf{s}_1, \mathbf{s}_2, t_0} \quad 1/t_0 \\
 \text{s. t.} \quad & t_0 \prod_{k=1}^K \left(\frac{g_k(\mathbf{s}_1, \mathbf{s}_2)}{\gamma_k} \right)^{-\gamma_k} \leq 1 \\
 & 2^{\bar{R}} \prod_{n=0}^{N-1} \prod_k \left(\frac{g_{nk}(\mathbf{s}_2)}{\gamma_{nk}} \right)^{-\gamma_{nk}} \leq 1 \\
 & \frac{1}{2} \|\mathbf{s}_1\|^2 \leq P, \\
 & \frac{1}{2} \|\mathbf{s}_2\|^2 \leq P.
 \end{aligned} \tag{4.17}$$

where $z_{DC}(\mathbf{s}_1, \mathbf{s}_2) = \sum_{k=1}^K g_k(\mathbf{s}_1, \mathbf{s}_2)$, $1 + \frac{s_{2,n}^2 A_{22,n}^2}{\sigma_{2,n}^2} = \sum_{k=1}^2 g_{nk}(\mathbf{s}_2)$, $\gamma_k, \gamma_{nk} \geq 0$ and $\sum_{k=1}^K \gamma_k, \sum_{k=1}^{K_n} \gamma_{nk} = 1$ for $n = 1, \dots, N$. Optimization problem in Eq. (4.17) is a standard GP problem that can be solved by using CVX. An iterative algorithm to solve Eq. (4.17) is given in Algorithm 1.

Algorithm 5 Iterative Algorithm for PC

```

1: Initialize:  $z_{DC}^{(0)} \leftarrow 0, i \leftarrow 0, \varepsilon, \mathbf{s}_1^{(0)}, \mathbf{s}_2^{(0)}$ 
2: repeat
3:    $i \leftarrow i + 1$ 
4:    $\gamma_k \leftarrow g_k(\mathbf{s}_1^{(i-1)}, \mathbf{s}_2^{(i-1)})/z_{DC}^{(i)}$ 
5:    $\gamma_{nk} \leftarrow g_{nk}(\mathbf{s}_1^{(i-1)})/\sum_{k=1}^2 g_{nk}(\mathbf{s}_1^{(i-1)})$  for  $n = 1, \dots, N$ 
6:    $\mathbf{s}_1^{(i)}, \mathbf{s}_2^{(i)} \leftarrow \arg \min \text{Eq. (4.17)}$ 
7:    $z_{DC}^{(i)} \leftarrow z_{DC}(\mathbf{s}_1^{(i)}, \mathbf{s}_2^{(i)})$ 
8: until  $|z_{DC}^{(i)} - z_{DC}^{(i-1)}| < \varepsilon$ 

```

The solution from the successive convex approximation method in Algorithm 5 converge to a point satisfying the KKT conditions [79, 80].

Remark 4.1: The achievable rate-energy region for the linear EH model and PC is given as

$$C_{\bar{R}-E}(P) \triangleq \{(\bar{R}, E) : \bar{R} \leq \sum_{n=1}^N \log_2 \left(1 + \frac{s_{2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2} \right), E \leq \sum_{n=1}^N (s_{1,n}^2 |h_{11,n}|^2 + s_{2,n}^2 |h_{12,n}|^2), \sum_{n=1}^N s_{k,n}^2 \leq 2P, s_{k,n} \geq 0, k = 1, 2\}. \quad (4.18)$$

By taking only the second order term in z_{DC} , Eq. (4.18) can be solved by using Algorithm 5. Note that, the optimization problem for Eq. (4.18) is convex such that the global optimum solution can be obtained as in Appendix A.

4.2.2 No-Cancellation

The achievable rate-energy region for non-linear model and NC is given as

$$C_{\bar{R}-\bar{z}_{DC}}(P) \triangleq \{(\bar{R}, \bar{z}_{DC}) : \bar{R} \leq \sum_{n=1}^N \log_2 \left(1 + \frac{s_{2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2 + s_{1,n}^2 |h_{21,n}|^2} \right), \bar{z}_{DC} \leq z_{DC}(\mathbf{s}, \phi), \frac{1}{2} \|\mathbf{s}_{1,1}\|_F^2 \leq P, \frac{1}{2} \|\mathbf{s}_{2,2}\|_F^2 \leq P\}. \quad (4.19)$$

Similar to PC, the optimum phase for WPT waveform that maximizes z_{DC} is given as $\phi_{1,n}^* = -\bar{\psi}_{11,n}$. Given the optimum phase ϕ_1^* , Eq. (4.19) can be written as a standard GP

Algorithm 6 Iterative Algorithm for NC

```

1: Initialize:  $z_{DC}^{(0)} \leftarrow 0, i \leftarrow 0, \varepsilon, \mathbf{s}_1^{(0)}, \mathbf{s}_2^{(0)}$ 
2: repeat
3:    $i \leftarrow i + 1$ 
4:    $\gamma_k \leftarrow g_k(\mathbf{s}_1^{(i-1)}, \mathbf{s}_2^{(i-1)}) / z_{DC}^{(i)}$ 
5:    $\gamma_{nk} \leftarrow g_{nk}(\mathbf{s}_1^{(i-1)}, \mathbf{s}_2^{(i-1)}) / \sum_{k=1}^3 g_{nk}(\mathbf{s}_1^{(i-1)}, \mathbf{s}_2^{(i-1)})$  for  $n = 1, \dots, N$ 
6:    $\mathbf{s}_1^{(i)}, \mathbf{s}_2^{(i)} \leftarrow \arg \min \text{Eq. (4.20)}$ 
7:    $z_{DC}^{(i)} \leftarrow z_{DC}(\mathbf{s}_1^{(i)}, \mathbf{s}_2^{(i)})$ 
8: until  $|z_{DC}^{(i)} - z_{DC}^{(i-1)}| < \varepsilon$ 

```

problem

$$\begin{aligned}
(P \ 4.3) \quad & \min_{\mathbf{s}_1, \mathbf{s}_2, t_0} 1/t_0 \\
\text{s. t.} \quad & t_0 \prod_{k=1}^K \left(\frac{g_k(\mathbf{s}_1, \mathbf{s}_2)}{\gamma_k} \right)^{-\gamma_k} \leq 1 \\
& 2^{\bar{R}} \prod_{n=0}^{N-1} (\sigma_{2,n}^2 + s_{1,n}^2 |h_{21,n}|^2) \prod_k^{K_n} \left(\frac{g_{nk}(\mathbf{s}_1, \mathbf{s}_2)}{\gamma_{nk}} \right)^{-\gamma_{nk}} \leq 1 \quad (4.20) \\
& \frac{1}{2} \|\mathbf{s}_1\|^2 \leq P, \\
& \frac{1}{2} \|\mathbf{s}_2\|^2 \leq P.
\end{aligned}$$

where $z_{DC}(\mathbf{s}_1, \mathbf{s}_2, \phi_{1,1}^*) = \sum_{k=1}^K g_k(\mathbf{s}_1, \mathbf{s}_2)$, $\sigma_{2,n}^2 + s_{1,n}^2 |h_{21,n}|^2 + s_{2,n}^2 |h_{22,n}|^2 = \sum_{k=1}^3 g_{nk}(\mathbf{s}_1, \mathbf{s}_2)$, $\gamma_k, \gamma_{nk} \geq 0$ and $\sum_{k=1}^K \gamma_k, \sum_{k=1}^{K_n} \gamma_{nk} = 1$ for $n = 1, \dots, N$. (23) can be solved by using CVX. The iterative algorithm to solve Eq. (4.20) is given in Algorithm 6.

The achievable rate-energy region for the linear EH model can be solved by using Algorithm 6 by taking only the second order term in z_{DC} .

Remark 4.2: The NIPT system is similar to one ID-user and one EH-user scenario in Chapter 3. Note that in Chapter 3, the transmitters are assumed to have the knowledge of local CSI and the energy harvesters are relying on the linear EH model. Since the waveform design in non-linear EH model favors to transmits at multiple subcarrier, the waveform design with non-linear EH model and limited knowledge of CSI is a problem of interest.

4.2.3 Summary for NIPT

From the simulation in Section (4.5), it was found that deterministic waveform for WPT can increase the achievable rate-energy region. Motivated by this observation, SIPT system is considered and the waveform optimization is discussed in the following section.

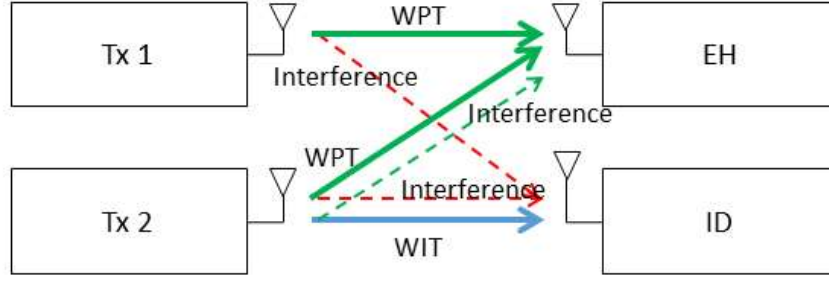


Fig. 4.2 Superposed Information and Power Transfer (SIPT)

4.3 Superposed Information and Power Transfer (SIPT)

As noted in the summary section for NIPT, in this section we expand the line of research and consider a system where the second transmitter transmits deterministic waveform for WPT along with the modulated waveform. Since the second transmitter has the knowledge of CSI, the deterministic waveform transmitted from the second transmitter can be perfectly optimized to enlarge the achievable rate-energy region. In this section, a system model for SIPT as shown in Figure 4.2 is discussed. The first transmitter transmits wireless power using N deterministic multisine waveform and the second transmitter transmits the superposition of N deterministic and modulated (OFDM) waveforms.

4.3.1 Transmit and Receive Signal - SIPT

Let us define $s_{k,m,n}$ and $\phi_{k,m,n}$ as the amplitude and the phase at k -th transmitter and subcarrier n with $m = 1$ and $m = 2$ refer to the deterministic and modulated signal, respectively. The transmit signal at the first transmitter is written as

$$x_1(t) = \Re\left\{\sum_{n=0}^{N-1} w_{1,1,n} e^{j\omega_n t}\right\}, \quad (4.21)$$

where $w_{1,1,n} = s_{1,1,n} e^{j\phi_{1,1,n}}$. The magnitudes and phases in Eq. (4.21) are collected into vectors $\mathbf{s}_{1,1}$ and $\boldsymbol{\phi}_{1,1}$. The transmit signal is subject to the transmit power constraint $\mathcal{E}\{|x_1|^2\} = \frac{1}{2} \|\mathbf{s}_{1,1}\|_F^2 \leq P$.

The transmit signal at the second transmitter consists in the superposition of deterministic and modulated at the same frequencies is given as

$$x_2(t) = \Re\left\{\sum_{n=0}^{N-1} (w_{2,1,n} + \tilde{x}_{I,n}) e^{j\omega_n t}\right\}, \quad (4.22)$$

where $w_{2,1,n} = s_{2,1,n}e^{j\phi_{2,1,n}}$ and $\tilde{x}_{I,n} = \tilde{s}_{2,2,n}e^{j\tilde{\phi}_{2,2,n}}$ are the deterministic and modulated at subcarrier n , respectively. The magnitudes and phases in Eq. (4.22) are collected into vectors $\mathbf{s}_{2,1}$, $\tilde{\mathbf{s}}_{2,2}$, $\boldsymbol{\phi}_{2,1}$ and $\tilde{\boldsymbol{\phi}}_{2,2}$. The transmit signal is subject to the transmit power constraint $\mathcal{E}\{|x_2|^2\} = \frac{1}{2}\|\mathbf{s}_{2,1}\|_F^2 + \frac{1}{2}\mathcal{E}\{\|\tilde{\mathbf{s}}_{2,2}\|_F^2\} = \frac{1}{2}\|\mathbf{s}_{2,1}\|_F^2 + \frac{1}{2}\|\mathbf{s}_{2,2}\|_F^2 \leq P$.

The received signal at the j -th receiver is given as

$$\begin{aligned} y_j(t) &= \sum_{k=1}^2 \sum_{n=0}^{N-1} s_{k,1,n} A_{jk,n} \cos(\omega_n t + \psi_{jk,n}) + \sum_{n=0}^{N-1} \tilde{s}_{2,2,n} A_{j2,n} \cos(\omega_n t + \tilde{\psi}_{j2,n}) \\ &= \Re\left\{ \sum_{n=0}^{N-1} \left(\sum_{k=1}^2 h_{jk,n} w_{k,1,n} + h_{j2,n} \tilde{x}_{I,n} \right) e^{j\omega_n t} \right\}. \end{aligned} \quad (4.23)$$

4.3.2 Energy Harvester - SIPT

From Eq. (4.4) and Eq. (4.23), z_{DC} for the non-linear EH model at the first receiver is given by

$$\begin{aligned} z_{DC} &= k_2 R_{ant} \mathcal{A}\{y_{1,P}(t)^2\} + k_4 R_{ant}^2 \mathcal{A}\{y_{1,P}(t)^4\} + k_2 R_{ant} \mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^2\}\} \\ &\quad + k_4 R_{ant}^2 \mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^4\}\} + 6k_4 R_{ant}^2 \mathcal{A}\{y_{1,P}(t)^2\} \mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^2\}\}, \end{aligned} \quad (4.24)$$

where

$$\mathcal{A}\{y_{1,P}(t)^2\} = \frac{1}{2} \sum_{n=0}^{N-1} \sum_{k_0, k_1}^2 s_{k_0,1,n} s_{k_1,1,n} A_{1k_0,n} A_{1k_1,n} \cos(\psi_{1k_0,1,n} - \psi_{1k_1,1,n}), \quad (4.25)$$

$$\begin{aligned} \mathcal{A}\{y_{1,P}(t)^4\} &= \frac{3}{8} \sum_{\substack{n_0, n_1, n_2, n_3 \\ n_0 + n_1 = n_2 + n_3}} \sum_{k_0, k_1, k_2, k_3}^2 \left[\prod_{j=0}^3 s_{k_j,1,n_j} A_{1k_j,n_j} \right] \cos(\psi_{1k_0,1,n_0} + \psi_{1k_1,1,n_1} - \psi_{1k_2,1,n_2} - \psi_{1k_3,1,n_3}), \\ \mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^2\}\} &= \frac{1}{2} \mathcal{E}\left\{ \sum_{n=0}^{N-1} \tilde{s}_{2,2,n}^2 A_{12,n}^2 \right\} = \frac{1}{2} \sum_{n=0}^{N-1} \mathcal{E}\{\tilde{s}_{2,2,n}^2\} A_{12,n}^2 = \frac{1}{2} \sum_{n=0}^{N-1} s_{2,2,n}^2 A_{12,n}^2, \end{aligned} \quad (4.26)$$

and

$$\begin{aligned} \mathcal{E}\{\mathcal{A}\{y_{1,I}(t)^4\}\} &= \frac{3}{8} \mathcal{E}\left\{ \sum_{\substack{n_0, n_1, n_2, n_3 \\ n_0 + n_1 = n_2 + n_3}} \tilde{s}_{2,2,n_0} \tilde{s}_{2,2,n_1} \tilde{s}_{2,2,n_2} \tilde{s}_{2,2,n_3} A_{12,n_0} A_{12,n_1} A_{12,n_2} A_{12,n_3} \right. \\ &\quad \left. \cos(\tilde{\psi}_{12,n_0} + \tilde{\psi}_{12,n_1} - \tilde{\psi}_{12,n_2} - \tilde{\psi}_{12,n_3}) \right\}. \end{aligned} \quad (4.28)$$

The linear EH model is obtained by taking the second order term in Eq. (4.24).

4.3.3 Information Receiver - SIPT

Perfect cancellation

The baseband received signal for perfect cancellation of WPT signal is given as

$$y_{2,n} = h_{22,n}\tilde{x}_{I,n} + v_{2,n} \quad (4.29)$$

and the achievable rate is

$$R(\mathbf{s}_{2,2}) = \sum_{n=0}^{N-1} \log_2 \left(1 + \frac{s_{2,2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2} \right), \quad (4.30)$$

where $\mathbf{s}_{k,m} \in \mathbb{R}^{N \times 1}$ for $k, m = 1, 2$ and $s_{2,2,n}^2 |h_{22,n}|^2$ is the received power of WIT from the second transmitter and n -th subcarrier.

No-cancellation

The baseband receive signal for no-cancellation of WPT signal is given as

$$y_{2,n} = h_{22,n}\tilde{x}_{I,n} + h_{21,n}w_{1,1,n} + h_{22,n}w_{2,1,n} + v_{2,n} \quad (4.31)$$

and the achievable rate is

$$R(\mathbf{s}_{2,2}, \mathbf{w}_{1,1,n}, \mathbf{w}_{2,1,n}) = \sum_{n=0}^{N-1} \log_2 \left(1 + \frac{s_{2,2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2 + |w_{1,1,n}h_{21,n} + w_{2,1,n}h_{22,n}|^2} \right), \quad (4.32)$$

where $|w_{1,1,n}h_{21,n} + w_{2,1,n}h_{22,n}|^2$ is the interference power from WPT signal.

4.4 Coordinated Waveform Optimization for SIPT

This section discusses the coordinated waveform optimization for SIPT with and without interference cancellation at the ID. The deterministic waveform at the first transmitter and the superposition of deterministic and modulated waveform at the second transmitter is optimized so as to maximize the rate-energy region of the system.

4.4.1 Perfect Cancellation

The achievable rate-energy region for PC and non-linear model is given by

$$C_{\bar{R}-\bar{z}_{DC}}(P) \triangleq \{(\bar{R}, \bar{z}_{DC}) : R \leq \sum_{n=0}^{N-1} \log_2(1 + \frac{s_{2,2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2}), \bar{z}_{DC} \leq z_{DC}(\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2}, \phi_{1,1}, \phi_{2,1}), \frac{1}{2} \|\mathbf{s}_{1,1}\|^2 \leq P, \frac{1}{2} \|\mathbf{s}_{2,1}\|_F^2 + \frac{1}{2} \|\mathbf{s}_{2,2}\|_F^2 \leq P\}. \quad (4.33)$$

In order to maximize the total DC current, it should be guaranteed that the $\cos(\cdot)$ in Eq. (4.25) and Eq. (4.26) to be equal to one. Since $\psi_{1k,n} = \bar{\psi}_{1k,n} + \phi_{k,1,n}$, the total DC current is maximized when the phase for WPT waveform is chosen such that $\phi_{k,1,n}^* = -\bar{\psi}_{1k,n}$ for $k = 1, 2$. Here, the phase for WPT waveform at each subcarrier is matched to the phase of the channel between the transmitter and the EH.

