A WIRELESSLY-POWERED SENSOR PLATFORM
USING A NOVEL TEXTILE ANTENNA

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DECLARATION OF ORIGINALITY

The research work presented in this thesis is the original work of the author conducted between October 2009 and Dec 2012. Parts researched externally have been duly referenced.

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ABSTRACT

This thesis describes the design and analysis of a novel wideband circularly-polarized textile antenna to power up a wearable wirelessly-powered sensor system operating in the 2.45 GHz ISM band (2.4-2.5 GHz) and the building of the whole system. The system is constructed using off-the-shelf components and it is shown that the wirelessly-powered sensor system is able to operate when just a few mW are transmitted from a base station at a distance over a metre.

Initially, standard linearly-polarized patch antennas are used for power transmission. However, the antennas have to be aligned perfectly for the best efficiency. Subsequently, a circularly-polarized antenna is proposed for enhanced wireless-power transfer due to the freedom of orientation. A wide-slot antenna without a ground plane has been chosen for its simplicity and wide impedance band. The geometry is firstly optimized for wide impedance and 3-dB axial ratio bandwidth on FR-4. The experimental and simulation results have been studied to analyse the characteristics of such an antenna.

The wideband circularly-polarized antenna is then constructed using a conductive textile and re-optimized for on-body applications. With a simple antenna geometry and only a single layer of conductive textile layer, the axial ratio and impedance bandwidths are wide enough to cover the whole 2.45 GHz ISM band with plenty of margin and are significantly wider than any other on-body circularly-polarized textile patch antennas which have been reported. The characteristics of this wideband circularly-polarized antenna under different conditions on the human body have been measured and then connected to the wirelessly-powered sensor system to demonstrate the effectiveness of power transfer to the human body.
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ABBREVIATIONS

AR  Axial Ratio
ARBW  Axial Ratio Bandwidth
CP  Circularly-Polarized
CPW  Co-planer Waveguide
EMI  Electromagnetic Interference
HB  Harmonic Balance
IC  Integrated Circuit
ISM  Industrial, Scientific and Medical
LED  Light-Emitting Diode
LHCP  Left Hand Circular Polarization
µC  Microcontroller
PCB  Printed Circuit Board
RF  Radio Frequency
RFID  Radio-Frequency Identification
RHCP  Right Hand Circular Polarization
S\textsubscript{11}  Input Port Voltage Reflection Coefficient
SMA  Sub Miniature Version A
UWB  Ultra Wide Band
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Chapter 1: Introduction and Background

1.1 Recent Developments of Far-Field Low-Power Wirelessly-Powered Systems

As the technology for transmitting energy by radio waves has been developed for more than 50 years, the early history of this technology has been well summarized in the literature [1-3]. Recently, this technology has been developed in relation to the field of RF energy recycling [2, 4-7]. Ambient RF energy harvesting using an antenna array has been a popular topic due to its environmental friendliness and its flexibility in a metropolitan area with a high concentration of RF energy emitting sources [8-10].

Another field of development of this technology is the very low-power wirelessly-powered sensors for on-body and off-body applications [11-15]. Instead of harvesting the RF power from random sources, there is an active power source emitting microwave power at a specific frequency and the remote transponder is a device with or without a battery receiving the microwave energy from the base station [6, 16, 17]. For on-body near-field wirelessly-powered systems, inductive coupling using coils has been commonly adopted due to its high efficiency and simplicity [18, 19]. However, inductive coupling is only efficient for a distance of a few centimetres [20]. Therefore, a far-field wirelessly-powered technology using radio waves at microwave frequencies can be adopted to overcome the distance limitation. The application of radio-frequency identification (RFID) or far-field wirelessly-powered systems in the biomedical field such as body sensor networks, bio-implanted sensors and wearable real-time body monitoring systems has become increasingly popular [11, 13, 21-23].
In the case of passive RFID or a wirelessly-powered system, the power requirement is strictly constrained because there is no battery in such a system and the energy comes from the RF power transmitted from a base station. According to the Friis transmission equation, the receiving RF energy will be reduced proportionally to the square of distance:

\[
\frac{P_r}{P_t} = G_t G_r \left( \frac{\lambda}{4\pi R} \right)^2
\]  

(1.1)

For far-field power transmission, an efficient voltage multiplier is needed to boost up the voltage level from the antenna to a usable level for standard CMOS logic to operate.

Figure 1.1: General system architecture for a wirelessly-powered sensor system.

As shown in Figure 1.1, a wirelessly-powered sensor system has several main components: an antenna, a rectifier, a voltage step-up circuit, a storage element, a microcontroller (µC), a sensor and a transmitter. The rectifier converts the RF energy into a DC voltage, and then the charge-pump IC carries out the voltage step-up and stores the energy in the capacitor. Once the
voltage of the capacitor is sufficiently high, the microcontroller and other electronics will start to operate. The detailed function of each block will be discussed in the following chapters.

Several papers on low-power microwave frequency voltage multipliers have been presented using custom ICs or Schottky diodes [13, 14, 24]. One approach is a hybrid system consisting of a thin lithium battery to power up the power converter for efficient RF energy harvesting in a way that there is no need for battery replacement [5, 6]. It has been reported that the system can harvest RF energy down to 1 µW level [5]. However, a thin film battery is needed for the system to operate as shown in Figure 1.2. The drawback of having a battery is that extra circuitry is needed to perform the voltage step-up to the specific voltage and energy is lost during the process. The lifetime of the battery is another issue, and so this type of hybrid RF/battery system is not considered in this thesis.

The other approach is a passive wirelessly-powered system without a battery and all the energy is scavenged from the RF energy emitted from a base station [11, 13, 14, 16]. An example of such a system utilizes 450 MHz for receiving the RF power and 2.3 GHz for the transmitting the signal back to the base station [25]. A lower input frequency does offer longer distance according the Friis free space equation. However, a 450 MHz antenna can be too big for body-worn sensor applications. A similar system using a 1 W 915 MHz RFID reader for sensing neural signals from an insect with a 1-metre range has been presented [14]. The RFID back-scattering technique used here has the advantage that one antenna is needed for both receiving RF energy and transmitting the sensor signal from and to the base station. However, the size of a 915 MHz antenna is still relatively big compared to the entire wirelessly-powered system itself.
A more recent 2.4 GHz wirelessly-powered glucose sensor in a contact lens as shown in Figure 1.3 has a size advantage of being smaller when compared to other lower frequency systems [13].

Figure 1.2: The 2.4 GHz wirelessly powered sensors system from the University of Colorado [6].

Figure 1.3: The 2.4 GHz wirelessly-powered glucose sensor in a contact lens from the University of Washington [13].

The antenna is an essential component not only affecting the overall size of the wirelessly-powered system but also the power efficiency and sensitivity of the system. There are many different types of antennas suitable for this application including monopole, dipole, patch, and spiral antennas [11, 14, 26]. The limitation of a linear polarized antenna such as a dipole is that
the orientations of both the transmitting and receiving antennas have to be aligned perfectly for maximum power efficiency. Otherwise there will be energy loss due to polarization mismatch. This requirement can be a problem in reality when the orientations of both antennas are unknown or changing over time such as the orientation of a body-worn antenna is changing over time due to body movement. One solution is using a dual-polarized patch antenna to receive energy from both orthogonal polarizations [5, 27, 28]. An example of such a system is shown in Figure 1.4a. Since the antenna can receive energy from an arbitrary polarization, there is no limitation on the orientation of the antenna pair. However, since the feed lines are on both edges of the antenna, the rectifier and matching network are needed to be duplicated on each edge [5, 8].

![Figure 1.4](image.png)

**Figure 1.4:** a) The integrated dual polarized patch antenna with rectifiers on both edges, b) Spiral antenna array [5, 8].

Another approach resolving the polarization mismatch between antenna pairs is to adopt a circularly-polarized antenna such as a spiral antenna [8]. With circular polarization the electromagnetic wave is constantly rotating in a clockwise or counter-clockwise direction and so
the polarization mismatch can be minimized if both transmitting and receiving antennas are circularly-polarized in the same direction. An array of spiral antennas and rectifiers has been built for ambient RF energy harvesting as shown in Figure 1.4b [8]. A circularly-polarized antenna has potential to be an efficient antenna for low power transmission and this will be discussed in more detail in Chapter 3.

The materials for the antenna are not limited to a traditional circuit board or ceramic substrate. Textile antennas have been a growing area of interest for body-worn antenna applications due to its light weight and ease of integration into clothing [29-35]. However, it is a challenge to design a robust textile antenna for use on a human body due to the coupling between the body and the near field of the antenna and this will be discussed in more detailed in Chapter 4 [36-38]. Therefore, this idea can be applied to a body-worn wirelessly-powered sensor system and a working system on a human body will be demonstrated in Chapter 4 and 5.

1.2 Research Objective

The main objective of this thesis includes the followings:

- Design a highly power sensitivity rectifier using off-the-shelf components.
- Design a novel wideband circularly-polarized antenna on a single-sided circuit board.
- Design a novel wideband circularly polarized textile antenna as a power receiving antenna on a wearable wirelessly-powered sensor system.
- Use very low microwave power to wirelessly power a temperature sensor system on a human arm over a metre range.

However, there are several issues and areas are not addressed in this thesis:
Despite a high performance wearable textile antenna is developed for the wirelessly-powered system in Chapter 4, the power transmission is impossible when the incoming electromagnetic wave is 90° to the z-axis of the power receiving antenna. However, this problem can be resolved using multiple antennas at different areas of a human body to widen the power receiving area coverage [31, 33].

In Chapter 2 and 5, the highly power sensitive wirelessly-power systems are developed using discrete components only and the designs are not further improved and implemented in a CMOS integrated chip. This is because the emphasis of this thesis is on the design of a novel textile antenna for a wirelessly-powered system and is not on the system itself. The purpose of designing such a system is to demonstrate the effectiveness of power transmission using the novel textile antenna developed in Chapter 3 and 4.

All the tests and measurements are carried out inside an anechoic chamber, therefore the effect of multipath or interference due to reflections or from the environment is not considered in this thesis.

1.3 Thesis Organization

The thesis is divided into 5 sections. The operation of the RF-DC conversion of the microwave rectifier will be first introduced. Then the design of the wideband circularly-polarized wide-slot antenna will be discussed. Later the wearable textile antenna for power transmission will be developed. The application of the textile antenna on the wirelessly-powered sensor system is then demonstrated and finally future plans will be discussed.
1.3.1 Microwave Rectifier

The RF energy is obtained from a base station and then it is transmitted through the antenna to the microwave rectifier, where the RF-DC conversion is carried out. A zero-bias and highly efficient Schottky diode is used in the design. Several different diode configurations have been simulated and the prototypes are built on FR-4 for measurement in the laboratory. In Chapter 2, the design methodology for a highly efficient microwave rectifier using the circuit optimizer inside the Microwave Office (MWO) is presented. The power sensitivity and efficiency are measured and compared to the other designs.

1.3.2 Antenna Design

In Chapter 3, a wide-slot wideband circularly polarized antenna is developed. This type of antenna has the essential characteristics such as wide impedance and 3-dB axial ratio bandwidth that are required for a flexible textile antenna for body-worn applications in the later chapters. The objective is to understand and design a circularly polarized antenna with a single layer of metal on a FR-4. The simulation is optimized using Microwave Office and CST and the prototype is built and measured in an anechoic chamber. The result is compared with standard circularly polarized patch antennas.

1.3.3 Textile Antenna for on-Body Application

In Chapter 4, the wide-slot antenna design from Chapter 3 is adopted and modified to become a wearable textile antenna. The textile antenna is hand-made using a conductive fabric and self-adhesive felt. The textile antenna is first simulated and measured without the consideration of a human body, and then the simulation model considers the proximity of a human body to closely
match reality. The effect of bending and distance from a body are considered and measured in the anechoic chamber.

1.3.4 A Sensor Platform with a Wearable Circularly Polarized Textile Antenna

As the energy is scavenged using the rectenna (Rectifying Antenna), a charge-pump IC is used to step up the voltage to 2.4 V for the microcontroller and sensor to operate. However, due to the very low voltage requirement (0.3 V), the conventional DC-DC converter is not functional at this voltage level. To solve this problem, a Seiko charge-pump IC is used to carry out this function and this enables the wirelessly-powered sensor system to operate at an ultra-low input RF power level (27 µW). In Chapter 5, a complete workable system will demonstrate how to scavenge weak RF energy using the textile antenna and transforms the energy into the required voltage for the microcontroller to sense the temperature and send the data back to the base station.

1.3.5 Conclusion and Future Work

In Chapter 6, the conclusion will be drawn and the possible directions for future projects will be discussed.

1.4 Conclusion

After brief discussion of the current development on wirelessly-powered technology, the basic concept has been introduced. Therefore, the design detail of each building block will be investigated in the coming chapters in order to build a low-power body-worn sensor system.
Chapter 2: Microwave Rectifier

2.1 Introduction

In this chapter, the simulation and implementation of the microwave rectifier centred at 2.45 GHz will be discussed. It will start with the basic concept of large signal $S_{11}$ and Harmonic Balance (HB) simulation in AWR Microwave Office® (MWO) and then the model of the microwave rectifier is drawn in schematic. The performance of several different diode configurations will be compared. Next, the implementation of the designs on PCB is used to verify the simulation result. At the end, the efficiency and power sensitivity of the design are measured and analyzed and a complete rectenna system is tested in the anechoic chamber.

