Enable In-Band Full-Duplex Communication: a) Backscatter Modulation, b) Self-Interference Cancellation

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THESIS
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To my mother, father and Farhad.
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I was fortunate enough to work under Prof. Besma Smida supervision as my PhD. adviser. She supported me through all these past three years and I can not thank her enough. Her supports for me are countless and I thank her for all she has done for me, both in my academic and personal life. She is one of the smartest people in the field who taught me how to think out of the box. I am glad to have a teacher with a kind heart as my mentor and my friend during my academic years and so on. She made it one of the most pleasant period of my life, to be remembered always.

I would also sincerely thank Prof. Danilo Erricolo, for all his supports and kind advises. He has been supporting me always and my research outcomes would not be achievable without his assist.

I would like to thank all the committee members, for their comments and feedbacks on my dissertation, from the proposal through the defense. I appreciate their time.

Last but not least, I thank my lab mate, Dr. Farhad Farzami who is my research partner and co-author of my research publications. We worked as a team for four years of research and now I can claim that we were the best two-person team ever who worked very well together. I am blessed that I have him as my beloved husband.

I thank all the staff members of the ECE department for their help to sort out all the academic and financial matters during my studies at UIC.

Finally, thanks to all the friends who made my time in Chicago not just productive but truly enjoyable.

SK
PREFACE

This dissertation is an original intellectual product of the author, S. Khaledian. All of the work presented here was conducted in the Networks Information Communications and Engineering Systems Laboratory (NICEST) and Andrew Lab at the University of Illinois at Chicago. The main project has been partially supported by National Science Foundation (NSF) CAREER award 1620902. The results of these works have previously appeared (or is appearing) as several article in IEEE Transactions on Communications, IEEE Transactions on Wireless Communications, IEEE Transactions on Antennas and Propagation, IEEE Transactions on Microwave Theory and Techniques, IEEE Antennas and Wireless Propagation Letters and IEEE Microwave, Wireless Components Letters, and several conference publications: USNC-URSI’15, IGARSS’15, USNC-URSI’16, ISIVC’16, USNC-URSI’17, IEEE APS-URSI’17, URSI-GASS’17, WAMICON’18, USNC-URSI ,ICRA’18. The copyright permissions for reusing the published materials have been presented in Appendix A.

Seiran Khaledian
March 14, 2019
CONTRIBUTION OF AUTHORS

In chapter 2 I introduce a novel configuration to enable in-band full-duplex communication systems. This configuration uses the concept of backscatter modulation in in-band full-duplex communication, called ReflectFX, to prevent strong self-interference signal caused by transmitter imposed on the receiver. Prof. Smida was the leader of this project who proposed this concept for the first time. I was responsible for deriving the throughput and ergodic capacity of ReflectFX and compare it with conventional in-band full-duplex systems. I proposed negative resistance configuration to be used in ReflectFX in order to amplify the backscatter signal and improve the backscatter range. The results of this work has been published in IEEE Transactions on Communications (1).

In Chapter 3 I introduce two configurations named parallel and series, as tools to build ReflectFX. I analytically discussed two parallel and series ReflectFX configuration and derived the down-link and up-link performance of each configuration. I was responsible for major idea of this project including proposing each configuration structure and the analytical derivations and visual illustrations of the each configuration performance. Dr. Hamza Soury helped with bit error rate derivation and writing the paper. Dr. Farhad Farzami and Prof. Erricolo helped with writing the paper. Prof. Smida was the leader of this project. The results of this chapter are published IEEE Transactions on Wireless Communications (2).

In Chapter 4 and 5 I designed, simulated and built negative resistance and bidirectional amplifier to build parallel and series configurations, respectively. I proposed one techniques to build negative resistance using tunnel diode. The results of tunnel diode based negative resistance has been published in 2017 IEEE International Symposium on Antennas and Propagation, USNC/URSI National Radio Sci-
CONTRIBUTION OF AUTHORS (Continued)

ence Meeting. In this project, I was responsible for designing, simulations and measurements. Dr. Farhad Farzami was responsible of fabricating the prototype and helped with writing the paper. Prof. Smida and Prof. Erricolo lead this project and help with writing the paper.

In chapter 5 I proposed a novel bidirectional amplifier structure to realize series ReflectFX configuration. By using two identical reflection amplifiers (built using negative resistors) and one branch line coupler, I designed a bidirectional amplifier. I was responsible for the major design, simulations and measurement of the proposed bidirectional amplifier. Dr. Farhad Farzami was responsible for fabrication and also helping in simulation and writing the paper. Prof. Erricolo helped with writing the paper and Prof. Smida leaded this project and also help with the manuscript. The results of this chapter is published in IEEE Microwave and Wireless Components Letters, 4.

In chapter 6 I implemented a two-way backscatter circuit as one example of ReflectFX. In this circuit I used On/Off key (OOK) backscatter modulation. I used transistor as negative resistance to have backscatter amplification. In forward path, I used a microstrip coupled line to couple a portion of received signal to the tag receiver. I was responsible for part of the design and Dr. Farhad Farzami helped with design and fabrication. Prof. Smida and Prof. Erricolo leaded this project. The results of this chapter is submitted in IEEE Microwave and Wireless Components Letters, 5.

In part II, chapter 7 I proposed a new method to cancel the self-interference signal in a single antenna full-duplex configuration at analog stage. I used an RF circulator to separate transmitted and received signals. Instead of estimating the self interference signals and subtracting them from the received signals, I used the inherent secondary self interference signals at the circulator, reflected by the
antenna, to cancel the primary SI signals leaked from the transmitter port to the receiver port. I was responsible for the main designing, simulating and measuring. Dr. Farhad Farzami was responsible for fabrication and some parts of simulations. Prof. Smida and Prof. Erricolo leaded this project and help with writing the paper. The results of this work is published in IEEE Transactions on Microwave Theory and Techniques and IEEE 19th Wireless and Microwave Technology Conference (WAMICON). \cite{6,7}.

In chapter \cite{8} I implemented self-interference cancellation at antenna stage. I proposed a robust passive SIC technique for IBFD system using one Tx and two Rx double stacked rectangular patch antennas with orthogonal linear polarizations. The Tx antenna is placed between two Rx antennas. The Rx antennas are rotated 90 degree to have orthogonal polarization compared to the Tx antenna. Thus, the dominant SI signal in this configuration is the Tx antenna x-pol electric field components which are induced on the Rx antennas. The proposed technique leverages the x-pol electric field component phase distribution characteristic of rectangular microstrip antennas. I was responsible for the main designing, simulating and measuring. Dr. Farhad Farzami was responsible for fabrication and some parts of simulations. Prof. Smida and Prof. Erricolo leaded this project and help with writing the paper. The results of this work is published in IEEE Transactions on Antennas and Propagation. \cite{8}. 
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<tr>
<td>AXR</td>
<td>Axial Ratio</td>
</tr>
<tr>
<td>BLC</td>
<td>Branch Line Coupler</td>
</tr>
<tr>
<td>BSC</td>
<td>Back-scatter Communication</td>
</tr>
<tr>
<td>DUT</td>
<td>Device Under Test</td>
</tr>
<tr>
<td>ER</td>
<td>Embedded Resonator</td>
</tr>
<tr>
<td>FBR</td>
<td>Front-to-Back Ratio</td>
</tr>
<tr>
<td>FET</td>
<td>Field Effect Transistor</td>
</tr>
<tr>
<td>IBFD</td>
<td>In-Band Full-Duplex</td>
</tr>
<tr>
<td>IC</td>
<td>Integrated Circuit</td>
</tr>
<tr>
<td>LHCP</td>
<td>Left-Handed Circular Polarization</td>
</tr>
<tr>
<td>NDR</td>
<td>Negative Differential Resistance</td>
</tr>
<tr>
<td>PEC</td>
<td>Perfect Electric Conductor</td>
</tr>
<tr>
<td>PMC</td>
<td>Perfect Magnetic Conductor</td>
</tr>
<tr>
<td>RCS</td>
<td>Radar Cross-Section</td>
</tr>
<tr>
<td>RDA</td>
<td>Retrodirective Antenna</td>
</tr>
<tr>
<td>RFID</td>
<td>Radio Frequency Identification</td>
</tr>
<tr>
<td>RHCP</td>
<td>Right-Handed Circular Polarization</td>
</tr>
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<td>Abbreviation</td>
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<tr>
<td>RWG</td>
<td>Rectangular Waveguide</td>
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<tr>
<td>Rx</td>
<td>Receiver</td>
</tr>
<tr>
<td>SIC</td>
<td>Self-Interference Cancellation</td>
</tr>
<tr>
<td>SIS</td>
<td>Self-Interference Signal</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal to Noise Ratio</td>
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<tr>
<td>Tx</td>
<td>Transmitter</td>
</tr>
<tr>
<td>TEM</td>
<td>Transverse Electromagnetic</td>
</tr>
<tr>
<td>UIC</td>
<td>University of Illinois at Chicago</td>
</tr>
<tr>
<td>VLER</td>
<td>Varactor Loaded ER</td>
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<tr>
<td>VLESRR</td>
<td>Varactor Loaded Embedded Split Ring</td>
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<tr>
<td>VSWR</td>
<td>Voltage Standing Wave Ratio</td>
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SUMMARY

One key limit to spectrum utilization efficiency involves the current practice of half-duplex communication, in which a node either transmits or receives a signal in a single channel usage. Frequency Division Duplex (FDD) and Time Division Duplex (TDD) are the two most commonly used techniques. Contrary to FDD and TDD, in in-band full-duplex (IBFD) communication systems, we can transmit and receive at the same frequency bandwidth, simultaneously. IBFD communication systems are one of the most promising ideas for the next-generations of wireless networks. IBFD communications have the potential of doubling the spectrum efficiency of communication systems as well as eliminating hidden terminals and the need of duplex filters, improving fairness, and network latency. However, the major challenge of IBFD systems is reducing self interference signals, that are imposed on the received (Rx) signals by the transmitted (Tx) signals. In this matter, I proposed a new IBFD at end-user, which use backscatter modulation concept to prevents from having any self-interference signals, called ReflectFX. ReflectFX offers a new concept for two-way wireless communication: rather than avoiding self interference as in half-duplex, or combating self interference as in conventional IBFD, nodes will re-use the received radio-carrier waves to transfer information. The electromagnetic waves are modulated and reflected by the same antenna that receives them. Since the backscattered electromagnetic waves experience the free path loss twice, the communication range will be limited to short rang applications. However, the transmission range of ReflectFX can be improved by adding negative resistance to the end-user load. By deriving an expression for the overall achievable throughput and ergodic capacity of ReflectFX, it can be shown that ReflectFX outperforms both conventional full-duplex and half-duplex.
In order to transform ReflectFX from a concept to a practical wireless communication system, hardware tools are needed. Thereby, I proposed two configurations which can implement ReflectFX, called parallel and series configurations. In parallel configuration, my tool is a reflection amplifier. Reflection amplifier can be built using negative resistance. Negative resistance can be implemented by different methods, in which using transistors and tunnel diodes are among the most well-known techniques. Tunnel diode based negative resistance is discussed, designed and fabricated. In order to implement series configuration, I designed a full-duplex bidirectional amplifier. Full-duplex bidirectional amplifiers, amplify the incoming signal in both directions simultaneously. They can be realized by using two identical one-port reflection amplifiers and a branch line coupler (BLC). Here, I proposed a configuration using tunnel diodes based reflection amplifiers, which have been integrated with a BLC to form the bidirectional amplifier. The proposed bidirectional amplifier structure has been analyzed theoretically, simulated and fabricated. For both parallel and series configurations, I used active loads in the end-user modulators to maximize the backscatter communication range and implement the desired backscatter constellation, subject to no data loss in down-link path. Contrary to most existing backscatter communication studies, I used Thevenin/Norton equivalent circuit, only to calculate the received power at end-user, while I derived the backscatter power using the antenna scatterer theorem. Moreover, a closed-form expression of the average bit error probability (BEP) at reader in Rician fading channel environment for both parallel and series configurations is derived. The last chapter of part I includes a implemented ReflectFX circuit with On/Off key backscatter modulation. In this chapter I proposed using transistor and microstrip coupled line to enable a two-way backscatter tag. I used an RF switch to modulate the backscatter signal and employ transistor as negative resistance to amplify the backscatter
As mentioned earlier, in conventional IBFD systems, self-interference signals are the main barrier to realize a practical wireless communication. This obstacle can be overcome by using self interference cancellation (SIC) techniques to decrease the power of self interference (SI) signals to the level of noise floor at the receiver. SIC could be realized through: i) antenna cancellation or isolation (also called passive interference cancellation technique) by preventing Tx antenna signals to leak to Rx antennas in the first place; ii) analog and digital SIC by subtracting a replica of SI signals at the Rx to cancel existing SI signals at the RF and digital stage. Here, I proposed novel self-interference cancellation techniques at analog stage, to be used in a single antenna IBFD system. I used an RF circulator to separate Tx and Rx signals. Instead of estimating the SI signals and subtracting them from the Rx signals, I used the inherent secondary SI signals at the circulator, reflected by the antenna, to cancel the primary SI signals leaked from the Tx port to the Rx port. I modified the frequency response of the secondary SI signals using a reconfigurable impedance mismatched terminal (IMT) circuit, which consists of two varactor diodes at the antenna port. The frequency response of the IMT circuit can be controlled by the varactor diodes bias voltages. The proposed analog SIC configuration is low-power consumption and low-profile and can be a good candidate for small form-factor wireless systems. By combing this technique with digital SIC technique, a sufficient amount of SIC can be achieved and a practical IBFD communication system can be implemented.

I also proposed a SIC implemented at antenna stage. I proposed a robust passive SIC technique for IBFD system using one Tx and two Rx double stacked rectangular patch antennas with orthogonal linear polarizations. I introduced a novel compact rectangular microstrip antennas configuration in IBFD.
SUMMARY (Continued)

system. This configuration includes one double stacked patch Tx and two double stacked patch Rx antennas. I exploited the polarization diversity technique combined with radiation characteristic of rectangular patch antennas to achieve maximum isolation between Tx and Rx ports. In this system, the SI signals from the Tx antenna to the Rx antennas are added destructively. I also kept to a minimum the distance among the antennas, which corresponds to less than $\frac{\lambda_0}{10}$ at 2.4 GHz. My proposed technique obtained more than 50 dB and 60 dB of SIC over 180 MHz and 120 MHz frequency bandwidth, respectively.

In summary, in this thesis, by proposing: a) new concept of IBFD system by means of backscatter modulation (ReflectFX), which is free of SI signals, and b) novel and efficient analog and antenna SIC techniques for conventional IBFD communication systems, I contributed to move one step closer to realize a practical IBFD system.
CHAPTER 1

INTRODUCTION

Given the ever-increasing demand for high-speed data services, modern wireless networks will require more efficient frequency spectrum usage strategies. Contrary to current practice of half-duplex communication, in which a node either transmits or receives a signal in a single channel usage, in in-band full-duplex (IBFD) communication systems, nodes can send and receive at the same frequency band, simultaneously. IBFD system can double the throughput of channel and directly affects the spectrum efficiency. My objective is to enable IBFD communication system, which has not been employed in the wireless communication systems, to date.

IBFD systems suffer from strong self-interference (SI) signals that transmitter (Tx) imposes on the receiver (Rx), at the same node. Since in a conventional wireless node, Tx signals are way stronger than Rx signals, even Tx leakage can mask the Rx signals completely. Thus, in order to enable IBFD system, we must either i) avoid SI signals or ii) cancel SI signals to the level of noise floor at receiver. In my thesis, I propose two methods to address both i and ii.

1.1 Thesis organization

1.1.1 Part I: ReflectFX

This part introduces ReflectFX, a novel concept of in-band full-duplex communication system by means of backscatter modulation. In this part, I also propose hardware implementation of ReflectFX
and illustrate the simulations and measurements results of the proposed prototypes.

Chapter 2: In this chapter, I proposed a novel concept of IBFD, which use backscatter communication to realize a free SI IBFD system, called ReflectFX. Contrary to conventional IBFD systems, in ReflectFX, nodes (here called end-user) do not generate their own radio-carrier waves, instead use a portion of received signals and backscatter it, to transmit its data back. By choosing from a set of impedance loadings, the end-user can map a sequence of digital symbols onto the backscattered waveforms. I accurately modeled and optimized backscatter modulation (BM) to provide sufficient power for the end-user demodulator while exhibiting maximum difference in the backscattered field. It is shown that ReflectFX can outperform half-duplex and conventional IBFB systems, in term of throughput and ergodic capacity.

Chapter 3: In order to transform ReflectFX from a concept to a practical communication system, hardware tools are needed. As mentioned earlier, ReflectFX uses a portion of received signals in backscatter communication and the rest are delivered to the receiver demodulator. Thus, ReflectFX needs a two-way backscatter modulator hardware at each node. In this matter, I propose two hardware configurations, parallel and series and analytically study them. In parallel configuration, a set of negative resistances are placed in parallel with the receiver demodulator. By choosing from this set of active loads, the backscatter signal is modulated and BM is implemented. Since negative resistors show reflection coefficient bigger than one, they act as reflection amplifier and improve the backscatter range. In series configuration, a bidirectional amplifier is
used to amplify the received and backscatter signal, simultaneously. In series configuration BM is realized by switching among different impedance matching circuits, bidirectional amplifier gain and phase shifter values, all placed in series. In order to have no data loss in down-link path, I set an additional requirement to both configurations design. The received power at the receiver demodulators should be at least equal to $P_{\text{match}}$. Where $P_{\text{match}}$ is the amount of power delivered to a conventional receiver demodulator when its load is conjugate-matched to its antenna impedance. The proposed BM configurations are analytically studied. By optimizing the set of modulator loads, that correspond to BM constellation symbols, I a) maximize the BM communication range, b) implement the desired BM constellation map such as QAM and phase shift keying (PSK) at up-link, and c) guarantee at least $P_{\text{match}}$ at the receiver demodulator. I proposed a new geometric representation for the optimization problem. Moreover, I obtain a closed-form expression of the average bit error probability (BEP) at reader in Rician fading channel environment for both tag configurations. By presenting this analytical study, I showed that the up-link and down-link performances of ReflectFX configurations are not independent of each other and are directly related to the values of BM load values.

Chapter 4: In this project, I designed, simulated and built reflection amplifier (negative resistance) to be used in parallel ReflectFX configuration. A reflection amplifier structure is introduced, using tunnel diode. For tunnel diode, we extracted the high frequency model using through, reflect and line (TRL) calibration method. Then we design matching circuit to obtain desired input impedance (reflection gain) at desired frequency band without any potential oscillation. This
reflection amplifier consumes only 0.2 mW DC power and its reflection gain is 17 dB for the incident signal power of -30 dBm

Chapter 5: In this chapter, a low-power, switchless, full-duplex bidirectional amplifier to be used in series ReflectFX configuration, is proposed. It is composed of two identical reflection amplifiers and one 90°, -3 dB Branch Line Coupler (BLC). The reflection amplifiers are designed using low DC power consumption tunnel diodes introduce in chapter 4. These reflection amplifiers are integrated with a miniaturized BLC to form a bidirectional amplifier that provides 9 dB of measured transmission gain and 22 dB of measured return loss.

Chapter 6: In this chapter, I implemented a ReflectFX circuit with On/Off key backscatter modulation, as an practical example. In this chapter, I proposed using transistor and microstrip coupled line to enable a two-way backscatter tag. I proposed a two-way reflection amplifier in backscatter tag which provides reflection and forward gain simultaneously. I used an RF switch to modulate the backscatter signal in On/Off key scheme and employ transistor as negative resistance to amplify the backscatter signal.

1.1.2 Part II: Self-interference cancellation

Chapter 7: In an conventional IBFD system, each node has its own separate source of transmitter, which imposes strong SI signals on the receiver. In this configuration we can not avoid SI signals, rather we have to cancel them. SIC can be implement in three stages: antenna, analog and digital stage. The amount of SIC value for an efficient IBFD system depends on the Tx signal power and the noise floor at the receiver. The SI signal level should be reduced to the same level as the
receiver noise floor. In this thesis, I propose a new analog SIC technique for IBFD transmission in single-antenna systems. I use an RF circulator to separate Tx and Rx signals. Instead of estimating the SI signals and subtracting them from the Rx signals, I use the inherent secondary SI signals at the circulator, reflected by the antenna, to cancel the primary SI signals leaked from the Tx port to the Rx port. I modified the frequency response of the secondary SI signals using a reconfigurable impedance mismatched terminal (IMT) circuit, which consists of two varactor diodes at the antenna port. This technique can also adjust the frequency band and the bandwidth by controlling the varactor diodes bias voltages. The IMT adjustability makes it robust to antenna input impedance variations and fabrication errors. Using this novel analog SIC technique, more than 40 dB cancellation over 65 MHz of bandwidth is achieved. This technique is independent of the RF circulator and antenna type and it can be applied to any frequency band. It is also very relevant to small mobile devices because it provides a simple and low-power and low-cost adjustable analog SIC technique.

**Chapter 8:** SIC can also be implemented at antenna stage. In this chapter, I proposed a new method to increase the isolation between Tx and Rx antennas in a IBFD system. This configuration includes one double stacked patch Tx and two double stacked patch Rx antennas. I used the polarization diversity technique. I combined the polarization diversity with radiation characteristic of rectangular patch antennas to achieve maximum isolation between Tx and Rx ports. In this chapter, I explained the detail design of the IBFD structure and the implementation process. I finished the chapter with simulation and measurement results of achieved SIC for proposed configuration.
All the proposed projects in this thesis have been designed, simulated and prototypes are fabricated and measured by myself and another PhD student Farhad Farzami under supervision of Prof. Besma Smiad and Prof. Danilo Erricolo in the EM Andrew lab at the University of Illinois at Chicago (UIC).
Part I
ReflectFX
CHAPTER 2

IN-BAND FULL-DUPLEX WIRELESS COMMUNICATION BY MEANS OF BACKSCATTER MODULATION: REFLECTFX

Parts of this chapter have been presented in [7]. Copyright © 2017, IEEE.

Given the ever-increasing demand for high-speed data services, modern wireless networks will require more efficient frequency spectrum usage strategies. Our objective is to create spectrum-efficient two-way wireless networks by means of reflected power. The urgency of this objective follows from three facts:

Communication is increasingly asymmetrical and becoming more so over time. Data traffic has changed greatly since mobile communication first emerged in the 1990s. From the viewpoint of traffic symmetry between uplink (mobile to base-station) and downlink (base-station to mobile), data traffic started as highly symmetrical two-way communication with voice services. Communication networks provide various services besides voice service and have conveyed various traffic such as data and video traffic in addition to voice traffic. Over time, however, communication has become markedly imbalanced, and the major growth of new multimedia services is characterized by asymmetric traffic, i.e., much higher downlink traffic. Overall traffic patterns are also becoming more asymmetrical, three to five times more in the downlink direction, because the largest bandwidth demand currently derives from downstream TV and Internet video signals [9, 10].

