Efficient Microwave Energy Harvesting Technology and its Applications

Dissertation

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Abstract

Using wireless transmitters to create an ocean of radio frequency (RF) energy and to power remote devices have been a dream since Nikola Tesla invented wireless communications. Recently, wireless power technology has been embedded into a plethora of consumer electronics to improve device reliability and extend battery life. Soon enough, miniature wireless sensors adapting microwave energy harvesting modules for power will be located at spots where it would be otherwise inconvenient, such as, to change their batteries. These spots include locations inside human body, within the steel or concrete of buildings, and in the dangerous innards of chemical plants. However today, even the most robust nodes can be counted on to last only a few years due to the existence of a battery. Ideally, engineers need a sensor that can last forever without external power sources or battery changes. According to research presented in this dissertation that dream is now within reach.

While solar energy harvesting has been widely used for years to power remote devices, several other types of energy-harvesting approaches have also emerged for micropower applications including vibration, thermal, mechanical, and RF. Of these technologies, RF energy is the only one that can provide either intentional or ambient power source for batteryless applications. Thanks to mobile and Wi-Fi networks, ambient RF energy is ubiquitous. Note that more than a million smartphones phones are activated every day, representing a large source of transmitters for RF energy
harvesting. Moreover, when more power or more predictable energy is needed than what is available from ambient sources, RF energy can be broadcasted in unlicensed frequency bands.

Of particular interest is the efficient harvesting of low power RF signals, which would possibly mean range improvements for dedicated microwave power transmission and considerably more DC power from ambient RF energy harvesting. Hence, in this dissertation, we propose a new class of microwave energy harvesting system which exhibits substantially improved conversion efficiency than the ones available off-the-shelf or in literature. The enhancement is due to novel rectifier circuits that can approach theoretical efficiency bounds by handling the best components on the market.

Although the developed energy harvester offers game-changing efficiency performance, maximum power supplied by a single module is usually not enough to energize most consumer electronics. Accordingly, a mathematical tool is presented to predict the optimal way of interconnecting multiple microwave energy harvesters. Subsequently, an energy harvesting array with nine elements is designed to power up a commercial thermometer and its LCD display using nothing more than ambient Wi-Fi signals in an office environment. In the end, the operational bandwidth of the designed ambient energy harvester is widened to include all cell phone bands and Wi-Fi to harvest more RF power from the environment.
Dedicated to my parents and my lovely sister...
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# Table of Contents

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Abstract</td>
<td>ii</td>
</tr>
<tr>
<td>Dedication</td>
<td>iv</td>
</tr>
<tr>
<td>Acknowledgments</td>
<td>v</td>
</tr>
<tr>
<td>Vita</td>
<td>vi</td>
</tr>
<tr>
<td>List of Tables</td>
<td>xi</td>
</tr>
<tr>
<td>List of Figures</td>
<td>xii</td>
</tr>
<tr>
<td>1. Introduction</td>
<td>1</td>
</tr>
<tr>
<td>2. Fundamentals of Microwave Energy Harvesting</td>
<td>11</td>
</tr>
<tr>
<td>2.1 Electrical Characteristics and Physics of Diodes</td>
<td>13</td>
</tr>
<tr>
<td>2.2 Schottky Detectors as Rectifier Circuits</td>
<td>18</td>
</tr>
<tr>
<td>2.3 Voltage Doubler as Rectifier Circuits</td>
<td>26</td>
</tr>
<tr>
<td>2.4 Dickson Charge Pump</td>
<td>29</td>
</tr>
<tr>
<td>2.5 Conclusion</td>
<td>32</td>
</tr>
<tr>
<td>3. Novel High Efficiency Rectifier Circuits for Microwave Energy Harvesting Applications</td>
<td>34</td>
</tr>
<tr>
<td>3.1 An Improved Voltage Doubler</td>
<td>38</td>
</tr>
<tr>
<td>3.2 Modified Greinacher Rectifier</td>
<td>41</td>
</tr>
<tr>
<td>3.3 A Planar Rectenna with a Modified Greinacher Rectifier</td>
<td>46</td>
</tr>
<tr>
<td>3.4 Conclusion</td>
<td>48</td>
</tr>
</tbody>
</table>
4. Investigation of Rectenna Array Configurations for Enhanced RF Power Harvesting ............................................ 50
  4.1 Analytical Approach ............................................. 52
    4.1.1 Outdoor Propagation .................................... 53
    4.1.2 Indoor Propagation ..................................... 54
    4.1.3 RF Combiner ........................................... 56
    4.1.4 DC Combiner .......................................... 57
    4.1.5 Rectenna Topology Indicator .......................... 57
  4.2 Rectenna Design Example ..................................... 59
    4.2.1 Rectifier Design .................................... 59
    4.2.2 Antenna Design ..................................... 62
  4.3 Rectenna Array Configurations .............................. 64
    4.3.1 Indoor Evaluation ................................... 64
    4.3.2 Outdoor Evaluation .................................. 68
  4.4 Conclusion .................................................. 70

5. Design of an Efficient Ambient WiFi Energy Harvesting System .................. 72
  5.1 Assessment of Ambient RF Signal Strength of WLAN ............. 74
  5.2 Integrated Rectenna Design ................................ 76
    5.2.1 Antenna Element Design ............................. 78
    5.2.2 Rectifier Circuit Design ............................ 80
    5.2.3 Array Design ....................................... 83
    5.2.4 Energy Storage and Management ...................... 86
  5.3 Ambient WiFi Energy Harvester ................................ 87
  5.4 Conclusion .................................................. 90

6. Development of a Novel Multi-band Ambient Microwave Energy Harvesting Module ............................ 91
  6.1 Ambient RF Power Density Characterization .................. 93
  6.2 Integrated Rectenna Design ................................ 95
    6.2.1 Rectifier Circuit Design ............................ 96
    6.2.2 Antenna Element Design ............................. 100
  6.3 Multi-band Ambient Microwave Energy Harvester ............ 104
  6.4 Conclusion .................................................. 105

7. Conclusion and Future Work .................................... 107

Bibliography ..................................................... 114
## List of Tables

<table>
<thead>
<tr>
<th>Table</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1 Parameters of the equation relating the input voltage to the output voltage of a Schottky detector. See also Fig 2.4(b).</td>
<td>20</td>
</tr>
<tr>
<td>4.1 Typical values for path loss exponents in different areas.</td>
<td>55</td>
</tr>
<tr>
<td>4.2 Power loss coefficient values, N, for the ITU Site-General indoor propagation model.</td>
<td>56</td>
</tr>
<tr>
<td>4.3 Floor penetration loss factor, $L_f(n)$, for the ITU Site-General indoor propagation model.</td>
<td>56</td>
</tr>
<tr>
<td>5.1 Performance ratings of the proposed microwave energy harvester.</td>
<td>88</td>
</tr>
<tr>
<td>6.1 Detailed dimensions of radiating element. All dimensions are given in millimeters.</td>
<td>102</td>
</tr>
<tr>
<td>6.2 Performance ratings of the proposed microwave energy harvester.</td>
<td>104</td>
</tr>
</tbody>
</table>
# List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.1</td>
<td>Simple half-wave rectifier (a) and microwave model of the simple rectifier with impedance matching at DC, the fundamental, and harmonics (b).</td>
<td>12</td>
</tr>
<tr>
<td>2.2</td>
<td>Cross-section view of a typical beamlead Schottky diode (a) and its equivalent circuit (b).</td>
<td>14</td>
</tr>
<tr>
<td>2.3</td>
<td>Measured effect of junction capacitance and series resistance on conversion loss, evaluated at 2 GHz [36].</td>
<td>17</td>
</tr>
<tr>
<td>2.4</td>
<td>Schottky detector circuit (a) and its simplified equivalent circuit (b) [39].</td>
<td>19</td>
</tr>
<tr>
<td>2.5</td>
<td>Measured and calculated output voltage (a) and RF-to-DC conversion efficiency (b) of a simple Schottky detector circuit.</td>
<td>23</td>
</tr>
<tr>
<td>2.6</td>
<td>Simulated and calculated output voltage (a) and RF-to-DC conversion efficiency (b) of a simple Schottky detector circuit.</td>
<td>25</td>
</tr>
<tr>
<td>2.7</td>
<td>Schematic of the voltage doubler circuit and its operation (a) and its evolution into a simple Schottky detector (b).</td>
<td>28</td>
</tr>
<tr>
<td>2.8</td>
<td>Schematic of a two-stage Dickson charge pump rectifier (a) and a sample design and its measured conversion efficiency (η) (b).</td>
<td>30</td>
</tr>
<tr>
<td>3.1</td>
<td>General relationship between microwave to DC power conversion efficiency and input power [51].</td>
<td>35</td>
</tr>
<tr>
<td>3.2</td>
<td>Schematic of a traditional voltage doubler (a) and the proposed efficient harvester (b).</td>
<td>39</td>
</tr>
</tbody>
</table>
3.3 Simulated (a) and measured (b) power harmonics at the output of the rectifier diode (D2) of the voltage doubler circuit. .......................... 39

3.4 Simulated power harmonics at the output of the harmonics harvester circuit (a) and measured conversion efficiency of the traditional and proposed rectifiers (b). ........................................... 40

3.5 Schematic of proposed single-stage modified Greinacher rectifier (a) and a two-stage Dickson charge pump (b). ........................................... 43

3.6 Measured conversion efficiency comparison between two different Dickson charge pumps and the proposed single-stage Greinacher rectifier at 2.45 GHz. ........................................... 44

3.7 Schematic of the second stage full-wave modified Greinacher rectifier. 45

3.8 Simulated and measured return loss (a) and realized gain (b) performance of the antenna element. ........................................... 47

3.9 PCB layout and the photograph of the fabricated version of the proposed rectenna (a) and the measurement setup used to evaluate the rectenna performance (b). ........................................... 47

4.1 Schematics of the investigated rectenna array configurations. $P_{RF}^i$ is the incident RF power impinging on the antennas, $P_{RF}^H$ refers to the harvested DC power by RF-combiner topology, and $P_{DC}^H$ denotes the harvested DC power by the DC-combiner topology. ........................................... 51

4.2 Layouts of the rectifier prototypes, printed on RO3206. $w_1 = 72$ mil, $w_2 = 15$ mil, $w_3 = 196$ mil, and $L_1 = 171$ mil. Fabricated samples are shown in the inset pictures.
* represents the shorting vias, ··· marks the location of zero-bias diodes (HSMS-2852), — marks the DC load, and --- marks the capacitors. ........................................... 60

4.3 Measured $\eta$ for the two presented Greinacher rectifiers, operating at GSM-1900 and ISM-2450. ........................................... 61

4.4 Geometry of a probe-fed shorted patch antenna for dual-band operation. The dimensions given in the figure are in millimeters. .................. 63
4.5 Measured return loss (a) and boresight realized gain of the probe-fed shorted patch antenna are shown in Figure 4.4. ................. 63
4.6 Top view of the fabricated 2×2 dual-band antenna array. ............ 65
4.7 Calculated and measured heatmap of the harvested voltage by the investigated rectenna array configurations in an office environment. 66
4.8 Normalized measured and calculated DC power (dBm - normalized) harvested by the two different rectenna topologies (a) and the outdoor environment (b). ................................................. 68
4.9 Measured and calculated RTI vs. θ (dB scale). Red line depicts the equilibrium line ($P_R^H = P_D^H$). .......................... 70
5.1 Typical office environment for RF energy harvesting. ............... 73
5.2 Ambient RF signal strength measured on one of the desks in the office environment depicted in Fig. 5.1(a) with a standard monopole antenna. 75
5.3 Block diagram of a rectenna array for ambient energy harvesting. Each element in the array is integrated with its own rectifier. The resulting DC outputs are combined and fed to power management electronics. 77
5.4 Antenna structure, its input impedance, and its realized gain performance at boresight. .................................................. 79
5.5 Layout (a) and $S_{11}$ (b) of the fabricated rectifier prototype. Substrate: Rogers RO3206, Diodes: SMS7630, Capacitors: 100 nF. ●: shorting vias, ○: location of the antenna excitation, —: zero-bias diodes, and ---: matching network. ............... 81
5.6 Simulated conversion efficiency of the proposed rectifier (a) and the measured input RF power vs. harvested DC voltage for various power harvesters (b). A picture of the proposed power harvester is given on top left in (b). P2110 performance is evaluated by connecting the load resistor in parallel to its supercapacitor terminal. .................. 82
5.7 Photographs of the fabricated RF power harvester. ................. 83
5.8 The ambient WiFi energy harvester stack-up. --- depicts the ground plane shared by the antenna and rectifier. ................................. 84

5.9 Circuit diagram of the power management unit [84-85]. A picture of the fabricated prototype (size: 1cm×1cm×1.27mm) is also given. . . 87

5.10 Proposed RF energy harvester powers a thermometer (including display) with harvested ambient WiFi power. ................................. 89

6.1 Measured ambient RF power density vs. time (from COST-281) [92] . 94

6.2 Simulated range of input impedances for two SMS7630 Schottky diodes (different packaging) as the operating frequency (0.75 GHz ≤ f_{inc} ≤ 3 GHz) and input power (-30 dBm ≤ P_{in} ≤ 10 dBm) are varied (Smith chart normalized to 50Ω). ................................. 97

6.3 Layout (a) and |S_{11}| (b) of the fabricated rectifier prototype. Diodes: SMS7630-SC70, C1: 56pF, C2: 100pF C3: 100nF. w_{1} = 1.83 mm, w_{2} = 0.30 mm, L_{1} = 5 mm, and A_{1} = 60° .......... 99

6.4 Simulated RF-to-DC conversion efficiency (\eta) of the proposed rectenna versus input RF power (-40 dBm ≤ P_{in} ≤ -20 dBm) and frequency (0.75 GHz ≤ f_{inc} ≤ 3 GHz). ................................. 99

6.5 Top view of the utilized planar monopole antenna (a), its measured impedance matching performance (b), and its simulated radiation efficiency (c). ................................. 101

6.6 Simulated 3D radiation patterns of the utilized planar monopole antenna at GSM900, DCS, and WiFi bands. The orientation of the antenna is also given for reference. ................................. 103

6.7 A photograph of the developed multi-band ambient RF energy harvesting module. ................................. 105
Chapter 1: Introduction

Mobile devices have revolutionized and become an inseparable portion of our lives. Thanks to the steady advances in semiconductor technology, energy required to achieve the same computing power with these devices is swiftly decreasing [1, 2]. Indeed, similar to Moore’s Law, which defines the trend of digital technology to double in transistor count every two years, an inverse trend occurs for power consumption per transistor. Roughly every 18 months, the power dissipated by transistors used in mobile systems is cut in half [3, 4]. These advancements in power efficiency have led to dramatic results for small, ultra-low-power microcontrollers (MCUs) and have paved the way for designs where battery life can exceed 10 years [5]. Batteries, such as lithium-ion cells, continue to be the default source for energy in most portable electronics as they have been for decades.

However, traditional batteries place hard restrictions on product usability, lifetime, and cost of ownership. While processing power roughly doubles every two years, battery technology advances at a much slower pace. Historically, battery capacity per unit volume has doubled every 10 years [6]. In addition to the slow growth in their energy capacity, traditional batteries have a limit to the total realizable energy density they can provide. Research has shown that it’s possible to increase energy density by tenfold within a few years [6]; however, even if this is achieved,
we must still consider practical safety concerns. Given improper use, batteries with extremely high energy densities can become dangerous, explosive devices [7].

For most battery-operated devices, the cost associated with owning and operating the device is rarely limited to the initial cost of manufacturing. In the long term, replacing the battery can have a significant impact on the overall cost of ownership. This is especially critical in applications where battery replacement is expensive or simply impractical. Take water meters that must be buried underground as an example. Accessing the water meter would require digging, which in colder climates may exceed four feet, and would cost $100 to $200 per water meter to replace the battery [8].

It is important to notice that the miniaturization of products has been an ongoing trend in most applications, though the major push has come from consumer electronics and medical devices. For consumer products, the demand for smaller and sleeker devices has driven innovation for more highly integrated electronics. While integration at the IC level has kept up with consumer demand, the power source has not seen the same level of miniaturization [9]. The allotted space for batteries is shrinking, expected lifetime is longer, and expected amount of power has increased. Despite these challenges, it’s still possible to maintain functionality with today’s rechargeable batteries.

Fortunately, recent technical developments have increased the efficiency of energy harvesting modules in converting trace amounts of energy from the environment into electricity [10]. In principle, energy harvesting has existed for thousands of years. The first waterwheels have been dated back as far as the fourth century BCE. The waterwheel effectively harvested the energy from flowing water and transferred it
to mechanical energy. Similarly, present-day wind farms or solar arrays all use the same principle of operation and usually provide power back to the main grid. These large-scale applications can be referred to as macro energy harvesting. On the other hand, micro energy harvesting, which this dissertation will be focusing on, is the principle that enables small, autonomous devices to capture and store energy from the environment. Although micro and macro energy harvesting are similar in principle, they differ in scope and applications.

Developments in micro energy harvesting and advancements in power efficiency have sparked interest in the engineering community to create more applications that utilize energy harvesting for power [11–13]. The idea is by harvesting miniscule amounts of wasted energy from the environment, low power systems can have near infinite up-time without a battery as their primary power source. As such, not only does energy harvesting enhance current applications by eliminating their dependency on the battery, but it also enables entirely new applications that weren’t feasible given the finite lifetime and size of batteries.

The sources of energy harvesting are numerous and more obscure systems continue to be introduced. Traditional energy harvesting methods include solar, piezo, and thermal. However, these sources share a common limitation of being reliant on sources far beyond their control [14]. Solar requires light, vibration requires motion, and thermal requires heat flow. On the other hand, a wireless power solution based on radio frequency (RF) energy harvesting overcomes this lack of control because power can be replenished (made dedicated) whenever it is desired.

Since 1893, when Nikola Tesla first proposed the concept of wireless power transmission [15], the idea of transmitting energy through air has excited public interest.
Today, RF energy is currently broadcasted from billions of radio transmitters around the world, including mobile telephones, handheld radios, mobile base stations, and television/radio broadcast stations. The ability to harvest RF energy, from ambient and/or dedicated sources, enables continuous charging of low-power devices and will possibly eliminate the need for a battery altogether. In both cases, these devices can be free of connectors, cables, and battery access panels with significant mobile freedom while charging and in use. The concept is particularly attractive given today’s ever-expanding universe of mobile devices and the constant risk of experiencing a dead battery in smartphones, tablets, or other portable data devices.

In an RF-based wireless powering system, a single transmitter or a network of them transmit power to multiple mobile devices, or nodes. Rectifying antennas (rectifier+antenna=rectenna) embedded inside these devices receive RF energy from the transmitters broadcasting radio waves at different frequency bands (controlled by the regulatory branches of governments). The receivers then convert the RF energy into direct current (DC) to power the mobile devices wirelessly. The roots of this idea can also be traced back to radio frequency identification (RFID) and World War II [16]. In recent years rapid development of RFID technology has resulted in a wide variety of applications and pushed the research efforts on wireless powering further. Currently, many RFID tags receive all of their operating power from an RFID reader and are not limited by battery life [16].