2.2 Recent Developments in Rectifier Design

The basic function of a microwave rectifier is to convert an RF signal into a DC voltage at microwave frequency. As shown in Figure 2.1, the full sine wave AC signal goes into the diode, and it only allows positive voltage larger than its forward bias voltage to get through and this is typical half-wave rectifier behaviour.

![Figure 2.1: Basic structure of a rectifier.](image-url)
The main challenge of a wirelessly-powered system is that the receiving RF power is attenuated rapidly over the distance in free space according to the Friis transmission equation. Therefore the receiving RF energy is not only needed to be rectified, but also needed to be step-up to the required voltage level to drive the electronics in the transponder. To enhance RF energy harvesting over a long distance, a multi-stage voltage multiplier is normally used as shown in Figure 2.2 and it can be implemented using a diode or diode-connected nMOS transistor [23, 24, 39]. A 36-stage floating-gate voltage multiplier has been proposed to reduce the threshold voltage of the CMOS devices to improve the sensitivity working over 44 metres [24]. However, with more stages of voltage multiplier, more energy will be lost in the diodes and so the power efficiency will be degraded as it will be shown in this chapter.

![Figure 2.2: Multi-stage voltage multiplier [39].](image)

Another method for improving efficiency is proposed based on LC power-matching network with a ground inductor and a floating capacitor to minimize the number of stages of the voltage multiplier and hence reduce the power lost in the transistors as shown in Figure 2.3 [40]. However, this method is limited by the quality of the inductor in standard CMOS process.
A further enhancement is proposed based on a microwave step-up transformer to improve the power efficiency over the LC power matching network as shown in Figure 2.4. The step-up transformer provides a higher input impedance compared to just a LC matching network and so the input voltage is increased and hence the overall power efficiency is increased [41]. However, its modified voltage multiplier can only generate differential DC voltage.

In order to minimize the power loss in the voltage multiplier circuit, one suggested that the voltage step-up process should be done at a lower frequency instead of at a microwave frequency.
as shown in Figure 2.5 [42]. Note that only the first stage rectifier is done at microwave frequency and then the DC generated is used to switch on the oscillator for the charge-pump circuit which operate at kHz range and hence the power efficiency is improved by 14% [42]. This circuit can be implemented easily using off-the-shelf components and the power efficiency is comparable to CMOS as it will be shown in this chapter.

![Diagram](image)

**Figure 2.5:** Improved voltage multiplier with a low frequency charge-pump circuit [42].

## 2.3 Rectifier Design

### 2.3.1 Choice of Diodes

PN junction diodes are commonly used for low frequency applications only due to the large junction capacitance [43]. PIN diodes are widely used in switches in microwave systems. However, its junction capacitance is still relatively large and it requires a forward bias current of 10-30 mA [43]. For energy harvesting application, the forward bias voltage must be very low for efficient RF-DC conversion and so a zero bias Schottky detector diode is the best candidate. Figure 2.6 shows the spice model of the Agilent HSMS-2850 Schottky diode [44].
Figure 2.6: Spice model and parameters of the Schottky Diode [44].

<table>
<thead>
<tr>
<th>SPICE Parameters</th>
<th>Units</th>
<th>HSMS-285x</th>
<th>SMS7630</th>
</tr>
</thead>
<tbody>
<tr>
<td>BV</td>
<td>V</td>
<td>3.8</td>
<td>2</td>
</tr>
<tr>
<td>Cj0</td>
<td>pF</td>
<td>0.18</td>
<td>0.14</td>
</tr>
<tr>
<td>Eg</td>
<td>eV</td>
<td>0.69</td>
<td>0.69</td>
</tr>
<tr>
<td>IbV</td>
<td>A</td>
<td>3.00E-04</td>
<td>1.00E-04</td>
</tr>
<tr>
<td>IS</td>
<td>A</td>
<td>3.00E-06</td>
<td>5.00E-06</td>
</tr>
<tr>
<td>N</td>
<td></td>
<td>1.06</td>
<td>1.05</td>
</tr>
<tr>
<td>Rs</td>
<td>Ω</td>
<td>25</td>
<td>20</td>
</tr>
<tr>
<td>Vj</td>
<td>V</td>
<td>0.35</td>
<td>0.34</td>
</tr>
<tr>
<td>XTI</td>
<td></td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>M</td>
<td></td>
<td>0.5</td>
<td>0.4</td>
</tr>
</tbody>
</table>

Table 2-1: Spice parameters of the HSMS-285x and SMS-7630 Schottky Diodes [44, 45].

The most important parameter for matching is the junction capacitance Cj0 as shown in Table 2-1. For such a small value, the matching network must require an external inductor to compensate the capacitance. The other choice of Schottky diode is the Skyworks SMS-7630.
which requires a slightly lower forward bias voltage [45, 46]. The lower forward bias voltage always yields better power sensitivity. The performance of the two diodes will be used in our rectifiers and the measured results will be shown in the later sections. The SC-79 (or SOD-523 for HSMS-285Y) packaging for SMS-7630 is chosen due to lower package parasitics and smaller footprint [46].

2.3.2 A Brief Background on Harmonic Balance Simulation

The Harmonic Balance (HB) method is a powerful technique for the analysis of high-frequency nonlinear circuits such as mixers, power amplifiers, and oscillators. HB simulation is frequency domain analysis. Unlike the time-domain simulator SPICE, it gives the transient behaviour and the steady state solution. The motivation behind using HB simulation over the transient simulator is the following:

- The distributed elements (e.g. transmission lines) in microwave circuits can be exclusively modeled and analyzed in the frequency domain which is very hard to do accurately and quickly in SPICE-like simulator.

- HB simulator can handle multi-tone analysis quickly of microwave circuits in the frequency domain. However, this is very difficult to handle in the time-domain. Imagine you have a 10 GHz carrier signal modulated with 1 MHz data. The time domain simulator must follow the slow envelope for several cycles to determine the steady state solution. In the other words, the time domain simulator must follow many cycles of the carrier to maintain the accuracy. It is obvious that this process is very time consuming and consumes a lot of computing resources or simply runs out of memory.
• There are a lot of microwave circuit which have very high-Q, which means the transient time can be very long. It is a time consuming to trace the whole transient response if only the steady state solution is in concern.

Therefore, a HB simulator is used for our microwave rectifier simulation and optimization. More detail on Harmonic Balance method can be found on the AWR web site [47].

2.3.3 The Circuit Model with Linear S-parameters Simulation

According to the datasheet, the suggested schematic for the single diode rectifier is shown in Figure 2.7 [48]. Note that the inductor before the diode is needed for impedance matching and it is the DC return path for the diode. The 65 nH inductor rotates the impedance of the diode to a point on a Smith Chart where the shunt inductor (the microstrip line to the ground) can pull it up to the centre. All the values of the dimensions and components are different because the central frequency is 2.45 GHz not 915 MHz and also the PCB has different thickness as suggested in the datasheet.

![Figure 2.7: Manufacturer suggested circuit for a single diode rectifier at 915 MHz [48].](image)

Before the simulation, the substrate specification has to be correctly specified in the simulator as shown in Figure 2.8.
Figure 2.8: Microstrip substrate specification.

The low-cost FR-4 has a dielectric constant of 4.2 and the height is 1.6 mm with 1 oz. copper (0.35 mm) [49]. Alternatively, a high quality RF substrate from Roger can be used in which the loss tangent is 10 times less than FR-4 [50]. However, the cost of the RF substrate is over £1400 versus £28 for FR-4 substrate and the gain in power sensitivity is less than 0.8 dBm from the simulation. Therefore, FR-4 is chosen due to its much lower cost with a reasonable performance. It is important to know the dimension of the microstrip line for the quarter-wavelength impedance matching and it is determined by using the TXLINE tool from MWO as shown in Figure 2.9. TXLINE will determine the correct dimension once the specification of the FR-4 is provided [47].
Figure 2.9: TXLINE is used to determine the quarter wavelength dimension of a microstrip line [47]. Note that the name of the material can not be changed due to the limitation of the tool.

The model is constructed similar to the one from the datasheet and it is shown on Figure 2.10. The optimizer is used to determine the right parameters for the schematic to achieve the $S_{11}$ input impedance matching and this provides good initial starting values for the more complicated models as shown in Figure 2.11. Note that the simulation is carried out using a linear simulator for a good initial guess and a more accurate result is obtained using nonlinear harmonic balance simulation during the next section.

Note that the input matching can be done without any lumped elements by using a microstrip alone [51, 52]. However, the footprint of the circuit will be significantly smaller using a chip inductor and it is possible to keep the trace as short as possible for less energy loss.
2.3.4 The Circuit Model with Harmonic Balance Simulation

As the concept of HB simulation has been discussed earlier in this chapter, HB simulation is only concerned about the steady-state solution. In our application, since the AC to DC conversion efficiency is at steady state, HB simulation is the best candidate for this purpose. Therefore, the schematic for the single diode configuration with a HB input port (PORT_PS1) for HB simulation is set up as shown in Figure 2.12. The lowest power of -14 dBm at 2.45 GHz is specified. This is important because junction capacitance changes with input power level, and the circuit optimization is based on this power level to obtain the best low power sensitivity of the microwave rectifier. The package inductance and capacitance is modeled according to the
datasheet of the Schottky diode for better matching to the real circuit [46, 53]. An open circuit $\lambda/4$ stub is also added at the cathode of the diode to short the fundamental frequency at the output [51]. A 4.7 nH inductor is needed for the input matching with HSMS285 diode as a result. The schematic for the SMS7630 is similar as shown in Figure 2.13.

![Figure 2.12: The single diode rectifier circuit with a HSMS285 diode.](image1)

![Figure 2.13: The single diode rectifier circuit with a SMS7630 diode.](image2)
2.3.5 The Circuit Model of the Voltage Doubler and 4X Voltage Multiplier

According to the datasheet [48], the voltage doubler as shown in Figure 2.14 has 2 advantages over a single diode detector. First, the voltage sensitivity is almost doubled by adding the voltages of the 2 diodes in series. Second, the RF impedance is in parallel and so reactive matching is easier. However, as explained later in this chapter this is only partially correct because energy is lost in the devices and the extra routing on the PCB in our energy scavenging system.

![Diagram of voltage doubler circuit](image)

*Figure 2.14: The voltage doubler circuit suggested in the datasheet [48].*

The matching network for the voltage doubler is not provided by the manufacturer, and so a different matching network is needed. First, it is noted that the shunt diode can provide the DC return path for the series diode, and so the shunt inductor to ground is not needed as in the single diode rectifier. Second, the shunt diode capacitance cancels out the parallel capacitance of the diode in series. It turns out that a small capacitor is needed for the matching network as shown in Figure 2.15 with HSMS285 diode and Figure 2.16 with SMS7630 diode. The result of the circuit layout is similar to the single diode rectifier circuit presented earlier. Note that a 3.6 or 3.9 pF
capacitor is needed for the voltage doubler input matching as compared to an inductor which is needed for single diode input matching, depending on which diode.

The same reasoning applies to the matching network of the 4X voltage multiplier as shown in Figure 2.17 and the sub-circuit is shown in Figure 2.18. The topology for the 4X voltage multiplier is the Dickson multiplier and it is just a cascade of the 2 voltage doublers [39]. Therefore, the input matching topology is the same as the voltage doubler. Although the number of stages can be increased by cascading more stage of the multiplier, but due to the loss of energy in components and longer copper traces, there is no benefit using 8X voltage multiplier and therefore it is not considered in this project. Note that the exact dimensions and component values of each schematic are obtained by using the MWO Optimizer and it will be discussed in the next section.

![Diagram of Voltage Doubler with HSMS-285](image)

**Figure 2.15:** The schematic of the voltage doubler with HSMS-285.
Figure 2.16: The schematic of the voltage doubler with SMS-7630.

Figure 2.17: The schematic of the 4x voltage multiplier with HSMS-285.
2.3.6 Optimization and Simulation Results

In order to have the best efficiency from the microwave rectifiers, tuning of the $S_{11}$ parameter to be at least less than -10-dB at the centre frequency of 2.45 GHz is needed. The interface for setting up the optimizer is shown in Figure 2.19 and the requirement of $S_{11}$ parameter below -15 dB in the range of 2.4 GHz to 2.5 GHz is specified. The optimizer will tune the length of the transmission line in the schematic to try to match the goal. The optimization is done individually for each of the diode configurations.

Figure 2.18: The Dickson multiplier sub-circuit within the 4X voltage multiplier.

Figure 2.19: Circuit optimizer in AWR MWO.
The optimized results are shown in Figure 2.20. There are 5 different rectifier circuits in this figure: the voltage doubler with HSMS-2850, the voltage doubler with SMS-7630, the single diode rectifier with HSMS-2850, the single diode rectifier with SMS-7630 and the 4x voltage multiplier with HSMS-2850. Note that a 16 kΩ load resistor is chosen because this is an approximation of the input resistance of the charge-pump IC for a voltage step-up from a minimum of 0.3 V. It will be discussed more detail in Chapter 5.

![Figure 2.20: The large signal S_{11} parameters for all different diode configurations.](image)

As shown in Figure 2.20, S_{11} is matched to about -12 dBm on average only for the single diode design because all the diodes are barely turned on with such a low voltage level and so only a small amount of energy can get through the diode due to its built-in barrier.