Given the phase for WPT $\phi_{1,1}^*, \phi_{2,1}^*$, the standard GP problem in Eq. (4.33) is given as

$$\begin{aligned} (P \ 4.4) \quad & \min_{\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2}, t_0} \quad 1/t_0 \\ \text{s. t.} \quad & t_0 \prod_{k=1}^K \left(\frac{g_k(\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2})}{\gamma_k} \right)^{-\gamma_k} \leq 1 \\ & 2^{\bar{R}} \prod_{n=0}^{N-1} \prod_k \left(\frac{g_{nk}(\mathbf{s}_{2,2})}{\gamma_{nk}} \right)^{-\gamma_{nk}} \leq 1 \\ & \frac{1}{2} \|\mathbf{s}_{1,1}\|^2 \leq P, \\ & \frac{1}{2} \|\mathbf{s}_{2,1}\|^2 + \frac{1}{2} \|\mathbf{s}_{2,2}\|^2 \leq P, \end{aligned} \quad (4.34)$$

where $z_{DC}(\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2}) = \sum_{k=1}^K g_k(\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2})$, $1 + \frac{s_{2,2,n}^2 A_{22,n}^2}{\sigma_{2,n}^2} = \sum_{k=1}^2 g_{nk}(\mathbf{s}_{2,2})$, $\gamma_k, \gamma_{nk} \geq 0$ and $\sum_{k=1}^K \gamma_k, \sum_{k=1}^{K_n} \gamma_{nk} = 1$ for $n = 1, \dots, N$. Eq. (4.34) can be solved by using CVX. The iterative algorithm to solve Eq. (4.34) is given in Algorithm 7.

Note that the optimal phase for WPT in non-linear model is also optimal in the linear model. Given the phase, the amplitude of WPT and WIT waveforms is optimized by using Algorithm 7, by taking only the second order term in z_{DC} .

Algorithm 7 Iterative Algorithm for PC in SIPT

-
- 1: **Initialize:** $z_{DC}^{(0)} \leftarrow 0, i \leftarrow 0, \varepsilon, \mathbf{s}_{1,1}^{(0)}, \mathbf{s}_{1,2}^{(0)}, \mathbf{s}_{2,2}^{(0)}$
 - 2: **repeat**
 - 3: $i \leftarrow i + 1$
 - 4: $\gamma_k \leftarrow g_k(\mathbf{s}_{1,1}^{(i-1)}, \mathbf{s}_{2,1}^{(i-1)}, \mathbf{s}_{2,2}^{(i-1)}) / z_{DC}^{(i)}$
 - 5: $\gamma_{nk} \leftarrow g_{nk}(\mathbf{s}_{2,2}^{(i-1)}) / \sum_{k=1}^2 g_{nk}(\mathbf{s}_{2,2}^{(i-1)})$ for $n = 1, \dots, N$
 - 6: $\mathbf{s}_{1,1}^{(i)}, \mathbf{s}_{2,1}^{(i)}, \mathbf{s}_{2,2}^{(i)} \leftarrow \arg \min \text{Eq. (4.34)}$
 - 7: $z_{DC}^{(i)} \leftarrow z_{DC}(\mathbf{s}_{1,1}^{(i)}, \mathbf{s}_{2,1}^{(i)}, \mathbf{s}_{2,2}^{(i)})$
 - 8: **until** $|z_{DC}^{(i)} - z_{DC}^{(i-1)}| < \varepsilon$
-

4.4.2 No-Cancellation

The achievable rate-energy region for NC and non-linear model is given by

$$C_{\bar{R}-\bar{z}_{DC}} \triangleq \{(\bar{R}, \bar{z}_{DC}) : R \leq \sum_{n=1}^N \log_2 \left(1 + \frac{s_{2,2,n}^2 |h_{22,n}|^2}{\sigma_{2,n}^2 + |w_{1,1,n} h_{21,n} + w_{2,1,n} h_{22,n}|^2} \right),$$

$$z_{DC} \leq z_{DC}(\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2}, \phi_{1,1}, \phi_{2,1}), \frac{1}{2} \|\mathbf{s}_{1,1}\|^2 \leq P, \frac{1}{2} \|\mathbf{s}_{2,1}\|_F^2 + \frac{1}{2} \|\mathbf{s}_{2,2}\|_F^2 \leq P\}. \quad (4.35)$$

The phase of WPT at both transmitter has a tradeoff between maximizing the performance of ID and EH. Let us assume the same phases as in SIPT with PC is used. Here, the phase of WPT at each transmitter is chosen such that it is matched to the phase of the channel between the transmitter and the EH given as $\phi_{k,1,n}^* = -\bar{\psi}_{1k,n}$ for $k = 1, 2$. It is noted that with this phase strategy, the maximum WPT and WIT can be obtained by optimizing the power allocation.

A new variable $t_{1,n}$ is introduced such that the achievable rate in Eq. (4.35) can be written as $\sum_n \log_2(1 + \frac{s_{2,2,n}^2 |h_{22,n}|^2}{t_{1,n}})$ with an additional constraint $\sigma_{2,n}^2 + s_{1,1,n}^2 |h_{21,n}|^2 + s_{2,1,n}^2 |h_{22,n}|^2 + 2s_{1,1,n} s_{2,1,n} |h_{21,n}| |h_{22,n}| \cos(\psi_{21,n} - \psi_{22,n}) = t_{1,n}$. Here, $t_{1,n}$ can be a signomial (i.e., the sum of monomials with the coefficients can be positive or negative). We then form a relaxation of this equality constraint with a lower bound inequality constraint $\sigma_{2,n}^2 + s_{1,1,n}^2 |h_{21,n}|^2 + s_{2,1,n}^2 |h_{22,n}|^2 + 2s_{1,1,n} s_{2,1,n} |h_{21,n}| |h_{22,n}| \cos(\psi_{21,n} - \psi_{22,n}) \geq t_{1,n}$. As $s_{1,1,n}$ and $s_{2,1,n}$ increase, z_{DC} and $t_{1,n}$ increases. On the other hand, the rate constraint is decreasing function of $t_{1,n}$. Therefore, it can be found that the new inequality constraint $\sigma_{2,n}^2 + s_{1,1,n}^2 |h_{21,n}|^2 + s_{2,1,n}^2 |h_{22,n}|^2 + 2s_{1,1,n} s_{2,1,n} |h_{21,n}| |h_{22,n}| \cos(\psi_{21,n} - \psi_{22,n}) \geq t_{1,n}$ is always tight. The lower bound inequality constraints is then transformed by using monomial approximation.

Algorithm 8 Iterative Algorithm for NC in SIPT

```

1: Initialize:  $z_{DC}^{(0)} \leftarrow 0, i \leftarrow 0, \varepsilon, \mathbf{s}_{1,1}^{(0)}, \mathbf{s}_{2,1}^{(0)}, \mathbf{s}_{2,2}^{(0)}$ 
2: repeat
3:    $i \leftarrow i + 1$ 
4:    $\gamma_k \leftarrow g_k(\mathbf{s}_{1,1}^{(i-1)}, \mathbf{s}_{2,1}^{(i-1)}, \mathbf{s}_{2,2}^{(i-1)}) / z_{DC}^{(i)}$ 
5:    $\gamma_{nk} \leftarrow g_{nk}(\mathbf{s}_{2,2}^{(i-1)}) / \sum_{k=1}^2 g_{nk}(\mathbf{s}_{2,2}^{(i-1)})$  for  $n = 1, \dots, N$ 
6:    $\gamma_{dk} \leftarrow g_{dk}(\mathbf{s}_{1,1}^{(i-1)}, \mathbf{s}_{2,1}^{(i-1)}) / \sum_{k=1}^{K_d} g_{dk}(\mathbf{s}_{1,1}^{(i-1)}, \mathbf{s}_{2,1}^{(i-1)})$  for  $d = 1, \dots, N$ 
7:    $\mathbf{s}_{1,1}^{(i)}, \mathbf{s}_{2,1}^{(i)}, \mathbf{s}_{2,2}^{(i)} \leftarrow \arg \min \text{Eq. (4.36)}$ 
8:    $z_{DC}^{(i)} \leftarrow z_{DC}(\mathbf{s}_{1,1}^{(i)}, \mathbf{s}_{2,1}^{(i)}, \mathbf{s}_{2,2}^{(i)})$ 
9: until  $|z_{DC}^{(i)} - z_{DC}^{(i-1)}| < \varepsilon$ 

```

Given the phase for WPT waveform $\phi_{1,1}^*, \phi_{2,1}^*$, Eq. (4.35) is transformed to a standard GP problem given as

$$\begin{aligned}
 (P \ 4.5) \quad & \min_{\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2}, t_0} \quad 1/t_0 \\
 \text{s. t.} \quad & t_0 \prod_{k=1}^K \left(\frac{g_k(\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2})}{\gamma_k} \right)^{-\gamma_k} \leq 1 \\
 & 2^{\bar{R}} \prod_k^{K_n} \left(\frac{g_{nk}(\mathbf{s}_{2,2})}{\gamma_{nk}} \right)^{-\gamma_{nk}} \leq 1 \\
 & f_{d1} \prod_k^{K_d} \left(\frac{g_{dk}(\mathbf{s}_{1,1}, \mathbf{s}_{2,1})}{\gamma_{dk}} \right)^{-\gamma_{dk}} \leq 1 \\
 & \frac{1}{2} \|\mathbf{s}_{1,1}\|^2 \leq P, \quad \frac{1}{2} \|\mathbf{s}_{2,1}\|^2 + \frac{1}{2} \|\mathbf{s}_{2,2}\|^2 \leq P,
 \end{aligned} \tag{4.36}$$

where $z_{DC}(\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2}, \phi_{1,1}^*, \phi_{2,1}^*) = \sum_{k=1}^K g_k(\mathbf{s}_{1,1}, \mathbf{s}_{2,1}, \mathbf{s}_{2,2})$, $\sigma_{2,n}^2 + s_{1,1,n}^2 |h_{21,n}|^2 + s_{2,1,n}^2 |h_{22,n}|^2 + 2s_{1,1,n}s_{2,1,n}|h_{21,n}||h_{22,n}|\cos(\psi_{21,n} - \psi_{22,n}) = f_{d1} - f_{d2} \geq t_{1,n}$, $1 + s_{2,2,n}^2 |h_{22,n}|^2 / t_{1,n} = \sum_{k=1}^2 g_{nk}(\mathbf{s}_{2,2})$, $t_{1,n} + f_{d2} = \sum_{k=1}^{K_d} g_{dk}(\mathbf{s}_{1,1}, \mathbf{s}_{2,1})$, $\gamma_k, \gamma_{nk}, \gamma_{dk} \geq 0$ and $\sum_{k=1}^K \gamma_k, \sum_{k=1}^{K_n} \gamma_{nk}, \sum_{k=1}^{K_d} \gamma_{dk} = 1$ for $n, d = 1, \dots, N$. Eq. (4.36) can be solved by using CVX. The iterative algorithm to solve Eq. (4.36) is given in Algorithm 8.

The achievable rate-energy region for linear EH model can be obtained by using Algorithm 8, by taking only the second order term in z_{DC} .

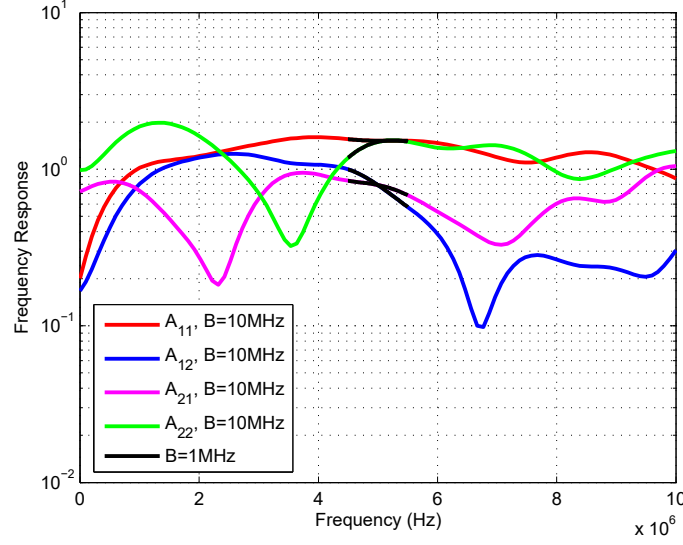


Fig. 4.3 Frequency response of the wireless channel for 1MHz and 10MHz bandwidths.

4.5 Simulation Results

This section presents the simulation results. The simulation consider a centre frequency of 5.18GHz, 36dBm EIRP, 2dBi receive antenna gain and 58dB path loss with the NLOS channel power delay profile obtained from model B [81]. The channel taps each with an average power β_l are independent, circularly symmetric complex random gaussian distributed and normalized such that $\sum_l \beta_l = 1$. This lead to an average received power of -20 dBm. The simulation is run over a channel realization with 1MHz and 10MHz bandwidths. The frequency response of the channel is illustrated in Figure 4.3. The SNR (defined as P/σ_n^2) at the ID receiver is 20 dB and the number of subcarrier is 8, otherwise stated.

Figure 4.4 and Figure 4.5 show the rate-energy trade-off for NIPT and SIPT with $B=1$ MHz and $B=10$ MHz bandwidths. The waveform is optimized relying on the linear and non-linear EH models discussed in Section (4.2) and (4.4). The z_{DC} on the y-axis refers to the DC component in Eq. (4.4) obtained based on the waveform optimization with linear and non-linear EH models. The global optimum solution for the linear model and NIPT with perfect interference cancellation at the ID receiver can be obtained in Appendix A (*Remark 4.1*). The simulation result by using GP and optimization in Appendix A is shown in Figure 4.6. Although the solution from reversed GP is not guaranteed to be global optimum [78], evaluation shows that the achievable rate-energy from Appendix A and by using reversed GP in Section (4.2) have a similar performance. This observation shows that reversed GP can provide close to the global optimum solution.

The extreme point on the x-axis (for NIPT and SIPT) refers to the maximum achievable rate obtained by using the waterfilling solution. The extreme point on the y-axis for NIPT refers to the maximum harvested energy when the second transmitter allocates its full transmit power to the modulated (i.e., OPDM/WIT) waveform. On the other hand in SIPT, the extreme point on the y-axis is obtained when the second transmitter allocates its full transmit power to the deterministic (i.e., WPT) waveform. Accordingly, the achievable rate at this point is equal to zero.

The shape of the rate-energy region depends on the frequency response of the channel realization. By looking at Figure 4.3, specifically A_{22} for the information transfer and A_{11} and A_{12} for z_{DC} , A_{22} , A_{11} and A_{12} for 1MHz bandwidth are not in a deep fade. Since the power allocation over the multiple carrier frequency increase the achievable rate and z_{DC} , 1Mhz bandwidth has a higher maximum rate and z_{DC} compared to 10MHz bandwidth that exhibits amplitude fluctuation and deep fade.

The main observation for NIPT in Figure 4.4 is that waveform design based on non-linear EH model has a larger achievable rate-energy region over the linear EH model for both frequency-flat ($B=1\text{MHz}$) and frequency-selective ($B=10\text{MHz}$) channels. An interesting observation is that for the perfect cancellation (PC), the non-linear EH model always has a larger rate-energy region over the linear EH model. For the linear EH model, by looking into Eq. (4.5) and Eq. (4.7), it is noted that modulated and deterministic waveforms have a similar RF-DC efficiency. Hence, the first transmitter with the linear EH model can be seen as transmitting the modulated waveform with no information to the ID receiver. A larger rate-energy region for the non-linear model implies that deterministic waveform has a better RF-DC efficiency compared to the modulated waveform. A better efficiency of deterministic waveform has also been observed in point-to-point SWIPT [19]. Motivated by this observation, a superposition of modulated and deterministic waveforms are considered at the second transmitter namely SIPT.

From Figure 4.4, for no-interference cancellation (NC) at the ID receiver and when the achievable rate is closed to the extreme point on the x-axis, the linear model get close performance to the non-linear model. Here, the second transmitter (with linear and non-linear EH models) allocates most of its power based on the waterfilling algorithm to maximize the information transfer. In addition, the first transmitter (with linear and non-linear EH models) transfer low power so as the interference at the ID receiver is low. Accordingly, it could be argued that the high order term would not distribute much to the z_{DC} for points close to the maximum rate.

Figure 4.5 and Figure 4.7 show the simulation result for SIPT with $B=1\text{MHz}$ and $B=10\text{MHz}$ bandwidths. It is observed that SIPT always outperform NIPT for both lin-

ear and non-linear models. First, we look at the linear EH model. The transmission of deterministic waveform at the first and second transmitters provide a beamforming gain that increase the z_{DC} . This helps enlarge the trade-off region for SIPT with the linear model for both PC and NC. In addition, by looking at the trade-off region of non-linear model for both PC and NC, SIPT has a larger region over the NIPT and the linear model. Since the information symbols of the modulated waveform changed randomly over symbol duration, it leads to uncontrollable modulated waveform and RF-DC inefficiency in the non-linear model. This show that the deterministic waveform at the second transmitter is beneficial to enlarge the rate-energy region.

The input distribution at the second transmitter for SIPT in Eq. (4.22) and on frequency n is given as $w_{2,1,n} + \tilde{x}_{I,n} \sim \mathcal{CN}(w_{2,1,n}, s_{2,2,n}^2)$. This waveform has a non-zero mean Gaussian distribution [9]. The magnitude of $|w_{2,1,n} + \tilde{x}_{I,n}|$ has a rician distribution with K-factor on subcarrier n given by $K_n = s_{2,1,n}^2 / s_{2,2,n}^2$. This observation implies that the transmit waveform at the information transmitter should be designed differently from the capacity achieving in the conventional WIT that has a zero mean Gaussian distribution. As the second transmitter only transmit OFDM waveform, the mean of the transmit waveform is zero. However, as the rate decreases, the second transmitter starts to transmit deterministic waveform such that the mean of the transmitted waveform become non-zero. Accordingly, it decrease the K-factor as depicted in Figure 4.7.

For PC in both NIPT and SIPT, the rate-energy region is convex such that it always outperform the time sharing (TS) between deterministic and modulated waveform. However, for NC in both NIPT and SIPT, the rate-energy region is concave-convex. At low rate, the rate-energy region outperform TS. At high rate, the convex hull is obtained by the TS.