RF power transfer based on magnetic resonance is hailed as one of the most promising wireless power technologies given their publicity [17]. This technology is based on non-radiating magnetic fields generated by resonating coils. Researchers at
Massachusetts Institute of Technology (MIT) have demonstrated 40% power transfer efficiency at 2 meters distance based on this technology [18]. In addition, they were able to continuously power a 45 inch flat panel television with wireless power. Although this technology has great potential, it’s intrinsically limited to extremely short distances because near-field, non-radiating magnetic power density attenuates at a rate proportional to the inverse of the sixth power of distance [19]. To put this in perspective, a magnetic resonance receive coil that achieves 40% efficiency at 2 meters distance has to be 244 times larger in surface area to achieve the same 40% efficiency at 5 meters distance, which is neither practical nor compact\(^1\).

On the other hand, early experimental studies have already demonstrated that dedicated microwave power transfer could be used to deliver several hundred Watts of power over great distances (e.g., remote powering of a helicopter for 10 hours [20]). The NASA Jet Propulsion Laboratory (JPL) has demonstrated a long distance wireless power transmission with a microwave beam over a distance of 1 mile with a DC output of 30 kW (Goldstone demonstration) [20]. Recently, NASA and The National Research Council (NRC) have started investigating the feasibility of microwave power transmission as a large-scale, green energy source for the terrestrial markets [21]. The idea (invented in 1968 by Peter Glaser) is to place a large solar power satellite in geostationary Earth orbit, collect sunlight, use it to generate an electromagnetic beam, and then transmit the energy to the Earth [20].

Similarly, several experimental studies (e.g., Nokia Morph [22]) have been conducted to demonstrate the possibility of ambient microwave energy harvesting. The

\(^{1}\text{Retaining the same efficiency means that the same power is delivered to the receiver at 5 m. Received power is equal to receiver aperture size } \times \text{ power density and it should remain the same. Power density changes from } 1/2^6 \text{ to } 1/5^6. \text{ This implies that the surface area for the receiver has to be } 244 = (1/2^6)/(1/5^6) \text{ times larger to receive the same power at 5 m.}
obvious appeal of harvesting ambient over dedicated is the utilization of “free” energy generated by radio transmitters, such as those for mobile base stations and handsets. The number of RF transmitters will continue to increase in conjunction with the number of mobile subscriptions [23]. Of particular interest is the recent momentum in the broadband mobile subscriptions, which has already reached 1 billion worldwide [24]. Mobile phones represent a large source of transmitters from which to harvest RF energy, and will potentially enable users to provide on-demand power for a variety of close range sensing applications. Additionally, consider the number of Wi-Fi routers and wireless end devices such as laptops and smartphones. Already, in some urban environments, one can detect more than ten Wi-Fi transmitters from a single location [25]. This ubiquitous ambient microwave energy can be used to charge or operate a wide range of low-power devices, including RFID tracking tags, wearable medical sensors, and consumer electronics.

It should be noted that at high power densities, such as those at or near the Federal Communications Commission (FCC) exposure limits, microwave harvesting systems can be relatively efficient, with some studies demonstrating conversion rates of up to 80-90% [26]. However, as power density drops, so does conversion efficiency. Simply walking a few meters further away from an RF source may drop conversion efficiency from 60% to less than five percent [27], depending on the frequency of operation. This is in addition to the drop in power density caused by the increased distance. Hence, early studies failed to develop operational microwave energy harvesters at very low RF power densities (i.e., ≤-20 dBm per cm²). However, efficient harvesting of low power RF signals would mean significant range improvements for dedicated microwave power transmission and considerably more DC power from ambient RF
energy harvesting. This dissertation is therefore focused on developing ultra-low RF power sensitive, high efficiency microwave energy harvesters and their applications in consumer electronics. Specifically the key contributions are:

1. Improved existing analytical bounds on efficiency of microwave energy harvesting and proposed recommendations to the semiconductor industry for optimal component design. Thus, we are able to pinpoint the contribution of diode parameters, component imperfections, and overall circuit design to harvesting efficiency.

2. Developed a new rectifier circuit that is more efficient than conventional designs when the available RF power density is low. Hence, ambient microwave power harvesting becomes more attractive and there is a significant increase in the maximum achievable transmission range for dedicated wireless power transmission systems.

3. Presented a robust investigation of rectenna (rectifier+antenna) array configurations to maximize the harvested DC power and verified the analysis with measurements in various environments. Consequently, we can predict which rectenna array topology will harvest the most DC energy in given conditions and design the harvesting system accordingly.

4. Designed and fabricated a novel, miniature ambient Wi-Fi (2450 MHz) energy harvesting system that powers a desktop thermometer (with its display). The presented harvesting module proves that sufficient energy can be extracted from ubiquitous ambient microwave signals to expel batteries from some consumer electronics altogether.
5. Developed a unique multi-band microwave energy harvesting module that can generate DC power from RF signals propagating at frequencies between 850 and 2500 MHz, thus increasing the power output, expanding mobility options, and simplifying installation.

The dissertation is organized as follows:

Chapter 2 presents a thorough background of the microwave power harvesting process. The goal of this chapter is to give readers an understanding of rectification and its limitations, mathematical tools for analyzing rectifier performance, and application examples in the state-of-the-art RF energy harvesters. Thus, Chapter 2 starts with an algebraic analysis of the simplest rectifier circuit, i.e., diode detector, using well-known diode models. A closed-form solution for RF to DC conversion efficiency is presented without making any simplifications to the nonlinear functions stemming from the diode. Next, we move into analyzing voltage doublers and traditional rectifier circuits (such as Villiard and Dickson) and discuss their limitations.

Chapter 3 begins by recognizing the fact that conventional rectifier designs not only generate DC, but also high order harmonics of the fundamental frequency. In traditional designs, the energy stored in the high order harmonics is filtered out by the circuitry and wasted. This chapter first presents a more efficient rectifier design based on harvesting energy from higher order harmonics generated by the diodes. Next, identifying the complexity associated with extra components, the conventional Greinacher rectifier is modified such that odd harmonics are not generated. Further, measurement results are shown to depict the performance improvements of the harmonic harvester and modified Greinacher rectifier over conventional designs. The
chapter closes with a simple single rectenna design that incorporates the modified Greinacher rectifier.

Typically, a single rectenna is not sufficient in supplying DC energy for reliable device operation. Alternatively, properly interconnecting several rectennas could provide microwave power for sufficient rectification. In Chapter 4, different rectenna array configurations are investigated and their advantages and disadvantages for enhanced microwave power harvesting are discussed. Chapter 4 begins with a presentation of a novel analytical approach that can evaluate the power harvesting performance of a given rectenna topology in different propagation environments (with help from commercial numerical solvers). Next, to confirm the validity of our approach, a miniaturized rectenna array that incorporates the proposed modified Greinacher rectifier is presented. In the end, the rectenna’s performance in indoor and outdoor settings are evaluated through measurements and results are compared to analytical predictions (for a 2×2 antenna array).

Chapter 5 introduces a novel, single-band, compact, and efficient rectenna array design (using the results of Chapter 4). The developed design harvests DC energy from very low level ambient microwave signals within an office to power a sensor system and its display. The ubiquity of Wi-Fi and the fact that it operates in the crowded 2450 MHz band (the same as Bluetooth, ZigBee, RFID, cordless phone, etc.) makes it the perfect candidate for single-band ambient microwave energy harvesting. Hence, this chapter first characterizes the ambient Wi-Fi signal strength in an ordinary office environment. It has been found that the ambient signal strength varies with time, reaching up to -15 dBm (with a 4 dBi antenna) during peak hours and decreasing down to -40 dBm during the night. To scavenge energy from this ultra-low
ambient power, efficient rectenna elements that integrate modified Greinacher rectifiers and small antennas are designed. Next, a low loss power management system is presented to minimize leakage and to provide uninterrupted regulated energy to the sensor system. Finally, the individual components are integrated into one complete microwave power harvesting module, which is used to drive low-power consumer electronics.

As can be surmised, increasing the operational bandwidth of the rectenna will enhance the amount of captured RF power, thus generating more DC energy. However antennas and power conversion devices are traditionally tuned to operate most efficiently at specific frequencies. An 850 MHz device designed to efficiently harvest ambient 3G energy is largely ineffective with Wi-Fi at 2500 MHz. In Chapter 6, a unique broadband rectenna element that can generate DC power from RF signals of frequencies between 825 and 2500 MHz is designed. That is, the developed rectenna element can efficiently harvest energy from both ambient cell phone signals and Wi-Fi routers. Overall, this chapter compares the energy harvesting performance of the proposed broadband rectenna to a network of individually tuned rectennas and discusses the pros and cons of utilizing a broadband microwave harvester.

The dissertation concludes with a summary of major contributions, potential future applications of microwave energy harvesting, and provides guidelines for future research topics in this century old research area.
Chapter 2: Fundamentals of Microwave Energy Harvesting

Microwave rectification has predominately been used and discussed in the context of energy harvesting rectennas (rectifying antennas). Simply, a rectifier is an electrical device that converts alternating current (AC), which periodically reverses direction, to direct current (DC), which flows in only one direction. The process is known as microwave rectification when the rate of alternation in the direction of current flow is between 300 MHz and 300 GHz [19].

In its simplest form, a rectifier is formed by a series diode and a Resistor-Capacitor (RC) circuit, as depicted in Fig. 2.1(a). The idea is to pass one-half of the AC-cycle to an RC circuit, where the time-varying content is filtered such that only the DC component appears across the load, $R_L$. Even under ideal conditions, such a half-wave rectifier is limited to 50% power conversion efficiency. However, at microwave frequencies, the rectifier circuit can be treated as a resonant circuit [28], containing a nonlinear element (i.e. shunt diode) which traps modes of the fundamental frequency and its harmonics (see Fig. 2.1(b)). If the circuit is matched at each frequency, the rectifier acts as a full-wave RF rectifier, even if only one diode is used [28].
RF-to-DC conversion by a rectifier is achieved from the diode’s non-linear voltage and current (I-V) relation, which is described by Richardson equation [29, 30],

\[ i_d = i_s \exp \left( -\frac{q\phi_B}{kT} \right) \left[ \exp \left( \frac{qV_d}{nkT} \right) - 1 \right], \quad (2.1) \]

where \( i_d \) is the current flowing on the diode, \( i_s \) is the saturation current for the diode, \( q \) is the magnitude of the electrical charge on the electron, \( k \) is the Boltzmann constant in Joules/K, \( T \) is the ambient temperature, \( n \) is the ideality factor, \( \phi_B \) is the diode barrier height, and \( V_d \) is the voltage across the depletion layer. Here, \( \phi_B \) is primarily controlled by choice of the barrier metal and the type of semiconductor. \( \phi_B \) is a very important parameter as it determines the amount of RF power required to drive the diode into its nonlinear region [31]. If there is limited RF power available, a low barrier diode would be used. Alternately, if more RF power is available, a higher barrier diode should be used to improve intermodulation distortion.

Indeed, the diode’s output current in (2.1) can be expressed as a summation of harmonics of fundamental frequency, \( f \), by assuming small-signal operation and then
applying the Taylor series expansion, De Moivre’s formula, Euler’s formula, and the
binomial formula [30,31]. That is,

\[ i_d = 2i_s (k_0 + k_1 \sin(2\pi ft + \theta_1) + \ldots + k_n \sin(2\pi nft + \theta_n) + \ldots), \] (2.2)

where \( k_n \) is a constant, \( f \) is the fundamental frequency, and \( \theta_n \) is the phase correction.

As seen in (2.2), in addition to harmonics, the nonlinear diode creates a DC
signal (first term of the bracket) which can be extracted without affecting the RF
characteristics of the resonant rectifier circuit. The ratio of this extracted DC energy
(\( P_{\text{DC}} \)) to the accepted RF energy (\( P_{\text{RF (acc.)}} \)) by the rectifier is known as RF-to-DC
conversion efficiency and defined as,

\[ \eta = \frac{\int_0^t P_{\text{DC}} d\tau}{\int_0^t P_{\text{RF (acc.)}} d\tau}. \] (2.3)

Accurate investigation of the time-varying relationship between \( V_d \) and \( i_d \) in a
rectifier circuit is key in the proper evaluation of \( \eta \). Therefore, a good understanding
of a diode’s nature, its limitations, and its operation is essential.

2.1 Electrical Characteristics and Physics of Diodes

A diode is a two-terminal electronic component with an asymmetric transfer char-
acteristic with low (ideally zero) resistance to current flow in one direction and high
(ideally infinite) resistance in the other. A semiconductor diode, the most common
type today, is a crystalline piece of semiconductor material with a junction connected
to two electrical terminals. Semiconductor diodes’ nonlinear I-V characteristic is tai-
lored by varying the semiconductor materials and introducing impurities into the
materials (doping). Doping is exploited in special-purpose diodes that perform many
different functions. In microwave energy harvesting applications, Schottky diodes are preferred over others because of their very low barrier height [32].

Schottky barrier diodes differ from ordinary junction diodes in that current flow involves only one type of carrier instead of both types. That is, in n-type Schottkys, forward current results from electrons flowing from the n-type semiconductor into the metal; whereas in p-type Schottkys, the forward current consists of holes flowing from the p-type semiconductor into the metal. Therefore slow and random recombination of n- and p- type carriers does not happen in Schottky diodes, which in turn helps Schottky diodes to cease conduction much faster. That is, if the forward voltage is removed, current flow stops “instantly” and reverse voltage can be established in a few picoseconds [30]. Unlike junction diodes, there is no delay effect due to charge storage in Schottky diodes. This accounts for the exclusive use of Schottky barrier in microwave power harvesters, where the diode must switch conductance states at the high local oscillator frequency [32].

![Cross-section view of a typical beamlead Schottky diode (a) and its equivalent circuit (b).](image_url)

Figure 2.2: Cross-section view of a typical beamlead Schottky diode (a) and its equivalent circuit (b).
Schottky diodes are fabricated by the deposition of a suitable barrier metal on an epitaxial semiconductor substrate (such as silicon or gallium arsenide) to form the junction. To improve the diode performance and reliability, silicon based Schottky diodes can be passivated with silicon dioxide, silicon nitride or both, as seen in Fig. 2.2(a) [30]. The choice and processing of materials result in low series resistance along with a narrow spread of capacitance values for close impedance control.

The equivalent circuit model of a Schottky diode is shown in Fig. 2.2(b), where \( C_P \) and \( L_P \) represent the package parasitics, \( C_{OV} \) is the overlay capacitance, \( C_j \) is the junction capacitance, and \( R_S \) is the series resistance. In fact, \( R_S \) is the sum of the resistance due to the epitaxial layer and the resistance due to the substrate, \( R_{epi} \) and \( R_{sub} \) in Fig. 2.2(a), respectively. Additionally, \( R_j \), junction resistance (also known as video resistance), in Fig. 2.2(a) corresponds to the “ideal diode” in Fig. 2.2(b). \( R_j \), a key diode parameter used by circuit designers, represents an effective resistance analogous to the radiation resistance of antennas. At fundamental frequency, during rectification, RF energy is “lost” to \( R_j \) while it is converted to DC. Consequently, \( R_j \) is a very important parameter when selecting the optimal load impedance.

In literature [33, 34], a rectifier diode is treated as a current generator across junction resistance. Hence, \( R_j \) is a function of the total current flowing through the diode (i.e., varies with the input RF power) and is given by [33],

\[
\frac{1}{R_j} = \frac{\partial i_d}{\partial V_d} = \frac{n k T}{q (i_{bias} + i_s)}, \tag{2.4}
\]

As can be inferred, all the circuit parameters in the equivalent model (Fig. 2.2(b)) should have minimum values to achieve maximum energy transfer to the ideal diode, or \( R_j \). Fortunately, the effects of \( C_P, L_P, \) and \( C_{OV} \) can be tuned out by a properly designed impedance matching network. Therefore, the focus is on minimizing \( R_S \) and
For Schottky diodes, $R_S$ is given by [33],

$$R_S = R_{epi} + R_{sub} = \frac{L}{q\mu_N N_d A} + 2\rho_s \sqrt{\frac{A}{\pi}}.$$  

(2.5)

In (2.5), $L$ is the thickness of the epitaxial layer, $q$ is the magnitude of the electrical charge on the electron, $\mu_N$ is the mobility of electrons in the dopant, $N_d$ is the doping density of the epitaxial layer, $A$ is the area of Schottky contact, and $\rho_s$ is the substrate resistivity.

The other critical parameter, junction capacitance ($C_j$) of a Schottky diode, is determined to a first-order approximation by the metal used, the silicon doping, and the active area [34]. Hence,

$$C_j = A \frac{q\varepsilon_r \sqrt{N_d}}{2 (V_d - kT/q)}.$$  

(2.6)

In this, $\varepsilon_r$ is the relative permittivity of the epitaxial layer and $V_d$ is the voltage across the depletion layer. Evidently, $C_j$ is a dynamic parameter that varies with the applied voltage, complicating its evaluation.

It is clear from (2.5) and (2.6) that reducing the epitaxial layer thickness, optimizing the dopant, and using low-loss substrate in the diode design will reduce $R_S$, while using a lower permittivity epitaxial layer will reduce $C_j$. It has been suggested that using arsenic, instead of popular phosphorous, as dopant in diode design results in significant reduction of $R_S$ [35].

However, minimizing $R_S$ and $C_j$ at the same time is still a challenge. As suggested by (2.5) and (2.6), both Schottky contact area and the doping density of the epitaxial layer have inverse effects on $R_S$ and $C_j$. Increasing the contact area reduces $R_S$ but increases $C_j$. Similarly, reducing the doping density reduces $C_j$ but increases $R_S$. Nevertheless, depending on the application, the rectifier efficiency might forgive
focusing on only one parameter. For instance, a low frequency, high RF power system’s rectification efficiency will depend significantly more on the value of $R_S$ than it will on $C_j$. However, for a high frequency (i.e., microwave frequencies), low RF power system, $C_j$ must be in the 0.1 to 500 femtofarads range to achieve reasonable rectification efficiency (“rule of thumb” is $C_j$ in pF < $1/f$ in GHz). A good approximation to the effect of junction capacitance and series resistance on conversion loss is given as [36],

$$L = 1 + \frac{R_S}{R_j} + \omega^2 C_j^2 R_S R_j.$$  \hfill (2.7)

![Graph showing conversion loss vs. RF input power for different junction capacitance and series resistance values.](image)

Figure 2.3: Measured effect of junction capacitance and series resistance on conversion loss, evaluated at 2 GHz [36].

It is apparent from (2.7) that conversion loss depends on operating frequency and rectification efficiency will swiftly deteriorate as the frequency increases. According to (2.7), $C_j$ has a more dominant effect on the conversion loss than $R_S$ when only
the frequency-dependant term is considered. However, in addition to frequency, the conversion loss also depends on the strength of the input RF power, since $R_j$ is a function of it. Fig. 2.3 shows the effect of $C_j$ and $R_S$ on rectification efficiency by studying three different diodes at 2 GHz under different input RF power levels [36]. As seen in Fig. 2.3, for high power cases ($P_{RF} > 0$ dBm), a low $R_S$ diode is a must for minimizing the conversion loss and for low power cases, a low $C_j$ diode should be considered. Further, the diodes that do not require biasing should be preferred over others. This is crucial for microwave power harvesting applications as even a few microamperes of bias current is difficult to generate.