The designs using the SMS-7630 show the best input matching due to a lower barrier, but the difference is not significant when compared to the HSMS-2850. The S_{11} matching seems better
with voltage doublers and multipliers as their impedance is in parallel and so reactive matching is easier, as stated in the datasheet. However, this doesn’t mean more energy gets transferred to the load because there is more energy loss in the diodes when compared to the single diode configuration, as shown in the experimental results section later. Due to a mismatch between the model and the circuit on the PCB, several iterations are made to match the model as close to the circuit on PCB as possible. It is found that the ground connection for the matching stub is about 1 nH and it must be added in the model for the better accuracy to match the reality. With an adjusted model, the $S_{11}$ matching between simulation and measurement is very close with only a small manual adjustment needed on the PCB to accommodate the process variation on each design.

### 2.3.7 Manufacturing of the Rectifiers

All the rectifiers are made using the facilities in the laboratory and the standard procedure is briefly described here. First, the track and surface-mount-device (SMD) pads are simulated using MWO as specified in the previous section. Then a mask is printed on a semi-transparent sheet and is placed on top of a double-sided FR-4 as shown in Figure 2.21. Then the UV light is turned on for 120 seconds and the portion uncovered by the mask will be exposed under UV light. Then the circuit board is rinsed with a photoresist developer, the exposed photoresist layer is washed away and so the copper portion will be exposed during etching. The circuit board is in the tank for etching for about 10 to 15 mins and the end result is shown in Figure 2.22. The final prototypes are built with all the components as shown in Figure 2.23.
Figure 2.21: The mask on top of a circuit board inside the UV light box.

Figure 2.22: The circuit pattern is developed after etching.
Figure 2.23: The 5 different rectifier designs. The top row from the left is the voltage doubler with HSMS-2850, SMS-7630 and 4X voltage multiplier with HSMS-2850. The bottom row from the left is the single diode rectifier with HSMS-2850 and SMS-7630.

2.4 Experimental Results

2.4.1 Large Signal $S_{11}$ Parameter Matching

The $S_{11}$ parameters are measured using an Agilent E8361A PNA. Each circuit was manually tuned to resonate at 2.45 GHz by iterating between the simulation and the measured results. For a small degree of tuning, adjusting the length of the matching stub is needed. The measured $S_{11}$ values are shown in Figure 2.24.
Figure 2.24: The \( S_{11} \) parameter measurements with -14 dBm signal power.

As shown in Figure 2.24, the voltage doubler always has better \( S_{11} \) matching than the single diode design, no matter which diode is used. This is because the shunt diode provides clamping for the series diode and so there is a DC voltage at the junction of the 2 diodes. This provides the forward bias voltage for the series diode and so the RF signal can get through the barrier easier. And so less energy gets reflected back, meaning the return loss is lower. However, though this may imply that more power can go through the series diode and more RF power has been converted to DC voltage for the load, this is not true for the voltage doubler. It is because a portion of the energy is consumed in the shunt diode to provide the DC bias voltage of the series diode, less energy gets transferred to the load, although the return loss is better than the single diode design. This phenomenon is more apparent in the case of the 2-stage voltage multiplier as
even more energy is consumed in the diodes and traces on the PCB, resulting in far less energy getting to the load for conversion into usable voltage.

2.4.2 Rectifier with a Microcontroller

The ultimate goal for building these rectifiers is to provide the power for the wirelessly-powered sensor system as will be shown in the later chapters. It is important to understand the power efficiency of the rectifier and also the efficiency of the overall system with each input power level.

Each rectifier is connected to an R&S model SML 03 signal generator directly and then the minimum input power levels required for the voltage multipliers to turn the microcontroller (µC) on to flash the LED without the Seiko charge-pump IC is measured as shown in Figure 2.25 and the results are summarized in Table 2-2.

Figure 2.25: The single diode rectifier with the µC connected to the signal generator.

The single diode rectifier and voltage doubler with HSMS-2850 require 1.0 dBm and -0.4 dBm input power to turn on the µC, respectively. While the single diode rectifier using SMS-
7630 requires only 0.7 dBm, the voltage doubler requires only -2.3 dBm. When using the HSMS-2850, the voltage doubler produces a 1.4 (1+0.4) dBm improvement on the power sensitivity and the voltage sensitivity is improved from 250.9 mV to 213.5 mV over the single diode rectifier configuration. The voltage doubler with SMS-7630 delivers 3.0 dBm power sensitivity improvement over the single diode design and the voltage sensitivity is improved from 242.4 mV to 171.6 mV. This result is due to the measured forward bias voltages of the HSMS-2850 and SMS-7630 being 0.18 V and 0.16 V, respectively. The advantage is the lower forward bias voltage is doubled when the diodes are in the voltage doubler configuration and this results in a 41.9 mV voltage sensitivity improvement using SMS-7630 versus HSMS-2850. The 4X voltage multiplier with HSMS-2850 requires -2.1 dBm to turn the µC on. It exhibits a 1.7 (2.1- 0.4) dBm improvement over the voltage doubler and the input voltage sensitivity is very close to that of the voltage doubler with SMS-7630. The addition of an extra stage results in a less than 2 dBm improvement for input power sensitivity. Furthermore, there appears to be no advantage using a 3-stage voltage multiplier in the simulation as the impedance match is harder to achieve and power loss in the components and PCB is greater.

<table>
<thead>
<tr>
<th>Different microwave rectifiers with the µC</th>
<th>Single diode (HSMS-2850)</th>
<th>Single diode (SMS-7630)</th>
<th>Voltage doubler (HSMS-2850)</th>
<th>Voltage doubler (SMS-7630)</th>
<th>4X Voltage multiplier (HSMS-2850)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Minimum required input power to turn on the µC to flash the LED</td>
<td>1 dBm</td>
<td>0.7 dBm</td>
<td>-0.4 dBm</td>
<td>-2.3 dBm</td>
<td>-2.1 dBm</td>
</tr>
</tbody>
</table>

Table 2-2: Measured minimum input power to turn on the µC to flash the LED for the different rectifier designs.
2.4.3 Rectifier with a Charge-Pump IC and a Microcontroller

As explained in the literature, voltage multiplication at microwave frequencies is very lossy. It was demonstrated by using a single stage microwave frequency rectifier and then followed by a low frequency charge-pump circuit. It was shown that the voltage multiplier improved the power efficiency by 14% over the conventional modified Dickson voltage multiplier [39, 42].

With the same circuit topology, a Seiko charge-pump IC is connected between the microwave rectifier and the µC as shown in Figure 2.26 to measure the lowest supply input power to turn on the µC. The function of the charge-pump IC is to step up the input voltage from 0.3 V to 2.0 V. The measured input resistance for the converter is about 16 kΩ for an input voltage of 0.29 V.

The measured required minimum input power levels for the single diode rectifier with HSMS-2850 and SMS-7630 are -15.2 dBm and -15.6 dBm respectively which shows that the power sensitivity is comparable with the results reported [27, 51]. This means that the voltage sensitivity is about 37.1 mV for a 50 Ω system and only 27.5 µW is required to turn on the µC to flash the LED with an SMS-7630. The performance difference between the two device types is not significant in real life as the process variations on the device and the PCB can easily offset the difference.

![Figure 2.26: The single diode rectifier with the Seiko charge-pump IC and the µC.](image)
The minimum required input power levels for voltage doublers using HSMS-2850 and SMS-7630 are -13.8 dBm and -15.5 dBm, respectively. It is clear from this that the voltage doubler does not have any advantage over the single diode rectifier when operated with the charge-pump IC. The 4X voltage multiplier requires -12.3 dBm input power to turn on the µC and the results are summarized in Table 2-3.

<table>
<thead>
<tr>
<th>Different rectifiers with Seiko charge-pump IC and µC</th>
<th>Single diode (HSMS-2850)</th>
<th>Single diode (SMS-7630)</th>
<th>Voltage doubler (HSMS-2850)</th>
<th>Voltage doubler (SMS-7630)</th>
<th>4X Voltage multiplier (HSMS-2850)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Minimum required input power to turn on the µC to flash the LED</td>
<td>-15.2 dBm</td>
<td>-15.6 dBm</td>
<td>-13.8 dBm</td>
<td>-15.5 dBm</td>
<td>-12.3 dBm</td>
</tr>
</tbody>
</table>

Table 2-3: Measured minimum input power for different rectifier designs with a Seiko charge-pump IC.

Voltage is boosted in the voltage doubler but since energy has to be conserved, less current is delivered to the load because of this. The rectifier has to provide sufficient voltage for the µC and power for the charge-pump IC to operate. Therefore the choice of diode configuration depends ultimately on the load resistance expected. The input resistances of the µC and the charge-pump IC just before turn-on are about 60 kΩ and 16 kΩ, respectively. For a lower resistive load, the single diode design is better because it can supply more current as long as the voltage supplied is just enough to turn the chip on. For high resistive loads, the voltage doubler or multi-stage voltage multiplier can perform better because the load draws less current. However, impedance matching is easier with a voltage doubler and so it has advantages where space is at a premium and the addition of a comparatively large microstrip matching stub would not be permissible.
2.4.4 Power Efficiency

The efficiencies for the single diode rectifier with SMS-7630, the charge-pump IC and the overall system at different power levels with load resistance 20 kΩ are measured and the results are summarized in Table 2-4. The 20 kΩ load is chosen because the µC has similar input resistance when it is turned on. The storage capacitor recharge interval represents the idle time of the µC because the charge-pump IC takes time to charge up the 1 µF storage capacitor to 2 V. A 1 µF capacitor is chosen because it has relatively short charging time and it provides sufficient energy for the operation of a µC in the next section. At the lowest operational input power level -15.6 dBm, it takes the charge-pump IC about 1.2 seconds to charge up the capacitor to 2 V to turn on the µC to flash the LED. However, at input power level -10 dBm, it only takes about 0.13 seconds to charge up the storage capacitor. The measurable efficiencies within the harvester are defined as follows:

Rectifier Efficiency = \frac{\text{Output Power from Rectifier}}{\text{Primary Input Power}} \quad (2.1)

Charge Pump IC Efficiency = \frac{\text{Output Power from Charge Pump IC}}{\text{Input Power to Charge Pump IC}} \quad (2.2)

System Efficiency = \frac{\text{Output Power from Charge Pump IC}}{\text{Primary Input Power}} \quad (2.3)

Note that although the output voltage level from the rectifier is continuous, the output voltage level of the charge-pump IC is a short pulse as shown in Figure 2.27. Note that the voltage is decreasing over time because the storage capacitor is discharging and the charge-pump IC will shut down the connection when the voltage drops below the threshold. The trace C1 shows that the separation of the output pulses from the charge-pump IC is about 1.2 seconds and it is
depending on the input power level. The trace Z1 shows that the pulse width is always about 6.6 ms because it is independent of the input power level. The charge-pump IC charges the storage capacitor up to 2 V and then releases the energy until it drops to 1.4 V. Therefore, the voltage output of the system is the following:

\[ V(t) = 2e^{-t/0.02} \]  

(2.4)

where \( V(t) \) is the voltage output of the system. The RC time constant is 0.02 because of the 1\( \mu \)C storage capacitor and 20 k\( \Omega \) resistor. The average power output is calculated as below where \( T \) is the period of the pulse and \( R \) is the load resistance:

\[
P = \frac{1}{TR} \int_{0}^{0.0066} V^2(t) \, dt = \frac{9.4}{T} \text{ mW.} 
\]  

(2.5)

Figure 2.27: Measured pulse separation and pulse width of the output of the charge-pump IC.
<table>
<thead>
<tr>
<th>Input Power (dBm)</th>
<th>Output Power of Rectifier (µW)</th>
<th>1 µF Storage Capacitor Recharge Interval (ms)</th>
<th>System Output Power (µW)</th>
<th>Rectifier Efficiency</th>
<th>Charge-pump IC Efficiency</th>
<th>System Efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>-15.6</td>
<td>3.74</td>
<td>1245</td>
<td>0.75</td>
<td>13.6%</td>
<td>20.2%</td>
<td>2.7%</td>
</tr>
<tr>
<td>-15</td>
<td>4.28</td>
<td>957</td>
<td>0.98</td>
<td>13.5%</td>
<td>22.9%</td>
<td>3.1%</td>
</tr>
<tr>
<td>-14</td>
<td>6.62</td>
<td>513</td>
<td>1.83</td>
<td>16.6%</td>
<td>27.7%</td>
<td>4.6%</td>
</tr>
<tr>
<td>-13</td>
<td>9.83</td>
<td>326</td>
<td>2.88</td>
<td>19.6%</td>
<td>29.3%</td>
<td>5.8%</td>
</tr>
<tr>
<td>-12</td>
<td>1.39</td>
<td>227</td>
<td>4.14</td>
<td>22.1%</td>
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<tr>
<td>-11</td>
<td>19.2</td>
<td>167</td>
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<td>24.2%</td>
<td>29.3%</td>
<td>7.1%</td>
</tr>
<tr>
<td>-10</td>
<td>26.2</td>
<td>128</td>
<td>7.34</td>
<td>26.2%</td>
<td>28.0%</td>
<td>7.3%</td>
</tr>
</tbody>
</table>

Table 2-4: Measured efficiencies for the rectifier, the charge-pump IC and the system at different input power levels with a 20 kΩ resistor.