The opportune moment for full-duplex communication. One key limit to spectrum utilization effi-
ciency involves the current practice of half-duplex communication, in which a node either transmits or receives a signal in a single channel usage. All currently deployed wireless networks require two separate channels to achieve bi-directionality. Frequency Division Duplex (FDD) and Time Division Duplex (TDD) are the two most commonly used techniques. FDD separates the downlink and uplink transmissions into different, sufficiently separated, frequency bands, while TDD involves downlink and uplink transmissions that take place in different, non-overlapping time slots using the same frequency band. This separation prevents self-interference which has frequently been seen as an insurmountable technical problem. In any case, this common practice results in a relatively inefficient use of the available frequency spectrum. Due to the popularity of small-cell wireless systems, recent works have provided experimental evidence and methodologies for full-duplex communication. In which a node transmits and receives signals at the same time and on the same frequency band. The self-interference can be effectively reduced by: 1) Physically separating the receiving antenna from the transmitting antenna, which decreases the interference power by propagation losses, cross-polarization, and/or antenna directionality; 2) Strategically placing the receiving and transmitting antennas, which splits the transmission signal among transmit antennas and strategically places them so that the self-interference signals add destructively at the receiving antennas; and 3) Estimating the self-interference and subtracting it from the received signal. Since the receiver knows the signal that has been transmitted, it can extract its own message after correcting for the channel effect. This can either be performed on analog signals using signal inversion based on a balanced/unbalanced (Balun) transformer or RF chain, and/or on digital signals. However, to date, all full-duplex demonstrations have been predicated on the principle that all nodes
must generate their own radio-carrier waves independently. In practice, this approach results in a high level of self-interference at both nodes, as illustrated in Fig. 1 (a). Contrary to previous full-duplex systems, within the framework of our concept the end-user receives a signal free of self-interference. Only the base-station receiver suffers from self-interference. At the base station we do not have severe space limitation so we can place the transmit and receiver antennas far apart or design them so that they have less coupling with each other.

There remains, therefore, a critical need to design highly reliable, full-duplex communication systems since, in the absence of such systems, the opportunities to meet the increasing demand for high-speed data services will not likely be realized.

**Backscatter modulation (BM) for power-efficient communication.** The history of BM can be traced to Stockman’s foundational paper published in 1948 (31). Stockman proposed a point-to-point communication, with the carrier power generated at the receiving end and the transmitter replaced by a modulated reflector. BM is based on the reflection of electromagnetic waves. BM takes advantage of the presence of reflection coefficient variations at the interface between the transmit antenna and the receiver input load. By varying the load, $Z_1$ or $Z_2$, the reflection coefficient changes as illustrated in Fig. 1 (b). This setup is capable of modulating data upon the original waveform, which can be received and decoded. The most prominent application of this technology is radio frequency identification (RFID). As can be seen in (32; 33; 34; 35; 36; 37; 38; 39), RFID tags are used as keys to open electronic locks, as a means of storing personal information in both credit cards and passports as well as tracking strategies in the shipping industry. BM is also implemented in a number of other areas, such as wireless sensor networks (40; 41; 42; 43; 44). With BM, the power limitations normally associated with tradi-
Figure 1: Illustrations of (a) conventional Full-duplex communication, and (b) Backscatter Modulation (BM).

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Conventional wireless sensor systems can be circumvented \(^{[45][46][47]}\). Experimental evidence shows that BM can reach up to 112 Mbps data-rate and use narrow-band, ultra-wideband or spread-spectrum signals \(^{[48][49][39][50][51][52]}\). This leads to relatively low-complexity, low-power, and low-cost communication terminals.

Despite these benefits, BM has not, to date, been seriously considered in two-way communication. In this respect, BM-based approaches have been prevented by one primary but relevant drawback: they are essentially restricted to short-range communications. This potentially serious challenge has two facets: first, the end-user does not generate its own radio wave; and second, most of the research studies related to BM have added additional power constraints, e.g., many systems rely on the assumption that nodes are passive (battery-less)\(^{[38][37][44][53]}\).
In contrast to the approach of the previously cited authors, we did not use BM to take advantage of its low-power requirements. Instead, we employed BM to allow the transmission of two-way flows of data over the same carrier at the same time. We therefore can significantly increase the range of BM by using negative resistance at the end-user transceiver. Furthermore, since the main objective of this work is to increase the spectrum efficiency, many of the proposed RFID modulations to achieve two-way communication are not relevant to our work. Here, we propose a modulation based on analog network coding that achieves in-band full-duplex, where the two flows of data simultaneously occupy the same bandwidth.

To our best knowledge, we were the first to show that it is beneficial to use BM for in-band full-duplex communication. Recently, the authors of (55) used a similar approach for passive sensing. The differences between our research and prior works are the following: (a) our full-duplex framework is based upon a backscatter modulation approach (ReflectFX); (b) the data can flow in two directions using analog network coding; (c) we accurately modeled and optimized BM to provide sufficient power for the demodulator while exhibiting maximum difference in the backscattered field; (d) we used negative resistances to improve the transmission range; (e) we derived achievable throughput and ergodic capacity expressions using constellation-constrained capacity expression; and (f) we compared the throughput and ergodic capacity of half-duplex, conventional full-duplex, and ReflectFX.

1In most RFID full-duplex systems (34, 35, 39, 48, 54), the transmit and received signals use two time slots or two different frequency bands to avoid self-interference.
2.1 Preliminaries - Antenna scattering

Since antenna scattering is of interest to both the wireless-communication community, which is primarily concerned with antennas, and to the Radar and RFID community – which is primarily concerned with scattering – two different terminologies have arisen from both groups of investigators. These two terminologies make it difficult to correlate results from the two groups. This is evidenced by several articles and comments that have appeared recently concerning the question of how much power is scattered and absorbed by a receiving antenna\cite{56, 57, 58, 59, 60, 61, 62, 63}. We will therefore provide a concise review of the antenna-scattering theory before introducing a general framework that allows transmission of two-way flows of data over the same carrier at the same time based on BM.

2.1.1 The general theory of antenna scattering

An antenna illuminated by an incoming wave scatters electromagnetic (EM) energy as a result of current induced by the incident field\cite{64}. The total amount and the radiation pattern of the backscatter power depend on the geometrical and EM properties of the object (size, shape, constitutive materials, etc.). In particular, the EM field scattered by an antenna is given by a relatively simple expression. For a given antenna in free space, the scattered electric field $E_{\text{scat}}(r, \theta, \phi | Z)$, when the antenna is loaded with impedance $Z$, at a particular point in space $(r, \theta, \phi)$ can be expressed as:

$$E_{\text{scat}}(r, \theta, \phi | Z) = \hat{E}_{\text{scat}}(r, \theta, \phi | Z_{\text{ant}}^*) - \hat{\Gamma}(Z) I(Z_{\text{ant}}^*) E_r(r, \theta, \phi),$$

(2.1)

where $E_{\text{scat}}(r, \theta, \phi | Z_{\text{ant}}^*)$ is the same field when the antenna is conjugate-matched ($Z = Z_{\text{ant}}^*$), $Z_{\text{ant}}$ is the antenna impedance, $\Gamma(Z) = \frac{Z - Z_{\text{ant}}^*}{Z + Z_{\text{ant}}^*}$ is the reflection coefficient, $I(Z_{\text{ant}}^*)$ is the current flowing in the
antenna when conjugate-matched, and $E_r(r, \theta, \phi)$ is the electric field generated by the antenna as a radiator when excited by a unit current source. Eq. (2.1) is an old and fundamental result in antenna scattering\cite{65, 66, 67, 68}. It can be proven by using the compensation and superposition theorems of circuit theory \cite{65}. An examination of Eq. (2.1) shows that the first term, called structure scattering, is independent of the load $Z$. The second term, called the antenna mode component of the scattering field, is load-dependent and vanishes when the antenna is conjugate-matched as the reflection coefficient becomes zero, $\Gamma(Z_{ant}) = 0$. Eq. (2.1) may be rewritten as:

\[
E_{scat}(r, \theta, \phi | Z) = I(Z^*_{ant})E_r(r, \theta, \phi)(A_{st} - \Gamma(Z)),
\]

(2.2)

where $A_{st} = \frac{E_{scat}(r, \theta, \phi | Z_{ant}) \cdot I(Z^*_{ant})}{I(Z^*_{ant})E_r(r, \theta, \phi)}$. Note that $A_{st}$ equals 1 for the special case of minimum scattering antennas (such as thin dipoles), but in general, $A_{st} \neq 1$.

2.1.2 Thevenin equivalent circuit – only for receiving antenna

To evaluate the power delivered to the receiver, we made use of the Thevenin equivalent model, as depicted in Fig. 2. The variable $Z_{mod}$ represents the modulation circuit and is controlled by the end-used micro-controller. The input impedance of all of the end-user’s IC components, except the modulator circuit, is modeled by $Z_{ant}^*$. Contrary to the prevailing wisdom, the Thevenin equivalent circuit can only be used to describe the current at the load and cannot be relied upon for calculating the scattered power \cite{69, 58, 61, 62, 63}. In other words, we cannot rely on this circuit model to calculate the scattered electromagnetic field. Therefore, in contrast to common practice \cite{70, 55}, we used Eq. (2.2) rather than
the Thevenin circuit model to evaluate more accurately the transmitted backscatter power as a function of the load $Z_{\text{mod}}$.

Figure 2: Receiving-antenna equivalent circuit. Copyright © 2017, IEEE.

2.2 Framework for full-duplex communication with backscatter modulation

In this section, we address how the base-station and end-user exchange messages at the same time and over the same frequency band using ReflectFX. For simplicity, we intentionally ignore some details that are not needed to understand the basic conceptual idea proposed in this work, e.g., error correction coding, up conversion, and down conversion. We also assume that the base-station has one transmitting and one receiving antenna, while the end-user has only one antenna.

2.2.1 Channel model for ReflectFX communication

We consider two nodes that intend to exchange data over a wireless channel as depicted in Fig. [3]. This model system, in which an end-user $m$ wishes to exchange messages with a central node, or base station $B$, is a model that captures the behavior of current and future cellular networks. The end-user $m$
has an independent message $w_{m,B}$ destined for base-station $B$, and the base-station has an independent message $w_{B,m}$ destined for end-user $m$. We consider that the base-station and each end-user transmit and receive simultaneously using exactly the same frequency band. We assume that the channel outputs are linear combinations of the transmitted signals scaled by random channel gains plus independent additive white Gaussian noise. The Rayleigh fading channels are named $h_{B \rightarrow m}$ (base-station to end-user), $h_{m \rightarrow B}$ (end-user to base-station), and $h_{B \rightarrow B}$ (base-station transmitter to base-station receiver).

**2.2.2 Forward link – self-interference free**

We will first focus on the forward link: the base-station to end-user link. The base-station transmitter carries out a digital-to-analog conversion (DAC) and sends the resulting symbol sequence to the transmission radio. In Fig. 3 signal $x_B(i,s)$ denotes the $i$-th symbol transmitted from the base-station during
packet $s$. A packet consists of $T$ consecutive transmitted symbols. We define $x_B(i, s) = \sqrt{E(s)}s_B(i, s)$ where $s_B(i, s)$ is the transmitted constellation symbol normalized to unit energy and $E(s)$ denotes the average symbol energy during packet $s$. The received symbol at the end-user ($i$-th symbol of packet $s$) is:

$$y_m(i, s) = \sqrt{\xi_{B\rightarrow m}(Z_{i,s})} h_{B\rightarrow m}(i, s)x_B(i, s) + n_m(i, s)$$

(2.3)

where $\xi_{B\rightarrow m}(Z_{i,s}) = (\frac{\lambda}{4\pi d})^2 G_B G_m p_B m (1 - |\Gamma(Z_{i,s})|^2)$ is the average power attenuation from the base-station transmitter to the end-user receiver\footnote{Note that in conventional communication, the load is matched to the antenna and hence $\Gamma(Z_{\text{ant}}^*) = 0$.}. $\lambda$ is the wavelength, $d$ is the distance between base-station antenna and end-user antenna, $G_B$ and $G_m$ are base-station transmitter and end-user antenna gains, $p_B m$ accounts for a polarization mismatch at the end-user antenna, $n_m(i, s)$ is the AWGN noise at end-user, and $\Gamma(Z_{i,s}) = \frac{Z_{i,s} - Z_{\text{ant}}^*}{Z_{i,s} + Z_{\text{ant}}^*}$. Note that the received symbol $y_m(i, s)$ varies with the receiver load $Z_{i,s}$ at the $i$-th symbol of packet $s$. Since the load is changing with the data, the receiving antenna cannot be conjugate-matched to the load; hence, the received power may be diminished when compared with conventional communication (see Figure 2). Therefore, we used negative resistance in the load to increase the received power. Note that during backscatter modulation, the end-user impedance is intentionally mis-matched, and hence the received power is reduced compared to conventional communication (which reduces the performance of the forward link). Contrary to most of the current full-duplex wireless networks\footnote{14, 15, 13, 25, 21, 71, 72, 20, 12} that are predicated on the principle that the base-station and the end-user must generate their own radio-carrier waves independently and hence both have to deal with very high self-interference, in this work the end-user will receive a self-interference-free signal.
Therefore, the base-station to end-user link is similar to the conventional half-duplex transmission. The end-user can extract its own message using conventional reception radio and analog-to-digital conversion (ADC). This will guarantee a very high data-rate toward the end-user and efficiently accommodate high-downlink traffic.

### 2.2.3 Backscatter Link

When the end-user antenna is illuminated by the EM field corresponding to the physical properties and impedance loading of the end-user antenna, a portion of the impinging EM wave will be backscattered from the end-user. By choosing from a set of impedance loadings, the end-user can map a sequence of digital symbols onto the backscattered waveforms. In this framework, we consider PSK modulation, so the backscattered signal becomes a network-coded combination of the messages sent by the end-user and the base-station.

For instance, if we use M-PSK modulation at both the base-station and the end-user then we set $x_B(i,s) = \sqrt{E}e^{j(l-1)\pi/M}$ and represent the $M$ symbols – at the end-user – on a circle of radius $\sqrt{E}$ centered at $c_{st} \in \mathbb{C}$ as $(A_{st} - \Gamma(Z_{i,s}))=\sqrt{E}e^{j(n-1)\pi/M+j\psi} + c_{st}$, where $l$ and $n \in \{1, \ldots, M\}$ and $M$ is the modulation constellation size. The backscattered signal $x_m(i,s) = x_B(l-t,s)(A_{st} - \Gamma(Z_{i,s})) = \sqrt{EE}e^{j(l-1)\pi/M+j(n-1)\pi/M+j\psi} + c_{st} \sqrt{E}e^{j(l-1)\pi/M}$ becomes an analog network-coded combination of the messages sent by the base-station and the end-user (modulated by $Z_{i,s}$), and $t$ accounts for the propagation delay between the base-station and the end-user. Recently, many researchers [49; 70; 71] have demonstrated the feasibility of building backscatter systems that support higher-order constellations. This opens up the possibility of coded modulation and other advanced modulation and coding combi-
Because of leakage from the base-station transmitting antenna, the signal received at the base-station receiver is given by

\[
y_B(i,s) = \sqrt{\xi_{m\rightarrow B}} h_{B\rightarrow m}(i,s) h_{m\rightarrow B}(i,s) x_B(i-t,s) (A_B - \Gamma(Z_{i,s})) + \sqrt{\xi_{B\rightarrow B}} h_{B\rightarrow B}(i,s) x_B(i,s) + n_B(i,s),
\]

where \( \xi_{m\rightarrow B} = \left( \frac{\lambda}{4\pi d_{B\rightarrow B}} \right)^4 G_{Bt} G_m^2 G_{Br} p_{Br} p_{mB} \) is the average power attenuation for the two-way link, \( G_{Br} \) is the base-station receiver antenna gain, \( p_{mB} \) accounts for polarization mismatch at the base-station receiver antenna, \( \xi_{B\rightarrow B} = \left( \frac{\lambda}{4\pi d_{B\rightarrow B}} \right)^2 G_{Br} G_{Bt} p_{BB} \) is the distance between the transmitter and receiver antennas at the base station, \( p_{BB} \) accounts for the polarization mismatch between base-station antennas, and \( n_B(i,s) \) is the AWGN noise at the base-station. \( h_{B\rightarrow B} \) is the channel between base-station antennas, which can be derived by sending pilot signals at base-station and measuring the channel gain.

At the first stage, since the base-station “knows” the strong self-interference signal it is sending \( x_B(i,s) \), it can extract the backscattered signal after correcting for channel effects. The base station is left with the backscattered signal \( x_m(i,s) \) that is a function of the messages sent by both the end-user and the base-station. Then at the second stage, knowing the message that was initially sent, the base-station can extract its own message after removing the structure-dependent term of \( x_m(i,s) \).

### 2.2.4 Practical consideration for ReflectFX

Based on the proposed ReflectFX framework, the end-user sends a sequence of symbols onto the backscattered waveforms by switching the impedance loads. An efficient fast-switch can be implemented using a switched pulse-injected oscillator. Note that this framework does not require the end-


user and the base-station to transmit at the same data-rate. We can modify the end-user data-rate by reducing the impedance load change (i.e. factor of the symbol rate), changing the end-user modulation constellation size, and/or changing the channel encoder rate. In this section, we assume a perfect symbol-level synchronization. In ReflectFX, the synchronization can be achieved using pilot-aided synchronization and/or global time synchronization. The channel estimation techniques used in half-duplex systems can be also employed in ReflectFX.

The impact of external interference may be significant. On one hand, the impact of the external interference on the ReflectFX downlink is identical to the half-duplex case. This link is more robust to external interference than conventional full-duplex systems. On the other hand, the external interference may have a combined impact on the ReflectFX uplink. The external interference effects on the downlink can propagate to the uplink because of the backscatter modulation. In addition, the impact of external interference on the ReflectFX downlink is identical to the case of half-duplex communication. However, external interference may be backscattered with the desired signal. This may aggravate the impact of external interference on the ReflectFX uplink. It is important to note that conventional full-duplex systems are also vulnerable to external interference for both uplink and downlink.

2.3 ReflectFX Load optimization: Negative resistance

The ReflectFX setting is different from the conventional setting in which a total power is imposed on the two nodes. Indeed, the end-user has no radio power available for transmission besides the received radio power. This guarantees no self-interference at the end-user, but it limits the backscatter link range. In addition, since the load is allowed to change, the receiving antenna is not conjugate-matched to that of the load. The performance of ReflectFX is mainly determined by how we manage the self-interference at
the base-station and how we optimize the power between demodulation and modulation at the end-user. A modulation-impedance design should provide sufficient power for the end-user demodulator while exhibiting maximal difference in the backscattered field, i.e. maximum distance between BM constellation points. To analyze these tradeoffs, we combined scattering theory with the Thevenin equivalent model to express the received as well as the backscattered powers at the end-user as a function of the modulation load \( Z_{\text{mod}} \). Next, we added negative resistance to the end-user load to improve the transmission range. Then, we derived our optimization method which we evaluated numerically. Finally, we proposed a negative-resistance circuit implementation.

### 2.3.1 Received power at the end-user

To evaluate the received power at the end-user, we made use of the Thevenin equivalent model, as depicted in Fig. 2. The variable \( Z_{\text{mod}} \) represents the modulation circuit and is controlled by the end-user micro-controller. The input impedance of all of the end-user’s IC components, except the modulator circuit, is modeled by \( Z_{\text{ant}}^* \). The power delivered to \( Z_{\text{ant}}^* \) is the power available for the demodulation at the end-user. As stated earlier, an optimum modulation-impedance design is needed to establish ReflectFX communication. We assume \( Z_{\text{mod}} = \alpha_n Z_{\text{ant}} \), where \( \alpha_n \in \mathbb{C}, \ n \in \{1,2,\ldots,M\} \) and \( M \) is the modulation constellation size. Under this model, the power available to the end-user demodulation is

\[
P_n = 4P_{\text{match}} \left| \frac{\alpha_n (Z_{\text{ant}}^* + Z_{\text{ant}})}{2Z_{\text{ant}}^* + 2\alpha_n (Z_{\text{ant}}^* + Z_{\text{ant}})} \right|^2. \tag{2.5}
\]

\( P_{\text{match}} \) is the power received at end-user when backscatter communication. The average available power for the downlink is then \( P_{\text{avg}} = \frac{1}{M} \sum_{n=1}^{M} P_n \).
2.3.2 Backscatter transmission

We assume M-PSK modulation at the base-station and at the end-user, so that the backscattered signal – a network-coded combination of the messages sent by the end-user and the base-station – can be decoded by the base-station. In this case, the antenna load $Z_{i,s} = Z_{mod} \parallel Z^*_{ant} = \frac{\alpha_n Z_{ant} Z^*_{ant}}{\alpha_n Z_{ant} + Z^*_{ant}}$ varies with the end-user data. We represent the M-PSK modulation constellation as $M$ symbols on a circle of radius $\sqrt{E'}$ centered at $c_{st} \in \mathbb{C}$, i.e.,

$$A_{st} - \Gamma \left( \frac{\alpha_n Z_{ant} Z^*_{ant}}{\alpha_n Z_{ant} + Z^*_{ant}} \right) := \sqrt{E'} \exp(j\theta_n + j\psi) + c_{st}, \text{ for } n \in \{1, 2, \ldots, M\}$$

(2.6)

where $\Gamma \left( \frac{\alpha_n Z_{ant} Z^*_{ant}}{\alpha_n Z_{ant} + Z^*_{ant}} \right) = \frac{Z^*_{ant}}{\alpha_n (Z_{ant} + Z^*_{ant}) + Z^*_{ant}}$. $\theta_n = (n - 1)\pi/M$ and $\psi$ is a rotation for all symbols. $E'$ represents the energy-gain generated by the antenna mode component of the antenna-scattering field.

We rewrite Eq. (2.6) as follows:

$$\alpha_n = \frac{Z^*_{ant} (1 + A_{st} - \sqrt{E'} \exp(j\theta_n + j\psi) - c_{st})}{(\sqrt{E'} \exp(j\theta_n + j\psi) + c_{st} - A_{st})(Z_{ant} + Z^*_{ant})}.$$

(2.7)

2.3.3 Load optimization – negative resistance

Most of the current RFID systems restrict the modulation impedance $Z_{mod}$ to non-negative resistive i.e., $Re(Z_{mod}) \geq 0$. In this work, we propose the usage of negative resistance to increase the backscatter power. Our approach was predicated on the fact that to increase the reflection coefficient $\Gamma \left( \frac{\alpha_n Z_{ant} Z^*_{ant}}{\alpha_n Z_{ant} + Z^*_{ant}} \right)$, the real part of $\alpha_n (Z_{ant} + Z^*_{ant}) + Z^*_{ant}$ should have a small positive value, while the imaginary part should be kept different from zero in order to reduce any undesired self-oscillations (64). The implementation of negative resistance is discussed in the following section. We propose here an optimum ReflectFX
modulation-impedance design that captures the tradeoffs between received and backscatter power. On one hand, to improve the performance of the uplink, we need to maximize the distance between the constellation points of the backscatter modulation by increasing $E'$ subject to $\text{Re}(Z_{\text{mod}}) = \text{Re}(\alpha_n Z_{\text{ant}}) \geq R_N$, where $R_N$ is the minimum negative resistance we can implement. On the other hand, to guarantee no loss in the downlink data-rate the $P_{\text{avg}}$ has to be at least equal to $P_{\text{match}}$. More specifically, our modulation-load optimization problem seeks to:

Maximize, $E'$

Subject to

$$\forall n \in \{1, 2, \ldots, M\}, \quad \text{Re}\left(\frac{Z_{\text{ant}}Z_{\text{ant}}^*(1 + A_{st} - \sqrt{E'} \exp(j\theta_n + j\psi) - c_{st})}{(\sqrt{E'} \exp(j\theta_n + j\psi) + c_{st} - A_{st})(Z_{\text{ant}} + Z_{\text{ant}}^*)}\right) \geq R_N$$

and

$$\frac{1}{M} \sum_{n=1}^{M} |A_{st} + 1 - \sqrt{E'} \exp(j\theta_n + j\psi) - c_{st}|^2 = |A_{st} + 1 - c_{st}|^2 + E' \geq 1,$$  \hspace{1cm} (2.8)

where we derived Eq. (2.8) by combining Eq. (2.7) and Eq. (2.5) and using the definition of $\theta_n = (n - 1)\pi/M$. We have optimized the modulation-impedance design through an exhaustive search. Based on this interpretation, we prove\(^1\) that for all $M \geq 4$ we can increase the constellation size without reducing either $E'$ or the available power for demodulation at the end-user. Once the optimum $c_{st}$ and $\psi$ are found numerically for QPSK, they can then be used for all $M > 4$.

\(^1\)Note that in the case of conventional half-duplex and full-duplex communication the received power equals $P_{\text{match}}$.