The effect of N and R to the performance of z_{DC} with 1MHz and 10MHz bandwidths is depicted in Figure 4.8 and Figure 4.9. The waveform design based on non-linear model outperform the linear model for all configurations except for $N = 4$ and $R = 20$. Here, for given channel realization and N , $R = 20$ is close to its maximum achievable rate. As observed in Figure 4.4 and Figure 4.5, the linear model has close performance to the non-linear model when the rate constraint is close to its maximum achievable rate for that channel realization.

The simulation result in Figure 4.8 and Figure 4.9 show that for a given rate constraint, the efficiency of waveform design based on non-linear model increases as N increases. The waveform design based on SIPT increase z_{DC} over NIPT for given R and N . This result highlights the importance for accounting the non-linearity of the rectifier for the coordinated waveform design in SWIPT. Moreover, for 10MHz bandwidth and small N , the linear model has close performance to the non-linear model. As N increase, the gain of non-linear model over the linear model is more apparent. Since the power allocation over multiple waveform

subcarrier increase z_{DC} in the non-linear model, the amplitude fluctuation of 10MHz channel limits the performance of z_{DC} . As N increase, the transmitter start to allocate the power over multiple subcarrier and accordingly increase z_{DC} .

4.6 Summary

This chapter has studied the coordinated waveform design for SWIPT in an interference channel. This chapter covers coordinated waveform in a two-user system with one energy user and one information user (with their corresponding transmitter respectively denoted as first transmitter and second transmitter). The perfect and no cancellation of WPT signal at the ID receiver were considered. The trade-off between harvested energy and achievable rate for information transfer was identified. The waveform was designed based on the non-linear EH model that accounts for the non-linearity of the rectifier as discussed in Section (2.3.1). Comparison was also been made with the linear EH model that ignore the non-linearity of the rectifier.

First, the coordinated waveform design for a Non-superposed Information and Power Transfer (NIPT) as shown in Figure 4.1 was discussed. The first transmitter transmits deterministic waveform for WPT and the second transmitter transmits modulated OFDM waveform for WIT. For NIPT with the non-linear EH model, the optimal phases for deterministic waveform were obtained in a closed form. On the other hand, the waveform optimization with the linear EH model was not a function of the waveform phase. Given the phase of the waveform, the amplitude of deterministic and modulated waveforms were optimized by using a Reverse Geometric Program (GP). The numerical simulation has shown that deterministic waveform can enlarge the trade-off region over the modulated waveform. This observation motivates the new design of the superposition of modulated and deterministic waveform at the second transmitter, named Superposed Information and Power Transfer (SIPT) as shown in Figure 4.2.

In SIPT system, the first transmitter transmits deterministic waveform for WPT and the second transmitter transmits the superposition of deterministic and modulated waveforms for WPT and WIT, respectively. The optimal phases for the deterministic waveform were obtained. Given the phase of the waveform, the amplitude of the waveform was optimized by using a Reverse GP. For NC, the phase of deterministic waveform should be chosen such that it maximizes the harvested energy at the EH receiver and the achievable rate at the ID receiver. This lead to a trade-off between these two objectives. The optimization can be solved by iteratively optimize the amplitude and the phase of the waveform, but it would lead to a high computation. Here, we adopt the same phase strategy used in PC (match filtering

w.r.t the phase of the channel). Note that with the proposed phase strategy, the maximum achievable rate or harvested energy can be achieved with an appropriate power allocation.

The numerical simulation has shown that SIPT always outperform NIPT for both linear and non-linear models. With the linear EH model, the transmission of deterministic waveform at the first and second transmitters in SIPT provide energy beamforming gain that increase the received power at the EH receiver. This helps to enlarge the trade-off region of SIPT for both PC and NC. In addition, by looking at the trade-off region of non-linear model for both PC and NC, SIPT has a larger trade-off region compared to NIPT. Since the information symbols of the modulated waveform changed randomly over symbol duration, the modulated waveform (amplitude and phase) was uncontrollable. For non-linear EH model, deterministic waveform is more RF-DC efficient compared to the modulated waveform. Hence, the transmission of deterministic waveform at the second transmitter with the modulated OFDM waveform helps to enlarge the trade-off region.

The transmitted waveform at the second transmitter in SIPT has a non-zero mean Gaussian distribution [19]. The numerical result has shown that SIPT provides a larger trade-off region compared to NIPT. This observation implies that the transmit waveform at the second transmitter should be designed differently from the capacity achieving in the conventional WIT that has a zero mean Gaussian distribution. Another observation is that, the non-linear EH model has shown to have a larger trade-off region compared to the linear EH model. This observation has shown that accounting for the non-linearity of the rectifier in the coordinated waveform design increase the resource efficiency compared to the waveform design obtained by ignoring the non-linearity of the rectifier.

It is interesting to extend the works in this chapter for the general K -user system. It is noted that for the general K -user system, the optimal phase cannot be obtained in a closed form. Therefore, the same optimization method proposed in this chapter cannot be used. The problem of limited CSI knowledge at the transmitter is also a problem of interest.

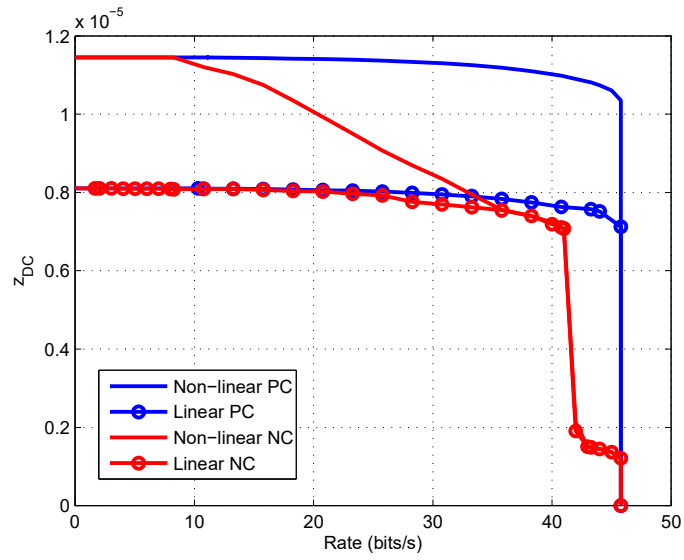
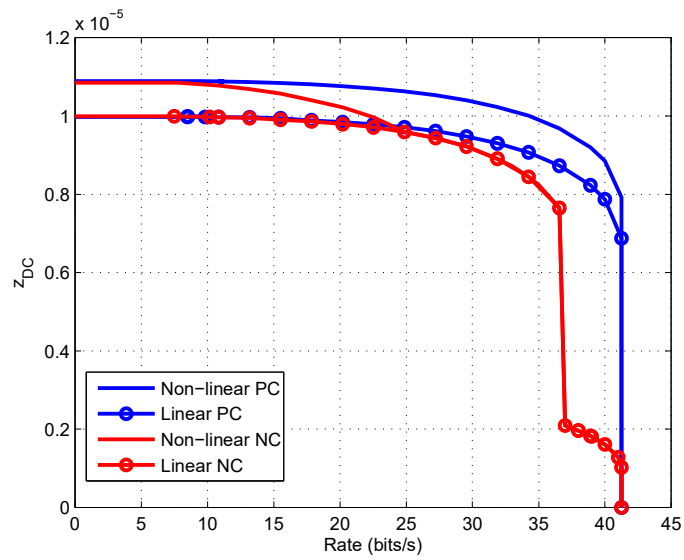
(a) z_{DC} with $B=1\text{MHz}$ (b) z_{DC} with $B=10\text{MHz}$

Fig. 4.4 Rate-energy trade-off for linear and non-linear EH models in NIPT.

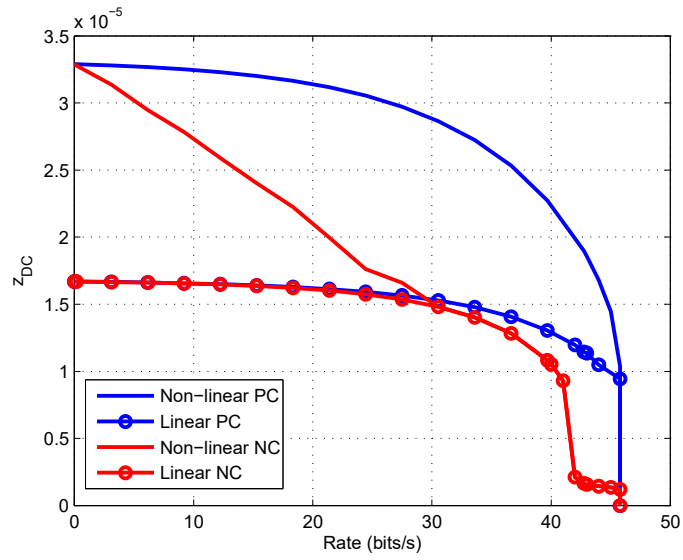
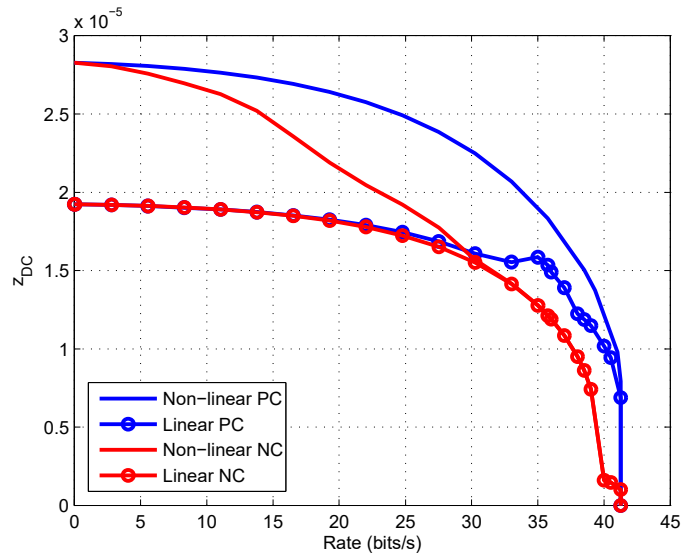
(a) z_{DC} with $B=1\text{MHz}$ (b) z_{DC} with $B=10\text{MHz}$

Fig. 4.5 Rate-energy trade-off for linear and non-linear EH models in SIPT.

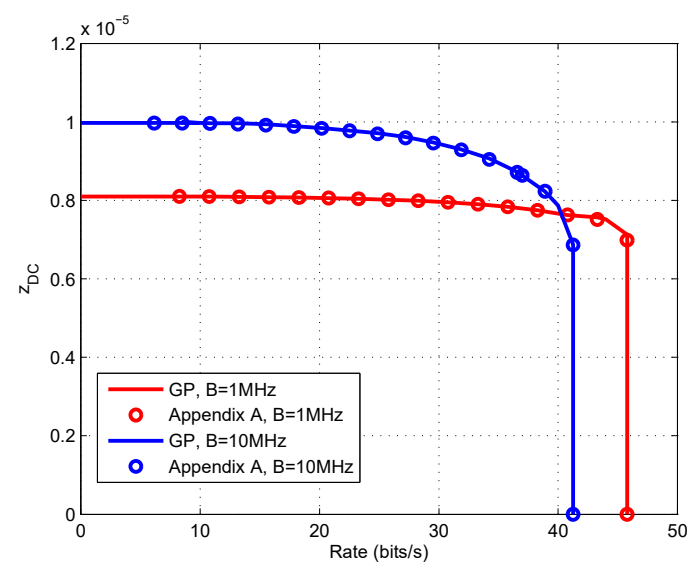


Fig. 4.6 NIPT with Linear EH.

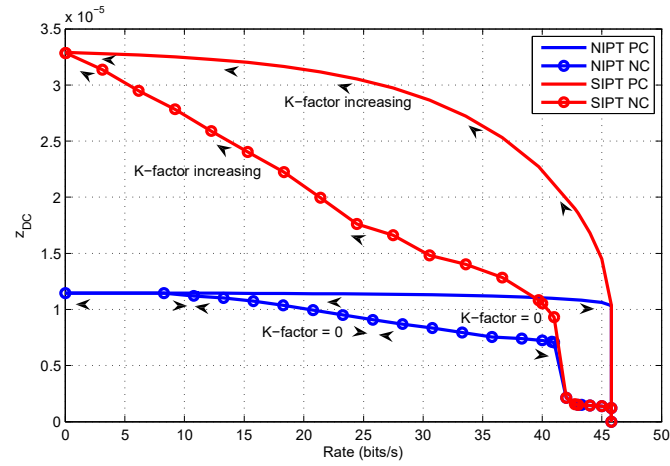
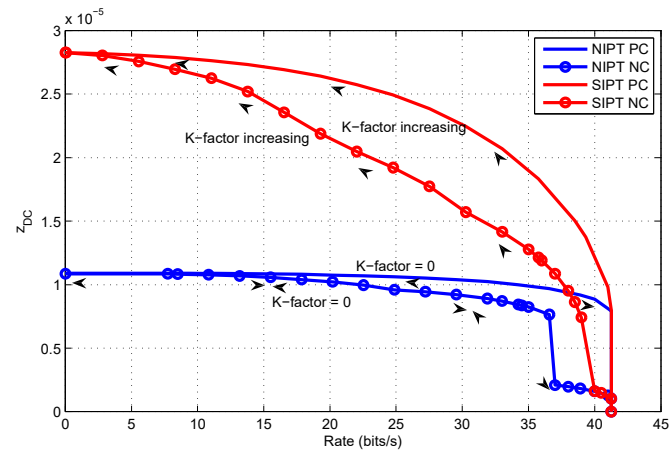
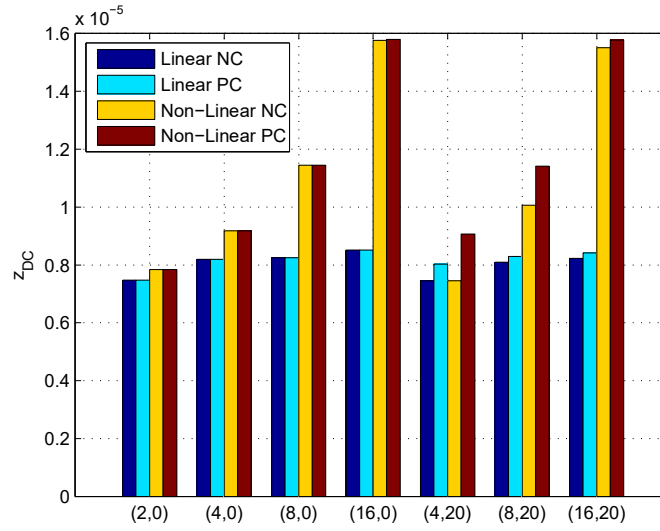
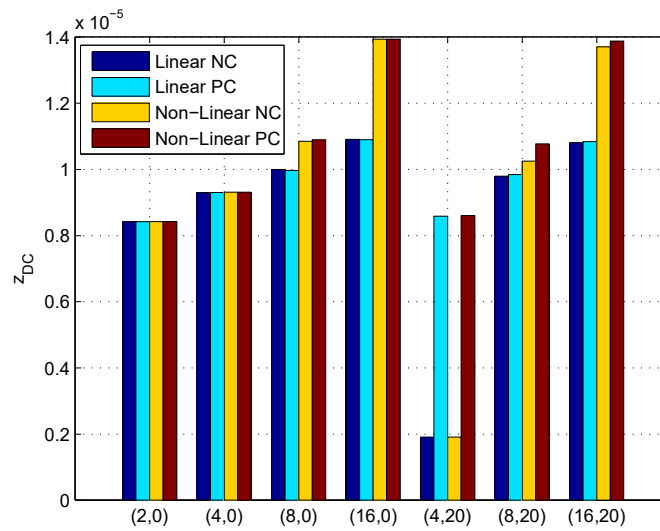
(a) z_{DC} with $B=1\text{MHz}$ (b) z_{DC} with $B=10\text{MHz}$

Fig. 4.7 Rate-energy trade-off for non-linear EH model.

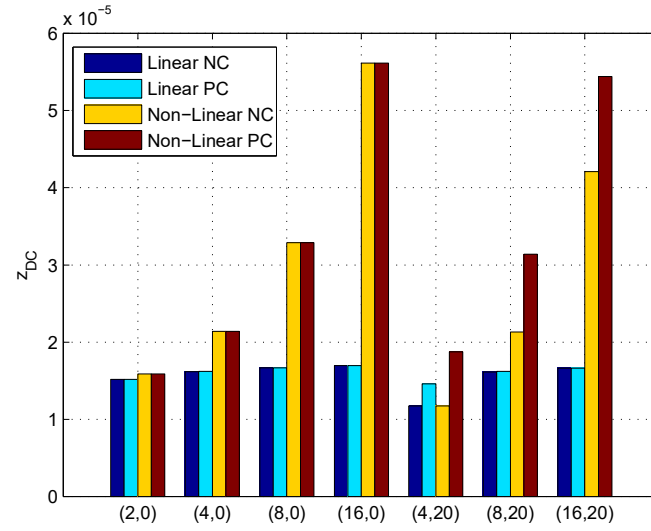
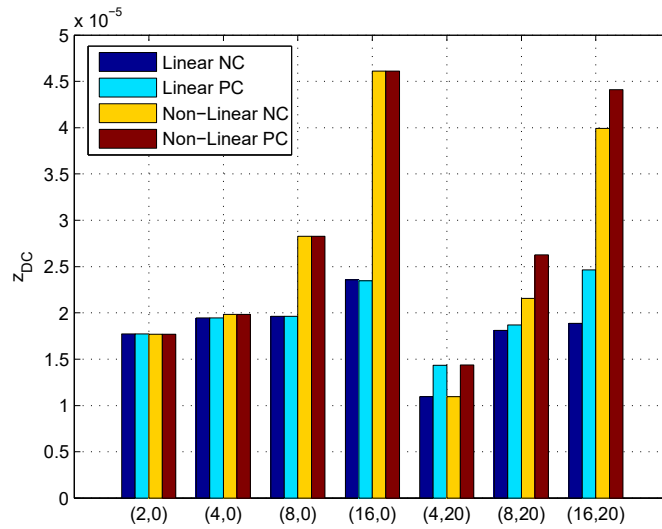


(a) B=1MHz



(b) B=10MHz

Fig. 4.8 z_{DC} as a function of (N, R) for NIPT.