The power efficiencies of different sections of the system based on Table 2-4 are plotted in Figure 2.28. The lowest operational input power level of the single diode rectifier with SMS-7630 is -15.6 dBm and the total system efficiency is 2.7%. The system efficiency increases with a decreasing rate as the power level increases beyond -13 dBm. It is because the charge-pump IC efficiency is saturated at 29.7% at -12 dBm input power level and the efficiency decreases as the power level increases beyond that point. On the other hand, the rectifier efficiency increases linearly when the input power level is higher than -15 dBm and the result is consistent with other studies at the specific power level [20, 21]. The total system efficiency finally reaches 7.3% when the input power is -10 dBm.

Note that the SMA connectors have very minor effect on the power efficiency of the system because only DC current is transmitted after the rectifier.
Figure 2.28: Measured efficiencies of the rectifier, the charge-pump IC and the overall system.

The comparison of this work to the other monolithic power harvesters is shown in Table 2-5. At 2.45 GHz, our power harvester performs as efficiently as the monolithic solution at -13.5 dBm. The RF sensitivity of our power harvester is -15.6 dBm and it is second best in the table due to the combination of a highly efficient Schottky diode and an ultra-low input voltage Seiko charge-pump IC. The 869 MHz monolithic design performs the best as the Schottky diodes are integrated in the chip and lower electromagnetic loss at the lower frequency. However, the size of an 869 MHz antenna is much bigger than a 2.45 GHz counterpart [23]. The output voltage of our power harvester is the highest in the table and it can be as high as 2.4 V by using a different model of the charge-pump IC [54]. With this output voltage, our power harvester can easily power up any general purpose µC, sensors and RF transceiver.

In general, an integrated CMOS chip has size and power advantages over the discrete component design such as for a biomedical implanted device where the size is strictly limited.
An integrated CMOS chip also has a cost advantage over discrete component design if it is produced in a large quantity. However, for the purpose of this thesis, only a few prototypes are needed to test out the idea in a very short turn around time and to test the effectiveness of the specialized antennas developed in Chapter 3 and 4 for low power transmission. Therefore, a discrete component design is more suitable to serve the purpose.

<table>
<thead>
<tr>
<th>References</th>
<th>This work [55]</th>
<th>[23]</th>
<th>[41]</th>
<th>[40]</th>
<th>[23]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency</td>
<td>2.45 GHz</td>
<td>2.45 GHz</td>
<td>3.85 GHz</td>
<td>920 MHz</td>
<td>869 MHz</td>
</tr>
<tr>
<td>Sensitivity (dBm)</td>
<td>-15.6</td>
<td>-13.5</td>
<td>-12.0</td>
<td>-14.1</td>
<td>-19.0</td>
</tr>
<tr>
<td>Maximum Output Voltage</td>
<td>2</td>
<td>1.5</td>
<td>1</td>
<td>1</td>
<td>1.5</td>
</tr>
<tr>
<td>System Efficiency</td>
<td>6.5% @ -12 dBm</td>
<td>5.7% @ -13 dBm</td>
<td>4.6% @ -14 dBm</td>
<td>2.7% @ -15.6 dBm</td>
<td>5% @ -13.5 dBm</td>
</tr>
<tr>
<td>Fabrication Process</td>
<td>discrete components on FR-4</td>
<td>0.5µm CMOS+ Schottky Diodes</td>
<td>0.18µm CMOS</td>
<td>0.18µm CMOS</td>
<td>0.5µm CMOS+ Schottky Diodes</td>
</tr>
</tbody>
</table>

Table 2-5: Performance comparison of different power harvesters.

2.4.5 Range Test for the Transponder

Instead of using a resistor in the previous section, a µC is connected to the system to verify the functionality of the whole wirelessly-power system. Lastly, the single diode rectifier with an SMS-7630 is incorporated to make a full rectenna system using simple FR-4 patch antennas attached to the R&S signal generator and rectifier circuit individually. With an RF power amplifier attached to the signal generator, it can radiate a total of 18.8 dBm (0.097 W) after power losses in the cabling to the anechoic chamber. The charge-pump IC and microcontroller are attached to the rectifier as shown in Figure 2.29.
Figure 2.29: The rectenna with 2.45 GHz patch antenna, the rectifier, Seiko IC and the microcontroller.

With this setup, the longest operational distance is measured from the antenna of the signal generator to the rectenna, as shown in Figure 2.30. The measured longest distance is 57 centimetres. Based on these measurements, according to the Friis transmission equation, 1W of transmitting power can easily power up the receiver over a 1 metre range.

Figure 2.30: The wirelessly-powered system inside the anechoic chamber for the measurement

2.5 Conclusion

The design of the highly power sensitive rectifier is constructed and able to demonstrate powering up the wirelessly-powered platform at 2.45 GHz using discrete components on a conventional FR-4 circuit board over a distance of 0.57 metres with just 100 mW and its performance is comparable to custom CMOS solutions with the help of the charge-pump IC. The
above work lays down a foundation building block for a low-cost and ultra-low-power battery-less wearable sensor platform which will be as shown in Chapter 5.

However, using a linear polarized patch antenna for power transmission as shown here has a very restricted limitation on the antenna orientation. As for a body-worn wirelessly-powered system, the human body is not always perfectly aligned to the power sourcing antenna at the base station. For example, if the antenna at the base station is in a vertical position and the antenna at the human body is in a horizontal position, the efficiency of the power transmission will be significantly degraded. Therefore, in the next chapter a better type of antenna topology is introduced to address this issue.
Chapter 3: Antenna Design

3.1 Introduction

In this chapter, a wideband wide-slot antenna topology will be introduced for robust off-body wireless power transmission. First, a standard linearly-polarized patch antenna is briefly described and then various methods to generate circular polarization are discussed. However, due to the bandwidth limitation or geometric complexity of a circularly-polarized (CP) patch antenna, it can be a problem when creating a low cost and simple body-worn antenna for microwave power transmission. Therefore, a simple wide-slot antenna topology is proposed to address this issue. The simulation shows that it has very wide impedance and axial ratio bandwidths when compared to a patch antenna. Finally, the performance of the antenna on FR-4 is measured and discussed.

3.2 Antenna Characteristics

Before designing a high performance antenna for the specific application, some important parameters have to be defined and then optimized during simulations.

3.2.1 Bandwidths

Bandwidth is defined as the frequency range that the antenna performs at an acceptable level [56]. There are two bandwidths that are of most concerned in this chapter for designing an antenna: impedance and axial ratio bandwidths. Impedance bandwidth is defined as the frequency range that the antenna can still maintain good impedance matching and having $S_{11}$ less than -10 dB is commonly considered as a good impedance matching at the present time [56, 57]. The axial ratio bandwidth is defined as the frequency range that the antenna is radiating in
circular polarization and an axial ratio with less than 3-dB is considered as good polarization purity in practice [26, 56, 58, 59]. The definition of axial ratio will be discussed in the later section.

3.2.2 Polarization

Polarization is defined as the orientation of the electric field coming from an antenna and it can be linear, circular or elliptical [56]. This is an important characteristic of an antenna as the polarization between the transmitting and receiving antennas has to match for them to communicate effectively. For example, a vertically polarized antenna cannot communicate with a horizontally polarized antenna. In a similar fashion, a circularly polarized antenna can only receive half of the power from a linear polarized transmitting antenna due to polarization mismatch [56].

Circular polarization can be left-handed circular polarization (LHCP) or right-handed circular polarization (RHCP) depending on the direction of the rotation of the electric field [56]. However, in reality circular polarization is difficult to achieve perfectly and in most cases it behaves as elliptical polarization with unequal orthogonal components in the electric field [56, 60, 61].

3.2.3 Axial Ratio

Axial ratio is a measure of the purity of the circular polarization of an antenna and it is defined as the absolute value of the sum of the electric fields divided by the difference of the electric fields from both polarizations and it is also defined by the ratio of the major axis to the minor axis of the polarization ellipse [31, 47, 56]:

60
Axial ratio = \(20\log \left| \frac{E_{RHCP} + E_{LHCP}}{E_{RHCP} - E_{LHCP}} \right| = 20\log \frac{E_{\text{Max}}}{E_{\text{Min}}}\)  

(3.1)

where \(E_{LHCP}\) and \(E_{RHCP}\) are the LHCP and RHCP electric field, respectively and \(E_{\text{max}}\) \(E_{\text{min}}\) are the major axis and minor axis of the polarization ellipse, respectively.

An axial ratio of 0-dB means that the polarization is perfectly circular. In reality, perfect circular polarization is hard to obtain and the so the axial ratio is usually bigger than 0-dB and 3-dB is currently considered as a good polarization purity in the antenna community. Therefore, 3-dB axial ratio bandwidth is commonly used as a figure of merit for the polarization performance of an antenna in most literature [59, 62-64].

3.2.4 Antenna Gain

Antenna gain is defined as the power of the radiation as a function of angles compared to an isotropic source and it is a function of efficiency and directivity [1, 56, 65]. A typical gain for a patch antenna is around 3-5 dBi due to variations of geometry and power loss in the material of the antenna [56].

3.3 Choices of Antennas

There are many types of antennas such as monopole, dipole, loop, horn, spiral, helical and patch antennas [65]. Monopole and dipole antennas are easily made using a simple wire and they are good choices for a transmitter radiating in all directions. However, the gain of a \(\lambda/2\) dipole is only 2.15 dBi and a monopole requires a ground plane that the whole structure will become 3-dimensional. A \(1\lambda\) loop antenna has a similar gain to a \(\lambda/2\) dipole and a large loop antenna has relatively high gain but it has a very large radiation resistance that makes input impedance matching difficult [65]. Horn and helical antennas have high gains and wide bandwidths, but the
physical sizes are too big to be portable. A patch antenna has higher gain than a dipole, but the bandwidth is very narrow. However, due to its simple planar structure, it can be manufactured with a standard etching processing for PCB. Therefore, it has been widely used in the industry such as mobile phones [56]. Due to its planar structure, it can be easily integrated into textile and so there is a growing interest to use wearable textile patch antennas for on-body communication [31, 66, 67].

### 3.4 Basics of a Patch Antenna

Microstrip or patch antennas are widely used in commercial applications because they are cheap, compact in size and easy to fabricate on a PCB using standard etching process [49]. A typical rectangular patch antenna is shown in Figure 3.1. A rectangular patch basically consists of a conductive patch and a ground plane with a dielectric in between. A patch is a piece of metal with length L and width W sitting on top of a substrate (circuit board) with height h:

![Figure 3.1: A standard linear polarized patch antenna.](image)

The frequency of operation is determined by the length of the patch and is approximately given by this equation [56]:

\[
\text{Frequency} = \frac{c}{2L}
\]
where $c$ is the speed of light, $L$ is the length of the patch antenna and $\varepsilon_r$ is the relative dielectric constant. The width $W$ controls the input impedance and radiation pattern. The permittivity $\varepsilon_r$ controls the fringing field which will affect the radiation efficiency. The directivity of patch antennas is approximately 5 dBi and the fields are linearly polarized.

### 3.5 Patch Antenna Feeding Techniques

There are many ways of feeding a patch antenna such as probe feed, inset feed, quarter-wavelength transformer feed and proximity feed. We will provide the brief background information of inset and proximity feedings in the following sections [68].

The first type of feed is an inset feed and since the edge of the patch antenna has high impedance, in order to match the 50 $\Omega$ line, the feed line goes toward the centre of the patch for lower input impedance as shown in Figure 3.2a. The exact match needs to be done experimentally on the length and the inset notch size [56]. The advantage of this feed is ease of fabrication as since it is a single layer it requires only standard etching process. The only disadvantage is that it requires manual tuning for the length of the inset and it suffers from spurious radiation from the feed line.

The second type of feed is when the feed line is not physically touching the patch antenna as shown in Figure 3.2b. The energy is radiated from the feed line to the patch through the substrate. But coupling exists between the feed line and the patch and this gives you extra freedom for tuning. The advantage is that it offers extra effective substrate thickness, and it will
give better radiation due to an increase of the fringing field. The disadvantage is that multilayer fabrication is required which means high cost of manufacturing.

The third type of feed is using a quarter-wavelength transformer to provide the impedance matching between a 50 Ω transmission line to the edge impedance of a patch antenna which is typically around 200 Ω as shown in Figure 3.2c. The disadvantage is the radiation pattern is slightly affected due to disturbance of the one radiating edge and it takes extra space for the transformer.

The fourth type of feed is a coaxial feed in which a transmission line directly feeds to the patch through the ground plane as shown in Figure 3.2d. The metal is removed from the ground plane to leave the space for the outer shield of coaxial line. This method provides the purest radiation pattern from all the above methods. But the disadvantage is that it requires drilling and locating the optimum 50 Ω point for feeding which in turn requires manual tuning.