\(^2\)The proof is straightforward from the figure.
2.4 ReflectFX performance analysis

In this section, we derive the ergodic capacity and throughput expressions for ReflectFX. Since ReflectFX utilizes load-change to modulate the signal, the input messages are constrained to a discrete and finite set. Hence, we use the constellation-constrained capacity expression for channel with a finite alphabet input. In addition, we account for the effect of self-interference at the base-station receiver. Our analysis is valid under the following idealized assumptions: 1) An infinite number of packets are available for each node; 2) the ACK/NACK feedback channel is delay-free and error-free; 3) wireless channels are frequency non-selective Rayleigh; 4) channel states remain static within the duration of a packet transmission, but become independent across transmissions (i.e., block fading); and 5) residual self-interference is a Rayleigh random variable.

2.4.1 Signal-to-Interference-plus-Noise Ratio (SINR) – Distribution

2.4.1.1 Forward link

The Signal-to-Noise Ratio (SNR) at the end-user receiver at the $s$-th transmission/packet is:

$$\text{SNR}_m(s) = \frac{E_{\xi_{B \rightarrow m}} |h_{B \rightarrow m}(s)|^2}{N_0}, \quad (2.9)$$

where here $\xi_{B \rightarrow m} = \left( \frac{\lambda}{4\pi d} \right)^2 G_B G_m$ is the power loss from the base-station transmitter to the end-user and $N_0$ is the variance of the Gaussian noise. We consider no polarization mismatch between the base-station transmit antenna and the end-user antenna ($p_{Bm} = 1$). The load-optimization, described in Section 2.3, guarantees that the average received power is at least equal to the $P_{\text{match}}$ – i.e., $(1 - |\Gamma(Z_{i,s})|^2) \geq 1$. For tractability, we assume here that the received power for all the loads to be equal to the minimum
$P_{\text{match}}$. We consider that the time-varying channel gain is a wide-sense-stationary narrow-band complex Gaussian process \( [75] \). Hence the SNR \( m(s) \) has an exponential distribution with mean \( \frac{E_{\xi_{B\rightarrow m}}}{N_0} \), and its Probability Density Function (PDF) and Cumulative Distribution Function (CDF) equal the following:

\[
 f_{\text{SNR}_m}(x) = \frac{N_0}{E_{\xi_{B\rightarrow m}}} \exp\left(-\frac{xN_0}{E_{\xi_{B\rightarrow m}}} \right), \quad \text{and} \quad F_{\text{SNR}_m}(x) = 1 - \exp\left(-\frac{xN_0}{E_{\xi_{B\rightarrow m}}} \right). \tag{2.10}
\]

### 2.4.1.2 Backscatter link

For the backscatter link, we account for the effect of self interference on the base-station receiver. The literature on the wireless interference cancellation is vast. Researchers \([14, 17, 76, 29, 23]\) have studied the impact of different self-interference cancellation mechanisms on the performance of full-duplex wireless communication systems, showing that when different interference-suppression mechanisms (passive, analog, and digital) are suitably combined, the average self-interference cancellation can reach up to 100 dB. The received SINR at the base-station at the \( s \)-th transmission/packet is:

\[
\text{SINR}_B(s) = \frac{E_{\xi_{m\rightarrow B}E'_{opt}|h_{B\rightarrow m}(s)|^2|h_{m\rightarrow B}(s)|^2}{E_{\xi_{B\rightarrow B}|\tilde{h}_{B\rightarrow B}(s)|^2 + N_0}}, \tag{2.11}
\]

here \( \xi_{m\rightarrow B} = (\frac{\lambda}{4\pi d})^4 G_B G_B G_m^2 \) is the power loss from the two-way base-station to the end-user, and end-user to base-station, \( E'_{opt} \) is optimized in Section \([2.3]\). We assume no mismatch polarization between the end-user antenna and the base-station receiver antenna \( (p_{mB} = p_{Bm} = 1) \). We also consider the base-station receiver is conjugate-matched to the antenna. \( \xi_{B\rightarrow B} = (\frac{\lambda}{4\pi d_{B\rightarrow B}})^2 G_BT G_BR \) is the interference power loss from the base-station transmitter to the base-station receiver. The random variable \( \tilde{h}_{B\rightarrow B}(s) \) represents the residual self-interference at the base-station receiver. We assume a completely sym-
metric base-to-mobile and mobile-to-base radio propagation channel. Hence, the time-varying channel
gain on a backscatter modulation link can be modeled by a product of two independent, identically-
distributed, wide-sense-stationary narrow-band complex Gaussian processes (77). The CDF and PDF
of the $SINR_B(s)$ are evaluated as:

$$F_{SINR_B}(x) = \Pr \left( \frac{E \xi_m \rightarrow B |h_{B \rightarrow m}(s)|^2 |h_{m \rightarrow B}(s)|^2}{E \xi_B \rightarrow B |h_{B \rightarrow B}(s)|^2 + N_0} \leq x \right),$$

$$f_{SINR_B}(x) = \int_0^x \frac{x(E \xi_B \rightarrow B y + N_0)}{E \xi_m \rightarrow B E'_{opt}} K_1 \left( 2 \sqrt{\frac{x(E \xi_B \rightarrow B y + N_0)}{E \xi_m \rightarrow B E'_{opt}}} \right) dy,$$

and

$$F_{SINR_B}(x) = \int_0^x \frac{e^{-\frac{x}{2\sigma^2 IC}}}{2\sigma^2 IC} (2 \sqrt{\frac{x(E \xi_B \rightarrow B y + N_0)}{E \xi_m \rightarrow B E'_{opt}}} ) K_0 \left( 2 \sqrt{\frac{x(E \xi_B \rightarrow B y + N_0)}{E \xi_m \rightarrow B E'_{opt}}} \right) dy,$$ (2.12)

where we used the PDF and CDF of the product of two independent exponentially distributed variables
(78). $\hat{K}_1$ and $\hat{K}_0$ are the modified Bessel function of the second kind of order 1 and 0, respectively, and
$\sigma^2 IC$ is the mean of the exponentially distributed residual self-interference $|\hat{h}_{B \rightarrow B}(s)|^2$.

### 2.4.2 Derivation of Throughput

To obtain the throughput we first needed an expression for average rate per transmission. This
depended on the probability of outage, i.e., that the target rate we are transmitting at, $R$, is above what
the channel may support. We defined the event $A_u^s := \{ C_u^s > R \} = \{ SINR_u(s) > SNR_0 \}$, where $C_u^s$ is the
capacity of the channel after the $s$-th transmission/packet to node $u$, and $u$ is either $m$ for end-user or $B$
for base-station. The $SNR_0$ is the required received SINR to achieve $R$, which is calculated based on the
constrained-constellation capacity equation. We consider that the receivers employ the re-transmissions using an Automatic Repeat-reQuest (ARQ). We assume here the simplest “basic ARQ” used in the ALOHA protocol (ALO); if the packet is not decoded, then the transmitter sends the same packet and the receiver discards the erroneous packets. We state the probability \( p_u(l) \) that the random sequence \( \text{SINR}_u(1), \text{SINR}_u(2), \ldots, \text{SINR}_u(l) \) of SINR at the user \( u \) decoder did not cross the level \( \text{SINR}_0 \) at the \( l \)-th step, \( p_u(l) = P_r\{A^u_1,A^u_2,\ldots,A^u_l\} \). We now derive the probability of outage for forward link as follows:

\[
p_m(l) = \prod_{s=1}^{l} \Pr(\text{SNR}_m(s) < \text{SNR}_0) = \prod_{s=1}^{l} (1 - e^{-\frac{\text{SNR}_0}{E_{\xi_m B}}}). \tag{2.13}
\]

Similarly for the backscatter link, we derive the probability of outage:

\[
p_B(l) = \prod_{s=1}^{l} \Pr\left(\frac{E_{\xi_B m} E_{\xi_B m}'|h_{B \rightarrow m}(s)|^2|h_{m \rightarrow B}|^2}{E_{\xi_B m}^2 |h_{B \rightarrow m}(s)|^2 + N_0} < \text{SNR}_0\right)
= \prod_{s=1}^{l} \left(1 - \int_{0}^{\infty} e^{-\frac{y}{2\sigma_B^2}} 2\sqrt{\frac{\text{SNR}_0(E_{\xi_B B} + N_0)}{E_{\xi_B B} E_{\xi_B m}^2}} \sqrt{\frac{E_{\xi_m B} E_{\xi_m B}'}{E_{\xi_m B} E_{\xi_m B}'}} \tilde{K}_1\left(2\sqrt{\frac{\text{SNR}_0(E_{\xi_B B} + N_0)}{E_{\xi_m B} E_{\xi_m B}'}}\right) dy\right). \tag{2.14}
\]

Under the assumption that all nodes always have packets to send and the feedback is delay-free and error-free, the event that the transmitter stops transmitting the current packet is recognized as recurrent event. Based on the renewal-reward theorem, we define the throughput as \( \nu^u := \frac{R_u}{T_u} = \frac{R(1 - p_u(N))}{\sum_{l=1}^{N-1} l(p_u(l-1) - p_u(l)) + N p_u(N-1)} = \frac{R(1 - p_u(N))}{\sum_{l=0}^{N-1} p_u(l)} \).
where, assuming maximum $N$ transmissions, $\mathcal{R}_u$ is the expected rate in bits/Hz/s, $\mathcal{T}_u$ is the expected number of transmissions per packet and $p_u(0) = 1$.

### 2.4.3 Ergodic Constellation-Constrained Capacity

The constrained capacity for finite-size quadrature amplitude modulation (QAM) constellations is derived in (81) as a function of the average signal-to-noise ratio (SNR), for $M$ equiprobable symbols, as:

$$C(SNR) = \log_2(M) - \frac{1}{M} \sum_{j=1}^{M} \frac{1}{\pi N_0} \int_{y \in c} e^{-\Delta^2(y,s_j)/N_0} \times \log_2 \left( \sum_{s=1}^{M} e^{(\Delta^2(y,s_j) - \Delta^2(y,s_s))/N_0} \right) dy,$$

(2.15)

where $s_j$ is the complex value of input symbol $j$, $\Delta^2(y,s_j) = |y - s_j\sqrt{SNR}/N_0|^2$ and $|.|^2$ denotes the squared Euclidean distance. The summation is of all the elements of the discrete input alphabet. The integration is over all possible values of the received signal. The ergodic constellation-constrained capacity is by definition (82):

$$E[C(SNR)] = \int_0^\infty x f_C(x) dx,$$

(2.16)

where $f_C(x)$ is the PDF of the constellation-constrained capacity. To make our analysis tractable, we used the piece wise linear (PWL) curve fitting of the constrained capacity. The long approximatively linear sections and large radii of the curvature of the constellation-constrained capacity function in Eq. 2.15

1 We stopped transmitting the same packet after $N$ attempts.

2 This expression is applicable to any QAM or PSK modulation type.
Eq. (2.15) made it a good candidate for PWL curve fitting (82). Using this approximation, Eq. (2.15) can be written as:

\[ C(SNR) \approx m_k SNR + b_k \quad \text{for} \quad c_{k-1} < C(SNR) \leq c_k, \]  

(2.17)

where the variables \( m_k \) and \( b_k \) are the slope and intercept coefficients for section \( k \) of the PWL fit. The coefficients are derived numerically for each modulation (see Table 1 in (82)). Using PWL, the ergodic constellation-constrained capacity in Eq. (2.16) thus became:

\[ E[C(SNR)] \approx \sum_{k=1}^{K-1} \int_{c_{k-1}}^{c_k} x f_C(x) dx + [1 - F_{SNR}(SNR_{max})] c_{max}, \]  

(2.18)

where \( c_0 = 0, \ c_{K-1} = c_{max} = \log_2(M), \ K \) is the number of segments in the PWL fit, and \( SNR_{max} \) is an approximation of the SNR needed to reach \( c_{max} \). The term \([1 - F_{SNR}(SNR_{max})]\) accounts for the probability that the received SNR exceeds \( SNR_{max} \), which is required to achieve \( c_{max} \). Since \( C(SNR) \) is a monotonic function of SNR, the PDF \( f_C(x) \) of the constellation-constrained capacity can be approximated for \( c_{k-1} < C(SNR) \leq c_k \) as:

\[ f_C(x) = \frac{f_{SNR}(C^{-1}(x))}{dC/dSNR} \bigg|_{SNR} = \frac{f_{SNR}(\frac{x-b_k}{m_k})}{m_k}, \]  

(2.19)

where \( dC/dSNR = m_k \) and \( C^{-1}(x) = \frac{x-b_k}{m_k} \).
2.4.3.1 Forward link

At the end-user, the SNR_{m}(s) has an exponential distribution with the mean \( \frac{E_{\xi_{B \rightarrow m}}}{N_0} \). Finally, we derived the ergodic constellation-constrained capacity at the end-user by replacing the PDF and CDF of Eq. (2.10) in Eq. (2.19) and Eq. (2.18) to get:

\[
E[C_m(SNR)] \approx K^{-1} \sum_{k=1}^{K-1} \int_{c_{k-1}}^{c_k} \frac{x}{m_k E_{\xi_{B \rightarrow m}}} \exp\left(-\frac{(x-b_k)N_0}{m_k E_{\xi_{B \rightarrow m}}} \right) dx + \exp\left(-\frac{\text{SNR}_{\text{max}}N_0}{E_{\xi_{B \rightarrow m}}} \right) c_{\text{max}}. \tag{2.20}
\]

2.4.3.2 Backscatter link

Similarly to the previous section, we again derive the ergodic constellation-constrained capacity at the base-station by replacing the PDF and the CDF of Eq. (2.12) in Eq. (2.19) and Eq. (2.18) to get:

\[
E[C_B(SNR)] \approx K^{-1} \sum_{k=1}^{K-1} \int_{c_{k-1}}^{c_k} \frac{x}{2\sigma^2_{\text{IC}}} \left( \frac{E_{\xi_{B \rightarrow B_0} + N_0}}{E_{\xi_{m \rightarrow B_0}^{\text{opt}}} E_{\gamma_{m \rightarrow B_0}^{\text{opt}}}} \right)^{\frac{1}{2}} \frac{1}{2} \left( \frac{x-b_k}{m_k E_{\xi_{m \rightarrow B_0}^{\text{opt}}}} + \frac{E_{\xi_{B \rightarrow B_0} + N_0}}{E_{\xi_{m \rightarrow B_0}^{\text{opt}}}} \right) dxdy + c_{\text{max}} \int_{0}^{\infty} e^{-\frac{x}{2\sigma^2_{\text{IC}}}} \left( \frac{\text{SNR}_{\text{max}}(E_{\xi_{B \rightarrow B_0} + N_0})}{E_{\xi_{m \rightarrow B_0}^{\text{opt}}} E_{\gamma_{m \rightarrow B_0}^{\text{opt}}}} \right)^{\frac{1}{2}} \frac{1}{2} \left( \frac{\text{SNR}_{\text{max}}(E_{\xi_{B \rightarrow B_0} + N_0})}{E_{\xi_{m \rightarrow B_0}^{\text{opt}}}} \right) dy. \]

2.5 Performance comparison and trade-offs

In this section, we numerically evaluate the throughput and the ergodic capacity obtained in the previous section for the case of two nodes which wish to exchange data over Rayleigh fading channel.

Transmit power normalization. For a fair comparison between ReflectFX, conventional full-duplex and half-duplex systems, the total energy transmitted by the ReflectFX base-station must be the same as
the total energy transmitted by the two half-duplex or the full-duplex nodes because the ReflectFX end-user does not generate its own radio wave. Since the energy is power times time then the power allocated for each half-duplex node is identical to the power allocated to the ReflectFX base-station. Analogously, when comparing the half-duplex and conventional full-duplex systems, the power allocated to the full-duplex node is half the power allocated to the half-duplex node. There is no doubt that at the ReflectFX, we have DC power consumption that we did not take into account in this comparison, since it is not in the scope of this chapter. I Chapter 3, we cover DC power consumption.

**Modulation-impedance design.** To guarantee that the ReflectFX downlink data-rate equal the half-duplex data-rate, the modulation-impedances are designed to meet the downlink requirement of \( P_{avg} \geq P_{match} \) while also being feasible to implement. To solve for optimum \( E' \), named \( E'_{opt} \), in Eq. (2.8), we exhaustively searched over all possible \( \{c_{st}, \psi\} \). Table I displays the optimal impedances and the optimal \( E'_{opt} \) for QPSK modulation. Our simulation results show that with negative resistance of -22.5, -24.5, -24.95 Ohm, we obtained a backscatter power gain of 8 dB, 22 dB, and 42 dB, respectively. Note that the transmitted power at the end-user varies with the transmitted power from the base-station, the distance between the base-station and the end-user, and the backscatter gain at the end-user. For instance, if the base-station transmit power is set to 15 dBm, the distance between the base-station and end-user node is \( d = 10 \) m, the wavelength is \( \lambda = 0.12 \) m (i.e., the path loss is \( -60 \) dB), the antenna gains are 0 dB, and the backscatter gain is 42 dB \((124^2)\), then the power transmitted at the end-user is \( -3 \) dBm.

**Throughput and ergodic capacity.** We set the target rate at \( R = 2 \) bits/Hz/s, this requires \( SNR_0 = 20 \) (13 dB) for the QPSK constellation. We assume \( \lambda = 0.12 \) m, \( d_{B\rightarrow B} = 1 \) m, \( G_{BT} = G_{BR} = G_m = \)
<table>
<thead>
<tr>
<th>Optimum modulation loads</th>
<th>$R_N$ (Ohm)</th>
<th>$\sqrt{E_{opt}^c}$</th>
<th>Optimum $\psi$</th>
<th>Optimum $\psi_{opt}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha_1 Z_{ant} = -18.89 + 11.48i$, $\alpha_2 Z_{ant} = -18.89 - 11.48i$, $\alpha_3 Z_{ant} = -19.96 + 1.99i$, $\alpha_4 Z_{ant} = -19.96 - 1.99i$</td>
<td>$-20$</td>
<td>$1.2$</td>
<td>$1.8$</td>
<td>$\frac{\pi}{4}$</td>
</tr>
<tr>
<td>$\alpha_1 Z_{ant} = -21.94 + 5.74i$, $\alpha_2 Z_{ant} = -21.94 - 5.74i$, $\alpha_3 Z_{ant} = -22.48 + 0.99i$, $\alpha_4 Z_{ant} = -22.48 - 0.99i$</td>
<td>$-22.5$</td>
<td>$2.4$</td>
<td>$3.1$</td>
<td>$\frac{\pi}{4}$</td>
</tr>
<tr>
<td>$\alpha_1 Z_{ant} = -24.47 + 1.2i$, $\alpha_2 Z_{ant} = -24.47 - 1.2i$, $\alpha_3 Z_{ant} = -24.5 + 0.2i$, $\alpha_4 Z_{ant} = -24.5 - 0.2i$</td>
<td>$-24.5$</td>
<td>$12.4$</td>
<td>$13.1$</td>
<td>$\frac{\pi}{4}$</td>
</tr>
<tr>
<td>$\alpha_1 Z_{ant} = -24.95 - 0.12i$, $\alpha_2 Z_{ant} = -24.95 + 0.12i$, $\alpha_3 Z_{ant} = -24.95 + 0.02i$, $\alpha_4 Z_{ant} = -24.95 - 0.02i$</td>
<td>$-24.95$</td>
<td>$124.5$</td>
<td>$126$</td>
<td>$\frac{\pi}{4}$</td>
</tr>
</tbody>
</table>

**TABLE I: OPTIMUM MODULATION IMPEDANCE FOR QPSK, $Z_{ant} = 50$ OHM, AND $A_{st} = 1$.**
1.6 (dipole antenna), and the maximum number of re-transmissions \( N = 3 \). We also set the path-loss exponent to 2, the polarization mismatch between antennas to 1, and \( A_{sl} = 1 \). For simplicity, we grouped the self-interference cancellation techniques into two classes: 1) passive cancellation; and 2) analog-and-digital cancellation. Our first assumption is that both the mobile and the base-station are using the same analog-and-digital self-interference cancellation technique denoted by \( \sigma_{IC}^2 \) (e.g., -30, -40 and -50 dB). Combining the analog-and-digital cancellation with the passive cancellation, generated by the propagation loss between the transmitting and receiving antennas, the total self-interference cancellation values are 70, 80 and 90 dB at the base-station. In the case of the conventional full-duplex, we assumed the distance between the transmitting and the receiving antennas at the end-user to be 0.1 m. The total self-interference cancellation values are hence 50, 60 and 70 dB at the end-user. All techniques that can be implemented at the end-user can be implemented at the base-station but not the other way around, due to space and processing limitations at the mobile.

**ReflectFX vs half-duplex.** We first compared the throughput of the ReflectFX and half-duplex systems as a function of the distance separating the end-user from the base-station. We set the maximum transmit power to 15 dBm and the noise level to -174 dBm, which make the half-duplex performance fixed at the shown distance range. Fig. 4 shows the throughput for different values of \( E'_{opt} \) and \( \sigma_{IC}^2 \). Note that the throughput of ReflectFX is the sum of the downlink throughput that is equal to the half-duplex throughput and backscatter link throughput. Consequently, it is always higher than the throughput of half-duplex systems. The backscatter link throughput itself could be calculated by subtracting the ReflectFX throughput from the half-duplex throughput. In the high SNR regime (i.e., small distance \( d \)) the half-duplex outage probability is very small and, therefore, its throughput reaches the maximum value.
of 2 bits/Hz/s. Using the negative values of the impedance load, allows the ReflectFX systems to double
the throughput. Fig. 5 shows the ergodic capacity for QPSK input constellation. Similar to the throughput
results, ReflectFX always outperforms the half-duplex system. The gain provided by ReflectFX
compared to the half-duplex increases with the self-interference cancellation and the backscatter gain,
as expected. In Fig. 6 we reduced the transmit power to -15 dBm and set the noise level to the practical
level of -90 dBm. Note that in the low SNR regime (i.e., large distance $d$), the end-user does not receive
enough power to transmit back to the base-station, and so ReflectFX performs as well as the half-duplex.
It is worth notice that the amount of self-interference cancellation can be increased by placing the base-station antennas far apart. There is no severe space restriction in base-station, so we can even reach higher interference cancellation.

*ReflectFX vs conventional full-duplex.* As expected, the performance of the conventional full-duplex depends largely on the amount of self-interference cancellation. In the high SNR regime (i.e., small distance $d$) the ergodic capacity and throughput of ReflectFX are higher than the conventional full-duplex. The gain provided by ReflectFX compared to the full-duplex increases with backscatter gain. In the low SNR regime (i.e., large distance $d$), ReflectFX performs as well as half-duplex but better than the conventional full-duplex with the worst performance because both nodes suffer from self-interference.
The conventional full-duplex outperforms ReflectFX when the amount of self-interference cancellation at the end-user is high and the backscatter gain is low.

To further analyze the impact of the self-interference cancellation at the end-user on the competitiveness of ReflectFX, we plot the throughput and ergodic capacity vs. the total self-interference cancellation at the end-user node. Here we set the maximum transmit power to 15 dBm and the noise level to -174 dBm. In Fig. 7, we analyzed the impact of not having the same analog-and-digital self-interference cancellation at the mobile and the base-station. We fixed the analog-and-digital self-interference cancellation at the base-station to 50 dB, generating a total self-interference cancellation of 90 dB at the base-station. We set the distance between the end-user and base-station to $d = 10$ m. We varied the
level of the analog-and-digital interference cancellation at the mobile from 20 to 70 dB, generating 40 to 90 dB total self-interference cancellation. As shown in Fig[7], the conventional full-duplex performance greatly depends on the value of the end-user interference cancellation. Fig[7] illustrates that ReflectFx outperforms conventional full-duplex in the case where the end-user cannot achieve a high self-interference cancellation because of space or processing limitations. Indeed, conventional full-duplex outperforms ReflectFX when the analog-and-digital cancellation techniques implemented at the mobile are better than those implemented at the base-station. It is also worth noticing that the RF/digital interference cancellation circuits are complicated, costly and take a lot of space in small devices.