The diode study presented in this section is crucial in understanding rectifier circuits. Next, the time-varying relationship between $V_{DC}$ and $i_d$ in a rectifier circuit can be analyzed and $\eta$ can be calculated. The analysis starts with the simplest rectifier circuit, the Schottky detector.

### 2.2 Schottky Detectors as Rectifier Circuits

A Schottky diode detector is constructed with a single diode, an RF impedance transformation circuit, and some low frequency components (see Fig. 2.4(a)). A simplified equivalent circuit for the same detector is given in Fig. 2.4(b). As seen, the equivalent circuit in Fig. 2.4(b) employs the diode model discussed in the last section (shown in Fig. 2.2(b)) and assumes that $C_P$, $L_P$, and $C_{OV}$ are tuned out by the impedance transformation network.

The behavior of Schottky detector circuits has been well studied in literature [33–35]. The Schottky detector is traditionally viewed as a device providing a DC output voltage that varies as the square of the input power at low power levels (the square-law
Figure 2.4: Schottky detector circuit (a) and its simplified equivalent circuit (b) [39].

region) and directly with the input power at high levels (the linear region). Examination of a typical detector manufacturer’s data sheets and experimental evidence [37] show that this accepted view is true only under restrictive conditions, which is often violated under practical operating conditions. Simplified analytical investigations of the behavior of diode detectors have been performed [38], but because of the perceived difficulty of treating the resulting nonlinear functions, truncated series approximations to the I-V characteristic have been employed. Unfortunately, the truncation process destroys possible insights into the details of the circuit behavior. Of course, numerical methods such as integration in the time domain or harmonic balance can be used to obtain solutions for specific circuit data, but being computationally intensive, such methods do not easily provide a global view of the circuit response.

In a previous study [39], it was shown that an all-analytical averaging method could be applied to obtain a closed-form solution without any need to simplify the nonlinear functions. In this dissertation, the theory presented in [39] will first be incorporated then improved by utilizing a more accurate diode model (including harmonics) to estimate the RF-to-DC conversion efficiency. As presented in [39], starting
\[
x = \frac{q}{nkT} V_g \cos(v\tau) \quad \quad a = \frac{q}{nkT} R_L i_s \quad \quad v = w R_L C_L \quad \quad \zeta = \frac{i_{\text{bias}}}{i_s}
\]
\[
y = \frac{q}{nkT} v_{\text{out}} \quad \quad b = \frac{R_g + R_S}{R_L} \quad \quad k = \frac{q}{nkT} R_S i_s \quad \quad g = \frac{C_j}{C_L}
\]

Table 2.1: Parameters of the equation relating the input voltage to the output voltage of a Schottky detector. See also Fig 2.4(b).

with a straightforward analysis of the circuit in Fig 2.4(b) results in the following differential equation,

\[
a \zeta + y' + y = a \left[ \exp \left\{ (x - y) - b(y' + y) - k \zeta \right\} - 1 \right] + g \left[ (x' - y') - b(y'' + y') \right], \tag{2.8}
\]

where the symbols ‘ and ′ indicate $\partial/\partial \tau$ and $\partial^2/\partial \tau^2$ respectively, with $\tau = t / (R_L C_L)$.

The other quantities are tabulated in Table 2.1.

The differential equation presented in (2.8) could be integrated numerically; however in the interests of obtaining a closed-form solution, the Ritz-Galerkin (RG) algebraic averaging method [40] used in literature is adopted here. To apply the RG method, the differential equation is represented in the form:

\[
\xi \left[ x, y, \tau, \partial/\partial \tau, \partial^2/\partial \tau^2 \right] \equiv 0, \tag{2.9}
\]

where $\xi$ is the nonlinear error operator. The exact solution can be approximated by an assumed solution,

\[
\bar{y}(\tau) = \sum_{k=1}^{N} a_k \psi_k(\tau), \tag{2.10}
\]

where the $\psi_k$ are $N$ linearly independent functions and the $a_k$ are $N$ adjustable constant coefficients. However, the assumed solution of (2.10) does not satisfy the differential equation. Therefore, the expression obtained by substituting (2.10) into
\[(2.9) \text{ is no longer identically equal to zero. Instead,}
\]
\[
\xi [x, y, \tau, \partial / \partial \tau, \partial^2 / \partial \tau^2] = \epsilon(\tau) \neq 0,
\]
\[(2.11)\]

where \(\epsilon(\tau)\) is called the residual and is a measure of the incurred error. It can be shown [40] that the error can be minimized by satisfying a system of \(N\) weighted residuals called the Ritz conditions:
\[
\int_{\tau_1}^{\tau_2} \xi [x, y, \tau, \partial / \partial \tau, \partial^2 / \partial \tau^2] \psi_k(\tau) d\tau = 0 \quad k = 1, \ldots, N. \quad (2.12)
\]

This procedure results in a system of \(N\) algebraic equations in \(N\) unknowns.

Currently, the output signal consists of a DC voltage \(V_0\) and a ripple voltage (see Fig. 2.1(b)), which the fundamental component is at the RF frequency \(w\) (see Table 2.1, \(v = wR_LC_L\)). Hence, one may assume a solution
\[
\ddot{y}(\tau) = \sum_{k=1}^{N} a_k \psi_k(\tau) = Y_0 + Y_1 \cos(v\tau + \theta_1) + Y_2 \cos(2v\tau + \theta_2) + \ldots, \quad (2.13)
\]

where \(Y_0 = q/(nkT)V_0\), \(Y_1\) and \(\theta_1\) are the amplitude and phase of the fundamental frequency, and \(Y_n\) and \(\theta_n\) are the amplitude and phase of the harmonic terms. In [39], a simplification to the procedure was suggested by neglecting the ripple, i.e. harmonics. Then all that remains is the single unknown
\[
\ddot{y}(\tau) = Y_0 \quad (2.14)
\]

with the following Ritz condition,
\[
\int_{\tau_1}^{\tau_2} \xi [x, y, \tau, \partial / \partial \tau, \partial^2 / \partial \tau^2] d\tau = 0 \quad (2.15)
\]

From (2.8) and (2.14), the following expression is obtained for the residual:
\[
\epsilon(\tau) = a \left[ \exp \left\{ X \cos(v\tau) - (1 + b)Y_0 - k\zeta \right\} - 1 \right] - vgX \sin(v\tau) - a\zeta - Y_0, \quad (2.16)
\]
where $X = q/(nkT) V_g$.

Carrying out the integration specified in (2.15) results in [39],

$$
I_0 \left( \frac{qV_g}{nkT} \right) = \left( 1 + \frac{i_{bias}}{i_s} + \frac{V_0}{R_L i_s} \right) \exp \left\{ \left[ 1 + \frac{R_g + R_S}{R_L} \right] \frac{qV_0}{nkT} + \frac{qR_S i_{bias}}{nkT} \right\}. \tag{2.17}
$$

Here, $I_0(x)$ is the zero-order modified Bessel function of the first kind and argument $x$ [41,42]. To solve (2.17), it is necessary to obtain the inverse of the modified Bessel function. Such an equation can be analyzed by a program such as Matlab, where the output voltage can be iterated to obtain a series of values for input voltage.

Note that (2.17) relates the amplitude of the input RF voltage ($V_g$) to the harvested DC voltage ($V_0$). To obtain RF-to-DC conversion efficiency, $\eta$, from (2.17) it can be assumed that, as in linear circuit theory, $P_{RF}$ is the power that would be absorbed by a conjugately-matched linear load. Therefore,

$$
\eta = \frac{P_{DC}}{P_{RF}} = \frac{V_0^2}{R_L} = 8 \frac{V_0^2 R_g}{V_g^2 R_L} \tag{2.18}
$$

The accuracy of (2.17) was verified in [43] by building the rectifier depicted in Fig. 2.1(b) and recording the harvested DC voltage levels when input RF power to the rectifier was varied. The measurement setup included an RF source that outputs a 100 MHz signal at various power levels, an Agilent HSMS-8201 Schottky diode, a 100 pF load capacitance, and a 1kΩ load resistance ($R_L$). Findings of [43] are depicted in Fig. 2.5, where the agreement between the measured data and the calculations verify (2.17).

(2.17) can also estimate the power lost to the series junction resistance of the diode ($R_S$) and the conversion inefficiency arising from the impedance mismatch between rectifier and the output DC load ($R_L$). Nevertheless, (2.17) does not tell the
Figure 2.5: Measured and calculated output voltage (a) and RF-to-DC conversion efficiency (b) of a simple Schottky detector circuit.

...complete story of microwave rectification since the critical term involving the capacitance ratio $g$ (see (2.8) and Table 2.1) has vanished in its calculation process. This is a consequence of ignoring the ripple component of the output voltage (see (2.14)), which in turn implies that the RF current flowing through $C_j$ and the harmonic terms generated by the nonlinear diode have been neglected. Thus, the projected relationship between input RF power and RF-to-DC conversion efficiency is precise only at very high frequency (VHF) bands (as depicted in Fig. 2.5). (2.17) loses its accuracy when the operating frequency is increased to the microwave bands since more current flows through $C_j$. Therefore, to maintain accuracy, at least one more term has to be included in (2.14) and (2.17) should be updated accordingly. Ergo,

$$\tilde{y}(\tau) = Y_0 + Y_1 \cos(v\tau + \theta_1)$$ (2.19)
From (2.8) and (2.19), the following expression is obtained for the updated residual:

\[
\epsilon(\tau) = a \left[ \exp \left\{ X \cos(\nu \tau) - (1 + b)(Y_0 + Y_1 \cos(\nu \tau + \theta_1)) + bvY_1 \sin(\nu \tau + \theta_1) - k\zeta \right\} 
- 1 - \zeta 
- g\nu[X \sin(\nu \tau) - Y_1 \sin(\nu \tau + \theta_1) - b(vY_1 \cos(\nu \tau + \theta_1) 
+ Y_1 \sin(\nu \tau + \theta_1))] 
- Y_0 - Y_1 \cos(\nu \tau + \theta_1) + vY_1 \sin(\nu \tau + \theta_1) \right] 
\]

(2.20)

In [39], the authors did not consider any harmonics in their calculations to avoid the cost of considerably increased algebraic complexity (as seen in (2.20)). However, some of this cost can be offset by omitting the external bias current, \(i_{bias} = 0\), from the calculations. This is a valid assumption since external bias is generally applied in RF detectors and omitted in most RF energy harvesting applications. Hence,

\[
\epsilon(\tau) = a \left[ \exp \left\{ X \cos(\nu \tau) - (1 + b)(Y_0 + Y_1 \cos(\nu \tau + \theta_1)) + bvY_1 \sin(\nu \tau + \theta_1) \right\} 
- 1 - \zeta 
- g\nu[X \sin(\nu \tau) - Y_1 \sin(\nu \tau + \theta_1) - b(vY_1 \cos(\nu \tau + \theta_1) 
+ Y_1 \sin(\nu \tau + \theta_1))] 
- Y_0 - Y_1 \cos(\nu \tau + \theta_1) + vY_1 \sin(\nu \tau + \theta_1) \right] 
\]

(2.21)

(2.21) can be simplified further via a multidimensional minimization method, e.g., a simplex method [44]. In [45], the authors demonstrated the effectiveness of the simplex method by developing a closed-form analytical expression for the input impedance of a diode at RF frequencies. Following their footsteps in [45], the following relationship between \(X\) and \(Y_1\) can be obtained (for \(\theta_1 = 0\)).

\[
Y_1 = \frac{X C_j (R_g + R_S)}{C_L R_L} = Xbg, \quad (2.22)
\]

Note that (2.22) supports the assumption that the signal strength of the first harmonic is much smaller than the input signal, i.e., \(X \gg Y_1\). With this, carrying out the integration specified in (2.21) results in (for \(\theta_1 = 0\), see [46], page 496),

\[
I_0 \left( \frac{qV_g}{nkT} \sqrt{1 - \frac{w^2C_j^2(R_g + R_S)^4}{R_L^2}} \right) = \left( 1 + \frac{V_0}{R_{L,i_s}} \right) \exp \left\{ \left[ 1 + \frac{R_g + R_S}{R_L} \right] \frac{qV_0}{nkT} \right\}.
\]

(2.23)
Similarly,
\[
\eta = \frac{P_{DC}}{P_{RF}} = \frac{\frac{V_0^2}{R_L}}{\frac{V_g^2}{V_g^2 R_g}} = 8 \frac{V_0^2}{V_g^2 R_L}
\]  

(2.24)

Again, to solve (2.23), it is necessary to obtain the inverse of the modified Bessel function. Similarly, Matlab can be used to analyze (2.23) by iterating the output voltage to obtain a series of values for input voltage. After establishing the relationship between $V_g$ and $V_0$ by (2.23), (2.18) can again be used to calculate the conversion efficiency, $\eta$.

![Diagram](image-url)

Figure 2.6: Simulated and calculated output voltage (a) and RF-to-DC conversion efficiency (b) of a simple Schottky detector circuit.

To verify the accuracy of (2.23), simulations (with Agilent ADS) were done to study the operation of a Schottky detector depicted in Fig. 2.1(b). The simulation recorded the harvested DC voltage levels when input RF power to the rectifier was
varied. The setup included an RF source that output a 2 GHz signal at various power levels, an Agilent HSMS-2852 Schottky diode, a 100 pF load capacitance, and a 1kΩ load resistance \( R_L \). Findings are depicted in Fig. 2.6, where the agreement between the simulated data and the calculations based on (2.23) shows significant improvement. Note that, compared to (2.17), (2.23) provides a more accurate projection of \( \eta \). Indeed, (2.23) could perform better without the \( X \gg Y_1 \) assumption, which is, of course, at the expense of increased computational complexity.

The analysis presented in this section helps us determine the fundamental bounds on conversion efficiency and pinpoint the contribution of individual diode parameters to harvesting inefficiency. This is crucial knowledge in building low-power efficient rectifier circuits. Now that it is established for Schottky detectors, we can continue our rectifier analysis with voltage doublers.

### 2.3 Voltage Doublers as Rectifier Circuits

In the majority of RF power harvesting applications, received power is relatively low, \( P_{RF} \leq 0 \) dBm. However, by using the received low RF voltage from the antenna, the full operating DC voltage must be provided to the load integrated circuit (IC). This is very challenging as typical ICs require 1 or 2 V to run and this has to be squeezed out of an antenna that is itself providing only about 0.2 V at a distance of a few meters away from a ultra high frequency (UHF) transmitter.

A very common approach to obtaining higher voltages from a rectifier is the use of a charge pump: a number of diodes connected in series so that the output voltage of the array is increased. The simplest sort of charge pump, a voltage doubler, is shown in Fig. 2.7(a). Two diodes (D1, D2) are connected in series, oriented so that
forward current must flow from the ground potential to the positive terminal of the output voltage $V_{DC}$. A capacitor (C1) prevents DC current from flowing between the antenna and the diodes, but stores charge and thus, permits high frequency currents to flow. A second capacitor (C2) stores the resulting charge to smooth the output voltage.

The operation of the voltage doubler is as following. When the RF signal is in negative cycle, the first diode (D1) is on (Fig. 2.7(a)). Current flows from the ground node through the diode, causing charge to accumulate on the input capacitor (C1). At the negative peak, the voltage across the capacitor is the difference between the negative peak voltage and the voltage on the top of the diode. At this instant, the output (right) plate of the capacitor is more positive than the RF input. When the RF input becomes positive, the first diode turns off and the second (output) diode (D2) turns on (Fig. 2.7(a)). The charge that was collected on the input capacitor (C1) travels through the output diode to the output capacitor (C2). The peak voltage that can be achieved is found by adding the voltage across the input capacitor, to the peak positive RF voltage and subtracting the turn-on voltage of the output diode. In the limit, where the turn-on voltage can be ignored (e.g. when the input voltage is very large) or in the case of zero-bias diodes, the output DC voltage is double the peak voltage of the RF signal, from which fact the circuit derives its name. The actual output voltage depends on the amount of current drawn out of the storage capacitor during each cycle, which depends on the value of the load resistance, $R_L$.

If the two diodes (D1 and D2) are contained in a single package, the cost impact associated with the addition of second diode (compared to the Schottky detector) is
Figure 2.7: Schematic of the voltage doubler circuit and its operation (a) and its evolution into a simple Schottky detector (b).
very small, making the doubler an interesting option for RF power harvesting applications. In Fig. 2.7(b), an equivalent circuit is described for the voltage doubler [43], where it can be analyzed relatively easily using theoretical approaches derived in the previous section, i.e., (2.23). It can be seen that the transfer curve for a voltage doubler can be predicted like a simple Schottky diode detector by doubling the value of $R_g$, halving the value of $R_L$, and doubling the calculated values of $V_0$.

Even though a voltage doubler rectifier enhances the harvested DC voltage, it is likely not high enough to turn on the majority of ICs in the market. Additional stages of voltage doublers are usually necessary to pump up the output DC voltage. The RF energy harvesters on the market (RFIDs, wireless sensors, etc.) experience this problem and the most popular way of solving it is to utilize a Dickson charge pump configuration [47], the next section’s topic.

### 2.4 Dickson Charge Pump

A common approach to boosting the voltages from a rectifier is to use a charge pump as part of the rectifier circuit. A charge pump incorporates a number of diodes and capacitors connected in series. Note that increased voltage levels are obtained in a charge pump as a result of transferring charges to a capacitive load, and do not involve amplifiers or regular transformers. A schematic of a two-stage Dickson charge pump is shown in Fig. 2.8(a). Each stage of the Dickson charge pump is formed using voltage doublers. Therefore, when the RF voltage is negative, D1 is on and the current flows from the ground through D1. Alternatively, when the RF input is positive, D1 turns off and D2 turns on to continue charging C2. As such, we expect the DC voltage on C2 (intermediate storage capacitor) to be twice the
supplied peak RF voltage. The operation of the second-stage Dickson charge pump is similar, resulting in an output voltage across C4 (main storage capacitor), double that of the voltage across C2, or quadruple of the supplied peak RF voltage. When higher output voltage is desired, additional stages of voltage doublers can be added to the Dickson charge pump [48] at the expense of reduced conversion efficiency.

Figure 2.8: Schematic of a two-stage Dickson charge pump rectifier (a) and a sample design and its measured conversion efficiency ($\eta$) (b).

In this study, a two stage Dickson charge pump was built (see inset of Fig. 2.8(b)) and its conversion efficiency performance at 2.45 GHz was studied when the input RF power was varied. Two zero-bias Schottky diode pairs (HSMS-2852) were chosen in this design (see Fig. 2.8(a)) as they have excellent performance at UHF and do not require external biasing. This is crucial, as even a few microamperes of bias current is difficult to generate. Further, these diodes have relatively low-barrier height and high-saturation current when compared to externally biased detector diodes. This results in higher output voltage at low-power levels. However, a drawback is their
higher series resistance, which inherently leads to higher losses. Further, the bypass capacitors (C1, C2) were chosen to be 100 pF and storage capacitors (C2, C4) are 100 μF, both from Panasonic Electronics and the load resistance, R_L, is 5 kΩ.