![Figure 3.2: (a) Inset feed, (b) Proximity feed, (c) Quarter-wavelength feed, (d) Coaxial feed.](image)

**3.6 Wideband Patch Antenna**

A linearly-polarized microstrip patch antenna is great for its low cost, high gain and easy manufacturing on a printed circuit board. However, a standard patch antenna has a typical limited bandwidth of less than 4% on a FR-4 board and so the resonant frequency can be shifted
due to manufacturing variations [56]. This will degrade significantly the performance of the microwave power transmission in the wirelessly-powered system. There are numerous ways to increase the impedance bandwidth of a patch antenna which involves: a) impedance matching network [56], b) increase the volume of the antenna such the thickness of the substrate [56], c) multiple feed-points [68] and d) variations of antenna geometry to introduce extra resonant frequencies [69]. It has been reported that the impedance bandwidths can go up to 32% and 69% at centre frequencies 2.4 GHz and 7.7 GHz, respectively [70, 71]. Some examples of these wideband patch antennas are shown in Figure 3.3. However, these structures would be too complicated to implement on a flexible substrate such as conductive textile as it may be bent on a human body as shown in the next chapter and so the performance will be degraded.

![Figure 3.3: Examples of the wideband patch antennas [70, 71].](image)

**3.7 Circularly-Polarized Patch Antennas**

In order to deliver robust power transmission to a human body, circular polarization is preferred due to the independence of orientation between the antenna pair. Therefore, a wide band circularly-polarized antenna structure is desirable for such an application. A standard patch
antenna can generate circular polarization by doing: a) diagonal feed point, b) cutting off corners or c) feeding with a 90° branchline hybrid. These methods can offer 3% to 20% 3-dB axial ratio bandwidths [72, 73]. However, a 3-dB axial ratio bandwidth can be up to 34% at centre frequency of 1.8 GHz using dual feeds and it can be as high as 81% using 4 feed lines at centre frequency of 1.5 GHz with 3-dimensional feeding structure as shown in Figure 3.4 [74, 75].

Figure 3.4: Example of the wide 3-dB axial ratio bandwidth antenna with multiple feed lines [74].

However, this multiple feed structure can be difficult and costly to implement on a conductive textile and the 3-dimensional feeding structure which is not resistant to bending due to body movement. The precise spacings within the structure are difficult to maintain under stress if the design is implemented on a flexible textile. Therefore, it is preferred to have a single fed planar design and at the same time featuring wide band circular polarization for the body-worn antenna that will be introduced next.

3.8 Wide-Slot Antenna

Recently, CP antennas have been getting more attention from the wireless community where the traditional patch is adapted by including notches or choosing the feed point along the diagonal of the patch with limited bandwidth as mentioned earlier. However, a simple geometry
and single layer square slot antenna can easily outperform standard CP patches but with lower cost as shown in Figure 3.5 [76].

Figure 3.5: Example of a wideband CP wide-slot antenna [76].

The advantage of a wide slot antenna is that the geometry is very simple and it doesn’t require an extra layer for a ground plane while it features wide impedance bandwidth compared to a patch antenna. The cost saving can be significant if a large array of antennas is needed for higher gain. The physical size can be more compact without a ground plane and it implies that it can be seamlessly integrated into clothings for body-worn textile antennas. There is a significant bandwidth advantage of the wide-slot antenna over a standard patch antenna with a just single layer of metal and substrate as shown in the later section. An impedance bandwidth of 110% with centre frequency of 3.5 GHz and 3-dB AR bandwidth of 85% has been recently reported using a single sided FR-4 substrate [77]. The missing ground plane means lower cost for manufacturing, but it also implies that the radiation can go in both the front and the back
directions. However, this is not a problem as the antenna is used as a passive device to absorb microwave energy from a base station as shown in Chapter 1. The coupling to the body will significantly affect the antenna performance, but as shown in the next chapter this is not a problem once the distance between the antenna and body is at a reasonable distance and the gain of the antenna is even amplified due to body reflection.

The frequency of operation is determined by the length of the slot and is approximately given by this equation [56]:

\[
f_r \approx \frac{300}{(L+W)\sqrt{\varepsilon_{r,\text{eff}}}}, \quad \varepsilon_{r,\text{eff}} = \frac{2\varepsilon_r}{1+\varepsilon_r},
\]

(3.3)

In this empirical formula, \( f_r \) is the resonant frequency of the fundamental mode in GHz, \( L \) and \( W \) are the inner width and length of the slot in millimetres and \( \varepsilon_{r,\text{eff}} \) is the effective dielectric constant.

### 3.8.1 Wide Slot Antenna Feeding Techniques

There are two common feeding methods for a wide slot antenna which are microstrip line feed and coplanar waveguide (CPW) feed as shown in Figure 3.6a and b. The advantage of the microstrip line feed is that there is more freedom on the shape of the slot and feed line as they can overlap but on different layers [63, 78-82]. However, the disadvantage is that it requires an extra metal layer as the slot and the feed are on the opposite sides of the substrate. The other method is using a CPW feed which has the advantage of using only 1 layer of metal as there is no ground plane for a wide slot antenna [59, 77, 83-86]. However, the disadvantage is there is more constraint on the shape of the feed line and the slot as they cannot overlap and the gap in the waveguide has to be small to match to 50 \( \Omega \). Even with these disadvantages, a CPW feed
antenna topology is chosen in this context due to ease of manufacturing on a single sided FR-4 board, and this characteristic is even more important for manufacturing a textile antenna as shown in the next chapter.

Figure 3.6: (a) Wide slot antenna with CPW feed, b) Wide slot antenna with microstrip line feed at the back of the antenna

3.8.2 Circularly Polarized Wide-Slot Antenna

Similar to a CP patch antenna, the principles to generate circular polarization on a patch antenna can be applied to a CP wide-slot antenna. The two most common and simple ways to generate wide band circular polarization on a wide-slot antenna are by perturbation structures and/or a bent feed line [63, 78, 87, 88]. An example of such wide band CP antenna with perturbation structures is shown in Figure 3.7a [85]. Similar to a CP patch, the perturbation structures are placed on the diagonal line to produce 2 orthogonal electric fields with a phase difference of 90° similar to the case of a patch antenna with notched corners discussed earlier [85]. Similarly, a wide-slot antenna loaded with a 45° slanted strip can generate two orthogonal resonant modes similar to the case of a patch antenna with a diagonal slot [60, 84].
The other way to produce circular polarization from a wide-slot antenna is by means of the feed line structure and an L-shaped feed line is one of most simple and common methods as shown in Figure 3.7b [78]. The surface currents form two orthogonal currents in horizontal and vertical directions over a specific frequency range due to the shape of the microstrip feed line, and the two orthogonal currents will generate two orthogonal radiations in horizontal and vertical directions. Since the electrical length of the horizontal part of the L-shaped feed line is about 90°, the phase of the current of the horizontal component lags behind that of the vertical components by about 90°. Therefore, there is approximately 90° out-of-phase relationship between the vertical and horizontal fields, which eventually generates the CP radiation in the far-field [63, 78, 89]. The shape of the slot is optimized accordingly to improve the impedance and 3-dB AR bandwidths [89]. As suggested in the literature, experiments have been done to combine both the perturbation structures with an L-shaped feed line in a wide-slot antenna to deliver even wider impedance and 3-dB AR bandwidths [59, 78].

Based on the above discussion, an L-shaped feed line wide-slot antenna topology has been chosen for the application of power transmission to the wirelessly-powered sensor system due to its simplicity and wide bandwidth advantage over a patch antenna. The design will be firstly optimized on FR-4 in this chapter and then the simplified antenna geometry will be applied to a flexible textile for the application of power transmission body-worn antenna in the next chapter.
Figure 3.7: Examples of CP wide-slot antenna with a) diagonal perturbation structures, b) L-shaped feed line [78, 85].

3.9 Wide-Slot Antenna Design Procedures

3.9.1 Linearly Polarized Wide-Slot Antenna Simulation

The first step in designing a linearly polarized wide-slot antenna is to decide the dimension of the ground plane. As shown from the references at about 2.4 GHz, a ground plane of 60 mm by 60 mm is a good starting point as if the size is too small as it will affect the 3-dB AR performance [78]. Also according to equation (3.1) concerning the resonant frequency of the wide-slot antenna, the slot size should be about 40 cm by 40 cm and this is consistent with the literature [84, 85]. The size of the CPW feed line is obtained from the TXLINE based on the substrate and centre frequency. The FR-4 substrate has dielectric constant 4.45 and the loss tangent is 0.02 [49]. The gap between the ground and the feed line is 0.3 centimetres for a reasonable width of feed line for 50 Ω matching. The length of the feeding is manually tuned to
the specific frequency. The final dimension of the antenna is shown in Figure 3.8. In the simulation, differential ports are used as the ground is on the same metal layer as the input. Ports ‘-1’ are connected the ground and port ‘1’ is connected to the feed line and therefore the simulator will understand the reference ground is the outer metal surrounding the slot.

![Figure 3.8: Dimension in mm of a linear polarized wide-slot antenna on a 16 mm FR-4 substrate.](image)

The initial result of a linear polarized slot antenna is shown in Figure 3.9 and a standard patch antenna performance is also shown for comparison. As shown in the figure, the -10-dB impedance bandwidth is 29% and it is significantly wider than 1.5% of the patch antenna shown in the same graph. This advantage is essential because it implies that it has a high tolerance for manufacturing variations and this is especially important for the flexible textile used in the next chapter. Moreover, the wide band characteristic makes it suitable for on-body power receiving textile antenna as the bandwidth is significantly degraded due to the proximity of the human body. Although the wide bandwidth may susceptible to interference, this is not the problem for the application of power harvesting. The gain of the patch antenna is higher due to the fact it
only radiates into +z direction whilst the wide slot antenna radiates into both +z and –z directions as shown in Figure 3.10.

![Graph showing S11 parameters and gains of the wide-slot antenna and the patch antenna.](image)

**Figure 3.9:** $S_{11}$ parameters and gains of the wide-slot antenna and the patch antenna.

Now the simple linearly polarized wide-slot antenna has been simulated and it resonates at the specific targeted frequency range, a circularly polarized wide-slot will be designed next based on this simple antenna topology.
Figure 3.10: Radiation pattern for a wide-slot antenna and a patch antenna. a) $\varphi = 0$, b) $\varphi = 90^\circ$. 
3.9.2 Circularly Polarized Wide-Slot Antenna Simulation

As discussed earlier, producing circular polarization on a wide-slot antenna is commonly done using an L-shaped feed line for a wide bandwidth 3-dB AR bandwidth. The diagonal perturbation structure inside the slot can be added to further enhance the performance [78]. Therefore, the first attempt is to modify the feed line of the linear polarized wide-slot to L-shaped as shown in Figure 3.11. The feed line has similar width as the linear polarized wide-slot antenna, but it is bent at 90° to create the necessary phase delay between the x and y components of the electric fields as suggested earlier. The exact ratio between the length of the horizontal and vertical sections is manually tuned and it is easier to start at the centre of the slot and move around to get the best 3-dB AR ratio.

Figure 3.11: Simple L-shaped feed circularly-polarized wide-slot antenna in mm.

As shown in the Figure 3.12b, for such simple antenna geometry it can offer very wide 3-dB AR bandwidth with very little effort. The 3-dB bandwidth is 50% from 2.26 GHz to 3.78 GHz.
and is much wider than the 2.45 GHz ISM band. Note that the gap between the feed line and the ground has a significant effect on impedance matching at the lower frequency range as it provides strong coupling to the ground [90, 91]. However, it cannot be smaller than 5 mm as it will be very difficult to hand-make on a conductive textile as will be shown in the next chapter.

Figure 3.12: Characteristic of the simple L-shaped feed wide-slot antenna a) $S_{11}$, b) 3-dB AR.

However, as shown in Figure 3.12a the -10-dB impedance bandwidth is not sufficient to cover the whole 3-dB AR band and the ISM band. This is mainly due to the low input impedance and this can be improved by increasing the width of the feed line as suggested [78]. The tapering technique for better input impedance matching has been commonly used in wideband antenna designs [77, 92]. It is found that 2-level tapering of the feed line will give sufficient impedance matching across the ISM band and this finding is consistent with the literature [78, 92]. There is
an additional tuning stub on the vertical section of the feed line to further improve impedance bandwidth towards the lower frequency end. The stub will also change the current flow in the feed and so it will affect the generation of circular polarization. The exact dimension of the feed is given in the Figure 3.13 and the dimension of the ground is kept the same, 60 mm by 60 mm, as before. The dimension of the feed line has to be carefully tuned as it affects both the impedance and 3-dB AR bandwidths at the same time.

![Figure 3.13: Dimensions of the tapered L-shaped feed wide-slot antenna in mm.](image)

As shown in Figure 3.14, the impedance bandwidth is 66% from 2.1 GHz to 4.1 GHz and it is significantly wider than the one from the original simple L-shaped wide-slot antenna and the
linear polarized wide-slot antenna. The 3-dB bandwidth is 57% from 2.2 GHz to 3.9 GHz and it is about a 14% improvement over the original design while the impedance band is significantly improved as shown in Figure 3.15. However, there is a slight degradation in the polarization purity at around 2.5 GHz to 3.2 GHz due to the change in the current flow of the tapered feed line but it has been tuned to minimize the effect. Above all, the antenna has enough impedance and 3-dB AR bandwidths to cover the 2.45 GHz ISM band which is needed for providing circular polarization for the wirelessly-powered system in the subsequent chapter and the extra bandwidth provides enough guard band for the manufacturing variations which is especially important for a flexible textile antenna. Note that there is a resonant point at 3.7 GHz as it supports higher order modes similar to a patch antenna [73].

![Figure 3.14: The simulated S11 of the wide-slot antennas and the linear polarized wide-slot antenna is shown for reference.](image-url)
Figure 3.15: The simulated axial ratios for the wide-slot antennas.