Figure 7: ReflectFX and conventional full-duplex performance comparison vs the total self-interference cancellation at end-user (passive cancellation = 20 dB). (a) Throughput and (b) Ergodic capacity. Copyright © 2017, IEEE.
2.6 Conclusion

In this chapter, we proposed an in-band full-duplex wireless communication system that uses reflected power, named ReflectFX. Our concept is based on backscatter modulation, where electromagnetic waves are modulated and reflected by the same antenna that receives them. We considered two nodes, a base-station and an end-user, that intended to exchange data over a wireless Rayleigh fading channel with additive white Gaussian noise. We derived an optimum and accurate design that provides sufficient power for the end-user demodulator while exhibiting maximal difference in the backscattered field. Throughput and ergodic capacity expressions were derived and numerically evaluated. The simulation results allowed us to conclude that there are potentially significant benefits to be gained from including ReflectFX in indoor communication systems.
3.1 Introduction

As explained in the previous chapter, ReflectFX works by varying the end-user loads ($Z_1$ to $Z_n$), to modulate the reflection coefficient and backscatter signal, as illustrated in Fig. 8. Please note that in this chapter we call end-user as tag, to have more correlation with existed backscatter manuscripts.

![Backscatter modulation illustration](image)

Figure 8: Backscatter modulation illustration. Copyright © 2019, IEEE.

In other word, ReflectFX use backscatter modulation (BM) concept to enable IBFD, since it propose a SI free end-user node. Up to date, BM has mostly been used in one-way communication, where all
the received power is used to establish backscatter communication and we can not write data back on the BM tag. Also most BM applications are limited to short-rang communication, due to passive loads used in their structure, such as radio-frequency identification (RFID) tags (83). However today, with the growth of the global distributed network, known as the Internet of Things (IoT), BM tags have potential to serve not only in mere identification, but also as a two-way communication node, to write data back onto the multiple types of devices, comprising the IoT, (84; 85; 86; 87). BM tags can also be integrated with sensor nodes, which make them capable of even a larger variety of applications including both industrial (88) and medical fields (89; 90). Here, we intend to exploit BM in two-way wireless communication systems such as IoT circuits, wireless sensor networks and specifically ReflectFX (91; 92; 1; 93).

In contrast to the approach of passive and conventional one-way RFID tags, we did not use BM to take advantage of its low-power requirements. Instead, we employed active BM to allow the transmission of two-way flows of data over the same carrier at the same time with improved communication range. Accordingly, we need to overcome one-way and short-range communication limitation associated with conventional BM tags due to the two-way path loss associated with.

Recently, several prototypes of BM tags succeed to increase the backscatter communication range by using reflection amplifier that amplifies the reflected signal (94; 95; 96; 97; 98; 99). Recent researches also proposed to use appropriate signal processing, system design and multiple antennas for passive backscatter communications to extend the communication ranges, (100; 101; 102; 103; 104). These techniques can be added to active tags to further improve the range.
In a two-way BM tag, demodulation and modulation are implemented at down-link (from reader
to BM tag) and up-link (from BM tag to reader), respectively. Thus, we need to optimize the BM
tag modulation loads to provide sufficient power for the tag demodulator while exhibiting maximal
difference in the backscattered field, i.e. maximum distance between BM constellation points. In \cite{105, 106} an analytical study has been carried out to optimize the value of loads in a passive RFID tag to yield
a four quadrature amplitude modulation (4-QAM) BM constellation, maximize the available power in
tag and achieve a bit error rate (BER) that is no greater than a predetermined threshold value.

In this chapter, we propose two BM tag configurations with active loads that implement two-way
wireless communications. In order to have no data loss in down-link path, we set an additional require-
ment to the BM tag configurations design. The received power at the BM tag demodulators should be at
least equal to \( P_{\text{match}} \). Where \( P_{\text{match}} \) is the amount of power delivered to a conventional receiver demodula-
tor when its load is conjugate-matched to its antenna impedance. The proposed BM tag configurations,
which we name parallel and series configurations, are analytically studied. By optimizing the set of
modulator loads, that correspond to BM constellation symbols, we a) maximize the BM communication
range, b) implement the desired BM constellation map such as QAM and phase shift keying (PSK) at
up-link, and c) guarantee at least \( P_{\text{match}} \) at the BM tag demodulator. We proposed a new geometric
representation for the optimization problem.

For the parallel BM tag configuration, we used negative resistance as the modulator load in parallel
with the BM tag demodulator. Negative resistance implementation process has been thoroughly covered
in \cite{107, 4, 93, 99, 98, 97, 96}. BM communication is realized by switching among different values of
negative resistors. We used the Norton equivalent circuit (NEC) of the BM tag and the antenna scatterer
theorem (AST) to derive the received and backscatter power at the BM tag, respectively. By using the proposed geometric representation of backscatter symbols, we can set the desired BM constellation map, minimize the average bit error probability (BEP) at the reader and have at least $P_{\text{match}}$ delivered to the BM tag demodulator in an oscillation-free condition.

At the series BM tag configuration, we used a bidirectional amplifier, an impedance matching circuits ($IMC_t$) and a digital phase shifter in series with the BM tag demodulator. The bidirectional amplifier simultaneously amplifies both received and backscattered signals at the BM tag. Different types of bidirectional amplifiers have been proposed in literature and their implementation process have been covered. (107, 108, 109, 110). By switching among different $IMC_t$'s, bidirectional amplifier gain and phase shifter values, the desired BM constellation map is achieved. We used Thevenin equivalent circuit (TEC) of the BM tag and the AST to derive the received and backscatter power at the BM tag, respectively. Like the parallel configuration, we used the proposed geometric representation of backscatter symbols, to set the desired BM constellation map, minimize the average BEP at the reader and have at least $P_{\text{match}}$ delivered to the BM tag demodulator in an oscillation-free condition. As an example, We also implemented 4-QAM constellation to meet all the mentioned optimization criteria for both parallel and series tag configurations.

To conclude this study, We derived a closed-form expression of the average BEP at reader in Rician fading channel environment for both tag configurations. The simulation results show that the series tag configuration provides better average BEP at the reader, at the expense of demanding more complex circuit elements with larger size and higher power consumption.
The remaining of the chapter is organized as follows. Section 3.2 presents preliminaries, including the AST, NEC and TEC. In Section 3.3 and 3.4, we study the parallel and series BM tag configurations, respectively, and derive the analytical approach toward their loads optimization. In Section 3.5, we derive closed-form expression of the average BEP at the reader. Some selected simulation results are provided in Section 3.6, while Section 3.7 concludes the chapter.

Notations: For the rest of this chapter, we specify the parameters of parallel configuration with superscript \( p \) and series configuration with superscript \( s \). We also use superscript \( n \) to specify the parameters associated to each loads value in the parallel configuration. \( \text{Re}(\cdot) \) and \( \text{Im}(\cdot) \) denote the real and imaginary parts of the given parameter, respectively, while \( x^* \) denotes the conjugate of the complex number \( x \) and \( i \) is the imaginary unit i.e. \( i^2 = -1 \).

3.2 Receiving and scattering properties of antenna

To derive the receiving power at the tag antenna, we use the TEC (NEC) to model the receiving antenna by a voltage source and its series impedance, current source and its shunt admittance. In that case, the voltage source is the voltage induced on the antenna at its feed point when it is open-circuit \( (V_{OC}) \), the source impedance is the antenna input impedance \( (Z_{ant}) \) and the load impedance is the impedance of the receiver connected to the antenna \( (Z_L) \). The current source \( (I_{SC}) \) is the current induced on antenna port when it is short circuit, \( Y_{ant} \) is the antenna input admittance and \( Y_L \) is the receiver admittance, as all shown in Fig. 9.

There are number of articles, comments and responses to comments trying to answer the question of how much power is scattered by a receiving antenna (111; 112; 113; 114; 115; 116). Contrary to the prevailing wisdom, the TEC/NEC can only be used to describe the current at the load and cannot
be relied upon for calculating the scattered power. The AST equation should be used to derive the backscatter power at the tag antenna interface (117). Therefore, in contrast to common practice (105; 106; 55), we used AST rather than TEC/NEC to evaluate more accurately the tag backscatter power. We use the AST to derive the scattered electric field $E_{\text{scat}}(r, \theta, \phi|Z_L)$ of an antenna in free space and particular point of $(r, \theta, \phi)$, loaded with impedance $Z_L$, as brought in Eq. (2.1).

3.3 BM Tag Parallel Configuration

In this section, we introduce the first proposed active two-way BM tag configuration, named parallel configuration. We first present the proposed BM tag structure and then derive the received and backscatter power expressions. Based on the geometric representation of the backscatter symbols, we optimize the BM tag modulator loads to implement the desired BM constellation map while maximiz-
ing the backscatter power and guaranteeing at least \( P_{match} \) at the demodulator. We also consider an oscillation-free condition in the BM tag design.

### 3.3.1 Parallel configuration schematic

The proposed parallel BM tag configuration includes three parts: i) the antenna, ii) the modulator loads \( (Y_{mod}) \), and iii) the BM tag demodulator \( (Y_{tag}) \), as shown in Fig. 10. \( Y_{tag} \) indicates the BM tag demodulator admittance which consists of both real and imaginary parts in general.

In this section we use the NEC of Fig. 10 to make equations simple and better illustrate each part of the circuit. Fig. 11 describes the NEC in parallel configuration.

In Fig. 11, the load \( Y_{tag} \) has complex value, where its real part is the conductance \( G_{tag} \), and the imaginary part is the susceptance \( B_{tag} \). The antenna admittance is given by \( Y_{ant} \) that has real and imaginary...
parts equal to $G_{ant}$ and $B_{ant}$, respectively. Without loss of generality, we set $Y_{p,n}^{mod} = \beta_n G_{tag} + B_{mod} + G_{mod}$ with $\beta_n \in \mathbb{C}$. $B_{mod}$ and $G_{mod}$ are chosen so the following equations hold

$$B_{ant} + B_{mod} + B_{tag} = 0, \quad \text{and} \quad G_{mod} + G_{ant} = G_{tag}. \quad (3.1)$$

Thus the modulation load has extra passive parts $B_{mod}$ and $G_{mod}$ which are added parallel to $\beta_n G_{tag}$ to cancel out the imaginary parts of the antenna and tag and match the real part of antenna to the real part of tag. In this way, we added the matching parts to the circuit and included them in our model to maximize the received signal in non-backscattering mode. Having Eq. (3.1), leads to simplify Fig. 11 to Fig. 12.

Note that $1 \leq n \leq M$, where $M$ is the total number of the backscatter symbols (i.e. $M$ is the constellation size).
The BM is realized via switching among different values of $\beta_n$. The admittance for the whole load connected to the antenna is $Y_{p,n}^{p,n}$

$$Y_{p,n}^{p,n} = \beta_n G_{tag} + G_{tag}.$$ (3.2)

The power reflection coefficient seen at the antenna port is

$$\Gamma_{p,n}^{p,n} = -\frac{\beta_n}{\beta_n + 2}.$$ (3.3)

### 3.3.2 Received power at the BM tag demodulator

The received power at the BM tag demodulator, denoted by $P_r^{p,n}$, is calculated using the NEC of the proposed parallel configuration as,

$$P_r^{p,n} = \frac{1}{2G_{tag}} |I_{tag}^{p,n}|^2,$$ (3.4)
where $I^{p,n}_{tag}$ is the current at the demodulator. $I^{p,n}_{tag}$ is calculated as $I^{p,n}_{tag} = \frac{I_{SC}}{2 + \beta}$. Thus, the demodulator received power corresponding to each $\beta_n$ is derived as

$$P^{p,n}_{r} = \frac{1}{2} \frac{|I_{SC}|^2}{(2 + \beta)^2 G_{tag}}.$$  \hfill (3.5)

### 3.3.3 Backscatter electric field and power

The backscatter power at the BM tag interface can be derived from the backscatter electric field as follows

$$P^{p,n}_{scat} = \frac{1}{2z} |E^{p,n}_{scat}|^2,$$  \hfill (3.6)

where $z$ is the characteristic impedance of the transmission medium. In free space $z = 377$ Ohm. $E^{p,n}_{scat}$ is the BM tag backscatter electric field corresponds to each modulator load and is calculated using the AST expression of Eq. (2.1), as

$$E^{p,n}_{scat}(r, \theta, \phi | Z_{L}) = I(Z_{ant}) E_{r}(r, \theta, \phi) (A_{st} - \Gamma^{p,n}_{L}),$$  \hfill (3.7)

where $A_{st} = E_{scat}(r, \theta, \phi | Z_{ant}^*) / (I(Z_{ant}) E_{r}(r, \theta, \phi))$ is the antenna structural mode and $\Gamma^{p,n}_{L}$ is given by Eq. (3.3) for the parallel configuration. Note that $A_{st}$ has a complex value, which depends on the material and geometry of the antenna, a method to compute the value of $A_{st}$ has been proposed in [118]. Substituting Eq. (3.3) in Eq. (3.7) and setting $E_0 = I(Z_{ant}) E_{r}(r, \theta, \phi)$, we have

$$E^{p,n}_{scat}(r, \theta, \phi | Z_{L}) = E_0 \left( A_{st} + \frac{\beta_n}{\beta_n} \right).$$  \hfill (3.8)
The backscatter electric field can represent the backscatter constellation symbols in BM communication. $E_0$ has a fix value for each antenna that does not change with different loads. Thus and without loss of generality, we define the backscatter constellation symbols as follows

$$S_n = \frac{E_{\text{scat}}(r, \theta, \phi|Z_{\text{L}}^{p,n})}{E_0} = A_{st} + \frac{\beta_n}{\beta_n + 2}. \quad (3.9)$$

$A_{st}$ has a fixed value for all backscatter symbols, while the second term changes with the modulator loads and takes different values for each backscatter symbol.

3.3.4 Optimization of loads – geometric representation

3.3.4.1 Tag demodulator received power constraint

In BM communication systems, the loads connected to the antenna ($Z_{\text{L}}^{p,n}$) are not conjugate-matched to the antenna impedance and a portion of the received power is reflected toward the reader. This results in downlink data loss compared to a conventional receiver, when the load is conjugate-matched to the antenna. However, by using active loads in BM tag configurations, we can amplify the received power and compensate the downlink power loss. Thus, we set the following requirement to the received power at the BM tag demodulator

$$P_{r}^{p,n} \geq P_{\text{match}}, \quad \forall n \in \{1, \ldots, M\}, \quad (3.10)$$

where $P_{\text{match}}$ is the power delivered to the BM tag demodulator when the modulator is open circuit and the demodulator is conjugate-matched to the antenna (conventional receiver circuits). It is given by

$$P_{\text{match}} = \frac{1}{8} \frac{|I_{SC}|^2}{G_{\text{tag}}}. \quad (3.11)$$
Eq. (3.10) guarantees that the received power at demodulator is always larger or equal to $P_{\text{match}}$ for all the BM symbols and modulation load values. Substituting Eq. (3.11) and Eq. (3.5) into Eq. (3.10), the demodulator received power constraint is derived in terms of the modulator loads as

$$\frac{1}{2|2 + \beta_n|^2} \geq \frac{1}{8}, \quad \forall n \in \{1, \ldots, M\}. \quad (3.12)$$

Since each value of $\beta_n$ maps to its corresponding backscatter symbol (see Eq. (3.9)), Eq. (3.12) can be rewritten as follows

$$|\frac{S_n^p - A_{st}}{1 - S_n^p + A_{st}} + 1| \leq 1, \quad \forall n \in \{1, \ldots, M\}. \quad (3.13)$$

Note that backscatter symbols have real ($x_n^p$) and imaginary ($y_n^p$) values in the In-phase/Quadrature (IQ) constellation plane, $S_n^p = x_n^p + iy_n^p$, and the antenna structural mode also has a complex value, $A_{st} = a + ib$. Thus, Eq. (3.13) can be mapped to the IQ constellation plane as follows

$$(x_n^p - (a + 1))^2 + (y_n^p - b)^2 \geq 1. \quad (3.14)$$

Eq. (3.14) represents the area outside the unit circle centered at $(a + 1, b)$, shown by the hatch area in Fig. 13a. For passive loads, we have $\text{Re}(\beta_n) \geq 0$, which maps to the inside area of the circle defined by $(x_n^p - (a + 0.5))^2 + (y_n^p - b)^2 \leq 0.25$, as illustrated by dot pattern in Fig. 13a.
3.3.4.2 Tag free-oscillation constraint

Any active circuit has the potential to become unstable and start to oscillate. This potential increases as gain increases (119). Based on Eq. (3.8), as $\beta_n$ moves toward -2, the backscatter gain and the risk of
having oscillation increase. Thus, in order to stabilize the circuit and guarantee no oscillation, a tighter bound on $\text{Re}(\beta_n)$ should be applied by setting a new offset, $\varepsilon > 0$, as

$$\text{Re}(\beta_n) \geq -2 + \varepsilon, \quad \forall n \in \{1, \ldots, M\}. \tag{3.15}$$

Eq. (3.15) is mapped to the area inside the circle centered at $C_{st}^p = (a + 1 - \frac{1}{\varepsilon}, b)$ with radius $\frac{1}{\varepsilon}$, in the constellation plane, as

$$(x_n^p - (a + 1 - \frac{1}{\varepsilon}))^2 + (y_n^p - b)^2 \leq \left(\frac{1}{\varepsilon}\right)^2. \tag{3.16}$$

Note that for $\text{Re}(\beta_n) < -2$, the net resistance of circuit is negative, causing an unstable oscillation.

### 3.3.4.3 Optimization through geometric representation

The new bound of $(\beta_n)$ is the union of the tag demodulator received power constraint (Eq. (3.14)) and the tag free-oscillation constraint (Eq. (3.16)) areas, as shown by the hatch area in Fig. 13b. This area is inside the circle centered at $(a + 1 - \frac{1}{\varepsilon}, b)$ with radius $\frac{1}{\varepsilon}$ and excluding the area limited by the circle centered at $(a + 1, b)$ with radius 1. Each backscatter symbol $S_n^p$, being represented as

$$S_n^p = C_{st}^p + \sqrt{E_n^p} e^{j(\theta_n^p + \varphi)}, \quad \forall n \in \{1, \ldots, M\}, \tag{3.17}$$

can be placed on the circle with center $C_{st}^p = (a + 1 - \frac{1}{\varepsilon}, b)$ and radius $\sqrt{E_n^p}$. Note that $C_{st}$ is the DC term due to backscattering from the antenna (the structure part of antenna). Generally, this term plays in calculating average backscatter energy. However, in average BEP derivation, only the distance between backscatter constellation points matters and this term should be omitted. In other words, $E_n^p$
directly affects the BEP at the reader, not $C_{st}$. $\theta_p^n$ is the phase of each symbol and $\psi_p$ is the rotation of all symbols. By choosing the backscatter symbols on the circumference of the hatch circle, $E_p^n$ is maximized and both the demodulator received power and oscillation-free constraints are guaranteed.

### 3.3.5 4-QAM example

In this section, a 4-QAM BM constellation scheme is implemented as an example. We set the constellation symbols in the feasible area of Fig. 13b and calculate the value of modulator loads corresponding to each symbols. The 4-QAM constellation symbols are $S_1^p, S_2^p, S_3^p$, and $S_4^p$, shown in Fig. 13b. These symbols are chosen for maximum distance among the backscatter symbols. For each of these symbols, the corresponding modulator load admittance is calculated and given by $Y_{mod}^{p,n}$ in Table I when $\varepsilon = 0.06, \psi_p = 0, A_{st} = 1$ and $B_{mod} = G_{mod} = 0$ (the antenna and demodulator are conjugate-matched). Otherwise, $B_{mod}$ and $G_{mod}$ should be added to $Y_{mod}^{p,n}$. We also report the corresponding values of $\beta_n$ for $G_{tag} = 1/50$ Ohm. The average received power at the BM tag demodulator is calculated as

---

**TABLE II: 4-QAM Constellation Symbols Example and Their Corresponding $\beta_n$ and Modulator Loads $Y_{mod}^{p,n}$**

<table>
<thead>
<tr>
<th>$S_p^n$</th>
<th>$\beta_n$</th>
<th>$Y_{mod}^{p,n}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$S_1^p$</td>
<td>$-26.4+11.7i$</td>
<td>$\beta_1 = -1.94+0.024i$</td>
</tr>
<tr>
<td>$S_2^p$</td>
<td>$-2.88+11.7i$</td>
<td>$\beta_2 = -1.94+0.144i$</td>
</tr>
<tr>
<td>$S_3^p$</td>
<td>$-2.88-11.7i$</td>
<td>$\beta_3 = -1.94-0.144i$</td>
</tr>
<tr>
<td>$S_4^p$</td>
<td>$-26.4-11.7i$</td>
<td>$\beta_4 = -1.94-0.024i$</td>
</tr>
</tbody>
</table>
\[ P_{\text{ave}}^n = \frac{1}{4} \sum_{n=1}^{4} P_{r,n}^n = \frac{1}{4} \sum_{n=1}^{4} \frac{4}{|2 + \beta_n|^2} P_{\text{match}}, \quad \forall n \in \{1,\ldots,4\}, \tag{3.18} \]

which is plotted as a function of \( \varepsilon \), as shown in Fig. 14a. As \( \varepsilon \) gets smaller values (moves toward the oscillation point), the average power delivered to the tag demodulator increases.

The average power of backscatter symbols is calculated as

\[ E_{\text{ave}} = \frac{1}{4} \sum_{n=1}^{4} E_n^p = \left( \frac{1}{\varepsilon} \right)^2, \quad \forall n \in \{1,\ldots,4\}, \tag{3.19} \]

where \( E_n^p \) is the square of each symbol amplitude, as shown in Fig. 13b. For a 4-QAM BM constellation map and maximum distance between backscatter symbols, \( \sqrt{E_n^p} \) has the same value for all symbols, equals to the radius of the hatch circle in Fig. (b) \( E_{\text{ave}}^p \) as a function of \( \varepsilon \) is plotted in Fig. 14b. As \( \varepsilon \) gets small, the average backscatter symbols power increases. The average received power at the BM tag demodulator and the backscatter symbols power show the same behavior. Thus, as we move toward oscillation point, both demodulator received power and backscatter symbols power increase.

### 3.4 BM Tag Series Configuration

In this section, we introduce the second proposed active two-way BM tag configuration, named series configuration. Like the previous section, we first present the proposed BM tag structure and derive the received and backscatter power. Based on the geometric representation of the backscatter symbols, we optimize the BM tag modulator loads to implement the desired BM constellation map, maximize the backscatter power and guarantee at least \( P_{\text{match}} \) at the demodulator.
3.4.1 Series configuration schematic

The proposed series BM tag configuration includes six parts: (i) the antenna, (ii) the antenna impedance matching circuit \((IMC_a)\), (iii) the digital phase shifter, (iv) the bidirectional amplifier (Bi-Amp), (v) the impedance matching circuit of tag demodulator \((IMC_t)\), and (vi) the BM tag demodulator \((Z_{tag})\), as shown in Fig. [15]. Bidirectional amplifiers are two-port amplifiers, which can amplify the incoming signal at both directions. Digital phase shifters are used to change the phase of electromagnetic waves, controlled by digital bits. The digital phase shifter, bidirectional amplifier, and tag demodulator have standard input impedance equals to \(R_{match}\) (usually 50 Ohm), and the antenna is matched with them using \(IMC_a\). To implement BM link, we make a deliberate mismatch \((\Gamma_t)\) using \(IMC_t\) at the tag de-
modulator interface. $\Gamma^s_t$ is the reflection coefficient seen at the BM tag demodulator interface. IMC_t is a passive loss-less impedance matching circuit which transfers $Z_{\text{tag}}$ to $KR_{\text{match}}$, where $K$ is a positive real number. The value of $K$ and bidirectional amplifier gain $G$, determine the amplitude of backscatter symbols. In order to modulate the amplitude of backscatter signal, we vary $K$, using RF switch at the IMC_t port. This is easier than changing the gain of bidirectional amplifier, since commercial bidirectional amplifier might have a fix gain. The symbols phase are set using the phase shifter. Therefore, to implement PSK constellation, $K$ and $G$ are fixed. To generate QAM constellation, we can change the amplitude of the backscatter symbols by varying the values of $K$ via switching among different IMC_t structures.