An impedance matching network is essential in providing maximum power transfer from the antenna to the rectifier circuit. One approach is to design the matching network as proposed in [49]. According to [49], the component models used in the simulations were not accurate enough to include all of the circuit parasitics. Therefore, it was concluded that the modeling of the rectifier circuit should be based on experimental characterization. This can be done by measuring the input impedance (S_{11}) of the rectifier circuit (consisting of the zero-bias diode pairs, capacitors, load resistance, and wires for output DC voltage) without a matching network. The results from the experimental characterization can then be used as a black box for the impedance-matching circuit design.

The conversion efficiency, η, of the two-stage Dickson charge pump for RF power levels varying from -25 to 10 dBm is shown in Fig. 2.8(b). It is apparent that even a typical two-stage Dickson rectifier can realize an RF-to-DC conversion efficiency greater than 60% over a wide range of power levels, with a maximum efficiency close to 70%. However, the good conversion efficiency performance overshadows the fundamental problems with the topology of the Dickson rectifier. In Dickson rectifiers, additional stages are added on top of the existing voltage doublers. Therefore, the reverse DC voltage experienced by the diodes (after rectification, before discharge) at the higher stages is much higher compared to the ones on the lower stages. This puts a significant restriction on component selection and limits the number of stages that the circuit can have. In addition, the circuit delivers all of the harmonics generated by
the diode to the DC load, a problem that also exists in Schottky detectors. Depending on the application, the harmonics can be fatal to the operation of IC that the circuit drives. However, introducing filters after rectification deteriorates the good efficiency performance depicted in Fig. 2.8(b). Therefore, a new topology that fixes these problems with Dickson rectifier and provides more efficient rectification at even lower power levels would be a game-changing technology.

2.5 Conclusion

The chapter started with presenting a thorough background and history of the microwave power harvesting process. Subsequently, the rectification concept was discussed and its fundamental limitations are highlighted. This was followed by recommendations to the semiconductor industry for optimal component design. This chapter showed that:

• To maximize RF-to-DC conversion efficiency that a rectifier can achieve, all the circuit parameters in the diode equivalent model should be minimized.

• Minimizing two of these parameters, i.e., $R_s$ and $C_j$, at the same time is a challenge since a Schottky diode’s contact area and the doping density of its epitaxial layer have inverse effects on these two.

• Depending on the application, the conversion efficiency can be improved by focusing on minimizing the diode’s $R_s$ or $C_j$.

Next, without making any major simplifications to the nonlinear functions stemming from the diode, an improved closed-form solution for RF-to-DC conversion efficiency was presented and its accuracy was investigated. The presented closed-form analysis

32
provides insights on how much of input power lost to $R_s$, what percentage of the input RF energy bypassed the diode through the junction capacitance, and the inefficiencies arising from the impedance mismatch between the rectifier and the output DC load. In the final sections of this chapter, specific rectifier topologies used in the state-of-the-art RF energy harvesters were studied and their weaknesses were noted. Chapter 3 will introduce novel, low-power efficient rectifiers, which do not need output filters to kill the diode harmonics.
Chapter 3: Novel High Efficiency Rectifier Circuits for Microwave Energy Harvesting Applications

Finite battery life is encouraging the industry and academia to develop innovative ideas and technologies to power mobile devices for an infinite or enhanced period of time. Energy harvesting has been hailed as one of the most promising solutions to the finite battery problem [2]. There are numerous sources from which energy harvesting can benefit from. Solar power is a key example since it has the highest energy density among other candidates [50]. However, solar harvesters can only operate when sunlight is present. Similarly, vibrational energy harvesting is feasible only when a constant motion exists. RF energy harvesting overcomes the reliance on the availability of ambient sources, because the power can be easily replenished when desired.

RF energy harvesting is done by a circuitry called a rectifier, in which a diode is the key electrical component. The RF-to-DC conversion efficiency of the nonlinear diode changes as the operating power level changes [51]. As discussed in Chapter 2, the loss incurred by series junction resistance ($R_S$) and junction capacitance ($C_J$) of the diode is the fundamental reason for the nonlinear efficiency performance. However, traditional RF power harvesting systems (Schottky detector, voltage doubler,
Dickson, etc.) have additional inherent inefficiencies at low and high input power regions (see Fig. 3.1 from [51]). The inefficiency in the low power region ($\leq -15$ dBm) stems from the weak input RF voltage swing, which is below or comparable with the forward voltage drop (barrier height) of the diode [51]. The efficiency increases as the power increases (medium power region) and levels off with the generation of strong higher order harmonics [51]. The efficiency sharply decreases in the high power region ($\geq 20$ dBm) as the voltage swing at the diode exceeds the reverse breakdown voltage ($V_{BR}$) of the diode [51]. Adaptive RF harvesters have been presented [52] to increase the power region where the conversion efficiency approaches to the theoretical bounds. These circuits can detect the available input RF power and then switch between individually optimized rectifiers to maximize the conversion efficiency. Nevertheless, these individually optimized rectifiers are still a form of traditional rectifiers and, therefore, inherit the problems associated with them.

![Figure 3.1: General relationship between microwave to DC power conversion efficiency and input power [51].](image-url)
In wireless power transfer (WPT) applications, where dedicated RF transmitters are utilized as energy sources, the wave impinging on the antenna has very high signal strength. For example, in the Goldstone demonstration, NASA JPL transmitted microwave power exceeding a mile and harvested a whopping DC output of 34 kW ($\approx 75$ dBm) with 72% efficiency [20]. As expected, in these WPT applications, the rectifiers operate in the high power region, where the conversion efficiency of the circuit is hampered by $V_{BR}$ of the diode and the creation of strong harmonics. By utilizing a diode with lower saturation current ($i_s$), it is possible to increase the $V_{BR}$ to a level where it is no longer problematic. However, the generation of strong harmonics still cannot be avoided. Conventional rectifiers employ lossy filters to remove the harmonics from the output signal. This chapter first proposes an improved circuit design that achieves higher conversion efficiency in WPT applications by harvesting the energy left in fundamental frequency and in harmonics.

Though given the ubiquity of RF transmitters, it may seem that installing dedicated RF transmitters is unnecessary, as transmissions from TV stations, cellular network towers, and Wi-Fi hot spots would bathe the average consumer in a steady source of untapped energy. However, the amount of RF power transmitted is limited by government regulations. Current FCC guidelines limit RF exposure to the general public to less than 1000 $\mu$W/cm$^2$ [53,54], which is the highest level of RF power one would normally expect to encounter. Currently, though, there is much less ambient power generated by cell towers, which are limited to 580 $\mu$W/cm$^2$ at ground level [54]. In addition, objects around an RF device reflect and absorb radio waves, causing fluctuations in received power. This is best seen through the reception of a mobile phone and how much it can vary over time and across location. Simply placing
a phone into a pocket, next to all of the RF absorbing water in the human body, can dramatically reduce the power available for harvesting. Also, portable devices are small. Therefore, most of the RF energy harvesting devices (rectenna) are also small. This implies a small antenna size and an even lower received RF power for harvesting i.e., -15 dBm or lower. Consequently, in most RF energy harvesting applications, the rectifier will operate at the low power region, where the conversion efficiency of the circuit is hampered by the forward voltage drop of the diode. By utilizing a diode with much higher saturation current, it is possible to reduce this forward voltage drop and hence, increase the efficiency of the rectifier in the low power region. However, note that increased $i_s$ comes at the expense of reduced $V_{BR}$ [30].

At the low power region, the rectifiers, by themselves, cannot harvest enough DC voltage to power up the ICs in the market. A common approach to boosting the output voltage from a rectifier is to use a multi-stage charge pump as part of the rectifier circuit (see Chapter 2). In traditional charge pumps, diodes experience different amounts of DC voltage at their output terminals, which forces the higher stage diodes to the breakdown region much faster than the lower stage diodes. Therefore, conventional rectifiers with charge pump units cannot utilize the diodes with the highest saturation current. Otherwise they cannot tolerate the reduced $V_{BR}$ at the medium power region, and hence cannot be very efficient at the low power region. In this chapter, we finally propose a novel rectifier circuit design (modified from the traditional Greinacher rectifier) that can employ these diodes and achieve higher conversion efficiency at the low power region, while maintaining the good conversion efficiency performance at the medium power region.
3.1 An Improved Voltage Doubler

A rectifier’s RF-to-DC conversion is owed to the diode’s non-linear I-V relation, viz., \( i_d = 2i_s e^{k \cos(2\pi ft)/(nd)} \) (simplified from (2.1)). Indeed, the diode’s output current can be expressed as a summation of harmonics of the fundamental frequency, \( f \). Specifically, the diode current can be expressed as (after applying the Taylor series expansion, De Moivre’s formula, Euler’s formula, and the binomial formula),

\[
i_d = i_0 + \sum_{n=1}^{N} \kappa_n \cos(nwt) + \zeta_n \sin(nwt), \tag{3.1}
\]

Here, \( i_0 \) is the DC portion of the current. \( \kappa_n \) and \( \zeta_n \) represent the amplitudes of the harmonic terms (operating at the radial frequency \( nw \)), which vary with diode parameters and load resistance as discussed in Chapter 2. A conventional rectifier design, such as the voltage doubler depicted in Fig. 3.2(a), harvests only the DC component, which dominates when the input RF power level is in the medium region (as discussed earlier). Hence, high conversion efficiency (70%) is feasible. However, as depicted in Fig. 3.3, when the same rectifier is driven into the high power region, a significant portion of the rectified energy (37% in 3.3(a)) is wasted in harmonics. Consequently, the overall RF-to-DC conversion efficiency performance drops to 58%, as depicted in Fig. 3.3. To recover the diode efficiency, the power that went into the harmonics must also be harvested.

In this chapter, a novel and improved Greinacher doubler circuit design is proposed (illustrated in Fig. 3.2(b)). In contrast to the traditional voltage doubler design, the proposed doubler introduces three new components: L1, C3, and D3. Assuming that the current flowing through the D2 diode terminals take the form in (3.1). The inductor L1 then rejects the RF components of the diode current \( i_d \) while still allowing
Figure 3.2: Schematic of a traditional voltage doubler (a) and the proposed efficient harvester (b).

Figure 3.3: Simulated (a) and measured (b) power harmonics at the output of the rectifier diode (D2) of the voltage doubler circuit.
the DC component to pass and power the load. Concurrently, the capacitor C3 and the Schottky diode D3 create an alternate current path for the rejected harmonics. As expected, C3 simply serves to block the DC current and force it to flow to the load through L1. The diode D3 rectifies the remaining RF energy stored in the harmonics created by D2 and rejected by L1. The harvested DC current from D3 flows to the load (adding to DC current flowing through L1) and increases the overall conversion efficiency up to 70%, as depicted in Fig. 3.4(a).

![Figure 3.4](image)

Figure 3.4: Simulated power harmonics at the output of the harmonics harvester circuit (a) and measured conversion efficiency of the traditional and proposed rectifiers (b).

Fig. 3.4(b) compares the measured efficiencies of the traditional and the proposed rectifier designs, where the plots are presented as conversion efficiency versus the input RF power at 2.5 GHz. As seen in Fig. 3.4(b), the developed rectifier topology increases
the conversion efficiency from 55% to 65% when the input RF power to the circuit is 10 dBm. It should also be noted that, the proposed rectifier has significantly expanded medium power region (where the circuit is efficient, i.e., \( \eta \geq 60\% \)) by harvesting the harmonic content instead of filtering. Of course, although faster for the voltage doubler, the conversion efficiency declines as the input RF voltage increases, since the DC voltage stored in the output capacitor reaches to a point where the diodes are forced into the reverse breakdown region.

The proposed rectifier circuit offers significant improvements in terms of RF-to-DC conversion efficiency over the conventional Greinacher voltage doubler. Specifically, the proposed circuit increases the conversion efficiency from 58% to nearly 70% in the high power region, which is achieved by adding a zero bias Schottky diode that rectifies the energy from the wasted harmonics. The performance of the proposed rectifier circuit topology is validated with measurements. The next section discusses another novel rectifier, one that can achieve higher conversion efficiency at the lower power region.

3.2 Modified Greinacher Rectifier

In the majority of microwave energy harvesting systems, RF energy must be extracted from the air at a very low power density, because as the distance from the energy source increases, the propagation energy rapidly drops [48]. In free space, the power density drops at the rate of \( 1/d^2 \), where \( d \) is the distance from the radiating source. With multi-path fading, the power density can drop dramatically in certain spots (see Chapter 4); thus, it is critical that the power conversion circuit operates at very low received power. However, when the available RF power to the receiver...
is under 100 μW (or -20 dBm), the available voltage for rectification in the RF-to-DC conversion system falls below 0.3 V, which is too small to power any consumer electronics [55].

In conventional designs, the voltage doubler rectifier is the basic building block for the RF-to-DC power conversion system for several reasons. The voltage doubler rectifies the full-wave peak-to-peak voltage of the incoming RF signal and is easily arranged in cascade to increase the output voltage, as is the case with the Dickson charge pump [55]. Cascading multiple doubler stages is necessary to boost the output voltage from a rectifier to a usable level. Keep in mind that the number of cascaded rectifier stages in the RF-to-DC conversion system has a significant effect on the rectifier performance, including the conversion efficiency and the input impedance. In general, cascading multiple rectifier stages in series cause the diode’s capacitive parameters to linearly increase with the number of stages while providing parallel paths for current that cause the resistive parameters to decrease [56]. Therefore, the input impedance of the rectifier, both the resistive and the reactive portions, decreases as the number of stages in the rectifier increases. In addition, as discussed earlier, the output voltage initially increases as more rectifier stages are added to the charge pump until an optimal point [57]. Then the harvested voltage swiftly reduces as the highest stage diode output voltage exceeds the reverse breakdown voltage of the diode. This section proposes a novel rectifier circuit design that can employ diodes with low reverse breakdown voltages and achieve higher conversion efficiency at the low power region.

Fig. 3.5(a) depicts the circuit schematic of the proposed high efficiency rectifier, and Fig. 3.5(b) depicts the conventional two-stage Dickson charge pump circuitry.
Figure 3.5: Schematic of proposed single-stage modified Greinacher rectifier (a) and a two-stage Dickson charge pump (b).

for comparison. The operation of the proposed modified Greinacher rectifier is as follows. First, the induced voltage at the RF port passes through the DC blocking capacitors (C1 and C3) and charges them. The “RF symmetric” voltage doublers rectify the incoming RF energy. The rectified current output is then pumped to the storage capacitors (C2 and C4), which supply a stable DC power to the load after the rectifier reaches its steady-state mode.

As seen in Fig. 3.5, the Dickson charge pump and the modified Greinacher rectifier use the same number of components, but a different circuit topology. Unlike the Dickson charge pump rectifier, the proposed modified Greinacher rectifier is RF symmetric. This way, every rectifying diode is excited with the same input signal amplitude, and the output diodes (D2 and D4 in Fig. 3.5) experience the same reverse bias voltage; thus, improving the conversion efficiency at the low power region. In addition, the symmetric structure does not generate the odd harmonics, which cancel at the output terminal and eliminates the need for a lossy harmonics filter present in most Dickson charge pumps.
Figure 3.6: Measured conversion efficiency comparison between two different Dickson charge pumps and the proposed single-stage Greinacher rectifier at 2.45 GHz.

Fig. 3.6 provides measured performance comparisons between two different Dickson charge pumps and the modified Greinacher rectifier at 2.45 GHz. In the first measurement setup, both circuits are built with the same components: Avago HSMS-2852 Schottky diodes, 100 nF Panasonic ECH series capacitors, and a 1 kΩ resistance as the DC load. As seen in Fig. 3.6, both circuits have similar RF-to-DC conversion efficiency at the low power range. However, as the input RF power approaches the medium power range, the proposed Greinacher rectifier outperforms the Dickson charge pump. The Dickson charge pumps has problems with the strong higher order harmonics, and more importantly, the reverse breakdown at D4 (in Fig. 3.5(b)). In the second setup, the components used to build the proposed Greinacher rectifier are the same, while the diodes used in the Dickson charge pump are changed to Avago HSMS-2862. The new diode, compared to the HSMS-2852, features lower saturation...
current, lower series resistance, higher diode barrier, and higher reverse breakdown voltage. As depicted in Fig. 3.6, the second Dickson charge pump and the modified Greinacher rectifier have the same peak conversion efficiency of 70%. However, the proposed Greinacher rectifier offers significant efficiency improvement over the Dickson charge pump at the ultra low power region, where efficiency is vital, and still maintains good efficiency performance ($\eta \geq 60\%$) at the medium power region.

![Figure 3.7: Schematic of the second stage full-wave modified Greinacher rectifier.](image)

Transitioning from single stage to two-stage rectification is not easy with the proposed modified Greinacher topology. Fig. 3.7 presents the circuit schematic of the two-stage rectifier, where the storage capacitors can be combined into one capacitor in each stage, $C_{s1}$ and $C_{s2}$. This way, the number of components in the system is reduced, which improves the overall system efficiency. More importantly, the low power input impedance of the proposed rectifier is smaller than the four-stage Dickson charge pump version. This is a clear advantage when the antenna-rectifier matching issue is addressed.
3.3 A Planar Rectenna with a Modified Greinacher Rectifier

This section presents a planar rectenna design to replace or recharge existing batteries in consumer electronics by scavenging the electromagnetic power from nearby RF devices operating at the 2.45 GHz ISM band. The designed rectenna structure combines the single-stage modified Greinacher rectifier circuit presented in Section 3.2 (see circuit schematic in Fig. 3.5(a) and performance in Fig. 3.6) with a miniature Koch shaped fractal patch antenna. The proposed rectenna achieves a relatively high realized gain (4 dBi) and good RF-to-DC conversion efficiency (up to 70%). The advantage of the modified Greinacher rectifier is its higher conversion efficiency at lower input RF power levels and the advantage of fractal element antennas, when compared to conventional antenna designs, center around size and bandwidth. The application of fractal geometry to conventional antenna structures optimizes the shape of the antennas in order to increase electrical length, thus reducing overall size.

Simulated and measured return loss and realized gain for the utilized antenna are given in Fig. 3.8, respectively. As seen, the simulations are in good agreement with measurements. It is also shown that the proposed antenna resonates at 2.45 GHz. However, the measured bandwidth is slightly narrower. Fig. 3.9(a) presents the PCB layout and a photograph of the final rectenna design, which is fabricated on a 50 mil thick low-loss RO3006 material \((\varepsilon_r = 6.15)\). Fig. 3.9(b) depicts the measurement setup created to evaluate the rectenna performance.

As seen in 3.9(b), the measurements are taken in an RF quiet environment with both the transmitter and the receiver placed on top of ten foot foam columns. As for the RF source, a commercial RFID interrogator transmitting 4W EIRP at 2.45 GHz ISM band is used. The interrogator’s antenna and the proposed rectenna are placed
Figure 3.8: Simulated and measured return loss (a) and realized gain (b) performance of the antenna element.

Figure 3.9: PCB layout and the photograph of the fabricated version of the proposed rectenna (a) and the measurement setup used to evaluate the rectenna performance (b).
such that both antennas are aligned in the direction of maximum power transmission. The measurements evaluated the effectiveness of the energy harvesting circuit by varying the transmitter-receiver separation and observing the LED, which is connected to rectenna’s output. A minimum voltage of 1.5 V is necessary to turn on the specific LED, which illuminates bright red when the interrogator-rectenna separation is at or within 4 meters.