(a) $B=1\text{MHz}$ (b) $B=10\text{MHz}$ Fig. 4.9 z_{DC} as a function of (N, R) for SIPT.

Chapter 5

Waveform Design for Wireless Powered Backscatter

Motivated by the interesting results from the previous chapter, in this chapter we study waveform design for wireless powered backscatter system. The waveform is designed based on non-linear EH model and adaptive to the CSI at the transmitter. For a point-to-point system, the optimization method from the previous chapter is revisited. Specifically, the optimal phases of the waveform are obtained in the closed form solution. Accordingly, the amplitudes of the waveform are optimized by using Reverse GP. However, for a multi-user system, the optimal phases cannot be obtained in the closed form. Therefore, a new optimization method will be proposed.

In this chapter, the trade-off between harvested energy and the reliability of backscatter communication will be investigated. The waveform is designed adaptive to the perfect CSI knowledge of forward and backward links available at the reader. A simple binary modulation is considered for the backscattering communication. The waveform is designed based on the non-linear EH model introduced in [17, 18]. Comparison is also been made with a linear EH model, that only accounts for the second order term in Eq. (2.21). This chapter covers a point-to-point and multi-user wireless powered backscatter.

5.1 System Model

A backscatter system with one reader (i.e., continuous waveform transmitter) and K backscatters (i.e., tags) is considered. The reader and backscatters have a single antenna. A slow fading environment is considered where the channels are assumed to be constant over the time period of interest. It is assumed that the reader has a perfect knowledge of forward and

backward channels (the channel from the reader and backscatters and the channel from the backscatter and reader, respectively).

5.1.1 The Transmit Waveform

The transmitter transmits the deterministic multisine waveform given by

$$x(t) = \sum_{n=0}^{N-1} s_n \cos(\omega_n t + \phi_n) = \Re\left\{ \sum_{n=0}^{N-1} w_n e^{i\omega_n t} \right\}, \quad (5.1)$$

where $w_n = s_n e^{i\phi_n}$. Here, s_n and ϕ_n are the amplitude and the phase of the n -th sinewave at frequency ω_n , respectively. The magnitudes and phases in Eq. (5.1) can be collected into vectors $\mathbf{s}$ and Φ . The transmit waveform is subject to transmit power constraint $\mathcal{E}\{|x|^2\} = \frac{1}{2} \|\mathbf{s}\|^2 \leq P$.

After multipath propagation, the received waveform at the j -th tag can be written as

$$y_j(t) = \sum_{n=0}^{N-1} s_n A_{j,n} \cos(\omega_n t + \psi_{j,n}) = \Re\left\{ \sum_{n=0}^{N-1} h_{j,n} w_n e^{i\omega_n t} \right\}, \quad (5.2)$$

where $h_{j,n} = A_{j,n} e^{i\psi_{j,n}} = \sum_{l=0}^{L-1} \alpha_{j,l} e^{i(-\omega_n \tau_{j,l} + \varepsilon_{j,l})}$ is the frequency response of the channel between the transmitter and the j -th tag at frequency component n and $\psi_{j,n} = \bar{\psi}_{j,n} + \phi_n$. $\tau_{j,l}$, $\varepsilon_{j,l}$ and $\alpha_{j,l}$ are the delay, phase and amplitude of the l -th path from the transmitter to the j -th tag, respectively.

5.1.2 The Backscatter Operation

We consider that each tag employs a simple binary modulation to transfer information to the reader as in [82]. Binary 0 refers to a perfect impedance matching that completely absorbs the incoming signal and binary 1 refers to a perfect impedance mismatch that completely reflects the incoming signal. The signal absorbed during binary 0 operation is conveyed to the rectifier for energy harvesting, while the signal reflected during binary 1 operation is backscattered to the reader. The reader performs information detection of sequence bit 0 and 1 from the incoming backscattered signal.

5.1.3 The Energy Harvester

The analytical EH model that accounts for the non-linearity of the diode has been introduced in Section (4.1.3). By ignoring the terms higher than 4, the DC component in Eq. (2.21) for

$$\mathcal{A}\{y_j(t)^2\} = \frac{1}{2} \sum_{n=1}^N s_n^2 A_{j,n}^2 \quad (5.5)$$

$$\mathcal{A}\{y_j(t)^4\} = \frac{3}{8} \left[\sum_{\substack{n_1, n_2, n_3, n_4 \\ n_1 + n_2 = n_3 + n_4}} [s_{n_j} A_{j,n_m}] \cos(\psi_{j,n_1} + \psi_{j,n_2} - \psi_{j,n_3} - \psi_{j,n_4}) \right]. \quad (5.6)$$

j -th backscatter can be written as

$$i_{d,j}(t) = \sum_{u=0}^4 k'_{u,j} R_{ant}^{u/2} \mathcal{A}\{y_j(t)^u\}. \quad (5.3)$$

Following [18], the maximization of Eq. (5.3) is equivalent to maximizing

$$z_{DC,j} = k_2 R_{ant} \mathcal{A}\{y_j(t)^2\} + k_4 R_{ant}^2 \mathcal{A}\{y_j(t)^4\}. \quad (5.4)$$

where $k_u = \frac{i_s}{u!(nv_t)^u}$. For $i_s = 5\mu A$, $n = 1.05$ and $v_t = 25.86mV$, typical values of $k_2 = 0.0034$ and $k_4 = 0.3829$. $\mathcal{A}\{y_j(t)^2\}$ and $\mathcal{A}\{y_j(t)^4\}$ are given in Eq. (5.5) and Eq. (5.6), respectively. By defining the weight for energy harvested at user j as $c_j \geq 0$, we are interested in a general metric consisting of a weighed sum of $z_{DC,j}$ given by $Z_{DC} = \sum_{j=1}^K c_j z_{DC,j}$.

5.1.4 The Backscattered Signal

The tag completely absorbs and reflects the incoming signal for binary information 0 and 1, respectively. The received signal at the reader is given as

$$z(t) = \Re\left\{ \sum_{n=1}^N \sum_{j=1}^K m_j h_{j,n} h_{j,n}^b w_n e^{i2\pi f_n t} \right\} + v(t) \quad (5.7)$$

where $m_j = 0, 1$ is the binary information for j -th tag, $h_{j,n}^b = A_{j,n}^b e^{i\psi_{j,n}^b}$ is the frequency response of the backward channel between the j -th tag and the reader at frequency component n and $v(t)$ is AWGN. After applying a product detector to each frequency component and assuming perfect low pass filtering, the baseband received signal at subcarrier n is given by

$$z_n = \sum_{j=1}^K h_{j,n} h_{j,n}^b w_n m_j + v_n. \quad (5.8)$$

v_n is the complex white Gaussian noise at frequency component n with variance σ^2 . Let $\mathbf{z} = [z_1, \dots, z_N]^T$, the signal $\tilde{z}_j$ used for information detection of the j -th tag can be written as

$$\tilde{z}_j = \mathbf{g}_j^H \mathbf{z}, \quad (5.9)$$

where $\mathbf{g}_j^H \in \mathbb{C}^{1 \times N}$ is the receive combiner at the reader. It is noted that the SINR of each tag is invariant to the scaling of $\mathbf{g}_j$. Thus, without loss of generality, we define $\|\mathbf{g}_j\| = 1$. The SINR for information detection of the j -th tag can be obtained by

$$\rho_j(\mathbf{w}, \mathbf{g}_j) = \frac{|\mathbf{g}_j^H \mathbf{H}_j \mathbf{w}|^2}{\sigma^2 + |\mathbf{g}_j^H \tilde{\mathbf{H}}_j \mathbf{w}|^2} \quad (5.10)$$

where $\mathbf{w} = [w_1, \dots, w_N]^T$, $\mathbf{H}_j = \text{diag}(h_{j,1}h_{j,1}^b, \dots, h_{j,N}h_{j,N}^b)$ and $\tilde{\mathbf{H}}_j = \text{diag}(\sum_{u \neq j}^K h_{u,1}h_{u,1}^b, \dots, \sum_{u \neq j}^K h_{u,N}h_{u,N}^b)$.

5.2 Point-to-point backscatter

In this section, a point-to-point wireless powered backscatter as shown in Figure 5.1 is considered. For point-to-point backscatter, the subscript j in Eq. (5.4) can be removed such that

$$z_{DC}(\mathbf{s}, \Phi) = \frac{k_2}{2} R_{ant} \sum_{n=0}^{N-1} s_n^2 A_n^2 + \frac{3k_4}{8} R_{ant}^2 \left[\sum_{\substack{n_1, n_2, n_3, n_4 \\ n_1 + n_2 = n_3 + n_4}} \prod_{k=0}^3 s_{n_k} A_{n_k} \cos(\psi_{n_0} + \psi_{n_1} - \psi_{n_2} - \psi_{n_3}) \right]. \quad (5.11)$$

The interference is not present in a point-to-point backscatter. Therefore, the baseband signal on each frequency n can be written as

$$z_n = h_n h_n^b w_n m + v_n \quad (5.12)$$

where $v_n \sim \mathcal{CN}(0, \sigma^2)$. The SNR is maximized by applying the Maximum Ratio Combining (MRC) combiner at the reader such that

$$\rho(\mathbf{s}) = \frac{\sum_{n=0}^{N-1} |h_n h_n^b w_n|^2}{\sigma^2} = \frac{\sum_{n=0}^{N-1} A_n^2 (A_n^b)^2 s_n^2}{\sigma^2}. \quad (5.13)$$

The achievable SNR-harvested energy region is defined as

$$C_{\bar{\rho}-\bar{z}_{DC}}(P) \triangleq \left\{ (\bar{\rho}, \bar{z}_{DC}) : \bar{\rho} \leq \rho(\mathbf{s}), \bar{z}_{DC} \leq z_{DC}(\mathbf{s}, \Phi), \frac{1}{2} \|\mathbf{s}\|^2 \leq P \right\}. \quad (5.14)$$

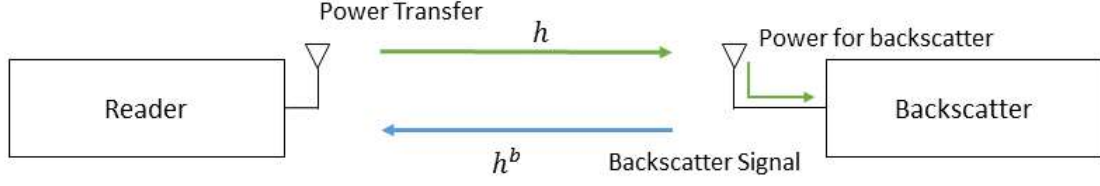


Fig. 5.1 Point-to-point backscatter system.

The optimization problem for Eq. (5.14) can be written as

$$\begin{aligned}
 (P\ 5.1) \quad & \max_{\mathbf{s}, \Phi} z_{DC}(\mathbf{s}, \Phi) \\
 & \text{subject to } \frac{1}{2} \|\mathbf{s}\|^2 \leq P, \\
 & \rho(\mathbf{s}) \geq \bar{\rho}.
 \end{aligned} \tag{5.15}$$

5.2.1 Feasibility Problem

Problem (P 5.1) is feasible if all constraints can be satisfied by some $\mathbf{s}$. The maximum feasible $\bar{\rho}$ for given transmit power constraint can be easily obtained as given by $\bar{\rho}_{max} = \frac{\max_n \{(A_n A_n^b)^2\} 2P}{\sigma^2}$.

5.2.2 Waveform Optimization

In this section, waveform optimization for a point-to-point backscatter is discussed. Note that the phase of the waveform ϕ_n changes the performance of z_{DC} but not the SNR. By looking at Eq. (5.11), it can be seen that matched filtering w.r.t. the phases of the channel is optimal. Therefore, the optimal ϕ_n is given as $\phi_n^* = -\psi_n$. Accordingly, Eq. (5.11) can be written as

$$z_{DC}(\mathbf{s}, \Phi^*) = \frac{k_2}{2} R_{ant} \sum_{n=0}^{N-1} s_n^2 A_n^2 + \frac{3k_4}{8} R_{ant}^2 \left[\sum_{\substack{n_1, n_2, n_3, n_4 \\ n_1 + n_2 = n_3 + n_4}} \prod_{j=0}^3 s_{n_j} A_{n_j} \right]. \tag{5.16}$$

For given ϕ_n^* , (P 5.1) can be written as

$$\begin{aligned}
 (P\ 5.2) \quad & \max_{\mathbf{s}} z_{DC}(\mathbf{s}, \Phi^*) \\
 & \text{subject to } \frac{1}{2} \|\mathbf{s}\|^2 \leq P, \\
 & \rho(\mathbf{s}) \geq \bar{\rho}.
 \end{aligned} \tag{5.17}$$

The optimization problem turns out to be maximizing the posynomial. Unfortunately, (P 5.2) is not a standard GP. Nevertheless, (P 5.2) can be formulated as a standard GP by using Reverse GP [78] as discuss in Chapter 4. By introducing t_0 , the objective function is changed into minimizing $1/t_0$ subject to $z_{DC} \geq t_0$. Then, the monomial approximation is applied to the lower bound inequality constraint. Let $g_k(\mathbf{s}, \Phi^*)$ and $f_n(\mathbf{s})$ are the monomial terms in the lower bound inequality constraint such that $z_{DC}(\mathbf{s}, \Phi^*) = \sum_k g_k(\mathbf{s}, \Phi^*)$ and $\rho(\mathbf{s}) = \sum_n f_n(\mathbf{s})$ with $f_n = A_n^2 (A_n^b)^2 s_n^2 / \sigma^2$, the lower bound inequality of a posynomial can be approximated with an upper bound inequality $\prod_k \left(\frac{g_k(\mathbf{s}, \Phi^*)}{\gamma_k} \right)^{-\gamma_k}$ and $\prod_n \left(\frac{f_n(\mathbf{s})}{\beta_n} \right)^{-\beta_n}$. Here, $\gamma_k = g_k(\mathbf{s}, \Phi^*) / z_{DC}(\mathbf{s}, \Phi^*)$ and $\beta_n = f_n(\mathbf{s}) / \rho(\mathbf{s})$ are chosen with $\sum_k \gamma_k = \sum_n \beta_n = 1$ and $\gamma_k, \beta_n \geq 0$. An iterative algorithm is used to tighten the upper bound approximation such that at each iteration, the standard GP is solved for an updated γ_k and β_n until convergence [18, 78]. Accordingly, the standard GP problem for (P 5.2) is written as

$$\begin{aligned}
 (P \ 5.3) \quad & \min_{\mathbf{s}, t_0} \quad 1/t_0 \\
 & \text{subject to} \quad \frac{1}{2} \|\mathbf{s}\|^2 \leq P, \\
 & \quad t_0 \prod_{k=1}^K \left(\frac{g_k(\mathbf{s}, \Phi^*)}{\gamma_k} \right)^{-\gamma_k} \leq 1, \\
 & \quad \bar{\rho} \prod_{n=0}^{N-1} \left(\frac{f_n(\mathbf{s})}{\beta_n} \right)^{-\beta_n} \leq 1.
 \end{aligned} \tag{5.18}$$

The iterative algorithm is detailed in Algorithm 9. Note that the successive approximation method used in the Algorithm 9, cannot be guaranteed to converge to the global optimal solution [78].

Algorithm 9 Point-to-point Backscatter Algorithm

- 1: **Initialize:** $\bar{\rho}, \Phi^*, \mathbf{s}, z_{DC}^{(0)} = 0$
 - 2: **repeat**
 - 3: $\check{\mathbf{s}} \leftarrow \mathbf{s}$
 - 4: $\gamma_k \leftarrow g_k(\check{\mathbf{s}}, \Phi^*) / z_{DC}(\check{\mathbf{s}}, \Phi^*), k = 1, \dots, K$
 - 5: $\beta_n \leftarrow f_n(\check{\mathbf{s}}) / \rho(\check{\mathbf{s}}), n = 0, \dots, N-1$
 - 6: $\mathbf{s} \leftarrow \arg \min (5.18)$
 - 7: $z_{DC}^{(i)} \leftarrow z_{DC}(\mathbf{s}, \Phi^*)$
 - 8: **until** $\left| z_{DC}^{(i)} - z_{DC}^{(i-1)} \right| < \varepsilon$
-

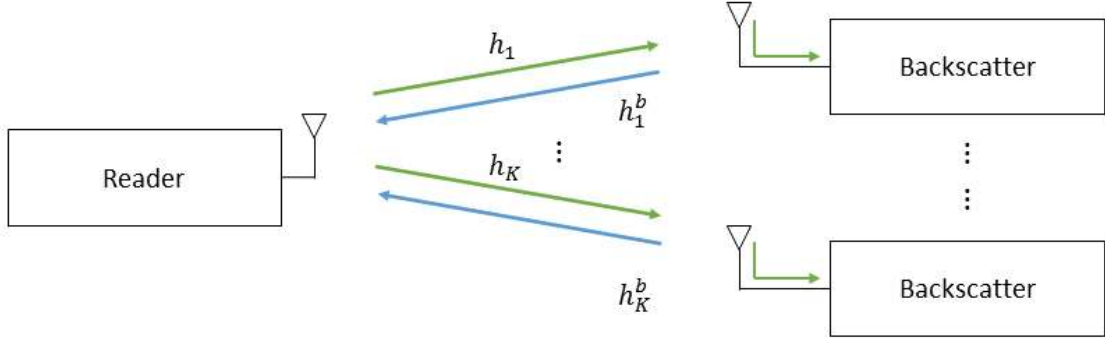


Fig. 5.2 Multi-user backscatter system.

5.3 Multi-user Backscatter

In this section, we discuss waveform optimization and characterize the trade-off between achievable SINR and energy harvested for a multi-user backscatter system as shown in Figure 5.2. The optimization problem is formulated as maximizing Z_{DC} given that an SINR constraint at each tag and transmit power constraint are satisfied. The optimization problem is given by

$$(P\ 5.4) \max_{\mathbf{w}, \mathbf{g}_j} \{Z_{DC} : \rho_j \geq \bar{\rho}_j, \|\mathbf{w}\|^2 \leq 2P, \|\mathbf{g}_j\| = 1, \forall j\}, \quad (5.19)$$

where $\bar{\rho}_j$ is the SINR constraint for the j -th tag. Note that, for a given transmit power constraint and channel realization, SINR constraints might not be satisfied with any $\mathbf{w}$ and $\mathbf{g}_j$. Therefore, we first should check the feasibility of problem (P 5.4) as discussed in the following section.