### 3.4.2 Received power at the BM tag demodulator

To make the equation more traceable, we fixed the values of $G$ and $K$ and switch between different values of phase shifter (M-PSK constellation). The received power at the BM tag demodulator, denoted
by $P_r^s$, has the same value for all different values of phase shifter and is derived using the TEC of the proposed series configuration, shown in Fig. 16. We assume to have a phase shifter with average insertion loss of $L_{ps}$, and loss-less impedance matching circuits. We also assumed that the bidirectional amplifier has the same forward and backward power gain $G$. Based on the TEC shown in Fig. 16 the received power at $KR_{tag}$ is defined as

$$P_r^s = \frac{1}{2} K R_{match} |I_L|^2 \text{ where } I_L = \sqrt{G L_{ps} \frac{V_{OC}}{R_{match} + K R_{tag}}}. \quad (3.20)$$

Thus,

$$P_r^s = \frac{1}{2} \left( \frac{V_{OC}}{K + 1} \right)^2 G L_{ps} K \frac{4K}{(K + 1)^2}. \quad (3.21)$$

Note that the RF switches also have loss, but since they are common part of all BM systems (passive, parallel and series configurations), they do not affect our comparative analysis.
3.4.3 Backscatter electric field and power

Like the parallel configuration, we use the AST to derive the backscatter power at the series BM tag interface

\[ P_{\text{scat}}^e,n = \frac{1}{2} |E_{\text{scat}}^e,n|^2 \text{ with } E_{\text{scat}}^e,n(r, \theta, \phi|Z_{S,n}^L) = E_0(A_{st} - \Gamma_{L,n}^s). \]  

(3.22)

As discussed in Section 3.3, we set the backscatter symbols to the normalized backscatter electric field, \( S_{n}^s = (A_{st} - \Gamma_{L,n}^s) \), where \( \Gamma_{L,n}^s \) is the power reflection coefficient seen at the input port of the phase shifter.

To derive the expression of \( \Gamma_{L,n}^s \), we first need to find the reflection coefficient of \( \Gamma_{B}^s \), seen at the bidirectional amplifier interface. \( \Gamma_{B}^s \) depends on the structure of the bidirectional amplifier. Using the bidirectional amplifier circuit proposed in (108), \( \Gamma_{B}^s \) can be expressed as

\[ \Gamma_{B}^s = -G \Gamma_{L}^s = -G \left( \frac{K - 1}{K + 1} \right), \]  

(3.23)

Therefore, \( \Gamma_{L,n}^s \) is given by

\[ \Gamma_{L,n}^s = \Gamma_{B}^s e^{-i2\theta_{ps}^n} = -G L_{ps} \left( \frac{K - 1}{K + 1} \right) e^{-i2\theta_{ps}^n}, \]  

(3.24)

where \( \theta_{ps}^n \) is the phase of the phase shifter for each backscatter symbol. Since the backscatter waves pass the phase shifter twice, we set the phase shifter delay as

\[ \theta_{ps}^n = -\frac{\theta_{s}^e}{2} = -\frac{2n - 1}{2M} \pi, \quad \forall n \in \{1,..,M\}, \]  

(3.25)
$M$ is the constellation size and $\theta_n^s$ is the phase of each backscatter symbol. Thus, the backscatter symbols, $S_n^s$, can be defined by

$$S_n^s = A_{st} + G L_{ps} \frac{K - 1}{K + 1} e^{i(\theta_n^s + \psi)} , \quad \forall n \in \{1, \ldots, M\} , \quad (3.26)$$

$\psi^s$ represents the rotation of all symbols.

### 3.4.4 Optimization of loads – the geometric representation

#### 3.4.4.1 Tag demodulator received power constraint

For the proposed series BM tag configuration, we set the same power requirement at the tag demodulator as the parallel configuration,

$$P_r^s \geq P_{\text{match}} . \quad (3.27)$$

We derive upper and lower bound of $K$, using Eq. (3.27) and Eq. (3.21), as

$$\left(\sqrt{G L_{ps}} - \sqrt{G L_{ps} - 1}\right)^2 \leq K \leq \left(\sqrt{G L_{ps}} + \sqrt{G L_{ps} - 1}\right)^2 . \quad (3.28)$$

Eq. (3.28) shows that to guarantee the tag demodulator power constraint, for a given value of $G$, $K \in [K_L, K_U]$. Also increasing the gain of bidirectional amplifier or decreasing the loss of phase shifter increase the range of $K$. Note that $K$ determines the value of $P_r^s$, see Eq. (3.21). For example, if $K = 1$ then $P_r^s = G L_{ps} P_{\text{match}}$, and if $K = K_L$ or $K = K_U$ then $P_r^s = P_{\text{match}}$. 


3.4.4.2 Tag oscillation-free constraint

For the proposed series BM tag configuration, the oscillation-free constraint is related to the gain of the bidirectional amplifier. There are several types of bidirectional amplifiers proposed in literature [110, 120, 121, 122, 108]. In this section we are not concerned about the oscillation criteria in the circuit, since $G$ is reported for the oscillation-free condition.

3.4.4.3 Optimization through geometric representation

Each value of $K$ maps to its corresponding backscatter symbol, see Eq. (3.26). Thus we can rewrite Eq. (3.28) in terms of $S_n$, to derive the geometric representation of feasible backscatter constellation area,

$$|S_n - A_u| \leq \sqrt{GL_{ps}^2 - GL_{ps}}, \quad \forall n \in \{1, \ldots, M\}. \quad (3.29)$$

For the complex constellation symbols in IQ plane, $S_n = x_n + iy_n$ and $A_u = a + ib$, Eq. (3.29) maps to the circle centered at $(a, b)$ with radius $\sqrt{GL_{ps}^2 - GL_{ps}}$, shown by hatch area in Fig. 17.

$$(x_n - a)^2 + (y_n - b)^2 \leq GL_{ps}^2 - GL_{ps}. \quad (3.30)$$

Note that each backscatter symbol $S_n$, being represented as

$$S_n = C_{st} + \sqrt{E_n} e^{i(\theta_s + \psi)}, \quad \forall n \in \{1, \ldots, M\}, \quad (3.31)$$

can be placed on a circle with center $C_{st} = (a, b)$ and radius $\sqrt{E_n} = GL_{ps} \left| \frac{K-1}{K+1} \right|$. 
Figure 17: Backscatter symbols constellation plane corresponding to $K_L \leq K \leq K_U$. Copyright © 2019, IEEE.

3.4.5 4-QAM example

In this section a 4-QAM BM constellation scheme is implemented, as an example. We use the geometric representation of Fig. [17] to choose the backscatter symbols $S_n$. The 4-QAM constellation symbols of $S_1$, $S_2$, $S_3$ and $S_4$ are picked, as shown in Fig. [17]. These symbols are chosen on the circumference of the hatch circle to maximize the distance among the backscatter symbols. As shown in Fig. [17] on the circumference of the hatch circle $K = K_U$ and $E_n^a = GL_{ps}^2 - GL_{ps}$. For $G = 13$ dB, $A_{st} = 1$, $\psi^a = 0$ and $L_{ps} = -3$ dB, backscatter symbols and their corresponding phase shifter values are given in Table [III]. The average received power at the BM tag demodulator for 4-QAM constellation is
TABLE III: 4-QAM CONSTELLATION SYMBOLS EXAMPLE AND THEIR CORRESPONDING PHASE SHIFTER VALUES.

| $S_i^*$ | $7.7+6.7i$ | $\theta_{ph}^1$ | $-22.5^\circ$ |
| $S_1$ | $-5.7+6.7i$ | $\theta_{ph}^2$ | $-67.5^\circ$ |
| $S_2$ | $-5.7-6.7i$ | $\theta_{ph}^3$ | $-112.5^\circ$ |
| $S_3$ | $7.7-6.7i$ | $\theta_{ph}^4$ | $-157.5^\circ$ |

$$P_{\text{ave}}^s = \frac{1}{4} \sum_{n=1}^{4} P_{r,n}^s = P_{\text{match}} G L_{ps} \frac{4K}{(K+1)^2}, \quad \forall n \in \{1,..,4\}, \quad (3.32)$$

For a fixed value of $G$ and $L_{ps}$, we plotted $P_{\text{ave}}^s$ as a function of $K$ in Fig. 18a. At $K = 1$, $P_{\text{ave}}^s = G L_{ps} P_{\text{match}}$ and at $K = K_L$ or $K = K_U$, $P_{\text{ave}}^s = P_{\text{match}}$. If $K \notin [K_L, K_U]$, then $P_{\text{ave}}^s < P_{\text{match}}$.

The average power of backscatter symbols for 4-QAM constellation is calculated as

$$E_{\text{ave}}^s = \frac{1}{4} \sum_{n=1}^{4} E_n^s = \left( GL_{ps} \frac{K-1}{K+1} \right)^2, \quad \forall n \in \{1,..,4\}, \quad (3.33)$$

$E_{\text{ave}}^s$ is plotted in Fig. 18b as a function of $K$ for a fixed value of $G$. At $K = 1$, $E_{\text{ave}}^s = 0$, and at $K = K_L$ or $K = K_U$, $E_{\text{ave}}^s = (G L_{ps})^2 - G L_{ps}$. Fig. 18 shows that, there is a tradeoff between the received power at tag demodulator $P_{\text{ave}}^s$ and backscatter symbols power $E_{\text{ave}}^s$. As the value of $K$ moves closer to $K_L$ or $K_U$, $E_{\text{ave}}^s$ increases but $P_{\text{ave}}^s$ decreases. This is contrary to the parallel configuration, where when $\varepsilon$ moves toward 0, both $E_{\text{ave}}^p$ and $P_{\text{ave}}^p$ increase.
3.5 Average Bit Error Probability

Since we guaranteed at least $P_{\text{match}}$ at the tag demodulators, the average bit error probability (BEP) at the reader will be used to compare the two proposed active BM tag configurations: Series and parallel. Due to the particularity of the channel between the BM tag and the reader, the distribution of the signal-to-noise ratio (SNR) should be studied before handling with the conditional BEP and then the average BEP.
3.5.1 SNR distribution

We define $P_p^d$ and $P_s^d$ as the received power at the reader for the parallel and series BM tag configurations, respectively. We derived their expressions using the AST for a fading channel as

$$
P_{d}^{p,s} = P_t G_d^{TX} L_{rt} |u_t . u_{inc}|^2 G_t^2 L_{tr} |u_r . u_{scat}|^2 G_d^{RX} 
\times |h_{rt} h_{tr}|^2 |A_{st} - \Gamma_l^{p,s}|^2 + N_t F_t |\Gamma_l^{p,s}|^2 L_{tr} |h_{tr}|^2 + N_r F_r,
$$

(3.34)

where $P_t$ is the transmitted power from the reader to the tag, $G_d^{TX}$ and $G_d^{RX}$ are the reader’s transmitter (TX) and receiver (RX) antenna gains, $L_{rt}$ and $L_{tr}$ are path-loss associated with signal from reader-to-tag and tag-to-reader, and $|u_t . u_{inc}|$ and $|u_r . u_{scat}|$ account for the possible polarization mismatch between (incident waves and tag antenna) and (backscattered waves and reader antenna), respectively. $h_{rt}$ and $h_{tr}$ are the normalized reader-to-tag and tag-to-reader channel gains. $N_t$ and $F_t$ are thermal noise power and noise figure at tag, amplified by tag backscatter gain and received at the reader through channel $h_{tr}$ with path loss $L_{tr}$. $N_r$ and $F_r$ are thermal noise power and noise figure at reader. It has been shown in [123], that the equivalent noise at reader due to tag thermal noise is much less significant than the reader’s own thermal noise. Thus, the instantaneous SNR at reader, $\gamma^{p,s}$, that is given by

$$\gamma^{p,s} = \gamma_0 |h_{rt} h_{tr}|^2 |A_{st} - \Gamma_l^{p,s}|^2,
$$

(3.35)
where $\gamma_0$ is the average SNR received at the reader, normalized by the backscatter symbol power, given by

$$\gamma_0 = \frac{1}{N_r F_r} P_t G_{d}^{TX} L_t |u_t u_{inc}|^2 G_{r}^{2} L_r |u_r u_{scat}|^2 G_{d}^{RX}.$$  \hfill (3.36)

The backscatter symbol, $S_{p,s} = A_{st} - \Gamma^{p,s}_{L}$, is given by Eq. (3.17) and (3.31) for parallel and series BM tag configurations, respectively. The offsets $C^p_{st}$ and $C^s_{st}$ do not contribute on the BEP derivation. Therefore, the contributed SNR in BEP, $\gamma_{p,s}^{p,s}$, for the parallel and series BM tag configurations can be written as

$$\gamma_{p,s}^{p,s} = \gamma_0 |h_{rt} h_{tr}|^2 E_{ave}^{p,s}.$$  \hfill (3.37)

The measurements in (124) showed that the backscattered signal has deeper fades than the signal from one link because the fading on the backscattered signal is the product of the downlink fading and the uplink fading. Thereby, the best model to fit the backscattered signal fading is the product Rician distribution. It is worth mentioning that some measurements of the product Rician distribution parameters were described in (124). Furthermore, it was proven in (125) that the Rician distribution is valid for LOS models, and that the product-Rician with independent segments is valid for local area, the measurements therein affirm that the backscatter channel can be approximated by a product Rician distribution with independent segments for monostatic and bistatic scenarios, where the reader transmitter and receiver antennas are separated and adequately spaced. In addition, in (126), the channel segments in RFID system were assumed independent, based on CDF measurements. Hence, the channel coefficients, $|h_{rt}|$ and $|h_{tr}|$, are modeled by Rician distribution with probability density function (PDF) given in (127) Eq. (2.15)) and distribution factors $k_{rt}$ and $k_{tr}$ for downlink and uplink, respectively. The PDF of the product
of two independent Rician random variables, $|h_r h_t|^2$, can be obtained in integral form. However, an infinite sum representation was derived in (128), where low number of terms can be considered with small truncation errors, depending on the Rician factors (e.g. for $k_{tr} = k_{rt} = 3$ dB, the truncation error reduces to $10^{-15}$ with only 600 terms). For that reason and by considering a change of variable, the PDF of the SNR can be derived using (128, Eq. (24)) as

$$f_{\gamma_{eff}}^{\eta_{p,s}}(\gamma) = \frac{2\kappa e^{-(k_{tr} - k_{rt})}}{\gamma_0 E_{ave}^{p,s}} \sum_{j,l=0}^{m} \frac{k_{tr}^j k_{rt}^l}{(j!l!)^2} \left( \frac{\kappa \gamma}{\gamma_0 E_{ave}^{p,s}} \right)^{\frac{j+l}{2}} K_{j-l} \left( \sqrt{\frac{4\kappa \gamma}{\gamma_0 E_{ave}^{p,s}}} \right),$$

(3.38)

where $\kappa = (1 + k_{tr})(1 + k_{rt})$ and $K_{j}(\cdot)$ is the $j$-th order modified Bessel function of the second kind (129, Eq. (9.6.24)).

### 3.5.2 Average BEP

Through an additive white Gaussian noise (AWGN) channel, the conditional BEP of 4-QAM constellation map is given in (127, Eq. (8.16)) by

$$\eta_{e}^{p,s} = Q\left( \sqrt{\gamma_{eff}^{p,s}} \right),$$

(3.39)

where $Q(\cdot)$ is the Gaussian $Q$ function (127, Eq. (4.1)). The average BEP is obtained by averaging the conditional BEP over the distribution of the SNR

$$\overline{\eta}_{e}^{p,s} = \int_{0}^{\infty} \eta_{e}^{p,s} f_{\gamma_{eff}}^{\eta_{p,s}}(\gamma) d\gamma.$$

(3.40)
Thus, by substituting Eq. (3.38) and Eq. (3.39) in Eq. (3.40), we get the average BEP as follows

$$\eta_{p,s}^e = \frac{2\kappa e^{-k_{tr}-k_{rt}}}{\gamma_0 E_{p,s}} \sum_{j,l=0}^{\infty} k_{tr}^j k_{rt}^l \int_0^\infty \left( \frac{\kappa \gamma_{p,s}}{\gamma_0 E_{p,s}} \right)^{j+l} Q(\sqrt{\gamma}) K_{j-l} \left( \sqrt{\frac{4\kappa \gamma_{p,s}}{\gamma_0 E_{p,s}}} \right) d\gamma. \quad (3.41)$$

To find a compact expression of Eq. (3.41), an integral of the form

$$\int_0^\infty x^{j+l+1} Q(x) K_{j-l}(ax) dx$$

should be solved for some positive number $a$. Using alternative expressions of the $Q$ function and the Bessel function in terms of the Meijer’s G function (MGF) (130, Eq. (2.9.1)), available in (131, Eq. (8.4.14.2)) and (130, Eq. (2.9.19)), respectively, the integral becomes an integral of the product of two MGFs, which can be solved using the integral identity (130, Eq. (2.8.4)). Therefore, the average BEP of 4-QAM BM scheme can be expressed as

$$\eta_{p,s}^e = e^{-k_{tr}-k_{rt}} \frac{2\kappa}{2\sqrt{\pi}} \sum_{j,l=0}^{\infty} \frac{k_{tr}^j k_{rt}^l}{(j+l)!} \int_0^\infty Q\left(\sqrt{\frac{4\kappa \gamma_{p,s}}{\gamma_0 E_{p,s}}} \right) d\gamma.$$

3.6 simulation results

This section shows some selected numerical results supported by Monte Carlo MATLAB® simulations to illustrate the average BEP at the reader. To compare the proposed tag configurations, we used $\gamma_0$ as the reference SNR and plotted the average BEP derived in Eq. (3.42), versus $\gamma_0$. While for each parallel and series BM tag configurations, the average backscatter symbols power, gets different values. We assume 4-QAM constellation scheme and the maximum backscatter power as an example.

3.6.1 Parallel BM tag configuration

For the parallel configuration, we run the simulations and plotted the average BEP at reader, for different values of $\text{Re}(\beta)$ and maximum backscatter power, shown in Fig. [19]. Note that the value of
average backscatter symbols power $E_{ave}^p$, can be derived from the value of $\text{Re}(\beta)$, based on Eq. (3.19). We assume Rician fading with Rician factors $k_{rt} = k_{tr} = 2.7$ dB (124). We set three different values of $\text{Re}(\beta)$: $-1.78, -1.75, \text{and} -1.69$. These numbers are practical values that set the circuit far from the oscillation point, since they correspond to reflection gain of 18.4 dB, 17 dB, and 14 dB, respectively.

Based on (132), tunnel-diode based reflection amplifier starts to oscillate for more than 20 dB gain. We also plotted the average BEP of passive BM tags to better highlight the impact of active loads on the reader average BEP. Like parallel and series BM tags, The maximum backscatter symbols power for a passive tag is the square of its constellation circle radius. The constellation circle of passive tags is shown with dot area in Fig. 13a. Thus, $E_{ave}^{\text{passive}} = (0.5)^2$.

The increased required power in active parallel configuration compared with passive configuration for the example of tunnel-diode based negative resistance is 0.2 mW (3). Note that since both passive and parallel configuration need RF switch to implement BM, we could wave the switches consumed power as a matter of comparison.

As expected, if we pick $\text{Re}(\beta)$ close to the oscillation point, the backscatter power increases and the average BEP at the reader decreases as a result. For the conventional passive BM tag, the reader needs 47 dB of $\gamma_0$ to guarantee an average BEP of $10^{-4}$. For the parallel BM tag configuration, we decreased this value to 31 dB, 29 dB and 28 dB for $\text{Re}(\beta) = -1.69$, $\text{Re}(\beta) = -1.75$ and $\text{Re}(\beta) = -1.78$, respectively.

Based on the approximate theoretical expression of (94, Eq. (19.20)), for 19 dB SNR gain at the reader, the up-link range increases up to 60%. It is noticeable that this range gain is achieved in expense of more complex, larger and battery-dependent circuit compared to passive tag circuit.
3.6.2 Series BM tag configuration

For the series BM tag configuration, we plotted the average BEP for different values of bidirectional amplifier gain, \( G = 8 \) dB, \( G = 11 \) dB, and \( G = 13 \) dB, and phase shifter insertion loss of 3 dB, as shown in Fig. [20]. Note that the average backscatter symbols power is derived from the values of \( GL_{ps} \) based on Eq. (3.33). We chose these values of gain based on (107), where oscillation starts after 15 dB gain for transistor-base bidirectional amplifier. The value of phase shifter insertion loss comes from Mini-Circuit phase shifters, which shows -3 dB insertion loss in average. Here the extra required power is determined by the type of bidirectional amplifier and the phase shifter. The transistor-based bidirectional amplifier in (107) consumes 2.4 mW DC power and for a 6-bit digital phase shifter from QORVO, with part number QPC2108, this number is 2 mW.

We set \( K = K_l \) to maximize the backscatter power and assume Rician fading with Rician factors \( k_{rt} = k_{tr} = 2.7 \) dB. Like the parallel configuration, we also plotted the average BEP at reader for passive BM tags. As expected by increasing the gain of bidirectional amplifier the average BEP at the reader decreases. Note that high-gain bidirectional amplifiers (generally all amplifiers) are prone to oscillate. Thus, there is a tradeoff between the gain and stability of the amplifier, which should be taken into account. Compared to passive tag, for \( G = 8 \) dB, \( G = 11 \) dB and \( G = 13 \) dB, maximum SNR gain of 15 dB, 22 dB and 25 dB could be theoretically achieved at reader for average BEP of \( 10^{-4} \). The SNR gains achieved in series BM tag configuration are larger than the SNR gains in parallel BM tag configuration. The reason lies in the BM tag configurations design. In the series tag configuration, we can set the receive power at tag demodulator to the least acceptable value \( P_{match} \), and backscatter the rest of power toward the reader. Also the backscatter signal is amplified two times by the bidirectional
amplifier, which further increases the backscatter power. Contrary to the series tag configuration, in the parallel tag configuration the received power at tag demodulator and the backscatter power increase together.

### 3.6.3 Parallel and series tag comparison

To compare the proposed parallel and series BM tag configurations, several factors should be weighted which includes but not limited to the circuit complexity, the power consumption, the size, the BEP at the reader, the received power at the tag demodulator and the feasibility of implementing the desired BM constellation map. These factors must be ranked by importance for individual applications. The comparison through all the mentioned factors between the two proposed configurations is beyond the scope of this chapter. In this section, we just compare the parallel and series BM tag configurations based on the average BEP at the reader.

In the series tag configuration, we consider the transistor-based bidirectional amplifier structure proposed in (107), which oscillates for $G > 15$ dB. To sit far from oscillation point, we consider $G=10$ dB (5 dB less gain). Also we use the average insertion loss of Mini – Circuit phase shifters, reported as -3 dB. In the parallel tag configuration, we used tunnel diode-based negative resistance parallel with the the tag demodulator as a one-port reflection amplifier. Based on (132), it starts to oscillate when its reflection gain is more than 20 dB. Again, to be in the safe side and away from oscillation, we picked 15 dB gain (5 dB less than oscillation border) of reflection amplifier which corresponds to $\text{Re}(\beta_n) = -1.7$. Here, we also add the plot for bidirectional amplifier gain of 15 dB, to compare the series and parallel configurations, when both have the same reflection gain.
The results are shown in Fig. 21, where we compared the average BEP at the reader for the parallel and series configurations. For the same oscillation-free constraint (5 dB less than oscillation border gain), the series configuration shows better performance in case of average BEP at the reader. For the same gain condition (G=15 dB and \(\text{Re}(\beta)=-1.7\)), the series configuration has maximum SNR gain of 12 dB is compared with parallel configuration at average BEP of \(10^{-4}\). The bidirectional amplifier and digital phase shifter used in the series configuration are complex circuits with higher power consumption compared with the parallel configuration elements, which is the price of better BEP performance at the reader.