3.4 Conclusion

This chapter first demonstrated that the RF-to-DC conversion efficiency of a rectifier changes with the operating power level. Subsequently, inherent inefficiencies of traditional rectifiers at low and high input power regions were discussed. This chapter showed that,

- Conversion efficiency performance of a conventional rectifier design deteriorates at the high power region due to the generation of higher order harmonics which is filtered out and wasted.

- The inherent efficiency of a conventional rectifier design in low power region stems from the trade-off between the reverse breakdown voltage and the forward voltage drop of the diode.

Accordingly, two novel circuits were presented in this chapter to exploit these properties. First circuitry presented in this chapter was a novel rectifier design that can harvest energy from the higher order harmonics and achieved an improved efficiency performance at the high power region. The second rectifier topology offered significant efficiency improvement over the traditional designs at the low power region, while maintaining the good efficiency performance at the medium power region.
The chapter concluded with a simple, proof-of-concept rectenna design that incorporates the presented low-power rectifier. The developed rectenna harvested enough energy from a commercial RFID interrogator that is four meters away (4W EIRP at 2.45 GHz ISM band) to power a 1.5 V LED, which is enough voltage to energize many commercial electronics.
Chapter 4: Investigation of Rectenna Array Configurations for Enhanced RF Power Harvesting

Recent advances in semiconductor technology and the introduction of passive RFIDs have provided additional impetus for power harvesting from radio waves [16]. Cell phone companies are already developing mobile devices that can be charged by harvesting ambient RF power [58]. Likewise, defense companies have been working on systems to power unmanned aerial vehicles (UAVs) in air by exploiting directed energy from microwave sources [59]. In practical applications, the power output from rectennas is determined by the power flux density, operating frequency, incident angle of incident microwave, and the rectifying circuit performance. The breakdown voltage of Schottky barrier diodes in each individual rectenna is also a limiting factor. Accordingly, devices operating at a low power level must intelligently manage the available power to meet user device requirements.

Typically, a single rectenna is not sufficient in supplying energy for reliable device operation because of its low incident microwave power. Alternatively, choosing a larger aperture and properly interconnecting multiple antennas could increase the available microwave power and provide sufficient rectification, respectively. However, aperture is a valuable resource and must be managed wisely. Hence, different rectenna
array configurations must be considered to maximize the power output from the RF harvester module.

![Diagram of RF and DC combining circuits](image)

Figure 4.1: Schematics of the investigated rectenna array configurations. $P_{RF}^i$ is the incident RF power impinging on the antennas, $P_{R}^H$ refers to the harvested DC power by RF-combiner topology, and $P_{D}^H$ denotes the harvested DC power by the DC-combiner topology.

In one configuration, multiple antennas can be arranged to channel the RF power to a single rectifier [60] (see Fig. 4.1(a)). In a wireless power transmission application, this configuration offers the most efficient power transfer scheme. In another approach, each antenna has its own rectifier that can separately harvest DC power (see Fig. 4.1(b)), which can then be combined in parallel, series, or a hybrid manner [61,62]. This setup is suitable for very large rectenna arrays (by avoiding complex feeds) or harvesting ambient RF power (by eliminating random polarization effects). However, in the case of mobile consumer electronics, the issues are different as the
transmitted energy is controlled and a broader reception is necessary. Therefore, an analytical method that can evaluate the performance of rectenna array configurations is useful.

In this chapter, the advantages and disadvantages of the two RF power harvesting configurations shown in Fig. 4.1 are discussed. First, an analytical approach that evaluates the power harvesting performance of the given rectenna topologies under different propagation conditions (indoor and outdoor) is presented. Then, an efficient rectification method using a miniaturized antenna is developed to better utilize the available aperture. Measurements are presented to evaluate the rectenna’s performance in urban and indoor environments (GSM-1900 and 2.45 GHz ISM bands respectively) and compared to analytical predictions (for a 2×2 antenna array).

4.1 Analytical Approach

RF-to-DC conversion efficiency, $\eta$, is of paramount importance for optimal wireless power transmission. Consequently, the conversion efficiency is defined in this chapter as

$$\eta = \frac{\text{Harvested DC Power}}{\text{Input RF Power to Rectifier}} = \frac{P_{\text{DC}}}{P_{\text{RF}}}.$$ (4.1)

As shown in Chapter 2, the nonlinear nature of diodes complicates the evaluation of $\eta$ via analytical means. Specifically, for most rectifier circuits, $\eta$ changes with RF input power, operating frequency, impedance matching, and diode properties (i.e., breakdown voltage, diode parasitics, etc.). In this study, the operating frequency is constant, and the diodes are identical, viz, $\eta = \eta(P_{\text{RF}})$. Given these assumptions, a simplified version of the rectifier model developed in Chapter 2 can be used to
estimate $\eta$. Hence,

$$\eta = \frac{V_{out}I_{out}}{\frac{1}{T} \int_0^T v_{in}(t)i_D(t)dt}.$$  \hspace{1em} (4.2)

In (4.2), $T$ is the period of the input RF signal, $v_{in}(t)$ is the input voltage to the rectifier, and $i_D(t)$ is the current flowing through the diode terminals. Also, $V_{out}$ denotes the DC voltage on the DC load, and $I_{out}$ is the current flowing through the load terminals.

Accurate calculation of $\eta$ necessitates precise estimation of the available RF power (to the rectifier) from antenna terminals. Many radio propagation models are available in literature to help with this calculation (for more detail, see [63]). Most of these models are derived using a combination of analytical and empirical methods. The empirical approach is based on fitting curves or analytical expressions that recreate a set of measured data. Over time, some classical propagation models have emerged incorporating the empirical approach [64–66] and are now widely used to predict the path loss in a given environment. This section begins by presenting modified versions of these propagation models (based on Friis transmission formula) to calculate the incident RF power impinging on the antennas, $P_{RF}^i$. Next, harvested DC voltage by the two topologies depicted in Fig. 4.1 is calculated with the mathematical tools developed in Chapter 2.

### 4.1.1 Outdoor Propagation

The mechanisms behind electromagnetic wave propagation are diverse, but can generally be attributed to diffraction, reflection, absorption, and scattering. In practice, the transmission path between the transmitter and the receiver can vary from a clear line-of-sight to one that is severely obstructed by buildings, mountains, forests,
etc. Due to multiple reflections from various objects, the electromagnetic waves may travel along different paths of varying lengths. As the distance between the transmitter and receiver increases, the strength of the waves will decrease, while the interaction between the waves can cause multipath fading at a specific location. Fortunately, large-scale propagation models (such as Okumura [67], Hata [64] or Lee) that incorporate the effects of these physical phenomena have been developed to estimate $P_{RF}^i$. In this chapter, a simplified version of the extended Okumura model is used.

\[
P_{RF}^i(\theta_t, \phi_t) = P_t G_t(\theta_t, \phi_t) \frac{\lambda^n}{\Delta} \left( \frac{1}{R} \right)^n e^{-\alpha R}. \tag{4.3}
\]

Here, $P_t$ is the input power to the transmitting antenna and $G_t(\theta_t, \phi_t)$ is the transmitting antenna’s realized gain in the direction $(\theta_t, \phi_t)$. Further, $\alpha$ denotes the effective decay coefficient in air ($\alpha = 0.001$), $m$ and $n$ are path loss exponents, and $\Delta$ is a constant. Typically, $n = m = 2$ in free space (unobstructed antennas, no multi-path) and takes a higher value in urban environments (no line-of-sight, strong multipath effects). Table 4.1 tabulates the typical values for $m$, $n$, and $\Delta$ in urban, suburban, and open areas. It should be noted that these values are approximated and valid under the assumption that the transmitter and mobile unit is 50 meters and 1.5 meters above the ground, respectively.

### 4.1.2 Indoor Propagation

Indoor propagation of electromagnetic waves is central to the operation of wireless LANs, cordless phones, and all other indoor systems that rely on RF communications.
Table 4.1: Typical values for path loss exponents in different areas.

<table>
<thead>
<tr>
<th></th>
<th>Urban Area</th>
<th>Suburban Area</th>
<th>Open Area</th>
<th>Free Space</th>
</tr>
</thead>
<tbody>
<tr>
<td>$m$</td>
<td>2.62</td>
<td>2.41</td>
<td>2.30</td>
<td>2.00</td>
</tr>
<tr>
<td>$n$</td>
<td>4.12</td>
<td>3.37</td>
<td>2.95</td>
<td>2.00</td>
</tr>
<tr>
<td>$\Delta$</td>
<td>12.20</td>
<td>2.19</td>
<td>7.16</td>
<td>$4\pi$</td>
</tr>
</tbody>
</table>

The indoor environment is considerably different from the typical outdoor environment and, in many ways, harsher [68]. Modeling indoor propagation is complicated by the diversity in building layouts, variability in construction materials, and the sensitivity of the RF environment to movement. For these reasons, deterministic models are often not used.

Indoor propagation depends upon reflection, diffraction, penetration, and, to a lesser extent, scattering. In addition to fading, these effects, individually and in concert, can degrade a signal. There are two general types of indoor propagation modeling: site-specific and site-general [69]. Site-specific modeling requires detailed information on building layout, furniture, and transceiver location(s). Not only is the knowledge of the building and materials limited in most environments, but the environment itself can change by simply moving furniture or doors. Thus, the site-specific technique is not commonly employed. Site-general models provide gross statistical predictions of path loss for link design and are useful tools for performing the initial design and layout of indoor wireless systems. In this chapter, a modified version of ITU’s indoor path loss model [70] is employed to calculate $P_{\text{RF}}^i$. Hence,

$$P_{\text{RF}}^i(\text{dB}) = P_t(\text{dB}) + G_t(\text{dB}) - 20 \log_{10}(f) + N \log_{10}(R) - Lf(n) - 28\text{dB}$$  \hspace{1cm} (4.4)
<table>
<thead>
<tr>
<th>Frequency</th>
<th>Residential</th>
<th>Office</th>
<th>Commercial</th>
</tr>
</thead>
<tbody>
<tr>
<td>900 MHz</td>
<td>–</td>
<td>33</td>
<td>20</td>
</tr>
<tr>
<td>1.8-2.5 GHz</td>
<td>28</td>
<td>30</td>
<td>22</td>
</tr>
<tr>
<td>5.2 GHz</td>
<td>–</td>
<td>31</td>
<td>–</td>
</tr>
</tbody>
</table>

Table 4.2: Power loss coefficient values, $N$, for the ITU Site-General indoor propagation model.

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Residential</th>
<th>Office</th>
<th>Commercial</th>
</tr>
</thead>
<tbody>
<tr>
<td>900 MHz</td>
<td>–</td>
<td>$9n$</td>
<td>–</td>
</tr>
<tr>
<td>1.8-2.5 GHz</td>
<td>$4n$</td>
<td>$15 + 4(n - 1)$</td>
<td>$6 + 3(n - 1)$</td>
</tr>
<tr>
<td>5.2 GHz</td>
<td>–</td>
<td>$16(n = 1$ only</td>
<td>–</td>
</tr>
</tbody>
</table>

Table 4.3: Floor penetration loss factor, $L_f(n)$, for the ITU Site-General indoor propagation model.

where $N$ is the distance power loss coefficient, $f$ is the frequency in MHz, $R$ is the distance in meters ($R > 1$m), $L_f(n)$ is the floor penetration loss factor, and $n$ is the number of floors between the transmitter and the receiver. Table 4.2 shows representative values for the power loss coefficient, $N$, as given by the ITU, and Table 4.3 gives values for the floor penetration loss factor, $L_f(n)$, as given by the ITU.

### 4.1.3 RF Combiner

Using the diode and radio propagation models, the total harvested DC power by the RF-combiner topology in Fig. 4.1(a) can be written as

$$P_R^H = P_{RF}^i G_r^{RF} \eta_r \xi.$$  \hspace{1cm} (4.5)
In 4.5, $\eta_r$ is the RF-to-DC conversion efficiency of the RF-combiner i.e., $\eta_r = \eta \left( P_{RF}^i G_{RF}^r \right)$. $\xi$ denotes the antenna polarization mismatch (if any), and $G_{RF}^r$ refers to the realized gain of the antenna array. An analytical calculation of $G_{RF}^r$ is possible for only a handful number of antenna types. Therefore, in this chapter, commercial numerical solvers are used to calculate the realized gain of the antenna array.

### 4.1.4 DC Combiner

The total harvested DC power by the DC-combiner topology shown in Fig. 4.1(b) can be calculated from

$$P_D^H = e_d \sum_{m=1}^{N} P_{RF}^i G_{m}^r \eta_d \xi. \quad (4.6)$$

In 4.6, $\eta_d$ refers to the RF-to-DC conversion efficiency of the DC-combiner i.e., $\eta_d = \eta \left( P_{RF}^i G_{m}^r \right)$. $N$ is the number of antennas in the array, and $G_{m}^r$ refers to the realized gain of the $m^{th}$ antenna element. $e_d$ denotes the efficiency of the DC combining circuit and may vary with the chosen DC combining topology (i.e., voltage-summing, current-summing, or hybrid).

### 4.1.5 Rectenna Topology Indicator

Rectenna Topology Indicator (RTI) function is introduced as a figure of merit in assessing the performance of rectenna array configurations as shown in Fig. 4.1. Specifically, the RTI function is defined as the ratio of final available DC power from the two topologies. Therefore,

$$RTI = \frac{P_R^H}{P_D^H}. \quad (4.7)$$
Substituting (4.5) and (4.6) into (4.7), an explicit expression for the RTI is obtained as,

\[
RTI = \frac{G_{RF}^{r}}{e_d \sum_{m=1}^{N} G_{r}^{m}} \frac{\eta \left( P_{RF}^{i} G_{RF}^{r} \right)}{\sum_{m=1}^{N} \eta \left( P_{RF}^{i} G_{r}^{m} \right)}. \tag{4.8}
\]

Further simplifications can be made to (4.8) depending on the antenna geometry. Assuming that the antenna elements are identical, i.e., \( G_a = G_m \), \( \forall m \), (4.6) becomes

\[
P_{H}^H = e_d N P_{RF}^{i} G_a \times \eta \left( P_{RF}^{i} G_a \right). \tag{4.9}
\]

Further, assuming that the rectenna topologies under investigation are planar arrays (having \( K \times L = N \) elements situated in the \( xy \)-plane) and no coupling exists among the array elements, \( G_r \) in (4.5) can be related to \( G_a \) in (4.9) via

\[
G_r = G_a \frac{\sin \left( \frac{K}{2} \psi_x \right)}{\sin \left( \frac{\psi_x}{2} \right)} \frac{\sin \left( \frac{L}{2} \psi_y \right)}{\sin \left( \frac{\psi_y}{2} \right)}, \tag{4.10}
\]

where

\[
\psi_x = k d_x \sin \theta \cos \phi \quad \text{and} \quad \psi_y = k d_y \sin \theta \sin \phi. \tag{4.11}
\]

As usual, \( k \) is the wavenumber, and \( d_{x,y} \) refer to interelement spacing in the \( x \) and \( y \) directions, respectively, and \((\theta, \phi)\) denote the spherical angles of the field incident onto the receiving antenna.

Under these assumptions, RTI takes the explicit form

\[
RTI(\theta, \phi) = \frac{e_r \sin \left( \frac{K}{2} \psi_x \right) \sin \left( \frac{L}{2} \psi_y \right) \eta_r}{e_d N \sin \left( \frac{\psi_x}{2} \right) \sin \left( \frac{\psi_y}{2} \right) \eta_d}. \tag{4.12}
\]

As seen 4.12, RTI is a function of the incident angle, \( \eta_r \) and \( \eta_d \). Note that \( \eta_r \) and \( \eta_d \) are just the rectifier efficiencies (i.e., \( \eta \)) that must be evaluated for different power levels. As can be surmised, for \( RTI > 1 \), the RF-combiner performs better than the DC-combiner. The converse is true for \( RTI < 1 \).
4.2 Rectenna Design Example

4.2.1 Rectifier Design

The efficiency of a rectifier design is critical for power harvesting; thus, in order to achieve high efficiency, two modified Greinacher rectifiers (introduced in Chapter 3) are utilized. One of the rectifiers is optimized for the GSM-1900 band (1850 MHz - 1990 MHz), while the other one is for the 2.45 GHz ISM band (2.40 GHz - 2.48 GHz). Fig. 4.2 depicts the layout of these rectifiers, along with inset pictures of the fabricated prototypes.

As seen in Fig. 4.2(a), a two-stage modified Greinacher rectifier (REC\textsubscript{GSM}) is designed to operate in the GSM-1900 band. Going back to the rectifier analysis in Chapter 3, parallel cascaded designs reduce the series junction resistance (reduced heat loss), but increase the total junction capacitance (increased harmonics loss) of the rectifier. Hence, a two-stage design in the GSM-1900 was found to provide a good trade-off between these two opposing factors.

However, as presented in Fig. 4.2(b), a single-stage modified Greinacher rectifier (REC\textsubscript{ISM}) is developed to operate in the 2.45 GHz ISM band. A single-stage design was chosen over the multi-stage design because of the higher reactive losses, introduced by the junction capacitances, are no longer tolerable. It should be noted that both REC\textsubscript{GSM} and REC\textsubscript{ISM} utilize four zero bias low barrier diodes in each stage. These diodes feature high saturation current and do not require additional biasing.

The impedance matching stage (also depicted in Fig. 4.2) is essential in maximizing the RF-to-DC conversion efficiency (by providing maximum power transfer from the antenna to the rectifier circuit). Designing the matching network is not straightforward, since the rectifier is a nonlinear load with a complex impedance that varies
(a) REC\textsubscript{GSM}, optimized for operation in GSM-1900 band

(b) REC\textsubscript{ISM}, optimized for operation in 2.45 GHz ISM band

Figure 4.2: Layouts of the rectifier prototypes, printed on RO3206. $w_1 = 72$ mil, $w_2 = 15$ mil, $w_3 = 196$ mil, and $L_1 = 171$ mil. Fabricated samples are shown in the inset pictures.

$\bullet$ represents the shorting vias, $\cdots$ marks the location of zero-bias diodes (HSMS-2852), $\cdots$ marks the DC load, and $\cdots \cdots$ marks the capacitors.
with frequency and input power level. One design approach is to model the rectifier circuit using experimental characterization at the minimum power level required by the application [49]. This can be done by measuring the input impedance (extracted from $S_{11}$) of the rectifier circuit (with all components) without a matching network at that power level. Using the impedance results from the experimental characterization (i.e., rectifier impedance) and assuming a 50Ω source load, the matching circuit design then becomes rather straightforward.

![Figure 4.3](image.jpg)

**Figure 4.3**: Measured $\eta$ for the two presented Greinacher rectifiers, operating at GSM-1900 and ISM-2450.

Fig. 4.3 depicts the measured $\eta$ of the fabricated prototype as a function of input RF power from -25 to +10 dBm. The measurements are recorded by utilizing a signal generator as the RF power source at the center frequencies and calculating $\eta$ by dividing the DC power dissipated on the 10 kΩ load by the input RF power. As plotted in Fig. 4.3, the measured $\eta$ increases monotonically with input power up to -5 dBm; then the performance deteriorates because of the increased current
flowing through diode terminals. Note that $REC_{GSM}$ performs similarly to $REC_{ISM}$ at low power levels and better at higher power levels. This can be attributed to the smaller reactive losses of $REC_{GSM}$ (since it operates at a lower frequency) and its two-stage design, which provides a smaller series junction resistance (due to reduced heat dissipation).