5.3.1 Feasibility Problem

For a given transmit power constraint and channel realization, problem (P 5.4) is feasible when the SINR constraint $\bar{\rho}_j$ for all tags can be satisfied. In the case that the problem has no solution, (P 5.4) is infeasible. Note that the waveform is only optimized when (P 5.4) is a feasible problem.

The coupling of the optimization variables $\mathbf{w}$ and $\mathbf{g}_j$ in the SINR constraints makes problem (P 5.4) non-convex. In this paper, we propose an algorithm based on alternating optimization [83] that optimizes $\mathbf{w}$ and $\mathbf{g}_j$ iteratively. As a result, we formulate the optimization problem as maximizing an auxiliary variable $\delta = \min_{j=1, \dots, K} \{\rho_j / \bar{\rho}_j\}$, which is used to check if

all SINR constraints are satisfied. Specifically, the feasibility problem is formulated as

$$\begin{aligned}
 (P\ 5.5) \quad & \max_{\mathbf{w}, \mathbf{g}_j} \quad \delta \\
 \text{s. t.} \quad & \frac{|\mathbf{g}_j^H \mathbf{H}_j \mathbf{w}|^2}{\sigma^2 \|\mathbf{g}_j\|^2 + |\mathbf{g}_j^H \tilde{\mathbf{H}}_j \mathbf{w}|^2} \geq \delta \bar{\rho}_j, \quad \forall j \\
 & \|\mathbf{w}\|^2 \leq 2P, \\
 & \|\mathbf{g}_j\| = 1. \quad \forall j
 \end{aligned} \tag{5.20}$$

Here, (P 5.5) is feasible if $\delta^* \geq 1$, which implies that $\rho_j \geq \bar{\rho}_j$ for all tags (i.e., the achievable SINR at all tags are larger than their corresponding SINR constraint). During the alternating optimization, δ is updated by using bisection search method as in [89] over $\delta_{\min} \leq \delta \leq \delta_{\max}$. $\delta_{\min}$ and $\delta_{\max}$ are chosen such that $\delta_{\min} < 1$ and $\delta_{\max} > 1$.

For a given $\mathbf{w}$, the optimal receive combiner that maximizes ρ_j in (P 5.5) is given in [86] as

$$\mathbf{g}_j^* = \frac{(\tilde{\mathbf{H}}_j \mathbf{w} \mathbf{w}^H \tilde{\mathbf{H}}_j^H + \mathbf{I} \sigma^2)^{-1} \mathbf{H}_j \mathbf{w}}{\|(\tilde{\mathbf{H}}_j \mathbf{w} \mathbf{w}^H \tilde{\mathbf{H}}_j^H + \mathbf{I} \sigma^2)^{-1} \mathbf{H}_j \mathbf{w}\|}. \tag{5.21}$$

For a given $\mathbf{g}_j$, as shown in [87, 88], we can adjust the direction of $\mathbf{w}$ such that $\mathbf{g}_j^H \mathbf{H}_j \mathbf{w}$ is real and non-negative without affecting the value of $|\mathbf{g}_j^H \mathbf{H}_j \mathbf{w}|$. Here, we assume that $N \geq K$ so that we can always get a $\mathbf{w}$ such that $\mathbf{g}_j^H \mathbf{H}_j \mathbf{w} = |\mathbf{g}_j^H \mathbf{H}_j \mathbf{w}|$ for all K tags. Accordingly, (P 5.5) can be recast as

$$\begin{aligned}
 (P\ 5.6) \quad & \max_{\mathbf{w}} \quad \delta \\
 \text{s. t.} \quad & (\mathbf{g}_j^H \mathbf{H}_j \mathbf{w})^2 \geq \delta \bar{\rho}_j (\sigma^2 \|\mathbf{g}_j\|^2 + |\mathbf{g}_j^H \tilde{\mathbf{H}}_j \mathbf{w}|^2), \quad \forall j \\
 & \|\mathbf{w}_j\|^2 \leq 2P, \\
 & \mathbf{g}_j^H \mathbf{H}_j \mathbf{w} = \Re(\mathbf{g}_j^H \mathbf{H}_j \mathbf{w}). \quad \forall j
 \end{aligned} \tag{5.22}$$

It can be found that (P 5.6) is a SOCP problem and can be solved by using CVX [90]. Specifically, at iteration l , δ is updated such that in case (P 5.6) is feasible, then $\delta_{\min}^{(l)} = \delta$. Otherwise, $\delta_{\max}^{(l)} = \delta$. Note that for a given $\mathbf{w}^{*(l-1)}$, Eq. (5.21) provides an optimal $\mathbf{g}_j^{*(l)}$ that maximizes ρ_j . However, it might not satisfy the SINR constraint $\bar{\rho}_j$ (i.e., $\delta^* < 1$). Therefore, $\mathbf{g}_j$ and (P 5.6) are optimize iteratively until convergence or $\delta_{\min} \geq 1$. The iterative algorithm is summarized in Algorithm 10.

Algorithm 10 Feasibility Algorithm

```

1: Initialize:  $\delta_{min}^{(0)} = 0, \delta_{max}^{(0)} > 1, \mathbf{w}^{(0)}, \varepsilon, l = 0$ 
2: repeat
3:    $l = l + 1, \delta_{min}^{(l)} = \delta_{min}^{(l-1)}$  and  $\delta_{max}^{(l)} = \delta_{max}^{(0)}$ .
4:   Update  $\mathbf{g}_j^{*(l)}$  by using (5.21) for  $j = 1, \dots, K$ .
5:   while  $\delta_{max}^{(l)} - \delta_{min}^{(l)} > \varepsilon$  do
6:      $\delta = (\delta_{max}^{(l)} + \delta_{min}^{(l)})/2$ .
7:     If (P 5.6) is feasible,  $\delta_{min}^{(l)} = \delta$ , store  $\mathbf{w}^*$ .
8:     Else  $\delta_{max}^{(l)} = \delta$ .
9:   end while
10:  Update  $\mathbf{w}^{*(l)} \leftarrow \mathbf{w}^*$  and  $\delta_{min}^{*(l)} \leftarrow \delta_{min}^{(l)}$ .
11: until  $\delta_{min}^{*(l)} - \delta_{min}^{*(l-1)} \leq \varepsilon$  or  $\delta_{min}^{*(l)} \geq 1$ .
12: Update  $\delta^* = \delta_{min}^{*(l)}$ .

```

According to Algorithm 10, for given $\mathbf{w}^{*(l-1)}$, $\mathbf{g}_j^{*(l)}$ is obtained by using Eq. (5.21). Subsequently, for given $\mathbf{g}_j^{*(l)}$, $\mathbf{w}^{*(l)}$ is optimized by using (P 5.6). Note that $\delta^* = \delta_{min}^{*(l)}$ (as in line 12). For convenience, we define $\mathbf{g} = (\mathbf{g}_1; \dots; \mathbf{g}_K)$ and the objective in (P 5.5) as the objective function $\delta = \delta_{min}(\mathbf{g}, \mathbf{w})$. Because for given $\mathbf{w}^{*(l-1)}$, $\mathbf{g}_j^{*(l)}$ from Eq. (5.21) is the optimal solution that maximize ρ_j and for given $\mathbf{g}^{*(l)}$, problem (P 5.6) is convex, it shows that $\delta_{min}(\mathbf{g}^{*(l-1)}, \mathbf{w}^{*(l-1)}) \leq \delta_{min}(\mathbf{g}^{*(l)}, \mathbf{w}^{*(l)})$. Hence, we can find that the sequence δ_{min} monotonically increases. As δ is upper bounded by δ_{max} , Algorithm 10 always converges. Since $\delta_{min}^{*(l)} \geq \delta_{min}^{*(l-1)}$, we set $\delta_{min}^{(l)} = \delta_{min}^{*(l-1)}$ in line 3 of Algorithm 10. It can be observed that $\delta_{min}(\mathbf{g}^{*(l-1)}, \mathbf{w}^{*(l-1)}) \leq \delta_{min}(\mathbf{g}^{*(l-1)}, \mathbf{w}^{*(l)}) \leq \delta_{min}(\mathbf{g}^{*(l)}, \mathbf{w}^{*(l)})$. As $l \rightarrow \infty$, it can be shown that Algorithm 10 converges to a stationary point of (P 5.5) [89, 91].

5.3.2 Waveform Optimization

In this section, we discuss waveform optimization in a multi-user system as formulated in (P 5.4). Recall that in a point-to-point system [82], exploiting reversed GP to optimize the waveform requires initial choice of phases before the magnitudes are optimized. In order to maximize Z_{DC} , waveform phases are chosen such that the $\cos(\cdot)$ in (5.6) is equal to 0. In a point-to-point system, the optimal phases can be obtained in closed-form [82]. However, closed-form solution cannot be obtained for a system with K tags. Therefore, we propose an algorithm that jointly optimized the phases and the magnitudes, by making use of the matrix formulation introduced in [20, 92]. However, this approach cannot be extended to a truncation order higher than 4.

$$\mathbf{M}_j = \begin{matrix} & \begin{matrix} \xleftarrow{k=-1} & & & & \end{matrix} \\ \begin{pmatrix} h_{j,1}^* h_{j,1} & h_{j,1}^* h_{j,2} & \dots & \dots & h_{j,1}^* h_{j,N} \\ h_{j,2}^* h_{j,1} & h_{j,2}^* h_{j,2} & \dots & \dots & \vdots \\ \vdots & \vdots & \ddots & \vdots & \vdots \\ h_{j,N-1}^* h_{j,1} & \vdots & \ddots & \ddots & h_{j,N-1}^* h_{j,N} \\ h_{j,N}^* h_{j,1} & h_{j,N}^* h_{j,2} & \dots & h_{j,N}^* h_{j,N-1} & h_{j,N}^* h_{j,N} \end{pmatrix} & \begin{matrix} \\ \\ \\ \xleftarrow{k=1} \\ \xleftarrow{k=0} \end{matrix} \end{matrix}$$

Fig. 5.3 An N -by- N matrix $\mathbf{M}_j$.

By letting $\mathbf{h}_j = [h_{j,1}, \dots, h_{j,N}]^T$, we define $\mathbf{M}_j = \mathbf{h}_j^* \mathbf{h}_j^T$. As shown in Figure 5.3, $k = 0$ is the index of the main diagonal of $\mathbf{M}_j$, $k \in \{1, \dots, N-1\}$ is the index of the k -th diagonal above the main diagonal and $k \in \{-1, \dots, -(N-1)\}$ is the index of the $|k|$ -th diagonal below the main diagonal. We define $\mathbf{M}_{j,k}$ by retaining the k -th diagonal of $\mathbf{M}_j$ while setting all other entries as 0. Thus, Z_{DC} can be written as

$$\begin{aligned} Z_{DC} &= \sum_{j=1}^K c_j \left(\frac{1}{2} \beta_2 \sum_n w_n^* h_{j,n}^* h_{j,n} w_n + \frac{3}{8} \beta_4 \right. \\ &\quad \left. \sum_{\substack{n_1, n_2, n_3, n_4 \\ n_1 - n_3 = -(n_2 - n_4)}} w_{n_3}^* h_{j,n_3}^* h_{j,n_1} w_{n_1} w_{n_4}^* h_{j,n_4}^* h_{j,n_2} w_{n_2} \right) \\ &= \sum_{j=1}^K c_j \left(\frac{1}{2} \beta_2 \mathbf{w}^H \mathbf{M}_{j,0} \mathbf{w} + \frac{3}{8} \beta_4 \mathbf{w}^H \mathbf{M}_{j,0} \mathbf{w} (\mathbf{w}^H \mathbf{M}_{j,0} \mathbf{w})^H \right. \\ &\quad \left. + \frac{3}{4} \beta_4 \sum_{k=1}^{N-1} \mathbf{w}^H \mathbf{M}_{j,k} \mathbf{w} (\mathbf{w}^H \mathbf{M}_{j,k} \mathbf{w})^H \right). \end{aligned} \quad (5.23)$$

where $\beta_2 = k_2 R_{ant}$ and $\beta_4 = k_4 R_{ant}^2$.

Z_{DC} in Eq. (5.23) is a quartic polynomial, which in general makes problem (P 5.4) NP-hard. In order to address the quartic polynomial problem, auxiliary variables t_k for $k = 0, \dots, N-1$ are introduced such that $\mathbf{w}^H \mathbf{M}_{j,k} \mathbf{w} = t_{j,k}$ [20, 92]. However, it is noted that for all j with $k \neq 0$, $\mathbf{M}_{j,k}$ is not a Hermitian matrix. Therefore, the term $\mathbf{w}^H \mathbf{M}_{j,k} \mathbf{w}$ is a bilinear function which may also lead to a NP-hard problem [92]. Hence, an auxiliary $\mathbf{X} = \mathbf{w} \mathbf{w}^H$ is defined as in [92] to linearize the bilinear term such that we can write $\mathbf{w}^H \mathbf{M}_{j,k} \mathbf{w} = \text{Tr}\{\mathbf{M}_{j,k} \mathbf{X}\} = t_{j,k}$. By letting $\mathbf{A}_0 = \text{diag}\{-3\beta_4/8, -3\beta_4/4, \dots, -3\beta_4/4\} \preceq 0$ and $\mathbf{t}_j = [t_{j,0}, \dots, t_{j,N-1}]^T$, we can

write $q_j(\mathbf{t}_j) \triangleq \mathbf{t}_j^H \mathbf{A}_0 \mathbf{t}_j = -3\beta_4/8 t_{j,0} t_{j,0}^* - 3\beta_4/4 \sum_{k=1}^{N-1} t_{j,k} t_{j,k}^*$. Therefore, Eq. (5.23) can be written as

$$Z_{DC} = \sum_{j=1}^K c_j \left(\frac{\beta_2}{2} t_{j,0} - q_j(\mathbf{t}_j) \right) \quad (5.24)$$

and Eq. (5.10) can be written as

$$\rho_j(\mathbf{X}, \mathbf{g}_j) = \frac{\text{Tr}(\mathbf{H}_j^H \mathbf{g}_j \mathbf{g}_j^H \mathbf{H}_j \mathbf{X})}{\sigma^2 + \text{Tr}(\tilde{\mathbf{H}}_j^H \mathbf{g}_j \mathbf{g}_j^H \tilde{\mathbf{H}}_j \mathbf{X})}. \quad (5.25)$$

Hence, problem (P 5.4) can be recast as

$$(P\ 5.7) \quad \min_{\gamma, \mathbf{t}_j, \mathbf{X}, \mathbf{g}_j} \quad \gamma \quad (5.26a)$$

$$\text{s. t.} \quad \sum_{j=1}^K c_j (-\beta_2 t_{j,0} + q_j(\mathbf{t}_j)) - \gamma \leq 0, \quad (5.26b)$$

$$\frac{\text{Tr}(\mathbf{H}_j^H \mathbf{g}_j \mathbf{g}_j^H \mathbf{H}_j \mathbf{X})}{\sigma^2 + \text{Tr}(\tilde{\mathbf{H}}_j^H \mathbf{g}_j \mathbf{g}_j^H \tilde{\mathbf{H}}_j \mathbf{X})} \geq \bar{\rho}_j, \quad \forall j \quad (5.26c)$$

$$\text{Tr}\{\mathbf{M}_{j,k} \mathbf{X}\} = t_{j,k}, \quad \forall j, k \quad (5.26d)$$

$$\text{Tr}\{\mathbf{M}_{j,k}^H \mathbf{X}\} = t_{j,k}^*, \quad \forall j, k \neq 0 \quad (5.26e)$$

$$\text{Tr}\{\mathbf{X}\} \leq 2P, \quad (5.26f)$$

$$\text{rank}\{\mathbf{X}\} = 1, \quad (5.26g)$$

$$\|\mathbf{g}_j\| = 1. \quad \forall j \quad (5.26h)$$

It can be seen that the non-convex quadratic constraint (5.26b), the coupled optimization variables in (5.26c) and the rank constraint (5.26g) make problem (P 5.7) intractable. Meanwhile, the necessary condition for optimizing γ , $\mathbf{t}_j$, $\mathbf{X}$ and $\mathbf{g}_j$ is nothing but to reduce the value of γ . Motivated by this necessary condition, we propose an iterative algorithm to solve problem (P 5.7). We first relax the rank constraint such that (P 5.7) can be written as

$$(P\ 5.8) \quad \min_{\gamma, \mathbf{t}_j, \mathbf{X}, \mathbf{g}_j} \quad \gamma \quad (5.27)$$

(5.26b), (5.26c), (5.26d), (5.26e), (5.26f), (5.26h).

As in [92], we then linearize (5.26b) to solve (P 5.8). Specifically, at iteration l , the non-convex term $q_j(\mathbf{t}_j)$ for $j = 1, \dots, K$ in (5.26b) is approximated at $\mathbf{t}_j^{(l-1)}$ (the optimal $\mathbf{t}_j^*$ achieved at iteration $(l-1)$) as a linear function by its first-order Taylor expansion given

as $\tilde{q}_j(\mathbf{t}_j, \mathbf{t}_j^{(l-1)}) = 2\Re\{[\mathbf{t}_j^{(l-1)}]^H \mathbf{A}_0 \mathbf{t}_j\} - [\mathbf{t}_j^{(l-1)}]^H \mathbf{A}_0 \mathbf{t}_j^{(l-1)}$. Thus, at the l -th iteration, the following approximate problem is solved

$$(P\ 5.9) \quad \min_{\gamma, \mathbf{t}_j, \mathbf{X}, \mathbf{g}_j} \quad \gamma \quad (5.28a)$$

$$\text{s. t.} \quad \sum_{j=1}^K c_j(-\beta_2 t_{j,0} + \tilde{q}_j(\mathbf{t}_j, \mathbf{t}_j^{(l-1)})) - \gamma \leq 0, \quad (5.28b)$$

$$(5.26c), (5.26d), (5.26e), (5.26f), (5.26h).$$

Since $-q_j(\mathbf{t}_j)$ is convex, it can be shown that $q_j(\mathbf{t}_j) \leq \tilde{q}_j(\mathbf{t}_j, \mathbf{t}_j^{(l-1)})$. Therefore, $\tilde{q}_j(\mathbf{t}_j^{(l)}, \mathbf{t}_j^{(l)}) = q_j(\mathbf{t}_j^{(l)}) \leq \tilde{q}_j(\mathbf{t}_j^{(l)}, \mathbf{t}_j^{(l-1)})$, which indicates that the solution of (P 5.9) always satisfy (5.26b).