3.7 Conclusion

In this section, we proposed two active two-way BM tag configurations. The first BM tag configuration, called parallel configuration, used negative resistance in parallel with the tag demodulator. The second BM tag configuration, called series configuration, used a bidirectional amplifier in series with the tag demodulator. Using AST and TEC/NEC, we accurately modeled and optimized each BM tag configuration to provide at least \(P_{\text{match}}\) at the tag demodulator while exhibiting maximum difference in the backscattered field and implementing the desired constellation map. For both BM tags, we proposed a new geometric representation of the backscatter symbols which gives the optimized tag modulator loads. We derived a closed-form expression of the average BEP at reader in Rician fading channel environment and used it to compare parallel and series configurations with a conventional passive BM tag. The simulation results allowed us to conclude that compared to a passive BM tag circuit, the proposed BM tag configurations can implement the desired constellation and considerably increase the communication range.
Figure 19: Average BEP at the reader for the parallel BM tag configuration. The lines denote the analytical results while the cross markers denote the Monte-Carlo simulations. Copyright © 2019, IEEE.
Figure 20: Average BEP at the reader for the series BM tag configuration. Phase shifter insertion loss=-3 dB.

The lines denote the analytical results while the cross markers denote the Monte-Carlo simulations. Copyright © 2019, IEEE.
Figure 21: The comparison between parallel and series tag average BEP at the reader. Phase shifter insertion loss=-3 dB. The lines denote the analytical results while the cross markers denote the Monte-Carlo simulations.

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CHAPTER 4

REFLECTION AMPLIFIER BY MEANS OF TUNNEL DIODE

Parts of this chapter have been presented in [4–6]. Copyright © 2017, IEEE.

4.1 Introduction

As explained in chapter [3], parallel configuration is one of the hardware implementations that can be used in long-range two-way BM systems, specifically ReflectFX, in this thesis. In chapter 3, parallel configuration was theoretically studied. It has been shown that by using negative resistance placed in parallel with tag demodulator, we can build a two-way BM system and improve the communication range. In this chapter, on the other hand, I design and fabricate negative resistance using tunnel diode. Negative resistances show reflection coefficient larger than 1, \(|\Gamma| > 1\), so potentially they can act as reflection amplifier. Actually, reflection amplifier is one of the most well-known applications of negative resistances. Thus, in this chapter we build negative resistance and exploit it to design and build reflection amplifier, as a more famous application. In this chapter, I propose tunnel diode-based reflection amplifier. Design, simulation and prototype measurement results are shown in this section.

4.2 Reflection amplifier by means of tunnel diode

4.2.1 Tunnel diode structure

Tunnel diode also called Esaki diode is a type of semiconductor consisting a p-n junction, that was introduced for the first time in 1957 by Leo Esaki, Yuriko Kurose and Takashi Suzuki at Tokyo Tsushin Kogyo, now known as Sony. Tunnel diode shows a non-usual relation between its current and voltage.
Figure 22: The I-V curve of AI201A tunnel diode measured by Keithley 2400. Copyright © 2017, IEEE.

(IV curve). Figure 22 shows the IV curve of a tunnel diode with part number AI201A Ga-As, where X-axis is the bias voltage applied on tunnel diode and Y-axis is the current flowing in tunnel diode due to the bias voltage. The I-V curve was measured with Keithley 2400.

This IV characteristic comes from the thin junction voltage barrier formed between two heavily-doped regions of tunnel diode. Usually, the forbidden energy gap of a p-n junction isolates the electrons on the two sides of the junction because their energy is not sufficient enough to pass this barrier. However, in tunnel diodes, this region is thin which makes it probable to electrons of the n-type region tunnel to the empty states of the valence band in the p-type region. Thus, by applying low forward voltage, tunneling effect happens and we have a increase in current, $I(V)$. As we increase the applied voltage, fewer and fewer electrons are able to tunnel through the barrier, which results in fig. 22 IV curve.
As can be seen from Fig. 22, at certain area of the IV curve (shown by dash red circle), as we increase the voltage applied to tunnel diode, the current decreases. The diode current decreases when the forward bias voltage increases from 0.1 V to 0.2 V. This behavior indicates that the equivalent differential resistance is negative.

In order to obtain an accurate model of the tunnel diode characteristic, the S-parameters of the diode has been extracted using the TRL (Through-Reflect-Line) calibration method, Fig. 23.

TRL is well-known among other calibration methods due to its accuracy. In the TRL technique, R stands for reflect, which can be both open or short. Note that only the sign of its reflection coefficient needs to be known. T stands for through which can either be a real through or a short transmission line. L stands for the line which must not be the same length as the through standard. The characteristic impedance of this line ($Z_0$) establishes the reference impedance for the measurement after calibration is
completed. The attenuation of this line need not be known, and the electrical length only needs to be specified within 1/4 wavelength. Note that line length cannot be the same length as the through. Difference between the through and lines must be between 20 degrees and 160 degrees. After having the S-parameters of T and R and L and also device under test (DUT), which is tunnel diode here, WINCAL XE software can be use to extract the DUT accurate S-parameters. For more information regarding TRL calibration technique, please visit Agilent E5070B/E5071B ENA Series RF Network Analyzers TRL/LRM calibration user manual.

4.2.2 Reflection amplifier design

By extracting the S-parameter of the tunnel diode, I designed a reflection amplifier to have reflection gain at the desired frequency band. The reflection coefficient of the tunnel diode is \( \Gamma_{TD} = \frac{(Z_{in}^{TD} - Z_0)}{(Z_{in}^{TD} + Z_0)} \) where \( TD \) denotes the tunnel diode, \( Z_{in}^{TD} \) is the input impedance of tunnel diode, and \( Z_0 \) accounts for the characteristic impedance of the connected load, which is usually 50 Ω. In the negative differential resistance (NDR) region \( Re(Z_{in}^{TD}) < 0 \) which leads to \( |\Gamma_{TD}| > 1 \). In order to achieve desired gain at desired frequency, an impedance matching circuit should be designed to convert \( Z_{in}^{TD} \) to the desired reflection amplifier input impedance \( Z_{in}^{RA} \). Active devices with input NDR are widely used in oscillator circuits. In order to avoid oscillation in tunnel diode circuit, the impedance matching circuit should be design in a way that reflection gain remains smaller than 20 dB, (132). Reflection gain can be derived as,

\[
G_{RA} = 20\log|\Gamma_{RA}| = 20\log\left|\frac{Z_{in}^{RA} - Z_0}{Z_{in}^{RA} + Z_0}\right| \tag{4.1}
\]
The circuit diagram of the tunnel diode based reflection amplifier is shown in Fig. 24. The measured input impedance of the tunnel diode at the bias voltage $V_{DC} = 0.117$ V is $Z_{in}^{TD} = -2.56 - j13 \Omega$ which is in the NDR region of the tunnel diode. This impedance is matched to $Z_{in}^{RA}$ to reach a desired reflection gain and avoiding oscillation in the circuit. The matching circuit in Fig. 24, consisting of a shunt short-ended microstrip, is designed to match the input impedance of the reflection amplifier $Z_{in}^{RA} \approx -35 - 20j \Omega$. This input impedance provides the reflection gain with a sufficient margin to prevent oscillation. The layout of the proposed reflection amplifier is shown in Fig. 25. Using a grounded co-planar waveguide (GCPW) transmission line helps to adjust the length of the stubs after fabrication. The proposed reflection amplifier has been fabricated and measured in Andrew Lab at the UIC, Fig. 25. Figure 26 shows the simulation and measurement results of the fabricated circuit. An Agilent Vector Network Analyzer PNA N5222A with source power level of -30 dBm is used to measure the reflection gain of the designed circuit. Differences between simulated and measured results are attributed to
Figure 25: Reflection amplifier layout. The dimensions in mm are: $l_1 = 7.5$, $l_2 = 5$, $l_3 = 5.6$, $l_4 = 4.5$, $l_d = 4.9$, $l_l = 3$, $l_b = 10$, $w_i = 1.8$, $w_s = 0.5$, $w_o = 1.1$, $w_d = 3.8$, $h = 0.812$ $w_{sub} = 37$ and $L_{sub} = 25$. $C_b = 33$ pF and $L_b = 40$ nH. Copyright © 2017, IEEE.

Fabrication errors as well as inaccuracies in the extracted values of the S-parameters of the tunnel diodes.

The power consumption is $178 \, \mu W$ for a 0.117 V DC bias voltage with a 1.5 mA forward current.

The reflection amplifier is also connected to an Agilent E4440A spectrum analyzer to monitor the output frequency spectrum and observe unwanted harmonics and spurs, Fig. 27. The proposed reflection amplifier is free from oscillation and the closest spur in the operation frequency band has only -85 dBm power. These low power spurs are important to avoid signal to noise ratio degradation in the reflected signal.
Figure 26: Simulated and measured return gain of the reflection amplifier. $V_{DC} = 0.117$ V, $I_{DC} = 1.5$ mA. Copyright © 2017, IEEE.

Figure 27: Reflection amplifier output spectrum. The spurious harmonics power are around -85 dBm. Copyright © 2017, IEEE.
4.3 Conclusion

A low-power reflection amplifier by means of a tunnel diode has been proposed. This reflection amplifier can be used in ReflectFX parallel configuration. Biasing the tunnel diode with DC voltage as low as 0.117 V, drives it into its NDR region. The proposed reflection amplifier is free from oscillation and the closest spur in the operation frequency band has only -85 dBm power. The measurement results show 15 dB of reflection gain for source power level of -30 dB.
CHAPTER 5

A FULL-DUPLEx BIDIRECTIONAL AMPLIFIER WITH LOW DC POWER

CONSUMPTION USING TUNNEL DIODES

Parts of this chapter have been presented in (4). Copyright © 2017, IEEE.

5.1 Introduction

In chapter 3, it has been shown that series configuration along with parallel configuration, can realize ReflectFX practically. In chapter 3, series configuration was studied theoretically and it has been shown that by using a bidirectional amplifier and phase shifter a long-range two-way BM structure can be implemented. In chapter 3 I did not go through bidirectional amplifier design, however in this chapter, I proposed a novel low-power full-duplex bidirectional amplifier to be used not only ReflectFX series configuration, but also in other applications including conventional half-duplex and full-duplex radio transceivers (133, 120, 91), retro-directive antenna arrays (121, 134), and read/write smart radio-frequency identification (RFID) tags (135). There are two types of bidirectional amplifiers: half-duplex and full-duplex (110). Half-duplex bidirectional amplifiers amplify the input signal in one direction each time. They can be implemented using a unilateral amplifier in each direction and selecting the path by RF switches (136) or transistor network configuration (133, 120, 121, 137). Full-duplex bidirectional amplifiers, on the contrary, amplify the incoming signal in both directions simultaneously. They can be realized by using two identical one-port reflection amplifiers and a BLC (134), using two identical unilateral amplifiers and two RF circulators.
or by implementing a network of amplifiers in monolithic microwave integrated circuit (MMIC) technology \((138)\). Transistor based amplifiers DC power consumption reaches up to 330 mW \((95)\) and complex design is needed. Using RF circulators to separate the signal paths also makes the circuit bulky. In this chapter, I present a low-power consumption and simple design bidirectional amplifier. It has been designed using two tunnel diodes as the reflection amplifiers, explained in chapter \(14\) which have been integrated by a BLC to form the bidirectional amplifier. The proposed bidirectional amplifier structure has been analyzed theoretically, simulated and fabricated. I achieved 9 dB transmission gain between the input/output ports with 22 dB return loss at the ports. The total power consumption is 356 \(\mu\)W which can be easily provided by ambient energy harvesting method, coin battery or a charge pump in fully passive circuits \((139)\). Table \(\text{IV}\) is a comparison between this work and similar bidirectional amplifiers.

5.2 Bidirectional amplifier design

Two identical tunnel diode base reflection amplifiers \((RA_1 \text{ and } RA_2)\) explained in the previous chapter are integrated with a four-port BLC to build a bidirectional amplifier as shown in Fig. 28. Port 1 and port 2 are RF in/output and port 3 and port 4 are connected to the reflection amplifiers. I first derived the forward gain from port 1 to port 2, defined by \(S_{21} = V_2^- / V_1^+ \bigg|_{V_2^- = 0}\). Figure 28 shows the diagram of the signals in a bidirectional amplifier. The incident signal \(V_1^+\) is divided between port 3 and port 4 with the same amplitude and a 90\(^\circ\) phase shift as \(V_3^-\) and \(V_4^-\). The waves coming to port 3 and port 4 are reflected back to the BLC by \(RA_1\) and \(RA_2\) with equal return amplitude gain \(\sqrt{G}\) as \(V_3^+\) and \(V_4^+\), respectively. Then \(V_3^+\) and \(V_4^+\) are divided between port 1 and port 2. As can be seen in Fig. 28 at port 1 the reflected signals are out of phase and cancel each other out while at port 2 they are in phase and add.
TABLE IV: COMPARISON OF THE PROPOSED BIDIRECTIONAL AMPLIFIER WITH SOME SIMILAR FULL-DUPLEx BIDIRECTIONAL AMPLIFIER.

<table>
<thead>
<tr>
<th>Work</th>
<th>Technique used</th>
<th>Power consumption</th>
<th>Gain</th>
<th>Size</th>
<th>Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>[7]</td>
<td>Two identical unilateral amplifiers + two RF circulators</td>
<td>750 mW</td>
<td>8 dB</td>
<td>λ/2.5 × λ/3.5</td>
<td>2.4 GHz</td>
</tr>
<tr>
<td>[10]</td>
<td>Network of amplifiers in MMIC technology</td>
<td>750 mW</td>
<td>27 dB</td>
<td>λ/12 × λ/16</td>
<td>10 GHz</td>
</tr>
<tr>
<td>This paper</td>
<td>Two tunnel diode-based reflection amplifiers + one BLC</td>
<td>356 μW</td>
<td>9 dB</td>
<td>(λ/4.3) × λ/7</td>
<td>0.9 GHz</td>
</tr>
</tbody>
</table>

constructively with amplitude gain $\sqrt{G}$. The backward gain from port 2 to port 1, $S_{12} = V_1^-/V_2^+|_{V_1^+=0}$, can be derived in the same manner. The bidirectional amplifier in Fig. 28 amplifies the signal coming from both port 1 and port 2 simultaneously with power gain $G$.

Figure 29 shows the bidirectional amplifier layout. Meander lines are used in the BLC branches in order to shrink the circuit size. To increase the return loss of the bidirectional amplifier, the two reflection amplifiers should be identical. Separate DC bias sources and tunable stubs may be used to achieve identical frequency responses for the reflection amplifiers. Thus, the fabricated circuit can be tuned by changing the length $l_t$ at the RA configuration, via a 33 pF capacitor connected to ground and tuning the bias voltage to optimize the in/output VSWR. Figure 30 shows the forward and backward transmission gains $(S_{21}, S_{12})$ of the proposed bidirectional amplifier for the input power of -30 dBm. The reflection
coefficient of port 1 and port 2, \((S_{11}, S_{22})\) is also shown in Fig. 30. Transmission gain of 9 dB and return loss of 22 dB at 915 MHz are measured.

In order to investigate the 1 dB compression point (P1dB) and the output power vs. the incident power, the measurements are done with a Keysight FieldFox N9916A microwave analyzer, using its minimum input power level \(P_{in,min} = -45\) dBm and a maximum input power level \(P_{in,max} = -17\) dBm. The transmission gain \(S_{21}\), phase \(\angle S_{21}\) and the reflection coefficient \(S_{11}\) of the proposed bidirectional amplifier are shown in Fig. 31. The operation bandwidth \((|\Gamma| > 0\ dB)\) is around 100 MHz and the total power consumption is \(356\ \mu\text{W}\). The bandwidth of the bidirectional amplifier mostly depends on the bandwidth of the tunnel diode, whose bandwidth is limited by its parasitic elements. Fig. 31 shows that by increasing the input power the gain decreases due to the tunnel diode whose behavior is greatly
affected by its bias point. Therefore, large input power values can cause a shift of the tunnel diode bias point. In turn, changes of the bias point can modify both the input impedance and the gain of the reflection amplifier and, consequently, the bidirectional amplifier gain.

5.3 Conclusion

A low-power reflection amplifier by means of a tunnel diode has been proposed. Biasing the tunnel diode with DC voltage as low as 0.117 V, drives it into its NDR region. Two reflection amplifiers with similar performance are integrated with a BLC to form a bidirectional amplifier. The outcome is a low-power switchless bidirectional amplifier that can be used in retro-directive arrays and RFID applications. The applied bias voltage is low enough to be extracted by a photo-voltaic cell. Note that the degree of similarity of the two reflection amplifiers has a major effect on the final performance.
Figure 30: The simulated and measured S-parameters of the proposed bidirectional amplifier. $P_{in} = -30\, dBm$,

$P_C (DC\, power\, consumed) = 356\, \mu W, \; V_{DC1&2} = 0.114\, V$. Copyright © 2017, IEEE.
Figure 31: The S-parameter of the bidirectional amplifier based on the incident power level. Copyright © 2017, IEEE.
CHAPTER 6

AN IMPLEMENTATION OF REFLECTFX WITH ON/OFF KEY (OOK)
BACKSCATTER MODULATION

In this chapter, I implemented a ReflectFX circuit which use On/Off key (OOK) modulation for the backscatter communication. I named this circuit backscatter tag (BSC). The most well-known passive BSC is radio frequency identification (RFID). The main characteristics of a conventional passive BSC tags can be highlighted as follows: i) instead of having a separate transmitter, they use backscattered signal to send their data, ii) they are used in one-way (read-only) applications rather than two-way (read/write) communication, iii) the amount of reflected power is limited, which restricts their applications to only short-range communication. By overcoming ii) and iii), we can leverage BSC in ReflectFX.

Recently, several prototypes of one-way BSC systems succeed to increase the communication range by using reflection amplifiers. Reflection amplifiers show reflection coefficient larger than one (\(|\Gamma| > 1\)) and amplify the reflected signal. They can be built using non-linear components with negative input differential resistance, such as tunnel diodes \(^4\) or using RF transistors \(^{99, 140}\). Reflection amplifiers also can be used in two-way BSC systems, where two of them are added in one branch line coupler structure to build bidirectional amplifier \(^{140, 4}\).

However, to the best of the author’s knowledge, the first and only two-way reflection amplifier which provides reflection and forward gain simultaneously, is reported in \(^{98}\). By changing the dc voltage bias
of the transistor, the authors could change the reflection gain and modulate the backscatter signal. In this case, the forward gain also varies by the bias voltage. Thus, the main drawback is that in high/low voltage regime both reflection and forward gains are high/low. It makes it difficult to maximize the distance between backscatter constellation points and maintain a fixed/high forward gain, at the same time.

In this chapter, I proposed a two-way BSC tag as one implementation of ReflectFX, with fixed forward gain and adjustable reflection gain and OOK backscatter modulation. The proposed circuit includes an RF single-port dual-terminal (SPDT) switch, a designed two-way reflection amplifier, an unilateral RF amplifier as a block gain and a band-pass filter (BPF). One switch is used to modulate the amplified backscattered signals in OOK while the fixed/high forward gain is always present. Compared to (98), I increased the communication range and decreased the bit error rate (BER) at the reader by maximizing the difference between reflection gain and maintaining a fixed/high forward gain.

The block diagram of the proposed two-way BSC circuit is shown in Fig. 32. There are two modes of operation according to the path of the input signal (RF_{IN}), which is selected by an RF SPDT switch.
In path 1, we have both reflection and forward gain, simultaneously. A portion of $\text{RF}_{\text{IN}}$ is amplified and reflected back toward the antenna, due to the reflection amplifier and the is amplified in forward direction and delivered to the RF amplifier, BPF and the receiver ($\text{RF}_{\text{OUT}}$). A band-pass filter is used to filter the output spectrum from the spurs and harmonics. This amplified signal can be received and demodulated at the receiver.

In the second mode, when path 2 is selected, we only have forward gain from the input to the output. In this mode, $\text{RF}_{\text{IN}}$ directly goes to the RF amplifier and the BPF. Therefore, the forward gain is always present in both operation modes. However, the reflection gain can be modulated between the maximum or minimum value (path 1/2), independently. The input matching circuit is designed to guarantee a stable output reflection coefficient ($|\Gamma_{\text{OUT}}| < 1$) to avoid oscillation and extra spurs and harmonic in the system. It also matches $\Gamma_{\text{IN}}$ to desired value in desired frequency bandwidth. The output matching circuit is designed to enable forward gain in both modes.

I investigated each part of the proposed circuit analytically and through simulations. I also built the circuit to validate the implementation.

6.0.1 Two-way reflection amplifier design

A reflection amplifier can be designed using a proper terminated RF transistor (140). At high microwave frequencies, the impedance and admittance parameters of the transistor cannot be directly measured. Thus, I used the scattering parameters to design the transistor-based amplifier (68). Here, I designed an unstable (but not oscillating) two-way reflection amplifier which shows reflection gain at its input port (port 1) $S_{11} \gg 0$ dB ($|\Gamma_{\text{IN}}| \gg 1$) along with forward gain from input to output port (port 1 to port 2) $S_{21} \gg 0$ dB. The output is connected to the receiver so it should be stable
Figure 33: Layout of the proposed two-way reflection amplifier.

and matched at port 2, $S_{22} << 0$ dB (\(|\Gamma_{OUT}| << 1\)). It also shows maximum output to input isolation $S_{12} << 0$ dB to avoid unwanted feedback, see Fig. 33. I used a super low noise figure and high associated gain GaAs FET transistor from Renesas with part number NE3509M04. The circuit substrate is Rogers 4003C with $\varepsilon_r = 3.55$ and $h = 0.8$ mm. The input and output matching circuits are designed to achieved the reflection and forward gain. The input matching consists of a microstrip line with length 8.5 mm and width 0.9 mm and a short-ended stub with length 4 mm and width 0.5 mm, with 0.2 mm gap. The output matching is an edge-coupled line which is open-ended at the narrower arm and terminated with a 50 $\Omega$ load at the wider arm. The length of the edge-coupled line is 16 mm. The narrow and wide arms width are 1.8 mm and 4 mm, respectively. The output matching provides the necessary $\Gamma_L$ and desired forward gain. Load and source stability circles of the transistor at 2.4 GHz including the effect of the source impedance and dc bias networks are shown in Fig. 34. The load and source stability circles are shown with red and blue lines, respectively. Small arrows show the region of stability which
are outside and inside of the load and source stability circles, respectively. To have $|\Gamma_{IN}| > 1$, $\Gamma_L$ seen at the output matching circuit, should be inside the load stability circle, $\Gamma_L = 0.95 e^{j27.5}$. The source reflection coefficient $\Gamma_S$ should be inside the source stability circle to have $|\Gamma_{OUT}| < 1$. $\Gamma_S$ is selected close to the optimum source reflection coefficient $\Gamma_{S\text{ Opt}}$, to have the noise figure (NF) close to $\text{NF}_{\text{min}}$. The NF≈0.3 dB for the selected $\Gamma_S$. The S-parameter results of the proposed two-way reflection amplifier is shown in Fig. 35. The reflection and forward gain $S_{11} = S_{21} = 11.5$ dB at 2.45 GHz. $S_{11}$ and $S_{21}$, show the same trend as we increase the transistor voltage bias, Fig. 36. At the low voltage regime ($0 \ V < V_b < 0.5 \ V$), there is no reflection and forward gain, while at high voltage regime ($V_b > 1V$), both reflection and forward gain reach to 11.5 dB. Thus, backscatter modulation via changing the bias

Figure 34: Load and source stability circles at 2.45 GHz.
Figure 35: S-parameters of the two-way reflection amplifier, $V_b = 1.5$ V.