4.2.2 Antenna Design

Critical to the practicality of most mobile applications is the utilization of a small size antenna with a broad radiation pattern (e.g., microstrip patch) [48]. In this study, a probe-fed, shorted patch antenna is proposed to reduce the size and achieve dual-band operation (see Fig. 4.4). A 6.5 mm thick foam substrate ($\epsilon_r = 1.45$) was used between the rectangular radiating patch and the ground plane. The rectangular patch has dimensions of $36 \times 16$ mm$^2$ with an 1 mm wide, 40 mm long, L-shaped slit. The slit was cut in the rectangular patch to achieve an additional operating band at 2.45 GHz band; the lower operating band at 1.9 GHz is mainly controlled by the dimensions of the rectangular patch. A 2.5 mm wide shorting strip short-circuits the patch to the ground plane. Measured return loss for this probe-fed, shorted patch antenna is shown in Fig. 4.5(a). The proposed dual-band antenna covers the entire GSM-1900 band (used by AT&T in the U.S.) and 2.45 GHz ISM band (used by WiFi, Bluetooth, etc.).

Radiation characteristics of the proposed linearly polarized antenna are also investigated. Fig. 4.5(b) presents the boresight measured realized gain across the wide impedance bandwidth in which the antenna operates. The peak realized gain is demonstrated to be 5.1 dBi in the 2.45 GHz band, and the gain variations within
Figure 4.4: Geometry of a probe-fed shorted patch antenna for dual-band operation. The dimensions given in the figure are in millimeters.

Figure 4.5: Measured return loss (a) and boresight realized gain of the probe-fed shorted patch antenna are shown in Figure 4.4.
the operating bandwidth are small and reasonable. Hence, this small dual-band antenna \((0.22\lambda_0 \times 0.10\lambda_0\), at 1900 MHz) can be used for the investigation of rectenna topologies depicted in Fig. 4.1.

4.3 Rectenna Array Configurations

Using the designed shorted, probe-fed patch antenna, a \(2 \times 2\) planar array with \(\lambda_0/2\) interelement spacing (based on \(f_c = 2450\) MHz) was constructed. This interelement spacing was chosen to combat fading, prevent aliasing, and avoid grating lobes. In addition, the chosen spacing greatly simplifies the calculation of the RTI at 2.45 GHz ISM band by allowing certain assumptions about the array (such as no coupling between antenna elements). Therefore, the chosen spacing enables in fast evaluation of the proposed analytical solution at 2.45 GHz. Further, one can notice that the electrical spacing between the antenna elements is smaller than \(\lambda_0/2\) at the GSM-1900 band. Accordingly, the assumptions made for 2.45 GHz are no longer valid in this band, essentially increasing reliance on numerical methods for the calculation of RTI.

Fig. 4.6 shows the top view of the fabricated array, comparing its size to a quarter dollar (a U.S. coin). It is important to note that the \(2 \times 2\) array is fabricated in such a way that it can support both of the configurations depicted in Fig. 4.1.

4.3.1 Indoor Evaluation

The indoor power harvesting capabilities of the \(2 \times 2\) array were evaluated in room 157 at the Ohio State University (OSU) ElectroScience Lab, where several boxes constituted the only furniture. For the RF source, a commercial linearly polarized dipole antenna (Cisco ANT2422SDW), transmitting 1W of CW signal at 2.45 GHz, was
mounted on the ceiling of room 257, one floor above the room 157. Both rooms are equal in size: 19 feet long, 28 feet wide, and 11 feet deep. The antenna array under test was placed in room 157 on a 4 foot foam block, with its boresight always pointing to the ceiling of the room. A four-way RF combiner (MiniCircuits ZB4PD-42+) and four 5” coaxial cables were utilized to feed the 2×2 array when in the RF-combiner configuration. The total insertion loss for the RF combining unit was 1.2 dB. The voltage-summing method was utilized for the DC-combiner configuration, and the efficiency of the combination scheme was measured to be 90%. Note that, the DC loads of the rectifiers were optimized separately for each array configuration (for the initial case, at the center of the room) for a fair comparison of the topologies in Fig. 4.1. The RF-combiner had a load impedance of 10kΩ, and the DC-combiner had a total combined load impedance of 9.6kΩ. The experiment was conducted by a computer-controlled multimeter, which was connected to the DC output terminal,
Figure 4.7: Calculated and measured heatmap of the harvested voltage by the investigated rectenna array configurations in an office environment.
while the 2×2 array was moved around the office. Note that a hundred measurements are averaged per foot and per configuration, i.e., 100×19×28×2 measurements are taken in total. The average DC voltage recorded by the multi-meter at each measurement was then used to calculate the measured DC power for each configuration. In the end the measured results were used to characterize the performance of the two rectenna array topologies in the indoor office environment.

Fig. 4.7 compares the measured and calculated harvested DC voltage from both configurations. As seen in Fig. 4.7, excellent agreement has been observed between the measurements and analytical predictions. Keep in mind that in calculating analytical predictions, the parameters were selected from Table 4.2 and Table 4.3 for the specific office environment, and the antenna radiation performance was calculated using numerical computation tools (Ansoft HFSS). One might notice a difference between the predicted voltage and the measured voltage at the edge of the office (width = 28 feet in Fig. 4.7), which corresponds to a metallic wall that was not properly accounted for in the analytical predictions. Another key observation about Fig. 4.7 is that the harvested voltage is maximized in the center of the office when the RF-combiner is used. This is because the RF-combiner has higher gain and captures more power per rectifier at the center of the office. With more power, the rectifier operates more efficiently and harvests more voltage. On the other hand, the DC-combiner offers a broader pattern and is less sensitive to the variations in the positioning of the device under test. Therefore, on average, the DC-combiner generates more voltage than the RF-combiner in this experiment.
4.3.2 Outdoor Evaluation

The outdoor power harvesting capabilities of the 2×2 array were evaluated at OSU’s west campus, where a 100 Watt AT&T cell phone tower\(^2\) that supports the GSM-1900 standard was chosen as the RF source. The cell phone tower employs six identical Commscope sector antennas, each of which is dual-polarized and has 18 dBi realized gain, but is omni-directional in the azimuth plane. The antennas’ long and narrow form gives them a fan-shaped radiation pattern, that is wide in the horizontal direction (66° HPBW) and relatively narrow in the vertical direction (10° HPBW). Typically, for base station antennas, there is a downward beam tilt in the radiation pattern to cover the immediate area more effectively.

![Diagram of radiation pattern and geographic area](image)

Figure 4.8: Normalized measured and calculated DC power (dBm - normalized) harvested by the two different rectenna topologies (a) and the outdoor environment (b).

\(^2\)This tower is installed by and registered under Cincinnati SMSA Limited Partners
Fig. 4.8 shows the physical location of the cell phone tower. As seen within the concentric red circle, this is a suburban area with large open areas, few houses, and dense traffic. The antenna array under test was placed on a 1.5 meter wooden block, with its boresight always pointing south. Again, a four-way RF combiner (MiniCircuits ZB4PD-42+) and four 5” coaxial cables were utilized to feed the 2×2 array when the RF-combiner configuration and the voltage-summing method has been utilized for the DC-combiner configuration. At 72 discreet angles, one thousand measurements per angle for each rectenna array configuration (a total of 72 × 1000 × 2 measurements) were taken along the red circle depicted in Fig. 4.8(b) (R =275 meters). The yellow star in Fig. 4.8 denotes the point where the first measurement was taken, which corresponded to 0 degrees. The harvested DC voltage (1000 measurements at each angle were collected and averaged) was used to calculate the DC power for each configuration. A TiePie Handyscope HS3 USB controlled multimeter was used to take the DC voltage measurements. To calculate the analytical predictions, the suburban radio propagation model was used and commercial numerical tools were utilized to estimate the antenna radiation characteristics.

Fig. 4.8(a) plots the normalized measured and calculated harvested DC power from both configurations. Also, Fig. 4.9 plots the measured and calculated RTI. As seen from Fig. 4.9, the RF-combiner offers better performance at boresight. However, the DC-combiner configuration performs better, as the array is rotating away from the normal incidence (greater than ±20°). The better performance is due to the broader radiation pattern of the individual elements, whereas for the array, the rectifier observes a narrower beam. Regardless, it is important to note that the measurements are in agreement with theoretical predictions.
Again, the first topology combined the RF signal from the antenna array to a single rectifier. This topology has the advantage of harvesting more power near the main beam, which is due to the higher power fed to a single rectifier, i.e., utilizing the diodes more efficiently. The other topology rectified the received RF signal of each antenna element prior to combining it at the DC output. This topology can harvest more power at angles away from broadside, as each rectifier is connected to the individual antenna elements and responds to the broad pattern of that element.

4.4 Conclusion

A method for comparing the harvested RF power by two different rectenna topologies under different propagation conditions (indoor and outdoor) was presented. This method was validated using a fabricated 2×2 antenna array employing small
(0.22\lambda_0 \times 0.10\lambda_0, at 1900 MHz), yet efficient (up to 70% conversion efficiency) rectenna elements.

The first topology harvested more power near the main beam, therefore utilizing the diodes more efficiently. On the other hand, the second topology harvested more power at angles away from broadside. Thus, the second topology is less sensitive to incidence angles. For a fair comparison of these configurations, the Rectenna Topology Indicator (RTI) parameter is introduced, where RTI is defined as the ratio of final available DC power from these two topologies. For RTI > 1, the RF-combiner performs better than the DC-combiner. The converse is true for RTI < 1. This chapter shows that:

- A single-element rectenna may not be sufficient in supplying the minimum required power for reliable device operation.
- Interconnecting several rectennas in a power efficient manner calls for innovative configurations of power-harvesting frontends.
- The angle where RTI intersects the equilibrium line is determined by \( \eta \) and \( N \), the number of antennas in the array.
- An increase in \( N \) will extend the region where DC-combiner performs better.
- \( \eta \), RF-to-DC conversion efficiency of the rectifier circuit, is a function of input RF power. Rectifiers that exhibit convex \( \eta \) will extend the region where RF-combiner performs better.

The method presented in this chapter can determine which configuration will perform best without actually building and testing the rectenna arrays.
Chapter 5: Design of an Efficient Ambient WiFi Energy Harvesting System

Research efforts to harvest ambient RF energy have gained impetus since the late 90s due to the growth of RF transmitting devices and the availability of low-power consumer electronics. Among previous works, Hagerty et. al. [61] presented a broad-band rectenna array (DC-combiner) that attempted to harvest ambient RF power over a frequency range of 2-18 GHz. Also, in 2009, Intel Research Seattle demonstrated ambient RF energy harvesting (RF-combiner) possible from 2.55 miles (≈4.1 km) away using a 960-kW TV broadcast station [71]. Powercast performed a similar demonstration in 2005 1.5 miles (≈2.4 km) away using a smaller power (5-kW AM) radio station [72]. However, these systems are typically only operate in the presence of physically large, very high gain antennas with a clear line-of-sight transmission, which significantly limits their mobility.

This chapter presents a novel, compact, and efficient microwave energy harvester that harvests very low level ambient energy to power a sensor system and its display from a typical WiFi router (transmitting 100 mW) within an office (depicted in Fig. 5.1(a)). The ubiquity of WiFi and its operation in the crowded 2.45 GHz band (used by Bluetooth, ZigBee, RFID, cordless phone, etc.) makes it the perfect candidate for ambient RF energy harvesting. In the following sections, the discussion
(a) 3D floor plan of the office. WiFi access point has been mounted on the ceiling of the office.

(b) 2D floor plan of the office.

Figure 5.1: Typical office environment for RF energy harvesting.
begins with the characterization of the ambient WiFi signal strength in an ordinary office environment (as shown in Fig. 5.1). Then, a DC-combiner rectenna array (using a miniaturized antenna) is developed to improve the harvesting of low-level ambient RF energy. Subsequently, a highly efficient power management system is presented to minimize leakage and provide uninterrupted regulated energy to the sensor. All these new components are integrated into a complete RF power harvesting system to drive low-power consumer electronics.

5.1 Assessment of Ambient RF Signal Strength of WLAN

Realistically assessing available ambient RF energy is essential to maximizing the performance of the harvesting circuitry. Of course, propagation characteristics heavily influence a mobile device’s received power, which significantly varies from location to location. This study is interested in RF energy harvesting while indoors and, therefore, begins with conducting an RF characterization (for WLAN) of the typical office environment depicted in Fig. 5.1.

Prior to presenting the details of the characterization, a basic understanding of WLAN (specifically, IEEE 802.11) is appropriate. The IEEE 802.11 standard specifies parameters for both the physical and medium access control (MAC) layers of a WLAN [73]. By combining existing measurement methods from other communication systems (such as GSM or UMTS) with the knowledge of the physical layers of a WLAN system, a spectral characterization method can be developed. Focusing on the modulation scheme used for 802.11b, it should be noted that 802.11b relies on a direct sequence spread spectrum (DSSS) with a chipping rate of 11 MHz [53]. However, other 802.11 schemes, such as 802.11g, use the hybrid complementary code
keying orthogonal frequency-division multiplexing modulation [74]. In any case, data transmission via the 802.11 protocol does not take place at a single frequency. Also note that the 802.11 protocol employs 11 transmission channels (13 in Europe, 14 in Japan), with the modulation spreading the data transmission over multiple channels for effective use of the frequency spectrum.

![Ambient RF Power Signal Strength](image)

Figure 5.2: Ambient RF signal strength measured on one of the desks in the office environment depicted in Fig. 5.1(a) with a standard monopole antenna.

In the far field, traditional RF equipment, such as an antenna with a spectrum analyzer as the receiver, can be used for characterization. For this study, the WiFi signal is measured from three orthogonal directions and from several positions within the office (depicted in Fig. 5.1) to observe and compensate for the fast fading. With this in mind, a quarter-wavelength monopole antenna (operating from 2.3 GHz to 2.5 GHz) is used, as well as an Agilent E4407B spectrum analyzer as the receiving end (see Fig. 5.1(b)). It should be noted that the room in Fig. 5.1(a) is adjacent to rooms
identical to itself; therefore, the traffic produced by smartphones and laptops from the room in Fig. 5.1(a) and the adjacent rooms fully use the WiFi spectrum. The monopole antenna listened to this environment while the spectrum analyzer recorded the power level of the captured RF signal. As expected, no RF signal is sent between the packages of data. Thus, the gaps between frequencies mean that full channel bandwidth will not be captured in one sweep. A way to circumvent this issue is to use the “max hold mode” option of the spectrum analyzer and to record the received RF signal over several sweeps. This approach provided a fair measure of the ambient RF power during the transmission of the data packages. Fig. 5.2 presents the measured WiFi signal strength taken over two minutes using this measurement approach. As seen in Fig. 5.2, the presence of heavy wireless traffic provides a considerable amount of ambient RF power available for harvesting.

5.2 Integrated Rectenna Design

A typical WiFi router transmits only 100 mW, almost ten million times less power than day-time TV broadcasting. Hence, efficient, low leakage, and compact RF harvesting at low power levels (≤-20 dBm) is of the utmost importance. Fig. 5.3 illustrates the block diagram of the proposed method for harvesting ambient RF energy in the 2.45 GHz ISM band while in the office space depicted in Fig. 5.1.

The first component of the RF-to-DC energy conversion system is the antenna. As seen in Fig. 5.3, an antenna array, instead of a single antenna, is used, since the incident WiFi power level is so low that a single antenna does not suffice. However, the array must still be small in size to make it practical. In this regard, the antenna element design and its miniaturization play an important role.
Figure 5.3: Block diagram of a rectenna array for ambient energy harvesting. Each element in the array is integrated with its own rectifier. The resulting DC outputs are combined and fed to power management electronics.

The next component of the RF harvesting circuitry is the rectifier. Once the RF signal is received, it must be rectified in the most efficient possible manner to generate DC power. Since the received WiFi signal is very low, to rectify the received RF signal, each antenna element in the array is integrated with its own rectifier. In the following section, the modified Greinacher rectifier using zero-bias Schottky diodes is discussed. The zero-bias diodes are important since the incoming signal is expected to be small. Therefore, the rectifier circuit should be turned on with the lowest power. Matching of the diodes to the rest of the circuitry and the antenna is critical to minimize reflections and therefore increase harvesting efficiency. However, doing so for the non-linear diode load is challenging.

Once the DC power is collected from all the array elements, the overall (harvested) DC voltage must be regulated by a power management block to ensure the delivery of a constant and on-demand DC voltage supply. A low-leakage capacitor was used
as the storage element in the management block, and a DC-to-DC converter start-
up IC was employed for stepping up low voltage levels. Below, the details of these
components are discussed.

## 5.2.1 Antenna Element Design

The use of a compact antenna is crucial in any mobile device. Planar patch anten-
nas are low-profile, conformal, lightweight, and easy to fabricate. However, they are
not essentially small in aperture. A popular solution for size reduction is to fabricate
the patch antenna on a high permittivity material (Rogers RO6010, $\epsilon_r = 10.2$, $d =
2.54$ mm has been chosen in this study). However, additional miniaturization can be
achieved by changing the patch design to a modified version of the Koch geometry.
The patch antenna design is shown in Fig. 5.4(a) and discussed in [75].

It is important to note that small size patches on high index materials are often
associated with degraded performance. This is likely due to the excitation of surface
waves [48,76]. Thus, patches on high dielectric substrates exhibit reduced efficiency,
degraded radiation patterns, and undesired coupling between the various elements
in array configurations. An approach to overcome these issues is to employ both
a substrate and a superstrate [77]. With this in mind, a superstrate layer (Rogers
RO6002, $\epsilon_r = 2.94$, $d = 0.5$ mm) is added to the Koch-shaped patch antenna (see
Fig. 5.4(b)), which also provides protection from the environment.

The antenna, with dimensions of $0.164\lambda_0 \times 0.162\lambda_0$, had an improved bandwidth
over a standard patch antenna through the use of a capacitively coupled probe feed.
Making the antenna dual polarized also allowed for reliable energy harvesting of the
ambient RF signals. Fig. 5.4(c) shows the magnitude of the measured $|S_{11}|$ and
(a) Fabricated prototype unit without the superstrate.

(b) Geometry of a probe-fed fractal patch antenna. The dimensions are in millimeters.

(c) Measured $|S_{11}|$ and total realized gain (boresight) of the proposed antenna element.

Figure 5.4: Antenna structure, its input impedance, and its realized gain performance at boresight.
realized gain of the designed antenna element operating at the 2.45 GHz ISM band. As seen in Fig. 5.4(c), the proposed antenna has a 6% bandwidth and a realized gain greater than 4.5 dBi at boresight.