### 3.9.3 Circular Polarization Mechanism

The basic idea of generating circular polarization using the L-shaped feed has been briefly described in an earlier section. The common way to verify this is to study the magnetic current flow quantitatively in a simulator [93]. As shown in Figure 3.16, the surface current is simulated using CST at 2.4 GHz and the current distribution is changing with the time. The majority of the current is going upward in the +y direction at 0° phase although there is a mixture of currents going in the different directions. At 90° phase, the majority of the current goes to the +x direction. At 180° and 270° phases, the current are going in the opposite directions as shown for 0° and 90° phases, respectively. Therefore, the current is turning in a clockwise direction and so this is left-handed CP as expected.
Figure 3.16: Simulated surface current distributions at different phases at 2.45 GHz.
3.9.4 Circular Polarization Distribution

Although the wide-slot antenna can offer wide 3-dB AR bandwidth in a simple geometry, the conical angle of the 3-dB AR becomes narrower as the frequency increases. This behaviour is visualized using CST 3D far-field simulator as shown in Figure 3.17. At 2.45 GHz, the AR distribution is fairly flat and smooth away from the centre covering a lot of area of the antenna surface at the far field. It implies that the circular polarization purity can be maintained at wider angles from the centre and more importantly the LHCP gain is not degraded significantly.

![Figure 3.17: AR distribution of the wide-slot antenna at 2.45 GHz.](image)

However, at 3.5 GHz the AR distribution has a lot of spikes around the centre and the AR rapidly increases even slightly away from the centre as shown in Figure 3.18. This limits the usefulness of the antenna in the high frequency range as the LCHP gain will decrease more steeply away from the centre of the antenna in the far field. This problem becomes more apparent during AR measurement in a later section.
Figure 3.18: AR distribution of the wide-slot antenna at 3.5 GHz.

### 3.9.5 Simulated Radiation Patterns and Gain

The simulated radiation patterns for the wide-slot antenna at 2.45 GHz are shown in Figure 3.19. It is noted that it produces LHCP in +z direction and RHCP in –z direction which is consistent with the surface current distribution analysis as shown before. It has a good circular polarization purity at the centre as it has good isolation between the LHCP and the RHCP in +z direction which is consistent with the 3-dB AR plot. The maximum gain is 3.3 dBi at \( \varphi = 10^\circ \) and \( \vartheta = 10^\circ \) and the gain at centre is 3.1 dBi which is only slightly off from the maximum.

Since the antenna radiates in both +z and –z directions, one suggestion is by adding a reflector at the back the antenna, the gain can be amplified due to the reflection of the –z direction radiation [36, 78].
Figure 3.19: Simulated radiation pattern at 2.45 GHz for the wide-slot antenna a) $\varphi = 0^\circ$, b) $\varphi = 90^\circ$.

The simulated LHCP gain at the centre of the antenna is shown in Figure 3.20 and it shows fairly flat response over the whole 2.45 GHz ISM band with the simulated gain over 3 dBi. The gain falls off quickly above 3 GHz and it is down to 0 dBi at 3.6 GHz.
3.10 Experimental Results for the CP Wide-slot Antenna

3.10.1 Measurement Methodology

The measurements are done using a calibrated horn antenna ETS-3115, which has a wide enough bandwidth to cover the frequency range of interest [94]. Note that the gain of the calibrated antenna varies slightly over the frequency range, and therefore compensation must be carried out for the gain measurement of the CP wide-slot antenna. LHCP and RHCP helix antennas are built to measure the radiation patterns for the wide-slot antenna and the dimension of the helix antennas are determined using a simple formula [65]. The gains for the helix antennas are determined using the calibrated horn antenna ETS-3115. The gain measurement for the wide-slot antenna is measured using the horn antenna in the vertical and horizontal positions because a linear horn antenna can only pick up half of the radiation from a CP antenna due to polarization mismatch. Therefore, the gain of a CP antenna is the total sum of the measured gain.
from both horizontal and vertical positions [65]. The measurement of the axial ratio is done by turning the linear polarized horn antenna to different degrees to obtain the maximum and minimum gains and the difference between the two is the corresponding axial ratio of the antenna under test [56, 65]. The horn antenna is the best candidate due to its purity of linear polarization and stable radiation pattern across a wide frequency range. All measurements are carried out with an Agilent E8361A PNA inside an anechoic chamber to minimize noise from the environment and reflections.

The tapered-L CPW-fed CP wide-slot antenna is manufactured in-house using the standard etching process which is the same as for making of the RF rectifier presented in the previous chapter and the final design is shown in Figure 3.21.

![Figure 3.21: The CP wide-slot antenna on the FR-4.](image)

### 3.10.2 Experimental Results

The measured $S_{11}$ of the CP wide-slot antenna is shown in Figure 3.22 and note that the simulated and measured results are closely matched. At the low frequency end, there is less than 0.1 GHz difference between the 2 results and at the high frequency end they are almost exactly
matched. The -10-dB impedance bandwidth is 61% (2.1 GHz to 4.1 GHz) and is able to cover the whole ISM band as expected.

![Graph showing S11 of the CP wide-slot antenna](image)

**Figure 3.22: The S11 of the CP wide-slot antenna.**

The measured axial ratio is plotted in Figure 3.23. The measured 3-dB AR bandwidth is 60% (1.8 GHz to 3.4 GHz) and the centre frequency is shifted by 0.4 GHz comparing to the simulated AR. The larger difference between the two results at higher frequencies may due to very limited angle of the 3-dB AR at the centre of the antenna at high frequencies. Therefore, when there is a small shift in the angle either due to the radiation pattern of the antenna or by the misalignment of the centre of the antenna under test to the centre of the calibrated horn antenna, the measured AR will be significantly degraded as shown earlier in Figure 3.16. Therefore, even though it may have good AR at the centre at a high frequency, but it is only available at a very limited angle and so it may be able to measure in reality.
The measured radiation patterns at 2.45 GHz are shown in Figure 3.24. The measured pattern is closely matched to the simulated pattern shown in Figure 3.19. With LHCP radiation in the +z direction, it radiates at RHCP at the –z direction as expected. The maximum gain is 3.1 dBi at $\phi = 10^\circ$ and $\theta = 10^\circ$ which is closely matched to what was predicted in the simulation.
Figure 3.24: Measured radiation pattern at 2.45 GHz at $\phi = 0^\circ$ and $90^\circ$. 
The measured gain for the wide-slot antenna is plotted in Figure 3.25. The LHCP gain at 2.45GHz is 2.7 dBiC and the maximum is 4 dBiC at 2.9 GHz. The overall trend of the measured gain is closely related to the simulated gain as shown.

![Figure 3.25: Measured LHCP gain of the wide-slot antenna.](image)

The performance of this wide-slot antenna is summarized in Table 3-1: Performance comparisons of different antenna designs. The designs from the other references with the similar physical area are also included in this table for comparison. The tapered-L feed wide-slot antenna performs significantly better than those with only simple straightened feed line in terms of 3-dB AR bandwidth. This work has a comparable 3-dB bandwidth to the one with L-shaped feed annular-ring wide slot antenna that requires extra metal layer for the microstrip feed line [78]. As the results suggest the two layer design may offer a more compact design in [78] because the feed is not constraint by the slot dimension. The only drawback of this work is that
the -10-dB impedance bandwidth is not able to cover the whole 3-dB AR bandwidth even though the whole 2.45 GHz ISM band is fully covered.

<table>
<thead>
<tr>
<th>Antenna</th>
<th>Area mm x mm</th>
<th>Impedance Bandwidth</th>
<th>3-dB AR Bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>This work</td>
<td>60x60</td>
<td>61%, 2.1-4.1 GHz</td>
<td>60%, 1.8-3.4 GHz</td>
</tr>
<tr>
<td>[76]</td>
<td>70x70</td>
<td>39.6%, 1.6-2.5 GHz</td>
<td>12.4%, 1.8-2.1 GHz</td>
</tr>
<tr>
<td>[85]</td>
<td>60x60</td>
<td>52%, 1.6-3.1 GHz</td>
<td>28.8%, 2.3-3.0 GHz</td>
</tr>
<tr>
<td>[84]</td>
<td>60x60</td>
<td>51.4%, 2.0-3.4 GHz</td>
<td>48.8%, 2.1-3.4 GHz</td>
</tr>
<tr>
<td>[78]</td>
<td>45x45</td>
<td>79%, 2.0-4.7 GHz</td>
<td>56.2%, 2.1-3.7 GHz</td>
</tr>
</tbody>
</table>

Table 3-1: Performance comparisons of different antenna designs.

### 3.11 Conclusion

The characteristics of the tapered-L feed wide-slot antenna has been investigated in this chapter and it has shown significant advantages in terms of impedance and AR bandwidth over a standard CP patch antenna while the geometry is still relatively simple. In the next chapter, this wide-slot antenna geometry will be implemented on a conductive textile for an on-body power transmission application.
Chapter 4: Textile Antenna Design

4.1 Introduction

Recently there has been growing use of textile antennas for body-centric applications such as healthcare, GPS and fire fighter personal communications [7, 29, 32, 33, 35, 67, 95-98]. A textile antenna is chosen for such applications due to its light-weight and easy integration into clothing. Designing a wide band CP wide-slot antenna on FR-4 has been successfully demonstrated in Chapter 3. Although the antenna design in Chapter 3 has a very wide bandwidth, that antenna design still may be too complicated to be implemented on a flexible textile. The complex geometry is hard to be maintained when the antenna is under physical stress on a body. For body-worn antenna application, a simpler and more robust antenna geometry is needed to be easily hand-made on a flexible conductive material. In order to translate the antenna geometry to a conductive textile, this chapter firstly starts with off-body simulation of the conductive textile with felt as the flexible substrate for the wide-slot antenna design. Then, the simulation model with body tissues is considered and the antenna is optimized in that environment. Once the simulation is complete, the antenna is built on a self-adhesive conductive textile with felt as the substrate. Finally, the antenna characteristics are measured under different conditions such as substrate thickness and bending.

4.2 Recent Developments of Textile Antenna

Patch or dipole antennas are the most common topologies for the textile antennas and there are some other special types in the forms of a belt, a zipper and a button as shown in Figure 4.1 [98-100]. A textile patch antenna with a ground plane is preferred because it can minimize the
effect of the proximity of a human body and retain most of the near fields within the patch and
the ground plane and so the radiation efficiency is preserved [36, 101]. Similar techniques
producing circular polarization in Chapter 2 can be also applied to a textile antenna. However,
due to the relatively narrow bandwidth of a patch antenna, bending on a textile patch antenna can
significantly degrade the antenna performance [29, 31].

Figure 4.1: a) Wearable textile slot patch antenna, b) Belt antenna, c) Button antenna [98, 100, 102].

The application of a textile antenna has been concentrated on data communication, but not
until recently it has been applied for low power RF energy transmission [7]. An antenna with
circular polarization is preferred over linear polarization for power transmission due to its
independence of orientation. However, to generate circular polarization using a patch antenna as
shown in Figure 4.1a on a human body results in a narrow axial ratio bandwidth and the centre
frequency can shift due to bending to match the curvature of the human body or manufacturing
variations [31, 36]. A wide band solution with a broadband phase shifter and power divider is
proposed to achieve circular polarization with simulation results [7]. However, the structure is
complicated to manufacture as shown in Figure 4.2 and there is no measured result to prove its power efficiency.

![Figure 4.2: Wearable multi-frequency RF energy harvesting textile antenna [7].](image)

In Chapter 3, a high performance wide-slot circularly polarized antenna has been proposed. A CPW-fed antenna features wide impedance and axial ratio bandwidths that are essential to body-worn textile antenna applications as it has a higher tolerance to different conditions on a human body. Also it only consists of a single layer of metalized textile without a separate ground plane. Therefore, a cost effective body-worn textile antenna can be designed based on a wide-slot antenna topology. However, due to the missing ground plane, the influence of the human body on the antenna’s near field has to be carefully simulated in MWO to ensure the effectiveness of operation. Also the antenna structure has to be simple and robust enough for bending on a human arm and is able to operate at various distances from a human body.
4.3 Textile Antenna Simulation

Before designing a textile antenna for efficient power transmission on-body, the material has to be chosen and the dielectric constant of the substrate has to be determined. Then the wide-slot antenna design from Chapter 3 can be optimized in the simulator within the proximity of a human body.

4.3.1 Textile Materials

A planar textile antenna consists of 2 layers, the top layer is conductive and the bottom layer is the substrate. Both materials have to be highly flexible but also rigid enough to conform to any surface of a human body. A Flectron™ self-adhesive EMI shielding sheet has been chosen for this purpose due to its flexibility and high conductivity (≤0.070 Ω/²) at very low cost. It is made of a thin metal coating of nickel on non-woven fabrics and it can be easily cut without a special tool [103]. A common 1 mm-thick 100% acrylic self-adhesive felt is chosen as the substrate for the antenna due to its flexibility and easy stacking for different thickness. The dielectric constant and loss tangent are determined as 1.5 and 0.02, respectively, by matching the experimental and simulation results of a simple patch antenna at a specific frequency [104]. Since the loss tangent of the felt substrate is similar to FR-4, therefore the efficiency of the textile antenna is similar to a design with FR-4 substrate.