Figure 36: S-parameters of the two-way reflection amplifier at 2.45 GHz vs. bias voltage of the transistor.

... voltage jeopardizes the forward gain. In order to implement BSC with fixed forward gain, I used an RF switch, as explained in the following section.

6.0.2 Two-way BSC circuit

I proposed a two-way BSC circuit with on/off reflection gain (OOK backscatter modulation), and a constant forward gain, Fig. [37]. In this circuit, an RF SPDT switch with part number VSWA2-63DR+ is...
controlled by voltage $V_c$ to select the reflection gain on (path 1, $V_c = 3$ V) or reflection gain off (path 2, $V_c = 0$ V). Thus, by changing switch voltage, we can implement OOK backscatter modulation. This is an absorptive switch, thus the edge-coupled line is always terminated even when the input RF is directed to reflection amplifier. The switch bias voltage is $V_{dd} = 3$ V. The short-ended stub of the input matching circuit and the length of the source impedance have distributed ground to provide adjustable length stubs for tuning after fabrication. The RF amplifier is a gain block with part number ECG001F-G from QORVO. The amplifier output spectrum is filtered by a SAW filter with part number SF2124E from Murata with 2.5 dB insertion loss. This filter can be replaced with lower insertion loss microstrip filters if there is enough space in the circuit layout. The measurement setup of the circuit in both modes are shown in Fig. 38a and 38b. The measurement and simulation results of the BSC circuit in both modes
Figure 38: (a) and (b) are the setup measurements for reflection gain on/off mode. The network analyzer is Keysight PNA-X N5242B.

are shown in Fig. 39. Fig. 39a shows a reflection gain $S_{11} = 20$ dB and a forward gain $S_{21} = 15$ dB, at the frequency bandwidth of 120 MHz, when path 1 is selected (mode on). The output return loss $S_{22}$ and port 2 to port 1 gain $S_{12}$ are below -15 dB and -20 dB, respectively. This avoids oscillation, harmonics and extra spurs, due to any potential impedance mismatch at receiver port. Fig. 39b shows that when path 2 is selected (mode off), there is no reflection gain as $S_{11} < -15$ dB and the forward gain is $S_{21} = 15$ dB. Similar to Fig. 39a $S_{22}$ and $S_{21}$ are below -12 dB and -23 dB respectively. As can be seen, the forward gain is present regardless of the operation mode and the reflection gain can be modulated with OOK modulation scheme. The input and output ports of the circuit is monitored with a spectrum analyzer. The output spectrum frequency at the ports in both operating modes is clear from
Figure 39: Measurement (M) and simulation (S) results of the proposed two-way BSC circuit in (a) reflection gain on, and (b) reflection gain off.

6.1 Conclusion

I proposed a new two-way (read/write) BSC tag that operates in two modes. There is a forward gain from the input port to the output port in both modes, but the reflection gain can be turned on or off by changing the mode. As a result, the proposed BSC tag is both writable (amplifies and delivers a portion of received signal to the tag receiver), and readable (implements OOK backscatter modulation by changing the reflection gain mode).
Part II

Analog and antenna self-interference cancellation
CHAPTER 7

ANALOG SELF-INTERFERENCE CANCELLATION

7.1 Introduction

In in-band full-duplex (IBFD) communication systems, nodes are able to transmit and receive simultaneously in the same frequency band. IBFD communications have the potential of doubling the spectrum efficiency of communication systems as well as eliminating hidden terminals and the need of duplex filters, improving fairness and network latency \(^{141,142,143,144,145,146}\). However, IBFD systems suffer from strong SI signals that are imposed on the Rx signals by the Tx signals. Thus, the major challenge that IBFD systems must confront is reducing SI signals. The amount of SI cancellation value for an efficient IBFD system, depends on the Tx signal power and the noise floor at the receiver. The SI signal level should be reduced to the same level as the receiver noise floor \(^{147,142}\). The main obstacle to cancel out SI signals lies in the fact that we do not know exactly them, we just know the base-band (digital) Tx signal. As the Tx signals go through the digital to analog converter, the up-convert mixer and power amplifier, we have harmonics, linear and nonlinear distortions, extra noise and frequency selective delays that are added to the Tx signals, making them totally different from the original Tx signals at the base-band.

Up to date, many works have proposed different SIC techniques. Recently the feasibility of a practical IBFD system has been proved \(^{147,148,149}\). SI cancellation can be accomplished in three
ways: passive cancellation, digital cancellation and analog (RF) cancellation. Passive cancellation uses electromagnetic isolation between Tx and Rx antennas. This method includes physical separation of antennas, directional isolation of antennas, absorptive shielding, polarization diversity, using band-gap structures, inductive loops, wave traps and slots on the ground plane (150; 151; 152). In digital cancellation, the base-band Tx signals are subtracted from the received signals all in digital domain (153; 154). Digital cancellation must be applied along with analog cancellation to achieve an efficient SIC. In the analog SIC, sample Tx signals with all transmitter impairments are tapped and manipulated through an estimator circuit to create a replica of the SI signals. This modified Tx sample is called cancellation signals and added to the received signals to cancel out SI signals (144).

Most of the previous techniques use multiple antennas (at least one Rx and one Tx antenna) (155; 156; 157; 158; 148; 159; 160; 161; 162; 163; 164). Using separate antennas significantly increases SI cancellation, but it has two main drawbacks: First, using multiple antennas prevents from dense integration of IBFD systems due to the required physical distance of the antennas (142). In other words, it does not satisfy the desired form-factor of most of today’s wireless communication systems, see Table 2 of (142). Second, multiple antennas configurations can be used in spatial duplexing system design rather than solving the duplexing problem in frequency or time domain as in IBFD systems (147; 141; 165). This warrants the study of single antenna IBFD systems.

Fewer works have studied single-antenna IBFD systems compared to multiple antennas configurations. To the best of our knowledge, the first single-antenna IBFD configuration was introduced in (141). It used two lumped-element circulators and two 3 dB quadrature hybrids to reduce SI. This configuration achieved up to 40 dB isolation between Tx and Rx channels over 25 MHz bandwidth in
the 900 MHz frequency band. Unfortunately, it is costly because it needs two RF circulators and the lumped-element microwave components are hard to implement at higher frequencies, due to low quality factor of inductors and capacitors. Recently, the authors of [147] used a complex analog SI cancellation circuit including multiple delay lines, tunable attenuator along with adaptive algorithm, which is not practical for small cellular devices. An electrical balance duplexer has been used to achieve high SI cancellation in single-antenna IBFD systems in [166; 142; 165; 167]. The main drawback of this technique is the large Tx insertion loss (half of the Tx power is lost). In [168] a simultaneous transmit and receive antenna (STAR) with two arm pairs (one pair as Tx and one pair as Rx antennas) in one platform has been proposed. A 37 dB isolation between Rx and Tx channels has been achieved by using two 3 dB 180° hybrids. However, the achieved isolation is limited to the proposed four-arms antenna and cannot be used in arbitrary IBFD systems with arbitrary antennas. A dual-polarized antenna has been used in [142; 169] and up to 60 dB and 47 dB isolation between Tx and Rx waves has been achieved, respectively. However, polarization diversity is considered as a method for doubling the capacity [147].

The novelty of this article is that in contrast to the approaches of the previously cited authors, we did not estimate the SI signal and subtract it from the received signal, instead we used the circulator inherent secondary SI signal, to cancel the primary SI signal leaked from the transmitter port. We modified the frequency response of the secondary SI using an adjustable IMT circuit at the antenna port. Accordingly, our contributions are the following: 1) We introduced a novel and efficient SI cancellation technique in analog domain which uses the circulator inherent SI signals. This decreases the complexity, cost and power consumption of the SIC circuits by eliminating many RF components (sampler, combiner, attenuator and phase shifter). 2) We proposed a reconfigurable IMT circuit by exploiting two varactor
diodes. This increases the system robustness to the antenna input impedance variations, component frequency response deviations and fabrication errors. 3) We fabricated a prototype of single-antenna IBFD system at 2.45 GHz. We achieved 40 dB SIC over a 65 MHz bandwidth. We also reached a 90 dB SIC for narrowband frequency systems.

The paper is organized as it follows: In Section 7.2, we describe the principle of operation of the proposed inherent SIC technique. In Section 7.3, we present how to build a robust reconfigurable IMT circuit. The simulation and measurement results of inherent SIC at 2.45 GHz are presented and discussed in Section 7.4. Section 7.5 concludes the paper.

7.2 principle of operation

7.2.1 SI signals in a single-antenna IBFD communication

Figure 40 shows the signal diagram of a single-antenna IBFD wireless communication system. A single antenna is used for transmission and reception. Tx and Rx channels are separated through an RF circulator. Port 1 of the circulator is connected to the Tx channel and port 3 is connected to the Rx channel. Practically, circulators do not provide perfect isolation between their port, thus some Tx signal leaks to the Rx channel, causing the primary inherent SI signal, illustrated by $I$ in Fig. 40. The secondary inherent SI signal is shown by $R_A$, comes from the Tx signal reflected at antenna due to the inherent impedance mismatch, added to the receiver through the circulator rotation property. Another SI signal shown by $R_C$ in Fig. 40 is due to environmental scattering.

These interference signals have different amplitude and phase, depending on the circuit frequency response and the environment parameters. However, generally $I$ and $R_A$ have the highest power because they don’t experience path-loss. As shown in Fig. 40, the desired received signal $R_x$, is combined with
major SI signals $I, R_A$ and $R_C$ at the receiver. Due to the free space path loss, Rx is significantly weaker than SI signals, makes it impossible to be extracted at the receiver. Thus, the interference signals must be effectively suppressed at the receiver.

### 7.2.2 Common Analog SI Cancellation Technique

In analog cancellation, we attempt to suppress SI signals of $I$ and $R_A$. $R_C$ is usually canceled with digital cancellation using pilot sequences and tones (142). The common analog cancellation technique is to tap the Tx signal, modify its amplitude and phase responses and then add it to the Rx channel. The modified Tx sample is called cancellation signal ($C$). The cancellation signals have the same amplitude and 180° phase shift compare to the SI signals. In other words, the cancellation signals cancel SI signals.
\( R_A \) and \( I \) out to have \( R_x + I + R_A + C = R_x \) at receiver. In the literature, different algorithms and techniques to implement this concept have been proposed and different values of achieved analog SI cancellation have been reported \cite{147,148,149}.

### 7.2.3 The proposed analog SI cancellation technique

Here we introduce a new analog SI cancellation technique. Instead of sampling from the Tx channel as in case of previous works, we use the SI signal of \( R_A \) as the cancellation signal. In other words, we take advantage of the reflected Tx signal at the antenna interface. However, the amplitude and phase of \( R_A \) must be modified to cancel \( I \) out at the receiver. To do so, we add a reconfigurable IMT circuit at the antenna interface to manipulate the amplitude and phase of \( R_A \). Figure 41 shows the new analog cancellation technique integrated in a single-antenna IBFD system. Here, the amplitude and phase of the cancellation signal is modified through the reflection coefficient seen by the IMT circuit \( \Gamma_T \). Thus, the IMT circuit is an impedance matching circuit, designed to provide the desired \( \Gamma_T \).

The cancellation signal is then added to the Rx channel using the same circulator used to separate Rx and Tx channels. Before investigating the possible configurations of IMT circuit, we need to determine the amplitude and phase of \( \Gamma_T \) to provide the proper cancellation signal.

### 7.2.4 \( \Gamma_T \) derivation

We used the signal-flow graph (SFG) of Fig. 41 illustrated in Fig. 42 to derive \( \Gamma_T \). The circulator SFG is drawn considering the reflection coefficient, insertion loss and isolation at each port. \( S_{11}, S_{22} \) and \( S_{33} \) are the return loss of port 1, 2 and 3 of the circulator, respectively. \( S_{21}, S_{32} \) and \( S_{13} \) are the insertion loss from port 1 to 2, port 2 to 3 and port 3 to 1, respectively. \( S_{12}, S_{31} \) and \( S_{23} \) are isolation between port 1 and 3, port 1 and 2 and port 2 and 3, respectively. \( Z_{in} \) is the input impedance seen by the IMT.
circuit interface. $a$ and $a'$ represent the incident and reflected Tx signals while $b$ and $b'$ are the incident and reflected Rx signals respectively. The value of $b'/a$ defines the amount of signal $a$ (Tx) leaking to the Rx port due to circulator leakage and $\Gamma_T$. The derivation of $b'/a$ is carried out using the Mason rule as follows

$$\frac{b'}{a} = \frac{S_{31}(1 - \Gamma_T S_{22}) + S_{21} \Gamma_T S_{32}}{1 - \Gamma_T S_{22}}. \quad (7.1)$$

Now the required $\Gamma_T$ can be determined by setting Eq. (7.1) equal to zero, results in

$$\Gamma_T = \frac{S_{31}}{S_{31} S_{22} - S_{21} S_{32}}. \quad (7.2)$$
From the definition of the reflection coefficient, we have

\[ Z_{in} = \frac{Z_0 + \Gamma_T Z_0}{1 - \Gamma_T}, \]  

(7.3)

where \( Z_0 \) is the input impedance (characteristic impedance) of the circulator seen at the port 2.

The IMT circuit is an impedance matching circuit which transfers the antenna input impedance to \( Z_{in} \). Since \( |\Gamma_T| \ll 1 \), transmission and reception losses are negligible.

### 7.3 Reconfigurable IMT Circuit

This section studies the IMT configuration that delivers the desired \( Z_{in} \). The value of \( Z_{in} \) is determined using Eq. (7.2) and (7.3). However, when it comes to the antenna, the following challenges appear: a) different types of antenna show different results of input impedance, b) even for a specific
antenna, its input impedance varies within its frequency bandwidth and c) the input impedance can be changed slightly due to the fabrication errors or mounting place.

Thus, an adjustable IMT design is needed to compensate for the antenna input impedance variation and also the fabrication errors and other RF components frequency response deviations. We propose a simple impedance matching circuit with two degrees of freedom as: a) a variable series capacitance and, b) a variable shunt capacitance. A matching circuit with these two degrees of freedom can match the antenna input impedance and tune potential circuit component frequency response deviations to the desired $Z_{in}$.

We used varactor diodes to achieve variable capacitance. Figure 43 shows the schematic of the proposed IMT circuit with one series and one parallel varactor diodes. In this figure, $Z_L$ represents the antenna input impedance and $C_p$ and $C_s$ are the parallel and series variable capacitance realized by
varactor diode. $V_s$ and $V_p$ are DC voltage biases of the series and parallel varactor diodes, respectively.

The varactor diode capacitance is a function of its DC bias voltage, $C_s = f(V_s)$ and $C_p = f(V_p)$. Since the input impedance of IMT circuit is a function of $C_s$ and $C_p$, we can tune $Z_{in}$ via $V_s$ and $V_p$. The inductor $L_c$ is connected to the ground to close the DC current path of the series varactor. We used this technique to build a single-antenna IBFD system at 2.45 GHz, described in the next section.

### 7.4 Single-Antenna IBFD Prototype at 2.45 GHz

#### 7.4.1 Components

This prototype consists of 3 parts: the antenna, the IMT circuit and the circulator. Each part is analyzed separately.
7.4.1.1 The antenna

We used an air filled stacked patch antenna since it shows a good matching bandwidth and it is easy to make. This type of antenna is also a successful model in the Wi-Fi transmitter market. Fig. 53 shows layout of the designed antenna at 2.45 GHz. The simulation result from HFSS and $S_{11}$ measurement of the antenna are shown in Fig. 45.

7.4.1.2 The circulator

We used a circulator with part number CR5358 from RFCI. This circulator S-parameters are extracted and substituted in Eq. (7.2) and (7.3) to derive $Z_{in}$. Figure 46 shows $Z_{in}$ in term of its real and imaginary parts ($Re(Z_{in}), Im(Z_{in})$), versus frequency.

Figure 45: Simulated and measured $S_{11}$ of the antenna shown in Fig. 53. Copyright © 2018, IEEE.
7.4.1.3 The reconfigurable IMT circuit

The principle of the reconfigurable IMT design has been explained in Section 7.3. Here, we designed the IMT circuit at 2.45 GHz. The layout of the IMT circuit is shown in Fig. 48. Varactor diodes should be chosen to have small insertion loss and proper capacitance range. We used SMV2019 varactor diode from Skyworks Inc. as the series and parallel variable capacitance. The capacitance value versus the applied reverse bias voltage is shown in Fig. 47. The varactor diode capacitance value also shows a slightly dispersive frequency response. The provided circuit model by Skyworks is used for the circuit simulation advanced design systems (ADS) from Keysight.

The proposed reconfigurable IMT circuit layout is shown in Fig. 48. The substrate is R04003C from Rogers with $\varepsilon_r = 3.55$ and $h = 0.8128$ mm (32 mil). DC bias capacitors are from Murata and
inductors are from Coilcraft. The IMT circuit simply includes microstrip transmission line with two varactor diodes in series and parallel configuration. The DC bias lines feed the varactor diodes and have been separated from the RF signal by proper RF choke. The meander line keeps the circuit low profile. To improve the simulation accuracy, electromagnetic (EM) co-simulator of ADS has been used. The circuit is designed by setting the DC bias voltages of the varactor diodes in the middle of their capacitance range for tuning purpose after fabrication. The simulated input impedance of the designed IMT circuit is shown in Fig. 46. The real and imaginary parts of the IMT circuit impedance are matched for the required $Z_{in}$ at 2.45 GHz.

### 7.4.2 Simulation and measurement results

We evaluate the final performance of the single-antenna IBFD system by measuring the achieved SI cancellation. Thus the amount of Tx signal leaked to the receiver ($S_{31}$) is measured using an Agilent
Figure 48: IMT layout connected to the circulator at 2.45 GHz. The parameter values in mm are: $l_1 = 19$, $l_2 = 3$, $l_3 = 9.5$, $l_4 = 15$, $l_5 = 12.5$ and $W = 1.8$. Copyright © 2018, IEEE.

N5222A PNA Vector Network Analyzer. However, since input the impedance of the IMT is a function of DC the voltage biases of varactor diodes, inherent SIC technique provides various SIC values for different sets of $V_p$ and $V_s$. Based on the application, we can set the DC biases to have a high amount of SI cancellation and/or a wideband spectrum SI cancellation. We also can adjust the operating frequency using $V_p$ and $V_s$. For example, for $V_p = 6.8V$ and $V_s = 2.2V$, 40 dB analog SI cancellation over 60 MHz frequency bandwidth has been achieved, as shown in Fig. 49. As discussed earlier, the input impedance of the antenna connected to the IMT circuit ($Z_{IMT}^{in}$), should match $Z_{in}$ over the desired frequency bandwidth. As shown in Fig. 46, $Z_{IMT}^{in}$ follows $Z_{in}$ at 2.45 GHz, over 60 MHz frequency bandwidth.

The inherent SIC performance is also measured for other DC voltage bias sets. Fig. 50 shows the achieved cancellation for extra three DC voltage bias sets. As shown in Fig. 50, up to 90 dB SI cancellation has been achieved by $V_s = 2 V$ and $V_p = 6.7 V$. A 65 MHz frequency bandwidth with
40 dB SIC is achieved using $V_s = 2.1 \, V$ and $V_p = 6.8 \, V$. We shifted the operation frequency from 2.45 GHz to 2.43 GHz using $V_s = 1.9 \, V$ and $V_p = 7 \, V$. The measured results of Fig. 50 shows that the final performance of the single antenna IBFD system, can be controlled with $V_p$ and $V_s$. Up to 90 dB cancellation could be achieved in narrowband applications while 60 MHz bandwidth is achieved in wideband applications. The insertion loss was measured as 0.75 dB for both the Tx to antenna path and the antenna to Rx path, which is negligible compared to other analog cancellation techniques used in previous works, as discussed in the introduction.

The measurement setup and fabricated version of the IMT circuit are shown in Fig. 51. The total area of the IMT circuit is $0.4\lambda_0 \times 0.2\lambda_0$ and the maximum DC power consumption of each varactor diode considering the maximum reverse current of 20 nA, is around 0.15 mW. It is noteworthy that the
amount of SIC achieved using this technique is enough to make sure that the Rx channel is not saturated by Tx signal. This technique should be combined with a digital cancellation stage to form a practical IBFD system.

7.5 conclusion

We presented a novel analog SI cancellation technique, named inherent SIC. This technique is used in single-antenna IBFD communication systems to cancel out the strong Tx interference imposed on Rx channel. Contrary to analog cancellation techniques used in other works, here we did not tap from Tx chain to generate the cancellation signal. We used the secondary inherent SI signal ($R_A$) to cancel out the primary inherent SI signal ($I$). This technique is implemented using an impedance mismatch...
terminal (IMT) between antenna and circulator. By using the inherent signal rotation path property of the circulator, we eliminate the RF combiner, RF attenuator and RF phase shifter. We also reduce the complexity and cost of digital attenuators and phase shifters control unit since we just need two analog voltages to control the systems. Inherent SIC circuit is simple with low profile configuration and significantly low power consumption (up to 0.2 mW).

Our design is also robust to antenna input impedance variation and can compensate for fabrication errors and component frequency response deprivations by tuning the varactor diodes DC voltage biases. We experimentally showed that our design can show flexible performance depends on the application.
As an example, we achieved 90 dB analog SI cancellation for narrowband applications and 40 dB for 65 MHz bandwidth.
CHAPTER 8

SELF-INTERFERENCE CANCELLATION TECHNIQUE FOR MICROSTRIP ANTENNA SYSTEMS

Parts of this chapter have been presented in [8]. Copyright © 2018, IEEE.

In In-Band Full-Duplex (IBFD) wireless communication systems, nodes can transmit and receive signals simultaneously in the same frequency band. The IBFD communication systems have the potential of doubling the spectrum efficiency of two-way wireless systems as well as eliminating hidden terminals and needs of duplex filters, improving the fairness and network latency [143, 144, 142]. However, a practical IBFD system suffers from strong Self-Interference (SI) signals that are superimposed on received signals by transmitted signals from adjacent Tx antennas or internal leakage. Due to propagation path loss and amplified transmitted signals, the received signals are significantly weaker than the SI signals, making it difficult to be detected at the Rx. Self-Interference Cancellation (SIC) could be realized through: i) passive cancellation or isolation (also called antenna interference cancellation technique) by preventing Tx antenna signals to leak to Rx antennas in the first place and ii) analog and digital SIC by subtracting a replica of SI signals at the Rx to cancel existing SI signals at the RF and digital stage [156, 157, 152, 142]. In practical IBFD systems, analog and digital SIC techniques are applied after the passive cancellation stage to reduce the SI signal as low as the receiver noise floor. Passive cancellation accounts for a large portion of the total SIC in existing full-duplex designs [171]. Thus, it is crucial
to decrease the complexity and requirements of analog and digital cancellation stages and prevent the analog to digital converter from saturation (147).

Several approaches have been proposed to increase the isolation between Tx and Rx antennas. These approaches include but are not limited to: a) decreasing the power of SI signals on Rx antennas. This can be realized by introducing path loss by physically separating Tx and Rx antennas, using non-aligned antennas or absorptive shields between Tx and Rx antennas (171); b) using two Tx and one Rx antennas or vice versa and strategically placing them so that the SI signals add destructively at the receiving port (172); c) modifying surface current distribution on the antennas ground plane to decouple the antenna elements (173; 174; 175); d) using electromagnetic bandgaps structures (176; 177); and, e) exploiting polarization diversity by assigning orthogonal polarizations to Tx and Rx antennas (156; 152; 142). In (156; 152), the authors combined the polarization diversity technique with an auxiliary port in the Rx antenna to achieve higher SIC. All the previously mentioned isolation techniques have some limitations: (a) requires large physical dimensions and may result in received signal loss due to misalignment of antennas or shielding; (b) is a narrow band technique and requires a very accurate geometry and power distribution among antennas. In addition, it requires large dimensions to achieve the required out-of-phase response. The method (c) degrades the radiation pattern. The method (d) usually requires periodic geometries and large dimensions, and the level of decoupling usually is not sufficient for having a practical IBFD communication system; and method (e) does not provide a large enough SIC using only a polarization diversity technique.