In the following, we show that the value of γ decreases over iterations. Unfortunately, (P 5.9) is non-convex with respect to $\mathbf{X}$ and $\mathbf{g}_j$. Therefore, we optimize $\mathbf{X}$ and $\mathbf{g}_j$ by using alternating algorithm. Specifically at the l -th iteration of SCA, we first optimize $\mathbf{g}_j^{(l)}$ for given $\mathbf{X}^{(l-1)}$. Then $\mathbf{X}^{(l)}$ is optimized with the updated $\mathbf{g}_j^{(l)}$.

For a given $\mathbf{X}$, problem (P 5.9) can be reduced to

$$(P\ 5.10) \quad \max_{\mathbf{g}_j} \quad \frac{\mathbf{g}_j^H \mathbf{H}_j \mathbf{X} \mathbf{H}_j^H \mathbf{g}_j}{\mathbf{g}_j^H (\sigma^2 \mathbf{I} + \tilde{\mathbf{H}}_j \mathbf{X} \tilde{\mathbf{H}}_j^H) \mathbf{g}_j}, \quad (5.29)$$

$$\text{s. t.} \quad \|\mathbf{g}_j\| = 1.$$

(P 5.10) can be transformed to $\rho_j(\tilde{\mathbf{g}}_j) = \frac{\tilde{\mathbf{g}}_j^H \mathbf{D} \tilde{\mathbf{g}}_j}{\tilde{\mathbf{g}}_j^H \tilde{\mathbf{g}}_j}$ where $\mathbf{D} = (\mathbf{C})^{-1} \mathbf{H}_j \mathbf{X} \mathbf{H}_j^H (\mathbf{C}^H)^{-1}$, $\mathbf{C} \mathbf{C}^H$ is the Cholesky decomposition of $\sigma^2 \mathbf{I} + \tilde{\mathbf{H}}_j \mathbf{X} \tilde{\mathbf{H}}_j^H$ and $\tilde{\mathbf{g}}_j = \mathbf{C}^H \mathbf{g}_j$. Here, the optimal $\tilde{\mathbf{g}}_j^*$ is the eigenvector corresponding to the largest eigenvalue of $\mathbf{D}$. Accordingly, $\mathbf{g}_j^* = (\mathbf{C}^H)^{-1} \tilde{\mathbf{g}}_j^* / \|(\mathbf{C}^H)^{-1} \tilde{\mathbf{g}}_j^*\|$.

For a given $\mathbf{g}_j$, problem (P 5.9) can be reduced to

$$(P\ 5.11) \quad \min_{\gamma, \mathbf{t}_j, \mathbf{X} \succeq 0} \quad \gamma \quad (5.30a)$$

$$\text{s. t.} \quad \sum_{j=1}^K c_j(-\beta_2 t_{j,0} + \tilde{q}_j(\mathbf{t}_j, \mathbf{t}_j^{(l-1)})) - \gamma \leq 0, \quad (5.30b)$$

$$\frac{\text{Tr}(\mathbf{H}_j^H \mathbf{g}_j \mathbf{g}_j^H \mathbf{H}_j \mathbf{X})}{\sigma^2 \|\mathbf{g}_j\|^2 + \text{Tr}(\tilde{\mathbf{H}}_j^H \mathbf{g}_j \mathbf{g}_j^H \tilde{\mathbf{H}}_j \mathbf{X})} \geq \bar{\rho}_j, \forall j \quad (5.30c)$$

$$\text{Tr}\{\mathbf{M}_{j,k} \mathbf{X}\} = t_{j,k}, \quad \forall j, k \quad (5.30d)$$

$$\text{Tr}\{\mathbf{M}_{j,k}^H \mathbf{X}\} = t_{j,k}^*, \quad \forall j, k \quad (5.30e)$$

$$\text{Tr}\{\mathbf{X}\} \leq 2P. \quad (5.30f)$$

Algorithm 11 Iterative Algorithm

-
- 1: **Initialize:** $l^{(0)} \leftarrow 0$, $\mathbf{X}^{(0)}$, ε , $\gamma^{(0)}$ and update $\{\mathbf{t}_j^{(0)}\}_{j=1}^K$ by (5.30d).
 - 2: **repeat**
 - 3: $l = l + 1$
 - 4: Find $\mathbf{g}_j^*$ in (P 5.10); $\mathbf{g}_j^{(l)} \leftarrow \mathbf{g}_j^*$.
 - 5: Solve SDP problem (P 5.11); $\mathbf{X}^{(l)} \leftarrow \mathbf{X}^*$, $\gamma^{(l)} \leftarrow \gamma^*$.
 - 6: Update $t_{j,k}^{(l)} \forall j, k$ in (5.30d) and (5.30e).
 - 7: **until** $|\gamma^{(l)} - \gamma^{(l-1)}|/|\gamma^{(l)}| \leq \varepsilon$.
 - 8: Find rank-one solution of $\mathbf{X}^*$.
-

(P 5.11) is a SDP problem and can be solved by CVX . Note that problem (P 5.11) has a solution if for given $\mathbf{g}_j$, (P 5.11) is a feasible problem. The solution from Algorithm 10 $\mathbf{w}^*$ can be used as the initial point in the iterative algorithm such that $\mathbf{X}^{(0)} = \mathbf{w}^* \mathbf{w}^{*H}$. Therefore, a feasible solution of (P 5.11) (and also (P 5.9)) can always be obtained.

The iterative algorithm to solve (P 5.9) is summarized in Algorithm 11 (from line 2 to line 7). Because $\mathbf{g}_j^*$ is the global optimal solution of problem (P 5.10) and problem (P 5.11) is convex, it shows that $\gamma^{(l-1)} \geq \gamma^{(l)}$. Note that the stopping criterion of Algorithm 11 is related to the convergence of γ (i.e., objective function of problem (P 5.9)). As the value of γ is monotonically decreasing and bounded, the iterative algorithm is guaranteed to converge. In Appendix B, we show that under the condition that the eigenvector of $\mathbf{D}$ can be uniquely obtained, the minimizers in Algorithm 11 may converge to a KKT point of problem (P 5.9).

The solution from the interior-point method in (P 5.11) usually yields to a high rank of $\mathbf{X}^*$ [93]. In the following, in order to obtain the rank-constrained solution that satisfy (5.26g), we derive a rank-one solution from the high rank $\mathbf{X}^*$ (line 8 of Algorithm 11).

Obtaining a Rank-One $\mathbf{X}$

We first show that if $K \leq 2$ (i.e., $j = 1, 2$), (P 5.8) can offer a rank-one optimal solution of $\mathbf{X}^*$ although the solution in (P 5.11) has a high rank of $\mathbf{X}^*$. From [92], (P 5.8) can be converted into an equivalent form as

$$\begin{aligned}
 (P\ 5.12) \quad & \min_{\mathbf{X} \succeq 0, \mathbf{g}_j} \quad \text{Tr}\{\mathbf{A}\mathbf{X}\} \\
 \text{s. t.} \quad & \frac{\text{Tr}(\mathbf{H}_j^H \mathbf{g}_j \mathbf{g}_j^H \mathbf{H}_j \mathbf{X})}{\sigma^2 + \text{Tr}(\tilde{\mathbf{H}}_j^H \mathbf{g}_j \mathbf{g}_j^H \tilde{\mathbf{H}}_j \mathbf{X})} \geq \bar{\rho}_j, \quad \forall j \\
 & \text{Tr}\{\mathbf{X}\} \leq 2P, \\
 & \|\mathbf{g}_j\| = 1. \quad \forall j
 \end{aligned} \tag{5.31}$$

where $\mathbf{A} = \mathbf{C} + \mathbf{C}^H$ is a Hermitian and $\mathbf{C} = \sum_{j=1}^K c_j \left(-\frac{\beta_2 + 3/2\beta_4 t_{j,0}^{(l-1)}}{4} \mathbf{M}_{j,0} - 3/4\beta_4 \sum_{k=1}^{N-1} [t_{j,k}^{(l-1)}]^* \mathbf{M}_{j,k} \right)$. Given an optimized $\mathbf{g}_j^*$ and by defining $\mathbf{B}_j = (\mathbf{H}^H \mathbf{g}_j^* \mathbf{g}_j^{*H} \mathbf{H} - \tilde{\mathbf{H}}^H \mathbf{g}_j^* \mathbf{g}_j^{*H} \tilde{\mathbf{H}})$, (P 5.12) can be recast as

$$\begin{aligned} (P\ 5.13) \quad & \min_{\mathbf{X} \succeq 0} \quad \text{Tr}\{\mathbf{A}\mathbf{X}\} \\ \text{s. t.} \quad & \text{Tr}(\mathbf{B}_j \mathbf{X}) - \sigma^2 \bar{\rho}_j \geq 0, \quad \forall j \\ & \text{Tr}\{\mathbf{X}\} \leq 2P, \end{aligned} \quad (5.32)$$

where $\mathbf{B}_j$ is Hermitian. According to [94], (P 5.13) is shown to be a separable SDP. By applying [94, Proposition 3.5], we can show that (P 5.13) can yield a rank one optimal solution when $K \leq 2$. Accordingly, a rank one $\mathbf{X}_{r1}^* = \mathbf{w}^* \mathbf{w}^{*H}$ can be obtained by using Rank Reduction (RR) [94]. Hence, we propose an algorithm that first iteratively optimizes $\mathbf{g}_j$ in (P 5.10) and $\mathbf{X}$ in (P 5.11). RR procedure is then performed to the high rank solution of $\mathbf{X}^*$ to find $\mathbf{w}^*$. The algorithm is summarized in Algorithm 11.

When $K \geq 3$, (P 5.13) may not have a rank-one optimal solution of $\mathbf{X}^*$. In this scenario, we solve (P 5.8) by iteratively optimizing $\mathbf{g}_j$ in (P 5.10) and $\mathbf{X}$ in (P 5.11) until convergence. Then, rank randomization method is performed to obtain T rank-one solution from higher rank of $\mathbf{X}^*$. By taking the eigenvalue decomposition (EVD) $\mathbf{X}^* = \mathbf{U} \Sigma^{1/2} \mathbf{U}^H$, we generate T random vectors $\mathbf{w}_t = \mathbf{U} \Sigma^{1/2} \mathbf{v}_t$ for $t = 1, \dots, T$ where $\mathbf{v}_t$ is a random vector with each complementary entry is from a circular uniform distribution such that $E[\mathbf{v}_t \mathbf{v}_t^H] = \mathbf{I}$. For each $\mathbf{w}_t$, we find their corresponding $\mathbf{g}_{t,j}$ from Eq. (5.21). Consequently, the best feasible rank-one solution satisfying constraint (5.26c) and (5.26f) is obtained as follows. Among T $\mathbf{w}_t$ and their corresponding $\mathbf{g}_{t,j}$, we find $\mathbf{w}'_t$ and $\mathbf{g}'_{t,j}$ that can satisfy constraint (5.26c) and (5.26f). Accordingly, the best feasible solution is $\mathbf{w}^* = \arg\max_{\mathbf{w}'_t} Z_{DC}(\mathbf{w}'_t)$ where $\mathbf{X}_{r1}^* = \mathbf{w}^* \mathbf{w}^{*H}$ and $\mathbf{g}_j^*$ is given as $\mathbf{g}'_{t,j}$ that associate with $\mathbf{w}^*$. The algorithm is summarized in Algorithm 11.

A Simplified Algorithm

A simplified iterative algorithm with a lower computational complexity than Algorithm 11 is proposed. Given an initial value of $\mathbf{X}$ with its corresponding $\mathbf{g}_j^*$ from (P 5.10), by fixing $\mathbf{g}_j^*$ at each iteration, (P 5.11) is optimized iteratively until convergence. Then, a rank-one $\mathbf{X}$ is obtained as discussed in section 5.3.2. The iterative algorithm is shown in Algorithm 12. For given $\mathbf{g}_j^*$ in line 2, the iterative algorithm converges to a stationary point of (P 5.11) [92]. Although this algorithm provide a better complexity compared to Algorithm 11, the performance of Algorithm 11 is expected to be better. This is because alternating optimization that optimizes $\mathbf{g}_j$ (in line 4 of Algorithm 11) and $\mathbf{X}$ iteratively in Algorithm 11 can reduce the value of γ over iteration.

Algorithm 12 A Simplified Algorithm

-
- 1: **Initialize:** $l^{(0)} \leftarrow 0$, $\mathbf{X}^{(0)}$, ε , $\gamma^{(0)}$ and update $\{\mathbf{t}_j^{(0)}\}_{j=1}^K$ by (5.30d).
 - 2: Find $\mathbf{g}_j^*$ in (P 5.10) for given $\mathbf{X}^{(0)}$.
 - 3: **repeat**
 - 4: $l = l + 1$
 - 5: Solve SDP problem in (P 5.11); $\mathbf{X}^{(l)} \leftarrow \mathbf{X}^*$, $\gamma^{(l)} \leftarrow \gamma^*$.
 - 6: Update $t_{j,k}^{(l)} \forall j, k$ by (5.30d) and (5.30e).
 - 7: **until** $|\gamma^{(l)} - \gamma^{(l-1)}|/|\gamma^{(l)}| \leq \varepsilon$.
 - 8: Find rank-one solution of $\mathbf{X}^*$.
-

Waveform Optimization based on the Linear EH Model

The linear model of EH has been widely used in the SWIPT literature [15]. The linear EH model only accounts for the second order term in Eq. (2.21). By substituting (5.26b) for $\sum_{j=1}^K (-\beta_2 t_{j,0}) - \gamma_1 \leq 0$, the waveform optimization with the linear EH model can be solved by using Algorithm 11.

5.4 Simulation Results

This section presents the simulation results. The simulation consider a system with a centre frequency of 5.18 GHz, 36 dBm EIRP, 2 dBi receive and transmit antenna gain at the backscatter and 2 dBi receive antenna gain at the reader is considered. The path loss for forward and backward link are 58 dB with the NLOS channel power delay profile is obtained from model B in [81]. The channel taps each with an average power β_l are independent, circularly symmetric complex random gaussian distributed and normalized such that $\sum_l \beta_l = 1$. This leads to an average receive power of -20 dBm at the backscatter. The SNR at the reader is defined as P/σ_n^2 . In the simulations, it is assumed that each tag has the same SINR constraint $\bar{\rho}$ and the weight for energy harvested $c_j = 1$ for all tags. The SINR (SNR for a point-to-point system) on the x-axis is the SINR constraint $\bar{\rho}$.

Figure 5.5 shows the simulation result for a point-to-point backscatter as discussed in Section (5.2.2). The simulation is run over a channel realization for $B=[1, 10]$ MHz bandwidths and the SNR is set to 10dB with the frequency response of the channel is illustrated in Figure 5.4. The maximum achievable SNR (the extreme point on the x-axis) is obtained by allocating full transmit power to a single waveform subcarrier corresponds to $\arg\max_n |h_n h_n^b|$. The maximum Z_{DC} can be obtained from [18, 20]. It is observed that the shape of the region depends on the frequency response of the channel. Specifically, for frequency response in Figure 5.4, since multiple waveform subcarrier increase the

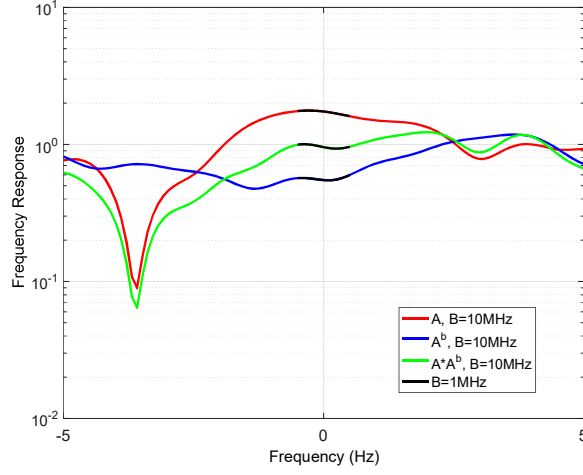
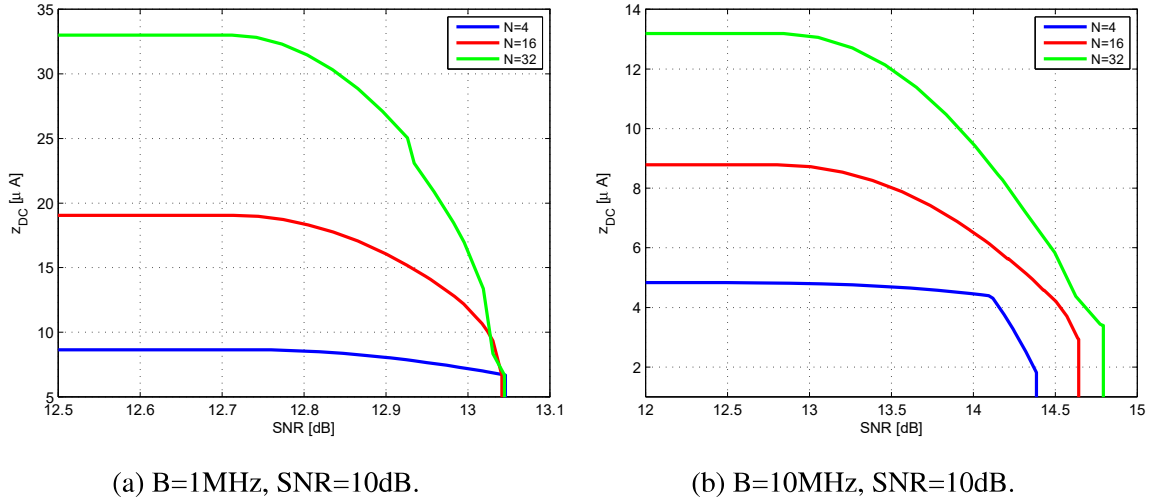
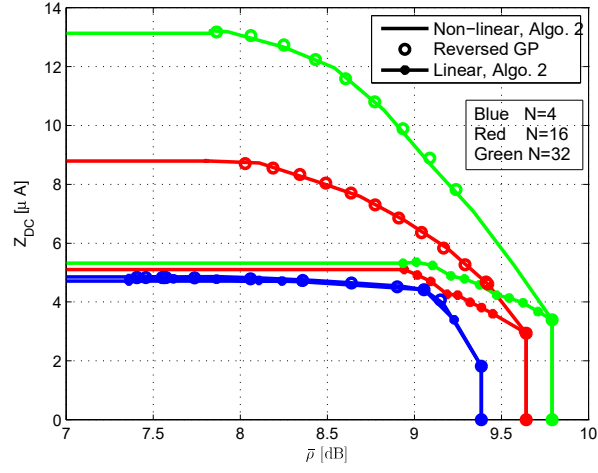


Fig. 5.4 Frequency response for a point-to-point backscatter.