4.3.2 Textile Wide-slot Antenna Simulation

The wide-slot antenna with wide impedance and axial ratio bandwidths has been designed in Chapter 3. However, the same design on FR-4 is needed to scale manually according to equation (3.1) in Chapter 3 due to the change in the dielectric constant of the substrate. Also note that the
gap within the feed line can’t be scaled due to the size of the mounting of the SMA connector. The outer dimension of the ground surrounding the slot is scaled up slightly from 60 mm X 60 mm to 76 mm X 76 mm with a substrate thickness of 1 mm for more aggressive area utilization and the antenna is shown in Figure 4.3. The model is constructed in MWO assuming that the conductive textile is a perfect conductor for simplicity. This simplification on modelling is acceptable because this wide-slot antenna features wide impedance and AR bandwidths, a slight shift in the centre frequency does not have significant degradation in performance in reality as shown in the experimental results later. However, the actual antenna gain will be smaller than in the simulation due to resistive losses in the conductive textile. The simulated $S_{11}$ and AR of the textile antenna are shown in Figure 4.4.

Figure 4.3: Scaled antenna geometry from Chapter 3 on a conductive textile with felt substrate.
The simulated $S_{11}$ of the textile antenna is very similar to the one shown in Figure 3.14 as expected and the -10-dB impedance bandwidth is 1.6 GHz (2.1-3.8 GHz). The AR bandwidth is wider than is shown in Figure 3.15 but it is very close to 3-dB at 3.3 GHz and the 3-dB AR bandwidth is 2.3 GHz (2.1-4.4 GHz). As shown in this simulation, the textile antenna performs in a similar way as the wide-slot antenna on FR-4 after size scaling. However, the simulation of a textile antenna on-body is necessary because the coupling of the human body can significantly affect the antenna performance [101].

### 4.3.3 Textile Wide-Slot Antenna On-Body Simulation

As the textile antenna is to be worn on a human body, due to the dielectric constant of a human body, the centre frequency of the textile is shifted to a higher frequency according to equation 3.1 in Chapter 3 [101]. The dielectric constants of human tissue from 10 Hz to 100 GHz...
have been studied in detail in the literature [105-108]. Therefore, the simulation model in MWO will be based on these dielectric constants and are listed in Table 4-1 [108].

<table>
<thead>
<tr>
<th>Tissue</th>
<th>Conductivity [S/m]</th>
<th>Relative permittivity</th>
<th>Loss tangent</th>
<th>Penetration depth (m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Skin</td>
<td>1.46</td>
<td>38</td>
<td>0.28</td>
<td>0.023</td>
</tr>
<tr>
<td>Fat</td>
<td>0.11</td>
<td>5.28</td>
<td>0.15</td>
<td>0.117</td>
</tr>
<tr>
<td>Bone</td>
<td>0.81</td>
<td>18.5</td>
<td>0.32</td>
<td>0.029</td>
</tr>
<tr>
<td>Muscle</td>
<td>1.73</td>
<td>52.7</td>
<td>0.017</td>
<td>0.022</td>
</tr>
<tr>
<td>Air</td>
<td>0</td>
<td>1</td>
<td>0</td>
<td>N/A</td>
</tr>
</tbody>
</table>

Table 4-1: Dielectric constants and penetration depth for human tissues at 2.45 GHz [108].

The thickness of each human tissue is based on the cross section of a human body from the Visible Human Project as shown in Figure 4.5 [109]. This cross section provides a good estimation for the tissue thickness of a human arm and body. Note that different people and different parts of a body will have different distributions of tissues as the model used in Figure 4.6. As shown in the model, skin, fat, muscle and bone are modelled with estimated thickness and the antenna is on top of the skin. The thickness of each layer is adjusted according to the observation from Figure 4.5. Note that only one side of a human body is specified because microwave at 2.45 GHz can’t go deep inside a human body according to the penetration depth in Table 4.1. The total thickness of the human tissue model already exceeds the penetration depth. Therefore, the textile antenna being under the air and on top of the skin, fat, muscle and bone in the model is sufficient for simulation accuracy.
Figure 4.5: Cross section of a human body on a 1 cm by 1 cm grid [109].

Figure 4.6: Human model with different layers of tissues in simulation.
The simulated results of the textile antenna with the human model are shown in Figure 4.7. As predicted by equation (3.1), the centre frequency is inversely proportional to the square root of the dielectric constant. Since the human body has much higher dielectric constant than air, therefore the centre frequency of both $S_{11}$ and AR of the textile antenna on-body are shifted to a lower frequency. The centre frequency is shifted significantly due to a missing ground plane when comparing to a patch antenna design with a ground plane. However, not only the manufacturing cost is lower without an extra layer of conductive material as a ground plane, but also a human body can serve as a reflector that will increase the gain of the antenna as shown later in this chapter. Based on this simulation, the textile antenna has to be redesigned to shift the frequency back to the 2.45 GHz ISM band for the intended application.

![Figure 4.7: Simulated characteristics of the textile antenna close to human body: a) $S_{11}$, b) AR.](image-url)
Another observation is that the distance between the antenna and the body has a significant effect on the antenna gain as shown in Figure 4.8. When the substrate thickness is 1 mm, the gain at 2.45 GHz is only -8.6 dBic. It is because the near field of the antenna is coupled to the human body and the energy loss is significant [32, 101]. As the substrate thickness is increased, the antenna gain is increased accordingly up to about 5 dBic when the thickness is at 20 mm. There is no significant increase in gain with the substrate thickness greater than 20 mm; however, with the felt thickness being 20 mm on a human arm, it may be too bulky to be bent and worn. Therefore, a 13 mm thick substrate is chosen to be the minimum thickness for future simulation and optimization, as the further increase in thickness will not offer significant gain. The thickness will be increased experimentally to obtain better antenna gain for the power transmission application with proximity to a human body.

![Figure 4.8: Antenna gains with different substrate thicknesses.](image-url)
4.3.4 Textile Wide-Slot Antenna On-Body Optimization

Due to the high dielectric constant of the human body, the centre frequency of the textile wide-slot antenna is shifted to a lower frequency. In order to increase the resonant frequency, the slot size can be reduced according to equation 3.1. The design process will be restarted from a simple L-shape feed line similar to the design process in Chapter 3. It is found that a bigger ground plane is preferred for better AR, both by simulation and as suggested [78]. Therefore, not only the slot size is reduced, it is moved away from the feed line as shown in Figure 4.9. The gap between the feed line and the ground is increased to 0.7 mm for better manufacturing tolerance and as a guard band to prevent shorting the feed line to ground when it is bent. The CPW feed line is also simplified when compared to Figure 4.3 for ease of cutting with little degradation in impedance matching. The outer dimension is kept the same as 76 mm x 76 mm for the same area utilization. The horizontal section of the feed line is placed in the middle of the slot initially and it is tuned for a better performance by iterations.

Figure 4.9: Basic L-shaped feed line wide-slot textile antenna on a body.
The simulated performance of the textile antenna is shown in Figure 4.10. The lower frequency end of the impedance band barely covers 2.4 GHz and the AR is around 4.5-dB to 6.5-dB which is less than the desired 3-dB. Although the AR is flat over the high frequency range, the AR needs to be lower for better circular polarization purity. Therefore, we can adopt the same performance enhancement techniques learned in Chapter 3 for the antenna design process on FR-4.

![Graph of Figure 4.10](image)

**Figure 4.10: Simulated performance of the textile antenna from Figure 4.7: a) S\textsubscript{11}, b) AR.**

By widening the width of the feed line, the input reactance is reduced and so it helps to widen the impedance bandwidth of the textile antenna [78]. At the same time, the AR is also improved due to refinement of the ratio between the horizontal and vertical sections of the feed line; it will affect the phase between the horizontal and vertical electric fields, which is the key factor in
generating circular polarization [63, 78, 89]. The ratio is tuned by moving the horizontal section of the feed line up or down to get the best result. It is found that by widening the width of the vertical and horizontal sections of the feed line to 16 mm and 25 mm, respectively, it can deliver the best impedance and AR bandwidths. The final dimension of the textile antenna is shown in Figure 4.11.

Figure 4.11: Final dimension in mm of the wide-slot textile antenna on body.

The antenna geometry is optimized not only for performance, but also optimized for ease of cutting by hand. The feed line geometry is reduced to a very basic L shape for ease of manufacturing comparing to the wide-slot antenna in Chapter 3. The gap between the feed line and the ground is kept at 0.7 mm, which is possible to achieve with a crafting knife. Even if the gap is slightly bigger than 0.7 mm, the impedance bandwidth is only slightly degraded.
The simulated performance of the wide-slot antenna on body is shown in Figure 4.12. The impedance bandwidth is 40% (2.1 – 3.1 GHz) and the 3-dB AR bandwidth is 25% (2.2 – 2.8 GHz). The maximum AR bandwidth achieved is degraded compared to the FR-4 counterpart which may be due to body reflection as suggested [78]. However, the goal for this textile antenna is to deliver robust power transmission within the proximity of a human body in the 2.45GHz ISM band and the simulation shows that the whole ISM band is well covered with this textile antenna.

Figure 4.12: The simulated performance of the textile antenna: a) $S_{11}$, b) AR.

The radiation patterns of the textile antenna at 2.45 GHz are plotted in Figure 4.13. Both patterns show very good separation between the RHCP and LHCP, which is consistent with the
AR shown in Figure 4.12 earlier. The maximum gain of the pattern is very close to the centre of the antenna which is desirable for power transmission.

Figure 4.13: The simulated radiation pattern of the textile antenna at 2.45 GHz: a) $\varphi = 0^\circ$, b) $\varphi = 90^\circ$. 
4.3.5 Textile Antenna Manufacturing

The textile antenna geometry has been optimized for ease of manufacturing due to its simple geometry. The mirrored antenna geometry is printed on a paper with 1:1 scale and then it is stuck on the back of the protective layer of a conductive EMI shielding textile as shown in Figure 4.14a [103].

It is then cut using a crafting knife with the help of a ruler to keep the knife running straight on the mask. Once it is finished, the protective layer is removed and then it is stuck on the felt substrate. The end-launch SMA connector is press-fit to the antenna by clamping the feet together and then conductive epoxy is applied to further secure the conductor to the textile. The advantage is that it is not only simple, but also the structure won’t be deformed due to the heat of the soldering temperature. The final product is shown in Figure 4.14b.

Figure 4.14: The hand-made wide-slot textile antenna: a) The mask on the back of the conductive textile, b) The completed textile antenna with a SMA connector.
4.4 Textile Antenna Experimental Results on a Human Body

The Agilent E8361A PNA with an ETS-3115 horn antenna is used for the $S_{11}$ measurements inside the anechoic chamber. The antenna is suspended securely from the SMA connector so that it lies straight during measurement as shown in Figure 4.15.

![Figure 4.15: The author with the textile antenna mounted on his body in an anechoic chamber.](image)

The simulated and measured on-body $S_{11}$ of the textile antenna are shown in Figure 4.16. The thickness of the substrate and the clothes on a human body are 11 mm and 2 mm, respectively. The unbent measured -10-dB impedance bandwidth is 981 MHz (39%, 1988-2969 MHz) and the simulated impedance bandwidth is 1024 MHz (39%, 2100-3124 MHz). The difference between the simulated and measured $S_{11}$ bandwidth can be due to manufacturing error and the variation of dielectric constants and thickness of human tissues.
This antenna will be fitted on a human arm as a power receiving antenna on the wirelessly-powered sensor platform in the next chapter. Therefore, the effect of bending on the impedance bandwidth has to be studied [38, 66, 110]. The antenna is bent in the x direction on a human arm with a radius of about 5 cm and this results in a shift of the impedance band (2183-3083 MHz) as shown in Figure 4.17. A human arm is chosen because it is the worst case for bending of the antenna on a body without breaking the antenna. Also the impedance matching is degraded across the whole band because bending in the x direction increases the size of the gap between the feed line and the ground, therefore the impedance is increased accordingly away from 50 Ω. However, due to the wideband characteristic of the textile antenna, the -10-dB impedance bandwidth is still able to cover the whole 2.45GHz ISM band. Therefore, it offers a safe guard
band for any shifting in resonant frequency due to bending and variation of dielectric constants of a human body.

![Graph showing measured S11 for bent and unbent conditions](image)

**Figure 4.17:** The measured bent and unbent $S_{11}$ of the textile antenna on a human body.

The axial ratio is an indicator of the quality of circular polarization. The simulated and measured axial ratios at the centre of the textile antenna are shown in Figure 4.18. The simulated and measured 3-dB axial-ratio bandwidths (ARBW) are 591 MHz (23%, 2237-2828 MHz) and 443 MHz (18%, 2226-2669 MHz), respectively. The small difference may be due to the curvature of a human body and mismatch in dielectric constant from the model in the simulator. However, the 3-dB ARBW is still more than enough to cover the whole 2.45GHz ISM band with good circular polarization purity and is significantly wider than a standard circularly polarized patch [31, 67, 111]. Bending the textile antenna on a human arm has a small effect on the 3-dB ARBW (413 MHz) and has shifted the central frequency slightly higher, and hence the wide impedance and AR bandwidths are still maintained on a human body.
Figure 4.18: The measured axial ratios of the textile antenna on a human body.