I propose a robust passive SIC technique for IBFD system using one Tx and two Rx double stacked rectangular patch antennas with orthogonal linear polarizations. The Tx antenna is placed between two
Rx antennas. The Rx antennas are rotated $90^\circ$ to have orthogonal polarization compared to the Tx antenna. Thus, the dominant SI signal in this configuration is the Tx antenna x-pol electric field components which are induced on the Rx antennas. The proposed technique leverages the x-pol electric field component phase distribution characteristic of rectangular microstrip antennas [178]. These components show out of phase distribution in the antenna H-plane. Thus, by placing Rx antennas at the Tx antenna’s H-plane sides, the induced SI signals on the Rx antennas have the same magnitude but $180^\circ$ phase shift with respect to each other. Therefore, they can be subtracted using a proper power combiner.

In this Chapter the proposed SIC technique provide 50 dB (60 dB) SIC over 180 MHz (120 MHz) bandwidth at 2.45 GHz. This level of SIC is achieved while the edge to edge distance of antennas is less than 11 mm ($< \lambda_0/10$).

In order to obtain a high level of SIC, the induced SI signals on the Rx antennas should perfectly match in terms of amplitude and phase. This means that any offset in the SI signals degrades the level of SIC. Thus, I proposed a phase-reconfigurable rat-race circuit by means of a varactor diode. Using this rat-race circuit, the SIC bandwidth increased to over 280 MHz (150 MHz) for at least 50 dB (60 dB) SIC, respectively. The phase-reconfigurable rat-race circuit also develops a robust and resilient system to fabrication errors and nearby scatterers, which impose unequal conditions on the SI signals. The proposed technique provides many advantages over other passive cancellation techniques by achieving a high level of SIC over a wide frequency bandwidth, realizing a robust circuit adjustable for potential phase deviation, maintaining a low-profile IBFD prototype and loosening the final performance from relying on the accurate position of antennas.
8.1 The X-pol Phase Distribution In A Rectangular Patch Antenna

In this section, the phase distribution of the x-pol electric field components of a double stacked rectangular patch antenna and its specific characteristic is investigated. I considered rectangular antennas because they are low-profile and low-cost which are commonly used in wireless systems [179]. However, using patch antennas without mutual coupling reduction techniques in IBFD systems is challenging. The interference between the Tx and Rx antennas is more severe in compact patch antennas configuration, since they share the same ground plane which increases the level of SI signal. For example, the coupling between two identical double stacked patch antenna ($S_{21}$) on an infinite ground plane is shown in Fig. 52 as a function of their distance over the free-space wavelength $s/\lambda_0$ at $f_0 = 2.45$ GHz. As expected, the mutual coupling is lower in x-pol alignment compared to the co-pol alignment. A
50 dB isolation could be achieved if the antennas are separated more than $2\lambda_0$ in the x-pol alignment. This distance prevents from having a compact and low-profile IBFD system. Therefore, in this thesis, my aim is to design a small and compact antennas geometry with high isolation over a wide frequency bandwidth.

I used a double stack rectangular patch antenna in my technique, since it shows wider operational frequency bandwidth compared to a single layer rectangular patch antenna, Fig. 53. In Fig. 54a, the $E_x$ phase distribution of this antenna is shown on the H-plane ($zx$ plane). This distribution shows a unique characteristic and it is $\pm 180^\circ$ phase shift between $E_x$ at $x > 0$ ($E_x^+$) and $E_x$ at its symmetrical point at $x < 0$ ($E_x^-$). To better highlight this phase difference, the antenna’s substrate surface, superstrate surface and top surface on the air are identified by A, B and C line indicators, respectively in Fig. 53. Then, $\angle E_x^+ - \angle E_x^-$ for these surfaces versus the distance from the antenna center is shown in Fig. 54b. The $\pm 180^\circ$ phase shift of antenna x-pol components have been maintained for various distances from the
center and is not limited to specific points. I use this characteristic of to further increase the isolation between the Tx and Rx antennas.

8.2 Antenna Systems Configuration

The proposed antenna system configuration for IBFD communication is shown in Fig. 55. In this figure, the Tx antenna co-pol components are in the \( \hat{y} \) direction, while the Rx antennas co-pol components are in \( \hat{x} \) direction. Thus, the dominant SI signals are the x-pol components of the Tx antenna which are induced on the Rx antennas, called \( E_x \). It should be noted from Fig. 55 that the Rx antennas themselves have a 180° phase shift with respect to each other due to their physical rotation. This rotation is necessary since by aligning the feed point at the same distance from the Tx antennas, the amplitude of

Figure 54: (a) Phase distribution of \( E_x \) on \( zx \)-plane for a double stacked rectangular patch antenna. (b) Phase difference of \( E_x \) on three lines that are indicated in Fig. 53 respect to the patch center. Copyright © 2018, IEEE.
Figure 55: (a) Antenna configuration; Copyright © 2018, IEEE. (b) Fabrication prototype. All dimensions in mm are: $l_c = 37.5$, $w_c = 30$, $l'_c = 37$, $w'_c = 30.75$, $l_s = 37.5$, $w_s = 32$, $l'_s = 37$, $w'_s = 31.25$, $l_{feed} = 12.5$, $l'_{feed} = 10.5$.

$E_x$ on Rx1 and on Rx2 are equal. This configuration is simulated by HFSS and the fabricated prototype is measured. The $E_x$ phase distribution of the Tx antenna is shown in Fig. 56. As can be seen, there is $180^\circ$ phase shift in the induced electric field of the Tx antenna on the Rx antennas. The coupling between Tx antenna and each Rx antennas ($S_{21}, S_{31}$) have the same value around $-25$ dB over 2.4 GHz to 2.5 GHz, Fig. 57.

In order to validate the frequency response of the Tx antenna on the Rx antennas, the measured phase difference $\Delta \varphi$ and amplitude difference $\Delta A$ of the antennas coupling are plotted in Fig. 58. The
$\triangle A < \pm 0.13$ dB and $\triangle \varphi < \pm 1.6^\circ$ over 150 MHz at center frequency 2.45 GHz. The phase difference of two coupled or SI signals in Rx antennas is zero instead of $180^\circ$ since, one $180^\circ$ phase shift of Tx antenna x-pol components and another $180^\circ$ phase shift of Rx antennas. By subtracting two SI signals $|S_{21} - S_{31}|$, there is less than $-45$ dB residual of the SI signal at the Rx port. It should be noticed, that an approximately $5^\circ$ offset caused by fabrication errors in the phase response of the SI signals prevents to perfectly eliminate the SI signals at the Rx ports.

8.2.1 Subtracting SI Signals

As mentioned earlier, the induced $E_x$ components on Rx1 and Rx2 antennas have the same phase and amplitude. Thus, I used a $180^\circ$ hybrid ring coupler known as rat-race circuit to subtract SI signals in the Rx antennas. The design procedure of a rat-race circuit can be found in (180). Using a rat-race
By considering the antennas measurement results, I designed and built a rat-race circuit at 2.45 GHz, as shown in Fig. 59. The S-parameters results of the rat-race are illustrated in Fig. 60a, where port 2 and port 3 are the inputs, and port 1 is the output. The phase and magnitude deviation of the fabricated rat-race circuit are shown in Fig. 60b. The phase shift between the rat-race outputs is set to 175° instead of 180° to compensate for the 5° phase offset in the fabricated prototype, see Fig. 58.

Antennas Rx1 and Rx2 are connected to port 2 and port 3 of the designed rat-race circuit and their combined signals show up at port 1, Fig. 61a. The simulation and measurement results of this configuration is shown in Fig. 61b. Note that the Tx antenna is connected to channel 1 and port 1 of
the rat-race is connected to channel 2 of the network analyzer. More than 50 dB and 60 dB isolation between Tx and Rx over 180 MHz and 120 MHz bandwidth is achieved, respectively.

The radiation patterns of Tx and Rx antennas are shown in Fig. 62. Both Rx and Tx antennas have their maximum directivity in boresight direction. This guarantees an IBFD wireless communication with the same transmit and receive directions. Since the center to center distance between the two Rx antennas is about 0.75λ₀, there are nulls in the zx-plane at θ = ±40°. However, they might not be visible in Fig. 62d due to high dynamic range of the radiation pattern plots. The gain of Tx and Rx antennas are 8 dBi and 9.5 dBi with 98% and 96% radiation efficiency, respectively. The Rx antennas act as a

![Figure 58: Phase and amplitude difference of E_x of antennas set in Fig. 55. Copyright © 2018, IEEE.](image)
two-antenna-array, which enhances the antenna directivity. The HPBW of Tx and Rx antennas are 50° and 35°, respectively. Here, the antennas and rat-race circuit are connected via SMA connectors. The rat-race circuit can be implemented on the antennas back side to achieve one integrated planar structure.

### 8.3 Robust SIC Technique

The level and bandwidth of achieved SIC are constrained by the variance of $E_x$ amplitude and phase on Rx1 and Rx2 antennas. The fabrication errors and nearby objects easily change the amplitude and phase of the electromagnetic waves. These variations can be compensated up to a certain level by using phase/amplitude adjustable elements in the circuit. We observed that there is a phase deviation between the two $E_x$ components induced on Rx1 and Rx2 antennas in the fabricated prototype. The amplitude difference is not a concern since Rx antennas are close to each other and the signals amplitude differences can be neglected. Therefore, in this section, I introduce a phase-reconfigurable rat-race to
compensate the phase deviation of SI signals. The phase-reconfigurable rat-race circuit should be able to
finely tune the phase response of its input signals to compensate the unforeseen phase deviation. Here, I
used a series varactor diode in one of the rat-race circuit input arms as a phase tuning element, Fig. 63. By changing the bias voltage of the varactor diode, its capacitance and consequently its phase response
can be controlled. The equivalent model of the varactor diode is illustrated in Fig. 63. I used a low-loss
varactor diode from Skyworks Solutions, Inc. with part number SMV1270. The series resistance of
this varactor diode is only 0.7 Ω. This varactor diode shows a capacitance range from 2 pF to 25 pF
for reversed bias voltages from 0 V to 20 V. I used the varactor diode SPICE model provided by the
company to model the diode in the ADS simulator. To evaluate the insertion loss (IL) and return loss
(RL) of the varactor diode and its frequency response vs. its bias voltage, I fabricated an evaluation
board to measure the diode losses, Fig. 64. I designed a calibration kit with open, short, load and a

Figure 60: (a) S-parameters of the rat-race, and (c) Phase and amplitude differences of the rat-race output ports.

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Figure 61: (a) IBFD antennas configuration connected to the rat-race circuit and S-parameters measurement setup, (b) the S-parameters measurement and simulation results. Copyright © 2018, IEEE.

through line to remove fixture effects including microstrip lines and SMA connectors. The diode S-parameters are measured for various bias voltages from 0 V to 6 V with 0.25 V steps over the frequency bandwidth from 2.3 GHz to 2.7 GHz, Fig. 64c and Fig. 64d. The RL < −17.5 dB and IL ≈ −0.15 dB on average for various bias voltages. The IL and RL for voltages close to 5 V which are relevant values for our tuning are -0.075 dB and -25 dB, respectively.

I placed the varactor diode in the shorter arm of the rat-race circuit in order to compensate the higher insertion loss associated with its longer arm. The phase response from port 1 to port 2 of the rat-race circuit ∠S_{21} for four sample frequencies as a function of the voltage bias are measured, Fig. 65. The varactor diode provides up to ±7.5° of phase adjustment which is enough for the proposed application. It should be noticed; the analog tuning of the bias voltage maximizes the likelihood of the induced SI signals.
Figure 62: Normalized radiation pattern of the proposed antenna. Tx Antenna in (a) \(yz\)-plane, (b) \(zx\)-plane and Rx antennas in (c) \(yz\)-plane and (d) in \(zx\)-plane. Red (Blue) lines are simulations (measurements) and solid (dashed) lines are co (cross)-polarization radiation.

The antennas and proposed phase-reconfigurable rat-race circuit are connected together as shown in Fig. 66a. The isolation between Tx and Rx ports \(S_{21}\) is given in Fig. 66b. A high level of SIC over a wide frequency bandwidth is achieved by tuning the varactor diode voltage bias \((V_{dc})\). The proposed technique achieved more than 50 dB SIC over 280 MHz and 60 dB over 150 MHz at \(V_{dc} = 4.8\) V. To better demonstrate the effect of varactor diode phase tuning on the antennas isolation, the \(S_{21}\) at \(V_{dc} = 0\) V (before tuning) is depicted in Fig. 66b. In this figure, the reflection coefficient seen from Tx and Rx ports, shown by \(S_{11}\) and \(S_{22}\) have been reported, respectively. These results show that the Tx and Rx ports remain matched during the SIC process.

### 8.4 Robust SIC in presence of Scatterers

In this section I examine the value of achieved SIC in presence of metallic scatterers as the potential source of phase offset. Practical antenna systems see a lot of scatterers nearby themselves including
Figure 63: (a) Phase-reconfigurable rat-race schematic with adjustable phase response. All dimensions in mm are: \( r_{fan} = 10 \), \( l_{arm} = 29 \), \( l_{arm} = 27 \) and \( \lambda_g/4 = 20 \), (b) Fabricated prototype of the novel phase reconfigurable rat-race.

their holder structures. I placed several types of scatterers to verify the SIC level in the proposed IBFD configuration. These objects are placed in front of the antennas and not exactly on the same line to better model real cases in antennas installations. I randomly selected four metallic scatterers which are: i) a cylinder with diameter \( \lambda_0 \), ii) a rectangular plane of size \( 2.4\lambda_0 \times 4.8\lambda_0 \), iii and iv) corner shape objects in two opposite directions of size \( 2.4\lambda_0 \times 4\lambda_0 \) as shown in Fig. 67. These are the shapes close to practical cases such as metallic pipe or corners of metallic cabinets or walls. The amount of SIC achieved in the presence of these scatterers were measured before and after tuning \( V_{dc} \) and the results are depicted in Fig. 68. By tuning the varactor diode bias voltage in the rat-race circuit, the level of SIC is tuned and returned to the acceptable level. For the cylinder and outside corner shape objects, more
than 10 dB SIC improvement on average is achieved over 2.4 GHz to 2.55 GHz. This improvement for
the two other cases is more than 15 dB on average over the 2.4 GHz to 2.55 GHz frequency bandwidth.
The peak-to-peak SIC improvement is as high as 40 dB.

8.5 Linearity and Power Handling Limitation

Varactor diodes are inherently non-linear RF components. This feature is required in some applica-
tions such as mixers, but it introduces some limitations in practical wireless transceivers. In this part,
these limitations are investigated. In the small-signal region, a non-linear device can be approximated
with a linear interpolation. However, in the large-signal or high-power signals, non-linearity effects
are inevitable. It should be noticed, in the receiving mode of the proposed reconfigurable rat-race, the
received signal is very small and the non-linearity effects of the varactor diode are neglectable. How-
ever, the transmitted signal leakage to the Rx antenna, which is connected to the varactor diode, may
limit the linearity region of the system. A component with non-linear impulse response produces higher
order mixing harmonics of the input signals. The third-order intermodulation distortion $\text{IMD}_3$ and its
counterpart the third-order input intercept point $\text{IIP}_3(\text{dBm}) = P_m(\text{dBm}) - 0.5\text{IMD}_3(\text{dBc})$ are figures of
merit for measuring the linearity of a device.

To evaluate the varactor diode frequency response, I simulated the diode model in ADS Harmonic
Balance Simulator. Two fundamental tones with 1 MHz offset from the center frequency at 2.45 GHz
with 15 dBm input power under bias voltage 4.8 V which is the same as the voltage to achieve the
maximum SIC level. The $\text{IMD}_3$ is 68 dBc and $\text{IIP}_3 = 48.6$ dBm for the varactor diode.

In order to validate the simulation results, I also fabricated an evaluation board and made a setup
to measure the varactor diode $\text{IMD}_3$. I used two signal generators HP 83595C and Agilent E8267C to
produce two fundamental input tones with 1 MHz offset from 2.45 GHz and a power combiner to mix the input tones. The output spectrum of signal generators is monitored to see if it is clean from harmonics and spurs, Fig. 69a. The varactor diode output spectrum is measured, and it is shown in Fig. 69b. The varactor diode $IMD_3$ is 67.5 dBc and it shows a very good agreement between measurement and simulation results. The selected diode shows a low $IMD_3$ level and the measurement results for $IIP_3$ is around 48.5 dBm, which can be considered acceptable for wireless communication applications. The other limitation is the maximum absolute power handling of the diode. For SMV1270, the maximum input power before the explosion current happens is about 38 dBm. Thus, this is the upper level of input power limitation and by considering 8 dBm margin, 30 dBm can be considered for the maximum input power with the $IMD_3 \approx 43$ dBm. Thus, the selected varactor diode can handle up to around 55 dBm of transmitted signal power since there is 25 dB isolation between the Tx and Rx antenna. This is much higher than the requested output power for wireless communication applications, which normally does not exceed more than 30 dBm.

### 8.6 Conclusion On The SIC Technique

In this chapter, I introduced a novel compact rectangular microstrip antennas configuration in IBFD system. This configuration includes one double stacked patch Tx and two double stacked patch Rx antennas. I exploited the polarization diversity technique combined with radiation characteristic of rectangular patch antennas to achieve maximum isolation between Tx and Rx ports. In this system, the SI signals from the Tx antenna to the Rx antennas are added destructively. I also kept to a minimum the distance among the antennas, which corresponds to less than $\lambda_0/10$ at 2.4 GHz. My proposed technique obtained more than 50 dB and 60 dB of SIC over 180 MHz and 120 MHz frequency bandwidth, respec-
tively. In addition, I further improved the design with a phase-reconfigurable rat-race structure to adjust
the phase offset and tune the circuit accordingly. The improved IBFD antenna system with the phase-
reconfigurable rat-race circuit provides 50 dB and 60 dB SIC over 280 MHz and 150 MHz frequency
bandwidth, respectively. Thus, the addition of the phase-reconfigurable rat-race circuit makes the final
performance less dependent upon fabrication tolerances and nearby scatterers, while still providing a
low-profile IBFD antenna.

I also verified isolation improvement in the presence of different scattering objects by using a phase-
reconfigurable rat-race circuit, integrated with IBFD antennas configuration. Finally, I would like to
emphasize that this SI cancellation technique is achieved solely through antenna design. Hence, higher
SIC level may be achieved through additional analog and digital SIC approaches.
Figure 64: (a) Fabricated prototype fixture calibration kit, (b) setup for the varactor diode IL and RL measurements. (c) and (d) are the varactor diode RL and IL for different bias voltages in the interested frequency band-width.
Figure 65: Phase response of the phase-reconfigurable rat-race for various varactor diode voltage biases and different frequencies. Solid lines are simulations by ADS and dashed lines are measurements. Copyright © 2018, IEEE.

Figure 66: (a) Measurement setup of the robust IBFD antennas configuration, (b) the measured S-parameters for $V_{dc} = 0$ V and $V_{dc} = 4.8$ V. Copyright © 2018, IEEE.
Figure 67: Robust IBFD antennas in presence of metallic objects. Copyright © 2018, IEEE.
Figure 68: Level of achieved SIC ($S_{21}$) in presence of (a) a cylinder, (b) rectangular plane, (c) corner shape in inside direction and (d) corner shape in outside direction. Dashed lines represent SIC level before and solid lines represent after tuning $V_{dc}$. Copyright © 2018, IEEE.

Figure 69: (a) Input signals and (b) output signals of the varactor diode $IMD_3$ measurement at 4.8 V bias voltage.
Appendices
Appendix A

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Author: Seiran Khaledian

Publication: IEEE Microwave and Wireless Components Letters

Publisher: IEEE

Date: Dec. 2017

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2018 IEEE 19th Wireless and Microwave Technology Conference (WAMICON)

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Appendix B

AUTHOR’S BIOGRAPHY

I am a women antenna/RF/microwave Engineer and I dedicate my time and study on these fields since the first year of my master. It is not common for women to study electrical engineer, let alone antenna and microwave engineering. However, I believe that you can succeed in whatever you are passionate about, no matter what. First I start learning about antenna and how they are converting electrical current to electromagnetic wave and send them out in air. I learned different types of antenna. Then I start to design them using software available like high-frequency structure simulator (HFSS) and CST MICROWAVE STUDIO. My first project was a microstrip antenna. Then I learned how to make them dual band or reconfigurable in frequency.

Next step for me was the circuit behind the antenna. So I start learning RF circuit. I learn the software first which for me was Advanced Design System (ADS). I designed a receiver circuit using ADS. It was very important for me to understand how different parts of a receiver from the base band to the antenna work. Thus, I combined the antenna design with the RF circuit design using both HFSS and ADS. In this case, I was able to see the effect each part can have on the final performance of the receiver.

I was thirsty to learn more about my major, so I decided to study PhD. I joined professor Smida team at the University of Illinois at Chicago (UIC) in 2015. Since then I had the opportunity to learn a lot from my adviser, my lab mates and other professors at UIC electrical engineering department. I start working on full duplex communication systems and all the challenges facing it to become a practical communication system. I quickly saw the potential behind full duplex system and all the ideas that
Appendix B (Continued)

could help it to overcome its barriers. I proposed two ideas regarding implementing full duplex systems.
I built my ideas and published the results in the most prestigious journal in electrical engineerin field, Institute of Electrical and Electronics Engineers (IEEE). I finished my PhD study in 2019. I learned alot during my research in UIC and spend plenty wonderful time with my mentors and friends and I would like to thank them all.

Seiran Khaledian,
CITED LITERATURE


89. Mongan, W. M., Rasheed, I., Ved, K., Vora, S., Dandekar, K., Dion, G., Kurzweg, T., and Fontecchio, A.: On the use of radio frequency identification for continuous biomedical monitor-


VITA

SEIRAN KHALEDIAN

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<tr>
<th>Education</th>
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<tr>
<td>Ph.D. Electrical and Computer Engineering</td>
<td>2015 – 2019</td>
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<tr>
<td>University of Illinois at Chicago</td>
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<tr>
<td>M.S. Electrical Engineering</td>
<td>2010 – 2012</td>
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<tr>
<td>Tarbiat Modares University</td>
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<tr>
<td>B.S. Electrical Engineering (Electronics)</td>
<td>2006 – 2010</td>
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<td>K. N. Toosi University of Technology</td>
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<th>Publications</th>
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Conference

F. Farzami; S. Khaledian; B. Smida; D. Erricolo, ”Ultra-Low Power Reflection Amplifier Using Tunnel Diode for RFID Applications,” 2017 IEEE AP-S Symposium on Antennas and Propagation and USNC-URSI NRSM, San Diego, CA, USA.

S. Khaledian, F. Farzami, B. Smida and D. Erricolo, ”A Power-efficient Implementation of In-Band Full-Duplex Communication System (ReflectFX),” 2016 ISIVC, Tunis, Tunisia.

F. Farzami, S. Khaledian, B. Smida, D. Erricolo, ”Tunable SIW Cavity Backed Active Antenna With Circular Polarization,” 2017 USNC-URSI NRSM, Boulder, CO, USA.

S. Khaledian, F. Farzami, B. Smida, D. Erricolo, ”Enhancement of Backscatter Tags Efficiency by Means of Low-Power Transistor-based Reflection Amplifier and QPSK Modulator,” 2017 USNC-URSI NRSM, Boulder, CO, USA.

Awards

scholarships from the Chicago Chapter of the IEEE Electromagnetic Compatibility Society

Memberships

IEEE Student Member (09)

Services

Journal Article Referee at IEEE

Internship

Energous Co., San Jose, CA, USA.
Antenna/Microwave Engineer.