5.2.2 Rectifier Circuit Design

Due to the low power transmission of a typical WiFi router (recall it only transmits 100 mW), power harvesting of ambient WiFi signals requires high efficiency circuitry. Towards this goal, the modified single-stage full-wave Greinacher rectifier (from Chapter 3) is employed. As shown in Fig. 5.5(a), four zero-bias low barrier Schottky diodes are used in the rectifier design, implying higher output voltage even though the ambient RF power is low. The only drawback of building the proposed rectifier with these diodes is the resulting higher series resistance, which implies that 100% conversion efficiency can never be achieved.

The operation of the single stage modified Greinacher rectifier is discussed in detail in Chapter 3. A modified Greinacher rectifier provides a significant conversion efficiency improvement over conventional rectifier configurations in the low power region, yet maintains excellent efficiency performance at the medium power region. While other rectifier configurations that can operate near theoretical performance bounds of power extraction are mentioned in literature [56, 78, 79], this dissertation uses a modified Greinacher rectifier because it can be built using discrete off-the-shelf components; viz. easier to fabricate.

The impedance matching stage of the RF harvesting circuit is vital in providing maximum power transfer from the antenna to the rectifier circuit. However, designing a matching network is challenging since rectifier diodes are nonlinear devices with
complex impedances that vary with frequency, input power level, and load resistance. Understandably, for optimal matching performance, these parameters must be determined prior to designing the matching network. The RF energy harvester designed in this chapter has the same frequency of operation as WiFi, 2.4 GHz to 2.48 GHz. Also, as indicated in Fig. 5.2, the anticipated input power is between -40 dBm and -20 dBm, affecting the design of the matching network. Further, the load resistance for this design was 10 kΩ, a decision that will be justified in the next subsection. With these parameters, a simulation model for the rectifier was created using Agilent ADS. Overall, the matching circuit design is rather straightforward given the simulated rectifier impedance and assuming a 50Ω source.
With the above in mind, a rectifier prototype was fabricated. Fig. 5.5 shows the measured reflection coefficient ($S_{11}$) of this rectifier as a function of frequency and input RF power level. As seen in Fig. 5.5, the rectifier is well matched when the input power is between -40 dBm and -20 dBm and the operating frequency between 2.4 GHz and 2.48 GHz. The power harvesting capabilities of the fabricated rectifier are depicted in Fig. 5.6. Note that, for the same load resistance, the newly designed RF power harvester generates approximately three times more voltage compared to the state-of-the-art technology. More specifically, at low power levels, the proposed rectifier circuit has a better conversion efficiency than any power harvester currently on the market.

Figure 5.6: Simulated conversion efficiency of the proposed rectifier (a) and the measured input RF power vs. harvested DC voltage for various power harvesters (b). A picture of the proposed power harvester is given on top left in (b). P2110 performance is evaluated by connecting the load resistor in parallel to its supercapacitor terminal.
5.2.3 Array Design

A single rectenna usually does not harvest sufficient energy to reliably power a device. Instead, multiple antennas can be arranged to capture a greater percentage of ambient RF energy and channel it to a single rectifier [80]. In a point-to-point RF system (pencil beam), this configuration offers the most efficient power transfer scheme. Alternatively, each antenna can incorporate its own rectifier to harvest DC power [81], which then be summed in parallel (current summing), series (voltage summing) or hybrid manner. These configurations are most suitable when dealing with large rectenna arrays (as they avoid complex feeds) or harvesting ambient RF power (as they eliminate nulling effects). Since the goal is to harvest the ambient RF power, the second configuration (voltage summing) is adapted in this chapter.

![Photographs of the fabricated RF power harvester.](image)

(a) The antenna array.  
(b) The rectifier array.

Figure 5.7: Photographs of the fabricated RF power harvester.
To generate sufficient DC power for the range of input power levels considered in this study, a $3 \times 3$ planar array of simple Koch-type patch antennas was designed. The interelement spacing for this array was chosen, such that mutual coupling between the antennas is low (i.e., $|S_{21}| < -10$ dB). Further, each antenna feed location was individually optimized to preserve the bandwidth performance (see Fig. 5.4(c)) of the single element. Fig. 5.7(a) shows a photograph of the fabricated antenna array (the superstrate layer is omitted). Note that the resulting physical size of the final antenna array is $9\text{cm} \times 9\text{cm}$ with the rectifiers built in the layer below the antenna array. Both the antenna array and the rectifier share the same ground plane. Fig. 5.7(b) shows a photograph of the fabricated rectifiers, and Fig. 5.8 provides the details of the layered design. The shared ground plane is marked with red dashed lines on Fig. 5.8.
A crucial design aspect of the rectifier array is to achieve optimal DC combining efficiency. Previous work [82] shows that predominantly parallel connections lead to smaller matched loads for the rectenna. However, for the series-connected array, as is case in this chapter, the matched load value is much higher. This is achieved by reducing the array to a combination of DC Thevenin sources with the matched load corresponding to the total source resistance. Regardless, the impedance of the energy management unit should be known for matching with the rectenna array. It turns out that the energy management unit designed in the next subsection has a resistance that varies between $85\text{k}\Omega$ and $105\text{k}\Omega$ when operational. Therefore, the series connected rectenna is a better choice since it performs well with these large loads [75]. The DC combining lines used in the design are depicted in Fig. 5.7(b).

As noted, the nonlinear performance of the diodes must be accounted for when considering the DC connections. The single rectifier properties may not be valid in an array setting without proper care. That is, the load used in the single rectifier may be different than the load within the array setting. In addition, the new DC lines may create unforeseen inductances, a potential design flaw. Here, the single rectifier was optimized for the $10\text{k}\Omega$ load, a reasonable assumption, as there are nine elements in the array and the total load varies between $85\text{k}\Omega$ and $105\text{k}\Omega$, implying an impedance of $9.4\text{k}\Omega$ to $11.6\text{k}\Omega$. Further, the DC lines were designed to be very thin and strategically placed to add minimal inductance to the load. With these precautions, it is safe to use the single element impedance.
5.2.4 Energy Storage and Management

Mobile devices operate in a variety of unknown conditions, with variations in the available ambient RF field and in rectenna characteristics. These variable conditions create significant challenges in maximizing the harvested RF energy and creates necessity for power management circuitry. The power management circuit serves to optimize the harvested power over a wide range of operating conditions independent of the load behavior. That is, power management circuitry delivers steady-state power to the load during harvesting. However, one concern is the duty cycle. Due to the sensors’ infrequent monitoring of physical quantities, their duty cycle of operation and average power requirement are low. For example, if a sensor system requires 3.3V at 30mA (100mW) while awake, but is only active for 1/100th of a second, the average power requirement is only 1mW. Further, if the same sensor only samples and transmits once every minute instead of once every second, the average power plummets under 20μW. Hence, with the help of a very low leakage, high capacity capacitor with a suitable energy management circuit, many sensor systems can be operational by harvesting ambient WiFi signals.

To realize the energy management circuit block, S882 and AS1310 integrated circuits are used to step up and regulate the output voltage from incoming harvesting circuits (see Fig. 5.9). AS1310 is a hysteric step-up DC-DC converter from Austria Microsystems that has ultra low quiescent current (< 1μA). Thus, it can even operate when the harvested DC voltage is as small as 0.7V [83]. On the other hand, S882Z from Seiko Instruments is a charge pump IC that improves the performance of any step-up DC-DC converter [84]. It is capable of stepping up the voltage to a level that allows AS1310 to start up. Next is the operation of the energy management
unit. When a voltage of 0.3 V or higher is harvested, the oscillation circuit inside S882 becomes operational and creates a clock signal. Subsequently, the CLK signal drives a charge pump circuit that steps up the harvested voltage and stores it in the CCPOUT capacitor, depicted in Fig. 5.9. When the voltage on CCPOUT reaches a certain prespecified level, the power begins to flow to AS1310. Ultimately, AS1310 converts the low input voltage to usable regulated output voltage (VOUT = 1.5V in this study) and powers the sensor. It should be noted that the CCPOUT is a low leakage capacitor that ensures enough energy is stored before the AS1310 chip feeds the sensor.

5.3 Ambient WiFi Energy Harvester

The aforementioned antenna array, rectifier array, and energy management units were combined to form the ambient WiFi energy harvester (see Fig. 5.8). This harvester was tested under a variety of operating conditions. To measure the performance, a monopole antenna and a spectrum analyzer were placed next to the power...
<table>
<thead>
<tr>
<th></th>
<th>Minimum</th>
<th>Typical</th>
<th>Maximum</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output Voltage</td>
<td>1.494 V</td>
<td>1.5 V</td>
<td>1.506V</td>
</tr>
<tr>
<td>Output Current</td>
<td>0</td>
<td>10μA</td>
<td>50μA</td>
</tr>
<tr>
<td>Time to Initialize</td>
<td>3 min.</td>
<td>5 min.</td>
<td>20 min.</td>
</tr>
<tr>
<td>RF Power Sensitivity</td>
<td>-40 dBm</td>
<td>-30 dBm</td>
<td>0 dBm</td>
</tr>
</tbody>
</table>

Table 5.1: Performance ratings of the proposed microwave energy harvester.

harvester to monitor the RF power sensitivity of the harvester. Table 5.1 summarizes the test results, where it is seen that the device can supply battery-like regulated voltage to the load.

The current generation capabilities of the device were evaluated by connecting a variable resistor and measuring the voltage across the resistor terminals. In a typical office environment and under the conditions depicted in Fig. 5.1, the device was able to supply a continuous current of 10μA to the load. The maximum measured output current under typical ambient WiFi conditions was 50μA (when ten WiFi devices were used to create wireless traffic). Although not the focus of this chapter, the device was also tested with dedicated RF, where an RFID interrogator served as the energy source (located 5 meters away from the harvesting device). In this scenario, the harvester generated 780μA.

One important parameter in Table 5.1 is the initialization time, which refers to the time needed for the storage capacitor to reach a certain voltage and the point where the management circuit powers the load. Typically, it takes five minutes for the harvesting device to initialize; however, as the WiFi traffic declines, the initialization time increases, reaching 20 minutes when only two WiFi devices were communicating with the WLAN access point. The shortest initialization time was observed when
The RFID interrogator was used as the energy source, ten seconds. As expected, an increase in the wireless traffic in the 2.45 GHz band yielded a reduction in the initialization time.

![Figure 5.10: Proposed RF energy harvester powers a thermometer (including display) with harvested ambient WiFi power.](image)

The ambient RF harvesting system was tested in a practical setting by powering a commercially available thermometer, which simultaneously measured both indoor and outdoor temperatures and indoor humidity [85]. By design, one 1.5V AAA battery is sufficient to power this device and its large LCD display. The thermometer was electrically characterized and measured to consume around 10μA at 1.5V (on average) from a laboratory power supply. Notably, about once every 30 seconds, its current consumption spiked to approximately 25μA, presumably when the sensor records an actual measurement.

The thermometer was still functional when the battery was removed because of the power supplied by the energy harvesting circuit. Further, the display read well
and temperature and humidity measurements were accurate (example provided in Fig. 5.10). In Fig. 5.10, four WiFi enabled devices were communicating (downloading files) with the access point mounted on the ceiling of the office. After the initialization period, the device operated uninterrupted. In another scenario, the number of WiFi communicating devices was reduced to two. After a 20 minute initialization period, the device started, but stopped working after 10 minutes. When there was no wireless traffic (i.e., router only transmits the beacon signal and no WLAN enabled devices are present), the device stopped operating altogether.

5.4 Conclusion

A highly efficient rectenna system that can generate battery-like voltage to run a variety of low-power consumer electronics was demonstrated. The RF harvesting module consisted of (a) 3×3 miniaturized antenna array, (b) novel rectifier circuitry efficient even at low power levels, and (c) power management circuitry. The final design generated DC voltage even when the received power was as low as -40 dBm. Thus, it could operate even when very small amounts of RF energy were available. A 9cm×9cm×1cm prototype was built and demonstrated that the RF harvesting design can deliver enough energy to power an off-the-shelf temperature and humidity meter with an LCD display. It was powered using nothing more than ambient WiFi signals in an office environment.
Chapter 6: Development of a Novel Multi-band Ambient Microwave Energy Harvesting Module

Supplying DC power through wireless transmission has been proposed and researched since the 1950s, mostly in the context of dedicated high-power beaming [20]. In dedicated microwave power transmission, the antennas have well-defined polarization and rectifiers have high rectification efficiency enabled by single-frequency, high microwave power densities incident on a rectenna array. Linearly, dual-, and circularly polarized receiving antennas were used for demonstrations of RF-to-DC conversion efficiencies ranging from around 85% at lower microwave frequencies to around 60% at the X-band and around 40% at the K_u-band [51]. Applications for this type of power transfer include helicopter powering [20], solar-powered satellite-to-ground transmission [86], intersatellite power transmission [51], mechanical actuators for space-based telescopes [87], small DC motor powering [86], and short-range wireless power transfer, e.g., between two parts of a satellite.

Over time, improvements in RF energy harvesting technology is expected lead to increased use cases and market extensions [88]. Currently, the technology is evolving from a paired system with the need for a dedicated transmitter to a single-sided system with the ability to fully capture radio waves emitted from existing and commonly used ambient RF energy sources, such as mobile base stations, TV and radio
transmitters, microwave radios, and mobile phones [61]. However, harvesting ambient RF energy is challenging as ambient microwave sources generate incident RF power densities that are orders of magnitude lower than those associated with the projects in the literature cited above. Therefore, received RF power level for harvesting is generally very low when compared to power-beaming applications. In addition to the drop in received power, a nonlinear decrease is observed in conversion efficiency at low RF power densities due to the nonlinear nature of the diode (studied in Chapter 2).

Overall, the harvestable power from ambient RF energy sources is very small, raising concerns regarding the usefulness of low-power rectification. Hagerty et. al. [61] refuted these concerns in 2004 by presenting a broad-band rectenna array that generates 100 nanowatts of DC power from ambient RF signals over a frequency range of 2-18 GHz. Furthermore, in 2009, Intel Research demonstrated harvesting 50 μwatts from ambient TV signals possible [71]. Similarly, in Chapter 5, the presented energy harvesting module generated 25 μwatts from nothing but ambient WiFi signals.

Ambient RF energy harvesting (i.e. power out of thin air) can enable wireless power to be as ubiquitous as wireless communications for micro-power applications. Ubiquity of power can make the “Internet of Things” a reality with untethered, autonomous, self-powered machine to machine (M2M) devices [72]. These devices (e.g. energy harvesting wireless sensors) could transmit data to local access points or through the mobile network with text messages. However, the RF energy harvesters mentioned above still cannot generate enough DC power and be mobile at the same time, thus cannot realize the “internet of things”. Intel’s system relies on large antennas and the RF harvester presented in Chapter 5 is limited to indoors and requires WiFi traffic. As expected, increasing the operational bandwidth of the rectifier can
enhance the amount of generated DC energy. Reportedly, Nokia Research is working on a rectenna which will be operational from 500 MHz to 10 GHz [58]. The ultimate goal is to get in excess of 20 miliwatts, enough power to keep a feature phone in standby mode indefinitely without having to recharge it.

This chapter presents a multi-band ambient microwave energy harvesting element that can recycle ambient GSM-800, GSM-900, digital cellular system (DCS), personal communications system (PCS), third generation (UMTS-3G), and WiFi signals. The developed rectenna can typically generate 12 $\mu$watts of DC power not only in indoors but also in outdoors. Of course, the designed harvester can be incorporated into a rectenna array to increase the output DC power. In comparison, the array presented in Chapter 5 was composed of nine elements and generated a total of 25 $\mu$watts, that is approximately 3 $\mu$watts per element. In the following sections, the discussion begins with literature review of ambient signal strength characterizations. Then, the design details of the novel, multi-band rectenna are presented. In the end, the proposed rectenna is integrated with the highly efficient power management circuit from Chapter 5 into a complete RF power harvesting system.

### 6.1 Ambient RF Power Density Characterization

Chapter 5 presented a study to realistically assess the available ambient WiFi energy for harvesting in a specific indoor environment. Since the characterization of the propagation environment in that case was site-specific and narrowband, the assessment was easily completed by a monopole antenna and a spectrum analyzer. However, this chapter requires the knowledge of ambient power density levels for RF signals from 800 MHz up to 2.5 GHz, both in indoors and outdoors. Further,
generating data on RF power density created by the base station antennas and mobile radios is a complicated and expensive process [89].

Figure 6.1: Measured ambient RF power density vs. time (from COST-281) [92]

Fortunately, a number of European Union initiatives were taken in the past to measure the RF exposure of public to the GSM and WLAN radiations as a result of a growing concern about a potential relation between non-ionising cell phone radiation and health-risks, such as cancer. The most important of these initiatives were COST-281: “Potential Health Implications from Mobile Communication Systems” and the “European Information System on Electromagnetic Fields Exposure and Health Impacts” [90]. COST-281 presents measured data on the RF power density levels generated by GSM base stations and WiFi routers in several European countries (Germany, Belgium, France, Hungary, and Italy) [91]. This chapter uses the averaged data from COST-281 (see Fig. 6.1) to predict the incident RF power impinging on the multi-band rectenna element. Keep in mind that the data used in this study were gathered in Europe, where the base station density is much higher.
than the U.S. Therefore, the actual received RF power by the rectenna element in U.S. will be less than the RF power estimated by COST-281.

6.2 Integrated Rectenna Design

As seen in Fig. 6.1(a), ambient RF power density is higher at GSM-800, GSM-900, DCS, PCS, UMTS-3G, and WiFi bands. Interestingly, the power density is found to be highest at the low bands and slightly lower at the high bands. Therefore, the predicted RF power for rectification from individual mobile communications bands would possibly be comparable if a fixed size antenna aperture is used to capture it. Detailed calculations show that the ambient RF power for harvesting would approximately be in the neighborhood of -30 dBm (vary between -40 dBm and -20 dBm) at each of these bands if the rectenna aperture is 40 mm wide and 70 mm long. Then, the estimated ambient RF power at individual mobile bands would be comparable to the available ambient WiFi power density depicted in Fig. 5.2. Hence, a similar rectenna topology can be used here.

As discussed earlier in this dissertation, the first component of the RF-to-DC energy conversion system is the antenna. In Chapter 5, an antenna array, instead of a single antenna, was used since the incident WiFi power level was so low that a single rectenna did not suffice. In this chapter, however, the goal is to harvest RF energy from multiple mobile communications bands, therefore, a single rectenna can possibly provide sufficient energy to power consumer electronics. Accordingly, this chapter focuses more on developing the individual rectenna element, not the array structure as was in Chapter 4 and Chapter 5.
Once the RF signal is captured, it must be rectified in the most efficient possible manner to maximize the generated DC power. To achieve the goals of this chapter, the rectifier should be capable of harvesting ambient RF energy at various frequencies, i.e., impedance matched at all aforementioned bands. However, designing a matching network is challenging since rectifier diodes are nonlinear devices with complex impedances that vary with frequency, input power level, and load resistance. The details of the impedance matching network is discussed in the next subsection.

Of course, once the DC power is harvested, it must be regulated by a power management unit to ensure the delivery of a constant and on-demand voltage supply. Fortunately, the energy management circuitry designed in Chapter 5 can be used in this chapter. Hence, design details of the energy management module is omitted in here (refer to Chapter 5.2.4).