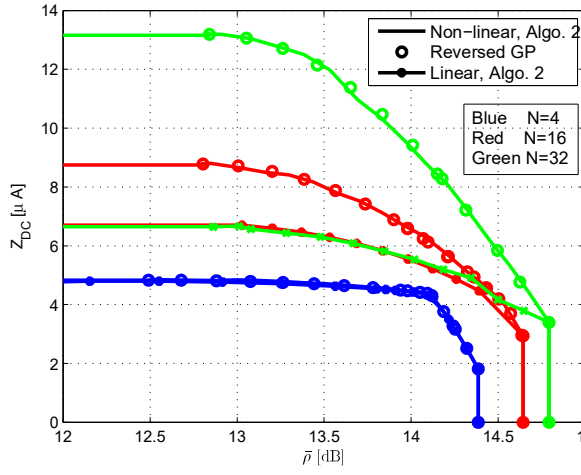
Fig. 5.5 ρ - z_{DC} region for a point-to-point backscatter from Section (5.2.2).

performance of Z_{DC} , $B=1\text{MHz}$ has a larger Z_{DC} over 10MHz bandwidth that exhibits deep fade channel. On the other hand, from Figure 5.4, the maximum $|h_n h_n^b|$ is observed for $B=10\text{MHz}$ at around 2MHz . Since a single subcarrier waveform maximize SNR, $B=10\text{MHz}$ has a larger maximum SNR than $B=1\text{MHz}$.

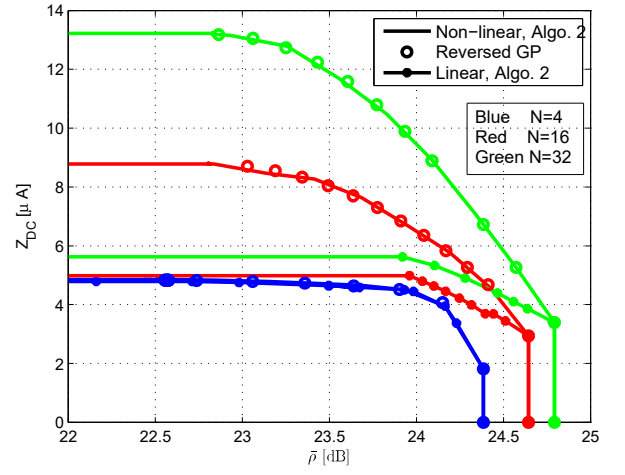
Figure 5.6 and Figure 5.8 show the simulation result for a point-to-point backscatter and $K = 3$, optimized based on matrix formulation as discussed in Section (5.3.2). The blue, red and green colours in the figures refer to $N = 4, 16$ and 32 , respectively. The solid lines in the figures indicate the result for waveform optimization based on non-linear EH model and obtained by using Algorithm 11. The simulation is run over a channel realization



(a) SNR=5dB.



(b) SNR=10dB.



(c) SNR=20dB.

Fig. 5.6 ρ - z_{DC} region for a point-to-point backscatter.

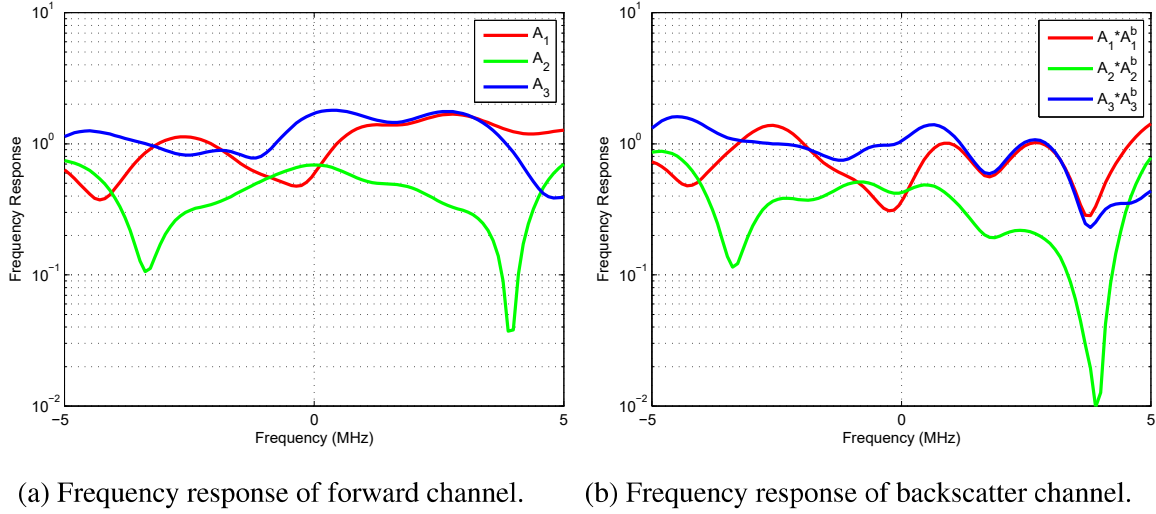
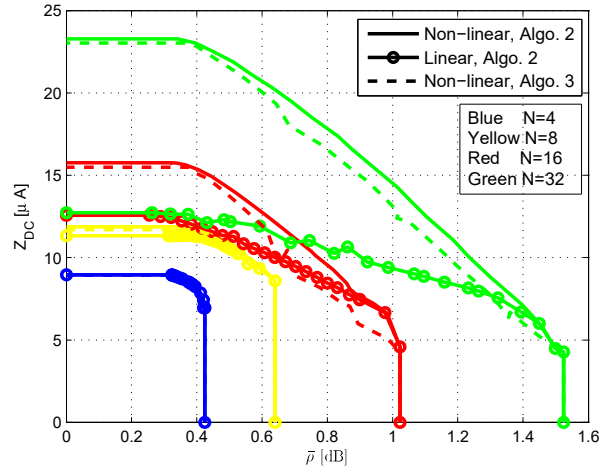


Fig. 5.7 Frequency response for three user.

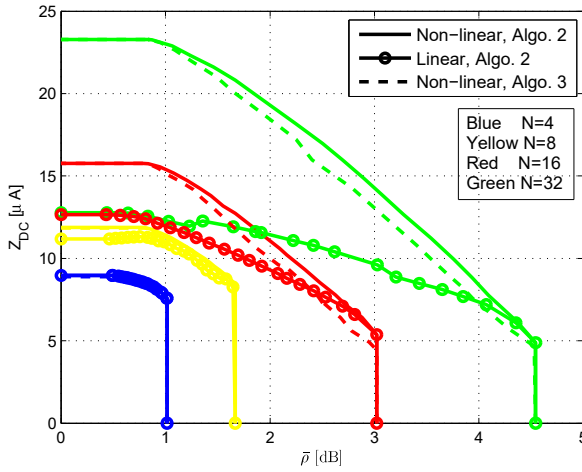
for three different SNRs (5, 10 and 20 dB) and $B=10\text{MHz}$ with the frequency response of the channel for point-to-point and multi-user system are illustrated in Figure 5.4 and Figure 5.7, respectively. The extreme point on the x-axis and y-axis for $K=3$ are obtained from Algorithm 10 and [92], respectively.

For a point-to-point system, because the phase of the transmit waveform can be obtained in a closed-form as in [82], the waveform can also be optimized by using Reversed GP [82] as described in Section (5.2.2). It is observed that the achievable $\bar{\rho}$ - Z_{DC} region obtained by using Reversed GP and Algorithm 11 are comparable as shown in Figure 5.6. Due to the non-convexity of the problem, the solutions from Reversed GP and Algorithm 11 cannot guarantee to converge to the global optimal solution. But both approaches converge to a KKT point. Table 5.1 compares the average computational complexity for waveform optimization based on Algorithm 11 and Reversed GP. Z_{DC} and the elapsed time are averaged over several channel realizations with $N=4, 8$. To draw the comparison, we employ the same initial point and $\epsilon=10^{-7}$ for both algorithms and the SNR constraint at the tag $\bar{\rho}=10\text{dB}$. The simulation is conducted by MATLAB R2013b on a computer with an Intel Core i7 processor at 3.3GHz, RAM of 16GB and Windows 10 Pro. The simulation results show that the average elapsed time for Algorithm 11 is significantly less than Reversed GP, while the Z_{DC} value from both algorithms is comparable.

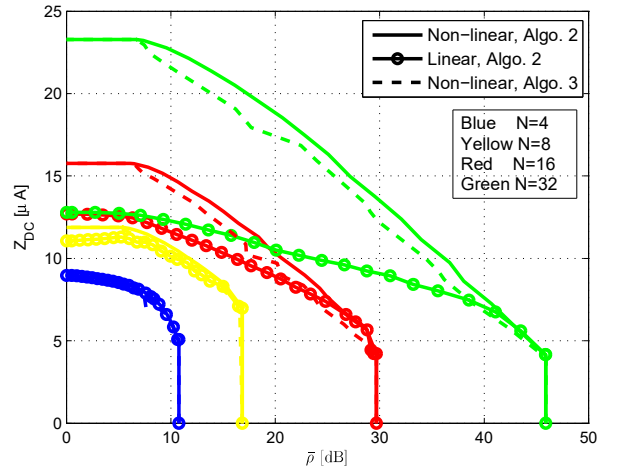
From Figure 5.6 and Figure 5.8, by looking at $\bar{\rho}$ - Z_{DC} region for non-linear EH model obtained from Algorithm 11, it is observed that the performance of wirelessly powered backscatter communication is subject to a trade-off between Z_{DC} and $\bar{\rho}$. That is, maximizing the Z_{DC} would result in a decrease in $\bar{\rho}$. The second observation is that for all N , as $\bar{\rho}$



(a) SNR=5dB.



(b) SNR=10dB.



(c) SNR=20dB.

Fig. 5.8 $\bar{p}$ - z_{DC} region for three-user backscatters.

Table 5.1 Elapsed running time: Algorithm 11 vs. Reversed GP

Algorithm	N= 4		N= 8	
	Average $Z_{DC}(\mu A)$	Average elapsed time(s)	Average $Z_{DC}(\mu A)$	Average elapsed time(s)
Algorithm 2	6.624	1.675	8.113	1.796
Reversed GP	6.617	118.72	8.17	215.215

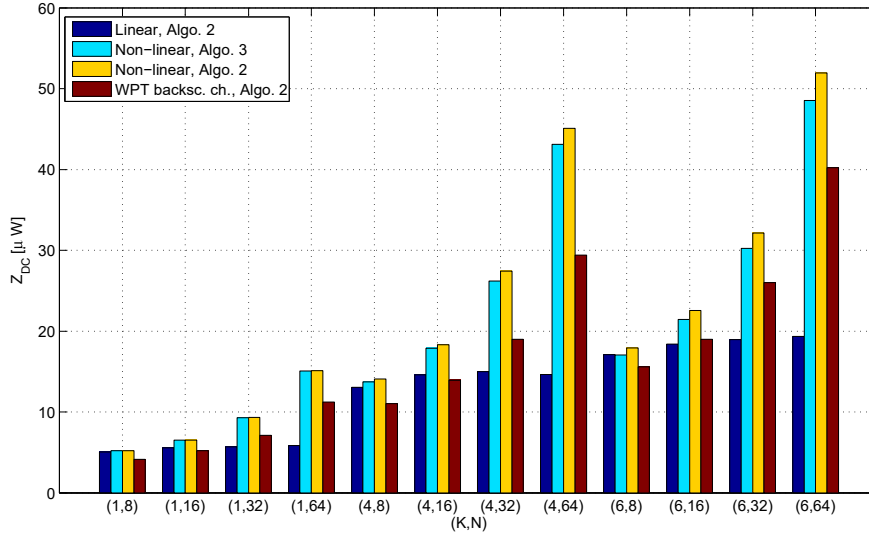


Fig. 5.9 z_{DC} with (K,N) and SINR constraint ($\bar{\rho}$) is set to 3dB.

increases, Z_{DC} decreases at a slower rate for higher SNR. Another observation is that, the trade-off region can be enlarged as the number of sinewaves in the waveform increases. This is because, power allocation across multiple sinewaves can benefit from exploiting a frequency diversity gain and the non-linearity of the rectifier.

We then study the performance of the optimized waveform based on the linear EH model which accounts only for the second order term in the Taylor expansion (i.e., the second order term in Eq. (2.21)). By taking only the second order term in Eq. (2.21), waveform optimization based on the linear EH model can be solved by using Algorithm 11. It can be drawn from Figure 5.6 and Figure 5.8 that for small N (e.g., $N = 4$), the linear model-based design benefits from channel frequency selectivity to get close performance to the non-linear model-based design. As N increases, the waveform design based on the non-linear model explicitly outperforms the linear model. The performance gap of the linear model-based design compared to the non-linear model-based design increases as N increases. Another observation is that as N keeps increasing ($N \geq 16$), the trade-off region for the linear model does not change much. This is because the linear model provides only a logarithmic increase of Z_{DC} with N [18]. These observations show the importance of accounting for the non-linearity of the rectifier to increase the efficiency of waveform design.

The performance of the waveform optimized based on the simplified algorithm as described in Algorithm 12 for $K = 3$ is illustrated in Figure 5.8. It is observed that $\bar{\rho}$ - Z_{DC} region offered by Algorithm 12 is slightly smaller than Algorithm 11, with a lower computational complexity. Table 5.2 compares the average computational complexity for waveform

Table 5.2 Elapsed running time: Algorithm 11 vs. Algorithm 12

Algorithm	(K,N)= (4,8)		(K,N)= (6,64)	
	Average $Z_{DC}(\mu A)$	Average elapsed time(s)	Average $Z_{DC}(\mu A)$	Average elapsed time(s)
Algorithm 2	14.073	12.277	51.979	263.967
Algorithm 3	13.729	7.4858	48.571	184.448

optimization based on Algorithm 11 and Algorithm 12. Z_{DC} and the elapsed time are averaged over several channel realizations with (K,N) is set to $(4,8)$ and $(6,64)$. To draw the comparison, we employ the same initial point and $\epsilon = 10^{-7}$ for both algorithms and the SINR constraint at each tag $\bar{\rho}$ is set to $3dB$. The simulation is conducted by using the same MATLAB version and computer as in Table 5.1. The simulation results show that the average elapsed time for Algorithm 12 is less than Algorithm 11, with the Z_{DC} achieved from Algorithm 12 is slightly lower than Algorithm 11.

Figure 5.9 shows the performance of average Z_{DC} under different N and K . SINR constraint at the reader for each tag is fixed to $3dB$. The observations from Figure 5.9 are as follows. *First*, waveform design based on the non-linear model has been found to outperform that based on the linear model for all configurations. *Second*, the waveform optimized by using Algorithm 12 has close performance to those obtained by using Algorithm 11. *Third*, waveform design benefits from a multiuser diversity gain to exhibit a significant increment of Z_{DC} with the number of users. Nevertheless, the average amount of energy harvested at each user decreases as the number of user increases. This is because, the waveform has to be designed to meet all SINR requirements, and therefore decreases the amount of energy harvested at each user. It is also observed that for given N , the performance gap between waveform optimized based on the non-linear model and the linear model increases as K increases. *Fourth*, waveform design benefits from a frequency diversity gain to exhibit a larger Z_{DC} with N . It can be observed that the inefficiency of the linear model is more severe as N increases irrespective of K . *Fifth*, we investigate the performance of Z_{DC} when the waveform is optimized by using backscatter channel to maximize the energy harvested as in [67]. Since CSI of forward channel is difficult to attain in a backscatter system, the transmit signal in [67] is optimized based on the backscatter channel to maximize the amount of energy harvested at the users. Here, (P 5.7) is optimized by substituting forward channel $h_{j,n}$ with backscatter channel $h_{j,n}h_{j,n}^b$ in Eq. (5.23). It is observed that waveform optimized by using backscatter channel to maximize the energy harvested leads to a smaller Z_{DC} than those optimized by using forward channel, the difference is severe as N increases.

5.5 Summary

This chapter studied waveform design in the wireless powered backscatter system. The waveform was designed adaptive to the perfect CSI knowledge of forward and backward links available at the reader. A simple binary modulation was considered for the backscattering communication. The waveform design was based on non-linear EH model introduced in [17, 18] with the 4-th order truncation in Eq. (2.21). It has been mentioned in [22] that the 4-th order truncation is the minimum order to sufficiently demonstrate the non-linearity of the rectifier process. Comparison was also made with a linear EH model, that only accounts for the second order term in Eq. (2.21).

The trade-off between harvested energy and the reliability of backscatter communication was investigated. The multisine transmit waveform was optimized so as to maximize the harvested energy subject to the transmit power and SINR constraint at all tags. First, a point-to-point wireless powered backscatter system was discussed. The optimization problem maximizing the harvested energy subject to the transmit power and SNR constraint is formulated. An optimal phase of the waveform in the closed form solution and the receive combiner were obtained. Accordingly, the amplitude of the waveform was optimized by using a Reverse GP. This optimization method (Reverse GP) can be generalized to solve any order of non-linear EH model in Eq. (2.22).

Then, a multi-user wireless powered backscatter was considered. Unfortunately, for a multi-user backscatter, the optimal phase of the waveform cannot be obtained in a closed form solution. Therefore, the harvested energy was then formulated in a matrix form. Accordingly, an iterative algorithm has been proposed to optimize the transmit waveform and receive combiner iteratively. However, this optimization method by using the matrix formulation cannot be generalized to a higher order than 4 of Eq. (2.22). Comparison was made between these two optimization methods (Reverse GP and matrix formulation) for a point-to-point backscatter and has been shown to have a similar performance.

Comparison with the linear EH model that has been widely used in the SWIPT literature has also been considered. The waveform based on linear and non-linear EH model were optimized in Section (5.3.2). Assuming all backscatters have the same SINR constraint $\bar{\rho}_j = \bar{\rho}$, the numerical simulation results were obtained. The performance of z_{DC} with different number of subcarrier N and user K were also identified.

The numerical simulation shows that the performance of wireless powered backscatter was subject to a trade-off between harvested energy and SINR. Another observation was that the region of non-linear model always outperform the linear model. The inefficiency of the linear model is more severe compared to the non-linear model as N increases. This observation highlights the importance of accounting for the non-linearity of the rectifier in

waveform design for wireless powered backscatter system. Comparison has been made with the waveform optimization that is based on the backscatter channel $\mathbf{h}_j \cdot \mathbf{h}_j^b$ as in [14, 61]. The result has shown that waveform optimization based on backscatter channel has a lower z_{DC} compared to the waveform optimization obtained based on forward channel. The inaccuracy of waveform optimization based on the backscatter channel is getting worse as N increases.