### 6.2.1 Rectifier Circuit Design

In low-power applications, as is the case for harvesting ambient microwave energy, there is generally not enough power to drive the diode in a high-efficiency mode (see Chapter 3). In addition, the diodes cannot be externally biased in these applications, therefore, it is critical to use a diode with very high saturation current and very low barrier height, i.e., low turn-on voltage. Unfortunately, the reducing the barrier height of a Schottky diode comes at the expense of reducing the reverse breakdown voltage [30]. Therefore, traditional rectifiers with charge pump units cannot utilize the diodes with the highest saturation current, otherwise they cannot tolerate the reduced reverse bias voltage at the medium power region (see Chapter 3). Consequently, the modified single-stage full-wave Greinacher rectifier (as in Chapter 5) is
employed in here. Modified Greinacher rectifier provides improvements in conversion efficiency over conventional designs in the low power region. Again, while other rectifier configurations that can operate near theoretical performance bounds of power extraction are mentioned in literature [56, 78, 79], this dissertation uses a modified Greinacher rectifier because it can be built using discrete off-the-shelf components.

Figure 6.2: Simulated range of input impedances for two SMS7630 Schottky diodes (different packaging) as the operating frequency ($0.75 \text{ GHz} \leq f_{\text{inc}} \leq 3 \text{ GHz}$) and input power ($-30 \text{ dBm} \leq P_{\text{in}} \leq 10 \text{ dBm}$) are varied (Smith chart normalized to $50\Omega$).

The impedance matching stage of the RF harvesting circuit is vital in providing maximum power transfer from the antenna to the rectifier circuit. In Chapter 5, the matching was comparatively easy as the operating bandwidth was very narrow, i.e., 3%. However, rectification over multiple octaves requires a different approach and detailed investigation of the diode behavior. Towards this goal, input impedance of
the Skyworks SMS-7630 diode was simulated with Agilent ADS over the frequency and input RF power range that the harvester module is expected to operate. Fig. 6.2 demonstrates the range of input impedances of the SMS7630 diode in two different packagings across the 0.75 to 3 GHz frequency range and from -30 to +10 dBm input RF power. The magnitude of the input impedance becomes smaller with increasing incident power. More significantly, the input impedance moves clockwise along a constant admittance circle with increasing frequency due to the junction capacitance.

A key observation from Fig. 6.2(a) and Fig. 6.2(b) is that the input impedance behavior of the same diode changes when a different packaging standard is used. Recall from Chapter 2 that a diode model includes package parasitics, represented by $C_P$ and $L_P$, and overlay capacitance $C_{OV}$. The effects of these parameters on the diode impedance must be minimized to achieve broader impedance matching performance from the rectifier. Fortunately, smaller packaging standards reduce the package parasitics of a diode at the expense of power handling. Therefore, in this chapter, diodes packed with SC-70 standard are preferred over the ones with SOT-23, the most popular industry standard [92].

With the above in mind, a rectifier prototype was fabricated (see Fig. 6.3(a)). Fig. 6.3(b) shows the measured reflection coefficient ($|S_{11}|$) of this rectifier as a function of frequency and input RF power level. As seen in Fig. 6.3(b), the rectifier is matched when the input power is between -40 dBm and -20 dBm and the operating frequency between 0.8 GHz and 2.5 GHz. The efficiency of the fabricated rectifier was simulated by Agilent ADS and depicted in Fig. 6.4 when the load resistance was 85kΩ. Note that the conversion efficiency declines as the operating frequency is increased and varies with the input RF power. The peak simulated conversion
Figure 6.3: Layout (a) and |S11| (b) of the fabricated rectifier prototype. Diodes: SMS7630-SC70, C1: 56pF, C2: 100pF C3: 100nF. 
\( w_1 = 1.83 \text{ mm}, \quad w_2 = 0.30 \text{ mm}, \quad L_1 = 5 \text{ mm}, \quad \text{and} \quad A_1 = 60^\circ \)

Figure 6.4: Simulated RF-to-DC conversion efficiency \( \eta \) of the proposed rectenna versus input RF power (-40 dBm \( \leq P_{in} \leq -20 \text{ dBm} \)) and frequency (0.75 GHz \( \leq f_{inc} \leq 3 \text{ GHz} \)).
efficiency is approximately 35% for -30 dBm input RF power, which is significantly better than any power harvester currently on the market.

6.2.2 Antenna Element Design

The rapid growth in mobile communication systems leads to a great demand in developing small antennas with multiband functions [93]. Planar antennas have several advantages over conventional monopole-like antennas since they are less prone to damage, compact in total size and aesthetic from the appearance point of view [94]. Hence, compact and low-profile structures have become increasingly attractive and popular in mobile applications [95,96]. Many new multiband designs based on planar folded monopole or planar inverted f-antenna (PIFA) concepts for achieving operation at the mobile bands discussed in this chapter have been reported in the open literature [94–96]. In here, a planar folded monopole antenna with a compact two-dimensional (2-D) structure is utilized (lightly modified from [97]) to capture the ambient RF energy in GSM, DCS, PCS, UMTS-3G, and WiFi bands.

The geometry of the utilized planar monopole antenna is shown in Fig. 6.5(a). The antenna was etched from Rogers a 25 mil thick TMM4 substrate ($\epsilon_r = 4.5$) and enclosed in an area of $38.5 \times 15$ mm$^2$ without the ground plane (the gray area in Fig. 6.5(a)). As depicted in Fig. 6.5(a), a long 50Ω microstrip line was feeding the antenna. Note that this microstrip line will be replaced by the rectifier circuit components in the final rectenna design. The tapered element in the antenna design was responsible from improving the impedance matching at the feed point and, as expected, was part of the radiating structure. Through investigation of the employed
Figure 6.5: Top view of the utilized planar monopole antenna (a), its measured impedance matching performance (b), and its simulated radiation efficiency (c).
Table 6.1: Detailed dimensions of radiating element. All dimensions are given in millimeters.

<table>
<thead>
<tr>
<th>Parameter:</th>
<th>a</th>
<th>b</th>
<th>c</th>
<th>d</th>
<th>e</th>
<th>f</th>
<th>g</th>
<th>h</th>
<th>i</th>
<th>k</th>
<th>m</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length (mm):</td>
<td>60.0</td>
<td>15.0</td>
<td>38.5</td>
<td>11.0</td>
<td>4.0</td>
<td>1.54</td>
<td>3.5</td>
<td>5.0</td>
<td>8.5</td>
<td>9.0</td>
<td>12.5</td>
</tr>
</tbody>
</table>

A prototype based on the presented design was fabricated and measured with Agilent PNA series Vector Network Analyzer. The measured return loss of the antenna is shown in Fig. 6.5(b). From an impedance matching point of view, the antenna was operational in all the mobile bands this chapter is interested in. However, since the antenna is very small, return loss performance does not tell the complete story. Hence, it is important to investigate the radiation efficiency of the antenna with respect to operating frequency. Note that in this study, the radiation efficiency is defined as the ratio between the realized gain and the directivity of the antenna. Ansoft HFSS was utilized to simulate the radiation efficiency and the results are depicted in Fig. 6.5(c). As seen, the compact folded monopole antenna realized an efficiency
of around 40% at the low-band and achieved more than 80% efficiency at higher frequencies. Fig. 6.6 shows the simulated radiation pattern of the utilized monopole at different frequency bands. The antenna had an omni-directional radiation pattern at the GSM and DCS-PCS bands. As the operating frequency reached to the WiFi band, the antenna became more directive.

Figure 6.6: Simulated 3D radiation patterns of the utilized planar monopole antenna at GSM900, DCS, and WiFi bands. The orientation of the antenna is also given for reference.
Table 6.2: Performance ratings of the proposed microwave energy harvester.

<table>
<thead>
<tr>
<th>Location</th>
<th>Minimum</th>
<th>Typical</th>
<th>Maximum</th>
</tr>
</thead>
<tbody>
<tr>
<td>1330 Kinnear Rd. (Room #150)</td>
<td>7.5μW</td>
<td>12.0μW</td>
<td>22.5μW</td>
</tr>
<tr>
<td>1330 Kinnear Rd. (Front door)</td>
<td>0</td>
<td>12.0μW</td>
<td>13.5μW</td>
</tr>
<tr>
<td>917 W. 10th Ave. (Basement)</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>1542 N. High St. (patio)</td>
<td>15μW</td>
<td>18μW</td>
<td>22.5μW</td>
</tr>
</tbody>
</table>

6.3 Multi-band Ambient Microwave Energy Harvester

The aforementioned antenna, rectifier, and the energy management unit from Chapter 5 were combined to form the ambient microwave energy harvester (see Fig. 6.7). The proposed multi-band energy harvester module was tested under a variety of operating conditions, both in indoors and outdoors. The testing began with the evaluation of current generation capabilities of the proposed device. This is done by connecting a variable load resistor to the device and measuring the DC voltage across the resistor terminals to make sure that it is still regulated to 1.5V. If the measured output DC voltage goes down to zero volts, then it means that the load resistance is trying to consume more current than the harvester module can provide. Typically, the device was found to supply a continuous current of up to 8μA to a resistive load. Of course the harvester module was able to supply more DC current when the density of the ambient microwave energy was higher, e.g., when in a busy office room or in a crowded downtown area. However, the device was not operational when placed in the basement of a house with very limited cell phone reception and very weak WiFi signal was available for harvesting. Table 5.1 summarizes the test results for the maximum harvested DC current in various environments.
One important parameter for the energy harvester is the initialization time, which refers to the time needed for the storage capacitor to reach a certain voltage and the point where the management circuit powers the load. Typically, it takes two minutes for the proposed harvesting device to initialize; however, as the ambient ambient microwave energy density declines, the initialization time increases, reaching 15 minutes when used in a suburban area.

6.4 Conclusion

This chapter presented a study on rectification of multi-band, statistically varying, ambient microwave radiation. The developed microwave energy harvesting module
consisted of (a) miniaturized multi-band antenna, (b) novel broad-band rectifier circuitry efficient even at low power levels (up to 35%), and (c) power management circuitry. The experimental results showed that the proposed harvester was able to provide stable DC voltage and generate sufficient DC current (up to 15μA) both indoors and outdoors, hence was truly mobile.

The motivation for considering the low-power multi-band rectification is applications in low-power battery-less sensors. An example could be a manufacturing environment, where a large number of sensors occasionally transmit data such as stress, temperature, pressure, and light level. A large number of such sensors with no batteries to be replaced could be powered with the ambient RF energy generated by the mobile radios carried by the employers. Another application example could be mobile patient monitoring. The results of this chapter show that it is possible to efficiently harvest ambient microwave energy and power such low-power batteryless sensors.
Chapter 7: Conclusion and Future Work

Portable electronic devices have intruded in our lives and have made their own unique stand in the way we interact with environment, causing a permanent change spearheaded by cell phones. Once considered as a luxury is now taken as much for granted as electricity or central heating. We do not even remember how life was before portable devices existed to a point where people seem to be born to have a mobile phone in their hands. Today, fulfilling the growing energy needs of the fast evolving must-have mobile technology lies on the shoulders of the traditional batteries. However, energy harvesting may enable wireless and portable electronic devices to be completely self-sustaining and batteries might eventually become obsolete.

Sources for energy harvesting are numerous and include mechanical vibrations, light, acoustic, airflow, heat, temperature variations, and electromagnetic sources, to name a few. Among these sources, radio frequency (RF) is the only one that can provide either an intentional or ambient energy source for harvesting. The concept is far from new. Tesla demonstrated wireless energy transfer via magnetic coupling almost 120 years ago at the 1893 World Fair in Chicago by providing power to a series of phosphorous light bulbs. In 1964, William Brown demonstrated a model helicopter that could fly by receiving power via a microwave beam over a distance of one mile. More recently, in 2008, Intel reproduced Tesla’s experiments by wirelessly
powering a light bulb and, in 2009, Sony demonstrated a wireless-powered TV at a range of 20 inches. Although this wireless energy transfer through magnetic coupling has great potential, it’s intrinsically limited to extremely short distances because near-field, non-radiating magnetic power density attenuates at a rate proportional to the inverse of the sixth power of distance. However, far-field methods permit longer range power transfers as they are subject to the inverse square law.

Regardless, increasing the distance between the source and receiver results in significant drop in received electromagnetic power. In addition, as power density drops, so does RF-to-DC conversion efficiency in traditional circuits. In some RF power transmission schemes, directional antennas were suggested on the broadcast side to reduce the effect of increased distance but this technique cannot be applied to mobile energy harvesting modules as portable devices do not have a fixed location or orientation. For the very same reason, ambient radio waves have also largely been ignored as a potential energy source. Though, efficient harvesting of low power RF signals can significantly reduce the effect increased distance, which in turn would mean significant range improvements for dedicated microwave power transmission and considerably more DC power from ambient RF energy harvesting. Therefore, in this dissertation, we propose a new class of microwave energy harvesting system which exhibits substantially improved conversion efficiency than the ones available off-the-shelf or in literature. This way, it is possible to power small electronic devices, such as wireless sensors installed in buildings and industrial machinery, using just “ambient” energy a ocean of existing radio waves produced by television, radio, internet, and mobile phone transmitters.

Specifically the new contributions can be summarized as follows:
1. Derivation of fundamental bounds on conversion efficiency of microwave energy to direct current (DC).

2. Introduction of two novel RF energy harvesting circuits that are more efficient than conventional designs.

3. Presentation of a robust study to predict the optimal way of interconnecting multiple microwave energy harvesters.

4. Design of a new, miniature ambient Wi-Fi (2450 MHz) energy harvesting system that powers a desktop thermometer (and its LCD display).

5. Development of a unique multi-band microwave energy harvesting module that can generate DC power from RF signals propagating at frequencies between 850 and 2500 MHz, thus increasing the power output and expanding mobility options.

Specifically, Chapter 2 introduced the fundamentals of microwave energy harvesting. The chapter started with presenting a thorough background and history of the harvesting process. Subsequently, the rectification concept was discussed in detail and a new closed-form solution was developed to highlight its limitations. The chapter concluded with the discussion of specific rectifier topologies used in the state-of-the-art RF energy harvesters and noted their weaknesses.

Chapter 3 presented two novel rectifier circuits that outperform conventional rectifier topologies in RF-to-DC conversion efficiency. The first rectifier significantly improved the efficiency in the high power region by harvesting the energy from the harmonics, a by-product of rectification that is mostly filtered and wasted in conventional designs. The second proposed circuit focused on harvesting energy in the
low power region and offered efficiency improvements over conventional charge-pumps with its RF symmetric structure. This was achieved by exploiting the trade-off between reverse breakdown voltage and the forward voltage drop of the diode. The chapter is concluded with the presentation of a proof-of-concept rectenna design.

Chapter 4 acknowledged that maximum power supplied by a single microwave energy harvesting module is usually not enough to energize most consumer electronics. Accordingly, a robust analysis is presented to predict the optimal way of interconnecting multiple energy harvesters. The presented analysis was verified with experiments conducted under various propagation conditions (indoors and outdoors).

In Chapter 5, an energy harvesting array (of $3 \times 3$ elements) is designed to power up a commercial thermometer and its LCD display using nothing more than ambient WiFi signals in an office environment. The chapter began by presenting a characterization study on the availability of the ambient RF energy generated by WiFi enabled devices. Then, a DC-combining rectenna array (using a miniaturized antenna) is developed to efficiently harvest low-level ambient WiFi energy. Subsequently, a highly efficient power management system is discussed to minimize leakage and provide uninterrupted regulated energy to the sensor. All these new components are integrated into a complete RF power harvesting system to drive low-power consumer electronics.

As can be surmised, increasing the operational bandwidth of the rectenna from just WiFi to include other mobile communications bands will enhance the amount of captured ambient RF power, thus generating more DC energy. Chapter 6 proposed a unique multi-band rectenna element that can generate DC power from RF signals of frequencies between 825 and 2500 MHz, where the majority of mobile devices operate at. The developed rectenna design can typically generate $12 \mu$Watts of DC
power both in indoors and in outdoors. Of course, the designed harvester can be incorporated into a rectenna array to increase the output DC power. In short, this dissertation demonstrated that low-power consumer electronics can be free of batteries and chargers just by harvesting the ambient RF energy that surrounds us.

A very promising future direction in RF energy harvesting is the utilization of adaptive rectifier circuits. These systems can detect the available input RF power and then switch between individually optimized rectifiers to maximize the conversion efficiency. Combining this concept with the research presented in this dissertation will significantly expand the medium power region (where the circuit is efficient, i.e., $\eta \geq 60\%$).

Another important performance aspect of an RF energy harvester is its ability to maintain high RF-to-DC conversion efficiency over many operating conditions, including wide output load resistance variations. Incorporating maximum power point tracking (MPPT) technique (as in some mechanical or solar power harvesting modules) into microwave energy harvesting modules can be very critical in extracting the maximum possible power under these varying output load conditions. The purpose of the MPPT system is to sample the output of the cells and apply the proper resistance (load) to obtain maximum power for any given environmental conditions. In one sense, the function of a MPPT unit is analogous to the transmission in a car. When the transmission is in the wrong gear, the wheels do not receive maximum power since the engine is running either slower or faster than its ideal speed range. The purpose of the transmission is to couple the engine to the wheels, in a way that lets the engine run in a favorable speed range in spite of varying acceleration and terrain. Likewise, the MPPT varies the ratio between the voltage and current delivered
to the battery, in order to deliver maximum power. If there is excess voltage available from the RF energy harvesting module, then it converts that to additional current to the output capacitor. Thus, with MPPT, the microwave energy harvester units presented in this dissertation can be scaled across many applications and devices that has different load conditions.

More importantly, a key research direction in moving this technology forward is the development of low-loss diodes with much higher saturation current. As discussed in Chapter 2, the high junction capacitance shorts the video resistance of the diode (resistance that models the RF-to-DC conversion) and high junction impedance reduces the amount of energy transferred to it. In addition, the large levels of parasitic capacitances in state-of-the-art semiconductor diodes lead to narrowband rectifier operation. In the literature, zero-biased antimonide based heterostructure backward diodes (Sb-HBDs) are shown to have extremely low junction parameters and superior I-V characteristics when the input power is very low. Utilization of these diodes in the presented modified Greinacher rectifier circuit has the great future potential to reduce the energy lost to the heat or reflected by the output filter. Moreover, utilization of these diodes will reduce the input impedance of the rectifier circuit, which in turn result in an easier and simpler matching network design for broadband operation.

Energy harvesting’s new frontier is an array of micro-scale technologies that scavenge milliwatts (not just μWatts) from solar, vibration, thermal and biological sources. Other than RF energy harvesting, the most promising energy harvesting technologies extract energy from vibration, temperature differentials and light. Traditionally, researchers focus only on one energy harvesting technique. However, these
harvesting technologies are not substitutes of each other, i.e., they can coexist. A
game-changing idea would be the introduction of a hybrid energy harvester, which
can harvest energy from vibration, temperature differences and RF power (ambient
and/or dedicated) simultaneously. These hybrid harvesters can be installed on
a wide range of different platforms, including but not limited to the human body,
aircrafts, railroads, and automobiles. Briefly, the hybrid energy harvester will inherit
the controllability of RF energy harvesting technology (since power can be replen-
ished whenever/wherever it is desired) and combine it with the ambient (and free)
energy harvesting techniques.
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