It is interesting to extend the works in this chapter for other system configurations such as K-user backscatters and multiple antennas. The waveform design for limited CSI knowledge at the reader is also very interesting to be studied.

Chapter 6

Conclusion and Further Work

The main objectives of this research are to design and optimize wireless information and power transfer networks with multisine waveform excitation. The analysis has been done and the main findings are summarised in this chapter.

6.1 Conclusion

6.1.1 Transmission strategy for SWIPT in a SISO K-user IFC

In chapter 3, the trade-off between achievable rate and harvested energy in SWIPT for SISO K-user OFDM IFC has been studied. The time switching receiver is considered where the receiver switch its operation mode between information decoding (ID) and energy harvesting (EH) over time.

First, it was assumed that the transmitters are non-cooperative and have perfect knowledge of the local CSI of the link between itself and all receivers. Therefore, the transmitters optimize their transmission strategy in a distributed manner. It has been found that the necessary condition for the optimal transmission strategy at high SNR is for each energy transmitter to transmit its signal by allocating its transmit power on a single subcarrier. Accordingly, one-subcarrier selection methods named maximum channel gain (MaxCG), minimum channel gain (MinCG), Signal to Leakage and Energy Ratio (SLER) and Signal to Leakage Ratio with Energy Constraint (SLREC) have been proposed. The optimization problem was then formulated as maximizing the achievable sum rate for information transfer subject to the transmit power and total harvested energy at the EH receivers. An iterative algorithm has then been proposed to optimize the power allocation at the EH transmitters and ID transmitters. From the simulation, it was observed that SLREC exhibits higher rate-energy

performance compared to the other subcarrier selection strategies (MaxCG, MinCG and SLER).

Another observation was that, depending on the *water-level*, information transmitters do not allocate their transmit power to the subcarrier with poor channel gain to noise plus interference ratio. If the subcarrier chosen at the energy transmitters is the subcarrier unutilized at the information transmitters, the transmit signal from the energy transmitters do not degrade the achievable sum-rate performance. Motivated by this observation, with some cooperation among the information transmitters to transfer a low rate feedback of the unutilized subcarrier indices to the energy transmitters, the achievable rate-energy region can be enlarged. Accordingly, a single subcarrier transmission strategy and two-subcarriers transmission strategy (one of them is the unutilized subcarrier at the ID users) was proposed. The simulation result shows that with a low-rate feedback of the unutilized subcarrier indices from the information transmitters, the rate-energy region was further enlarged. Finally, with full cooperation among the transmitters, the power allocation maximizing the trade-off between DoF at the information users and harvested energy has been proposed.

6.1.2 Coordinated Waveform Design for Wireless Information and Power Transfer

In Chapter 4, the coordinated waveform design for SWIPT with one-ID user and one-EH user has been studied. The time switching receiver is considered where the receiver switches its operation mode between ID and EH over time. It was assumed that the transmitters have perfect knowledge of global CSI of the link between all transmitters and receivers. The waveform design for perfect and no cancellation of the multisine Wireless Power Transfer (WPT) signal at the ID receiver was covered.

The problem of waveform design based on non-linear EH model that accounts for the non-linearity of the rectifier introduced in [18] was investigated. The optimization problem was formulated as maximizing the total harvested energy subject to the transmit power and achievable rate constraint. Comparison with the waveform design obtained from the linear EH model was also made. First, we consider a system with the energy and information transmitters respectively transmit deterministic and modulated waveform named NIPT as shown in Figure 4.1. The numerical simulation has shown that deterministic waveform can enlarge the trade-off region over the modulated waveform. This observation motivates the new design of the superposition of modulated and deterministic waveform at the information transmitter named SIPT as shown in Figure 4.2. This waveform design has a non-zero mean Gaussian distribution [19]. The numerical result has shown that SIPT provides a larger

trade-off region compared to NIPT. This observation implies that the transmit waveform at the information transmitter should be designed differently from the capacity achieving in the conventional WIT that has a zero mean Gaussian distribution. Moreover, the numerical simulation has shown that accounting for the non-linearity of the rectifier in the coordinated waveform design increase the resource efficiency that lead to a larger trade-off region than waveform design obtained by ignoring the non-linearity of the rectifier.

6.1.3 Waveform Design in Backscattering System

In chapter 5, waveform design in the wireless powered backscatter system has been studied. It was assumed that the reader has perfect knowledge of forward and backward links. The waveform design was based on non-linear EH model introduced in [17, 18] with the 4-th order truncation in Eq. (2.21). It has been stated in [22] that the 4-th order truncation is the minimum order to sufficiently demonstrate the non-linearity of the rectifier process. Comparison was also made with a linear EH model, that only accounts for the second order term in Eq. (2.21).

The trade-off between harvested energy and the reliability of backscatter communication has been investigated. The multisine transmit waveform was optimized so as to maximize the harvested energy subject to the transmit power and SINR constraint at all tags. First, a point-to-point backscatter system was considered. An optimal closed form solution for the phase of the waveform and receive combiner were obtained. Accordingly, the amplitude of the waveform was optimized by using a Reverse GP. This optimization method (Reverse GP) can be generalized to solve any order of non-linear EH model in Eq. (2.21). Unfortunately, for a multi-user backscatter, the optimal phase of the waveform cannot be obtained in a closed form solution. Therefore, the harvested energy was then formulated in a matrix form. Accordingly, an iterative algorithm has been proposed to optimize the transmit waveform and receive combiner. However, this optimization method (matrix formulation) cannot be generalized to a higher order than 4 in Eq. (2.21). Comparison was made between these two optimization methods (Reverse GP and matrix formulation) for a point-to-point backscatter and has been shown to have a similar performance. From the numerical simulation, it was found that the performance of wireless powered backscatter was subject to a trade-off between harvested energy and SINR. The numerical result has shown that accounting for the non-linearity of the rectifier in the wireless powered backscatter can increase the resource efficiency and lead to a larger trade-off region than the waveform design obtained by ignoring the non-linearity of the rectifier.

6.2 Further work

The following are the future works included in this research.

6.2.1 Backscattering system with the limitation of CSI knowledge

In WIPT, in order to maximize the trade-off between the RF-DC conversion efficiency and the reliability of communication, the amplitude and the phase of the waveform are optimized by adapting to the CSI. In practice, CSI acquisition especially in the power limited devices would be difficult and require additional overhead. The waveform design for limited CSI knowledge at the transmitter is a problem of interest.

6.2.2 Waveform design validation

It would be beneficial to validate the performance of waveform design in the coordinated waveform for WPT and WIT in Chapter 4 and the wireless powered backscatter in Chapter 5. It can be done with circuit analysis using PSpice and prototyping measurement.

6.2.3 The non-linearity of the rectifier

The non-linearity of the diode has been overlooked in the SWIPT literature. Motivated by the promising result of the waveform design accounting for the non-linearity of the diode in [19], Chapter 4 and Chapter 5, a future works consist in considering the WIPT architecture and waveform design for different system setup such as multiple antenna, relay and wireless powered beacon. Moreover, the coordinated waveform design in Chapter 4 can also be studied for the general K-user system.

6.2.4 Low complexity waveform optimization

A low complexity algorithm for WIPT where the performance is close to those obtained with the optimal solution is beneficial for practical application. It is an interesting problem to be considered.

6.2.5 High PAPR Issue

The multisine signal with high PAPR increases the RF-DC conversion efficiency. However, high PAPR reduces the efficiency of the power amplifier. The evaluation of the impact of the

PAPR is required to balance the trade-off between RF-DC conversion and power amplifier efficiency.

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Appendix A

Optimization for NIPT with Linear EH

The optimization problem in Eq. (4.18) can be written as

$$\begin{aligned}
 (P \ 4.6) \quad & \max_{\mathbf{p}_1, \mathbf{p}_2} \quad \sum_{n=1}^N (p_{1,n} |h_{11,n}|^2 + p_{2,n} |h_{12,n}|^2) \\
 \text{s. t.} \quad & \sum_{n=1}^N \log_2 \left(1 + \frac{p_{2,n} |h_{22,n}|^2}{\sigma_{2,n}^2} \right) \geq \bar{R} \\
 & \sum_{n=1}^N p_{k,n} \leq 2P, \\
 & p_{k,n} \geq 0 \text{ for } k = 1, 2.
 \end{aligned} \tag{A.1}$$

Since the interference from the EH transmitter is perfectly canceled at the ID, the transmission strategy at the EH transmitter should maximize $\sum_{n=1}^N p_{1,n} |h_{11,n}|^2$ subject to the transmit power constraint. It can be found that EH transmitter should allocate its full transmit power on a single subcarrier corresponding to $\bar{n} = \arg \max_{n=1, \dots, N} \{|h_{11,n}|^2\}$ such that $p_{1,\bar{n}} = 2P$ and $p_{1,n \neq \bar{n}} = 0$. Given $\mathbf{p}_1$, the Lagrangian of optimization problem (A.1) is given by

$$\begin{aligned}
 \mathcal{L}(\mathbf{p}_2, \lambda, \mu) = \\
 \sum_{n=1}^N (p_{2,n} |h_{12,n}|^2) + E_{11} - \mu \left(\sum_{n=1}^N p_{2,n} - 2P \right) + \lambda \left(\sum_{n=1}^N \log_2 \left(1 + \frac{p_{2,n} |h_{22,n}|^2}{\sigma_{2,n}^2} \right) - \bar{R} \right),
 \end{aligned} \tag{A.2}$$

where μ and λ are the Lagrangian multipliers for the transmit power constraint and the rate constraint, respectively and E_{11} is the total harvested energy from the EH transmitter given by $E_{11} = 2P |h_{11,\bar{n}}|^2$. The dual function is then defined as

$$g(\mu, \lambda) = \max_{\mathbf{p}_2 \geq \mathbf{0}} \mathcal{L}(\mathbf{p}_2, \lambda, \mu). \tag{A.3}$$

Algorithm 13 Optimization algorithm for the ID transmitter

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- 1: The maximization of $\mathcal{L}(\mathbf{p}_{2,2}, \lambda, \mu)$ over $\mathbf{p}_{2,2}$ for given λ and μ .
 - 2: The minimization of $g(\mu, \lambda)$ over λ and μ for given $\mathbf{p}_{2,2}$.
-

The dual problem is given by $\min_{\mu, \lambda} g(\mu, \lambda)$. The optimal solution for $\mathbf{p}_{2,2}$, λ and μ can be found by iteratively optimizing the steps in Algorithm 13.

The maximization of $\mathcal{L}(\mathbf{p}_2, \lambda, \mu)$ over $\mathbf{p}_2$ for given λ and μ in step 1 can be obtained by setting $\frac{d\mathcal{L}(\mathbf{p}_2, \lambda, \mu)}{dp_{2,n}} = 0$, such that

$$p_{2,n} = \left(\gamma - \sigma_{2,n}^2 |h_{22,n}|^{-2} \right)^+, \quad (\text{A.4})$$

where $\gamma = \lambda (\ln(2)(\mu - |h_{12,n}|^2))^{-1}$. Here, $\mu > |h_{12,n}|^2$ must hold so that the problem has an bounded optimal value. In step 2, μ and λ can be solved using the subgradient-based method, where the subgradient of $g(\lambda, \mu)$ is given by $(\sum_{n=1}^N \log_2(1 + \frac{p_{2,n} |h_{22,n}|^2}{\sigma_{2,n}^2}) - \bar{R}, 2P - \sum_{n=1}^N p_{2,n})$.

Appendix B

Converges of Algorithm 11

Let represent $\{\gamma, \mathbf{X}, \mathbf{t}_j\} = \Psi(\mathbf{g}_j)$ and $\mathbf{g}_j = \Theta(\mathbf{X})$ as optimization problem (P 5.10) and (P 5.11), respectively. It is noted that the value of γ is monotonically decreasing due to optimality of $\Psi(\cdot)$ and $\Theta(\cdot)$. Theoretically, the eigenvector of $\mathbf{D}$ is not unique such that the minimizers $\mathbf{X}$ and $\mathbf{g}_j$ may not converge. However, similar to [92] and [95], under the assumption that the eigenvector of $\mathbf{D}$ can be uniquely obtained, the sequence of minimizer $\{\gamma^{(l)}, \mathbf{t}_j^{(l)}, \mathbf{X}^{(l)}, \mathbf{g}_j^{(l)}\}_{l=0}^{\infty}$ converges to a limit point $\{\hat{\gamma}, \hat{\mathbf{t}}_j, \hat{\mathbf{X}}, \hat{\mathbf{g}}_j\}$. Alternatively, the stopping criterion of Algorithm 12 can also be related to the convergence of the minimizers such as $\|\mathbf{X}^{(l)} - \mathbf{X}^{(l-1)}\|_F / \|\mathbf{X}^{(l)}\|_F \leq \varepsilon$.

Note that problem (P 5.9) is similar to problem (P 5.11) except that $\mathbf{g}_j$ which is found in constraint (5.26c) and (5.26h) is optimized in problem (P 5.10). For brevity, here we discuss the KKT conditions associated with constraint (5.26c) and (5.26h). The remaining KKT conditions of (P 5.9) is not discuss since it can be easily found that the the remaining KKT conditions for problem (P 5.9) are also the KKT conditions for problem (P 5.11).

Let $\{\gamma, \mathbf{t}_j, \mathbf{X}, \mathbf{g}_j\}$ and Lagrangian multipliers $\{\mu, \nu_j, \xi_{j,k}, \eta_{j,k}, \varpi, \kappa_j\}$ associated with constraints $\{(5.28b), (5.26c), (5.26d), (5.26e), (5.26f), (5.26h)\}$ satisfy the KKT conditions of problem (P 5.9). Let $f'_{j,k}(\mathbf{X}, t_{j,k}) = \text{Tr}\{\mathbf{M}_{j,k}\mathbf{X}\} - t_{j,k}$, $f''_{j,k}(\mathbf{X}, t_{j,k}) = \text{Tr}\{\mathbf{M}_{j,k}^H\mathbf{X}\} - t_{j,k}^*$ and $f'''(\mathbf{X}) = \text{Tr}\{\mathbf{X}\} - 2P$. From the Lagrangian of problem (P 5.9), the KKT conditions

associated with constraint (5.26c) and (5.26h) are as follows

$$\begin{aligned} & \sum_{j=1}^K v_j \nabla_{\mathbf{x}} \rho_j(\mathbf{X}, \mathbf{g}_j) + \sum_{j=1}^K \sum_{k=0}^{N-1} \xi_{j,k} \nabla_{\mathbf{x}} f'_{j,k}(\mathbf{X}, t_{j,k}) + \\ & \sum_{j=1}^K \sum_{k=1}^{N-1} \eta_{j,k} \nabla_{\mathbf{x}} f''_{j,k}(\mathbf{X}, t_{j,k}) + \varpi \nabla_{\mathbf{x}} f'''(\mathbf{X}) = 0, \end{aligned} \quad (\text{B.1a})$$

$$v_j \nabla_{\mathbf{g}_j} \rho_j(\mathbf{X}, \mathbf{g}_j) = 0, \quad (\text{B.1b})$$

$$0 \leq v_j \perp \rho_j(\mathbf{X}, \mathbf{g}_j) - \bar{\rho}_j \geq 0, \quad \forall j \quad (\text{B.1c})$$

$$\|\mathbf{g}_j\| = 1. \quad \forall j \quad (\text{B.1d})$$

Based on problem (P 5.11), $\{\hat{\gamma}, \hat{\mathbf{t}}_j, \hat{\mathbf{X}}\}$ with Lagrangian multipliers $\{(\hat{\mu}, \hat{v}_j, \hat{\xi}_{j,k}, \hat{\eta}_{j,m}, \hat{\varpi})\}$ associated with constraints $\{(5.30b), (5.30c), (5.30d), (5.30e), (5.30f)\}$ must satisfy the KKT conditions of problem (P 5.11). The KKT conditions of problem (P 5.11) associated with constraint (5.26c) and (5.26h) are

$$\begin{aligned} & \sum_{j=1}^K \hat{v}_j \nabla_{\mathbf{x}} \rho_j(\hat{\mathbf{X}}, \hat{\mathbf{g}}_j) + \sum_{j=1}^K \sum_{k=0}^{N-1} \hat{\xi}_{j,k} \nabla_{\mathbf{x}} f'_{j,k}(\hat{\mathbf{X}}, \hat{t}_{j,k}) + \\ & \sum_{j=1}^K \sum_{k=1}^{N-1} \hat{\eta}_{j,k} \nabla_{\mathbf{x}} f''_{j,k}(\hat{\mathbf{X}}, \hat{t}_{j,k}) + \hat{\varpi} \nabla_{\mathbf{x}} f'''(\hat{\mathbf{X}}) = 0, \end{aligned} \quad (\text{B.2a})$$

$$0 \leq \hat{v}_j \perp \rho_j(\hat{\mathbf{X}}, \hat{\mathbf{g}}_j) - \bar{\rho}_j \geq 0. \quad \forall j \quad (\text{B.2b})$$

Based on problem (P 5.10), $\hat{\mathbf{g}}_j$ must satisfy the KKT conditions of problem (P 5.10). The KKT conditions of problem (P 5.10) for $j = 1, \dots, K$ are

$$\nabla_{\mathbf{g}_j} \rho_j(\hat{\mathbf{X}}, \hat{\mathbf{g}}_j) = 0, \quad (\text{B.3a})$$

$$\|\hat{\mathbf{g}}_j\| = 1. \quad (\text{B.3b})$$

Multiplying (B.3a) by $\hat{v}_j$, the KKT conditions in (B.3) are given as

$$\hat{v}_j \nabla_{\mathbf{g}_j} \rho_j(\hat{\mathbf{X}}, \hat{\mathbf{g}}_j) = 0, \quad (\text{B.4a})$$

$$\|\hat{\mathbf{g}}_j\| = 1. \quad (\text{B.4b})$$

By combining the KKT conditions of (B.2) and (B.4), we can obtain the KKT conditions as in (B.1). Hence, $\{\hat{\gamma}, \hat{\mathbf{t}}_j, \hat{\mathbf{X}}, \hat{\mathbf{g}}_j\}$ is the KKT point of problem (P 5.9).