When the textile antenna is worn on a human arm as a power receiving antenna, the transmitting power from a base station can arrive at an angle. Therefore, the change in axial ratio against angle is important and needs to be investigated. The axial ratio versus zenith (x axis) angle at 2.45 GHz on a human body is shown in Figure 4.19. Note that the unbent textile antenna 3-dB and 6-dB axial ratios can cover up to 90° and 110° zenith angles on a human body, respectively. Note also that the bent textile antenna is fitted along a human arm and has slightly wider zenith angles for 3-dB and 6-dB axial ratios. The axial ratio distribution is not symmetric around the centre due to its asymmetry in the antenna structure.
Figure 4.19: The measured axial ratios for unbent and bent textile antenna over zenith angle at 2.45 GHz on a human body.

The RHCP and LHCP gains versus zenith angles at 2.45 GHz on a human body is shown in Figure 4.20. The maximum RHCP gain is 3.5 dBic and is much higher than the LHCP gain as expected and it is consistent with the axial ratio shown in the previous Figure 4.19. Note that bending does degrade the centre RHCP gain by 1 dBic, but it still maintains a good purity of RHCP as the LHCP gain is much lower in the centre. Note also that the textile can switch to LHCP by simply flipping to the opposite side. The polarization in +z direction is RHCP and in –z direction is LHCP which is a common characteristic for a wide-slot antenna [76, 78].
The thickness of the substrate has a significant effect on the textile antenna gain especially as the ground plane is missing because the near field is coupled and absorbed into the body. As the distance between the antenna and a body is increased, the body will behave like a reflector to strengthen the antenna gain as shown in Figure 4.21 [95, 101, 112]. When the minimum distance between the conductive textile to a human body is at 3 mm (1 mm substrate + 2 mm of clothes), the gain is weak due to its near field significantly coupled to the human body [36, 101, 112]. However, as the distance increases by inserting the more layers of felt, the gain gradually increases up to maximum 4.9 dBi at the substrate thickness of 23 mm and the gain goes down as the resonant frequency is shifted away from 2.45 GHz. It is due to the fact that the human body serves as a reflector, part of the –z direction LHCP radiation is reflected and it becomes RHCP in +z direction and so the +z direction RHCP gain is strengthened due to the reflected
radiation. The other factor affecting the gain is due to the antenna efficiency as shown in Figure 4.21. As the antenna is further away from the human body, the coupling between the antenna and the body is weakened, so less energy is lost in the body [36]. Therefore, the improved efficiency further enhances the gain as the antenna further away from the body.

Figure 4.21: The RHCP gain and simulated efficiency with variations of the substrate thickness at 2.45 GHz.

The radiation pattern for the textile antenna is shown in Figure 4.22. As expected the measured radiation patterns are similar to the simulated radiation patterns shown earlier.
Figure 4.22: The measured radiation patterns at 2.45GHz with $\phi = 0^\circ$ and $\phi = 90^\circ$.

The performance comparisons of this CP textile antenna to the other CP textile antennas are summarized in Table 4-2. As shown in the table, all the other works are based on patch antenna
designs with a ground plane and a thin substrate which have very small impedance and AR bandwidths as expected from a patch antenna. This work has shown a significant advantage on manufacturing simplicity and bandwidths over a standard patch design on a human body.

<table>
<thead>
<tr>
<th>Antenna</th>
<th>Area mm x mm</th>
<th>Impedance Bandwidth</th>
<th>3-dB AR Bandwidth</th>
</tr>
</thead>
<tbody>
<tr>
<td>This work</td>
<td>76x76</td>
<td>39%, 2.0-3.0 GHz</td>
<td>18%, 2.2-2.7 GHz</td>
</tr>
<tr>
<td>[67]</td>
<td>50x50 (excluding ground plane dimension)</td>
<td>9.3%, 2.35-2.58 GHz</td>
<td>2.4%, 2.4-2.46 GHz (simulated)</td>
</tr>
<tr>
<td>[102]</td>
<td>110x110</td>
<td>5.2%, 1.605-1.690 GHz</td>
<td>1.2%, 1.6-1.62 GHz</td>
</tr>
<tr>
<td>[31]</td>
<td>65x65 (excluding ground plane dimension)</td>
<td>1.9%, 1.6-1.68 GHz</td>
<td>1.8%, 1.61-1.63 GHz</td>
</tr>
</tbody>
</table>

Table 4-2: The performance comparisons of CP textile antennas.

### 4.5 Conclusion

The design of a wideband circularly polarized textile on-body antenna is presented. The design has shown its flexibility in terms of geometry, material, impedance bandwidth and 3-dB ARBW. With 39% impedance bandwidth and 18% 3-dB ARBW on a human body, it has a huge guard band for manufacturing tolerance and variation of dielectric constants of the different human body. Its structure is simple as it doesn’t require a ground plane leading to lower cost and providing single-fed to produce circular polarization.

Now all the fundamental blocks for building a wirelessly-powered sensor platform are ready and so the complete sensor system will be constructed and discussed in the next chapter.
Chapter 5: Application on Wirelessly-Powered Sensor Platform

5.1 Introduction

From the previous chapters, the knowledge of the fundamental building blocks for a wirelessly-powered system has been acquired and now a more complicated sensor platform is ready to be made. The concept of a wirelessly-powered sensor platform has been briefly described in Chapter 1 and a simple wirelessly-powered system with a microcontroller controlling a flashing LED has been demonstrated in Chapter 2. In this chapter the technical detail of a complete wearable wirelessly-powered temperature sensor system will be demonstrated including C programming of an ultra low power microcontroller. Due to the weak RF energy received from the rectifier, a charge-pump IC is introduced for the voltage step-up to improve the power sensitivity of the system. Once all the temperature data is received at the microcontroller, a low-power RF transmitter is programmed to send the data back to a remote laptop for data acquisition. Finally, the system with a textile antenna worn on a human arm demonstrates the longest workable distance with only 50 mW transmitted power from a base station. This system is able to successfully transmit the temperature data from the transponder to a remote laptop in an anechoic chamber.

5.2 Wirelessly-Powered Sensor Platform Design

The top-level architecture is shown in Figure 1.1 in Chapter 1. Similar to other wirelessly-powered systems, weak RF power is received at the antenna and then the rectifier converts the
RF power into a DC voltage. The charge-pump IC steps up the voltage to 2.4 V and the energy is stored in a capacitor to provide stable system power for later use. The µC performs the timing and control tasks for the sensor and the RF transmitter. It collects the sensor data and converts it to a 10-bit number using its internal analogue-to-digital converter (ADC). The RF transmitter is used for sending the data back to a base station.

5.2.1 Rectifier

The detail of designing the rectifier can be found in Chapter 2. The Skyworks SMS-7630 is chosen over the HMS-2850 for higher sensitivity. It has been found that single diode configuration is the most power effective for the power harvesting application due to less energy loss on components and copper traces on a lossy substrate. The schematic for the single diode rectifier is shown in Figure 2.13 and FR-4 is chosen as the substrate due to its low cost.

5.2.2 Ultra Low-Power Charge-Pump IC

For the low-power RF energy scavenging application, the input level can be below -15 dBm (0.03 mW) as shown in Chapter 2. Therefore the voltage collected from the microwave rectifier is substantially below the normal operating voltage for a µC. The standard voltage multiplier can help to step up the low input voltage to a higher level. However, due to the energy loss in the substrate and the finite voltage barrier of the diode at microwave frequencies, putting in more than 3 or 4 stages of the voltage multiplier does not yield a higher voltage as shown in Chapter 2. Also, the µC draws current even before it is turned on at 1.8 V and so energy is wasted. Therefore, a solution is needed to provide a low input voltage and energy efficient voltage step-up.
The Seiko S-882Z charge pump IC is capable of stepping up at the extremely low input voltage of 0.3 V and the block diagram is shown in Figure 5.1 [54]. The S-882Z is turned on when the input voltage from the rectifier is at minimum of 0.3 V. The step-up voltage is stored in the external capacitor \( C_{CPOUT} \). Note that the OUT pin is disconnected from the storage capacitor and so there no leakage current going to the \( \mu \)C. Once the storage capacitor \( C_{CPOUT} \) reaches the threshold voltage 2.4 V, the gate of M1 will be on and the storage capacitor will discharge to the OUT pin. The function of pin VM is to shut down the IC when the voltage at pin VM is higher than \( C_{CPOUT} + 0.1 \) V. In this application, pin VM is grounded.

![Block diagram of the Seiko charge-pump IC](image)

Figure 5.1: Block diagram of the Seiko charge-pump IC [54].

### 5.2.3 Low-Power Microcontroller

In Chapter 2, a simple system with a \( \mu \)C flashing the LED at a regular time interval was demonstrated. The Microchip PIC12LF1822 is chosen as the \( \mu \)C for the sensor system due to its improved low-power consumption and easy-to-use programming interface [113]. It is an 8-bit
μC with a built-in 10-bit A/D converter with internal voltage references, comparators, timers and EEPROM operating on a supply voltage as low as 1.8 V with a typical power consumption 110 μW. The other common choice of μC can be TI MSP430F11 [114]. However, it doesn’t have internal voltage references and it is a 16-bit μC with a typical power consumption 352 μW that consumes too much power for the application in this thesis.

The function of this μC is to receive the analogue temperature data from the sensor and then the data is converted into a 10-bit number. That number is then formatted into packets and sent to the RF transmitter so that the remote RF receiver will receive that data and display the results on a laptop. Therefore, the μC has to coordinate all of the data traffic between chipsets with the correct timing under the restricted power constraint.

The μC is programmed in C under the MPLAB programming environment and the C compiler is free from HiC [115-117]. The complete program is listed in the Appendix and the important sections of the program will be briefly discussed here.

After assigning the function of each pin for the μC, the internal analogue-to-digital converter (ADC) needs to be configured so that it will convert the analogue data from the pin connected temperature sensor to a 10-bit number compared to a fixed internal 2.0 V voltage reference:

```c
// configure ADC
ADCON1 = 0b01000011; //use internal voltage reference
ADCON0 = 0b00001001; //use the 2\textsuperscript{nd} ADC and turn it on
FVRCON = 0b10000010; //enable fixed 2.048V internal reference
```
Before sending any packets to the RF transmitter, it needs to be configured whenever it is powered up and the detailed code for configuring the Nordic N2401G is shown in Appendix [118, 119]:

```c
__delay_ms(1); // wait 1.5 ms for the transmitter to be ready
__delay_us(500);
configure_transmitter(); // configure the transmitter
```

Since the ADC is already configured, now the μC starts the conversion and waits until the data is acquired:

```c
// sample analogue input
GO_nDONE = 1; // start conversion
while (GO_nDONE) // wait until done
;
```

Once the ADC operation is done, the data is stored in 2 registers called ADDRESH and ADRESL. Now the data is aligned and stored in the array called data_array and ready to be sent to the RF transmitter. The function of transmit_data() is to shift the data serially to the RF transmitter and signal the transmitter to send the data back to laptop at 2.52 GHz. The completed code can be found in Appendix:

```c
// 10 bits result
    data_array[0] = ADRESH;

// Top 8-bit ADC result
    data_array[1] = ADRESL;
__delay_ms(1);
```
\_\_delay\_us(500); //Wait for transmitter to be ready
transmit\_data(); //We send 4 bytes

Once the data is received at the RF receiver, the data is sampled using Saleae logic analyzer that is connected to a laptop through the USB port [120].

5.2.4 Temperature Sensors

Two different Maxim temperature sensors have been chosen for this platform to study the effect on system performance for different loads. One is a low current (15 µA) and low accuracy (±2°C) Maxim MAX6607 and the other one is a high current (140 µA) and high accuracy (±0.5°C) DS600 [121, 122]. Once the Seiko charge-pump IC charges up the storage capacitor, it will turn on the µC and the temperature sensor. The charge-pump IC will charge up the storage capacitor again and release the energy to the µC when it reaches 2.4 V and so the length of idle time between operations is determined by the level of the input power to the charge-pump IC required to charge up the storage capacitor. It is experimentally found that in the case of the higher power consumption temperature sensor DS600, a 67 µF capacitor is required to provide all of the energy needed for the µC, temperature sensors and transmitter. But for the lower power temperature sensor MAX6607, a 44 µF capacitor is only required and so the capacitor recharge time is reduced. In order to optimize the temperature resolution, the internal voltage reference from the 12LF1822 is utilized. The resolution of the ADC is 10-bit (2048 steps) and the internal voltage reference is 2.048 V. Therefore, the resolution of the sampled temperature data is more than enough for the application. The actual temperature resolution of this system will be tested over a wide temperature range in the later section.
5.3 The Transmission Link between the Transponder and the Base Station

In order to set up a reliable transmission link between the transponder and the base station, a robust and reliable transceiver pair is needed. Since the communication is done in the 2.45 GHz ISM band, it means that there is a chance of collision between the system’s signals and WiFi signals. The Nordic nRF2401A transceiver modules have been chosen due to its low pin count, low power consumption and CRC error check [118]. The chance of collision has been minimized using the ShockBurst™ technology and the detail can be found in the datasheet [118]. The μC has been programmed to interact with this transceiver. When the μC is on, it will sample the temperature sensor’s analogue signal and convert it into a 10-bit number using its internal ADC. Then the μC will set up the transceiver’s parameters for transmission, with regards to the carrier frequency, data rate, the address and data packet. Once the configuration for the transmitter is done, the μC will serially shift the data to the transmitter and the transmitter will compute the 16-bit CRC checksum for data integrity and packet the data automatically for transmission at 2.52 GHz. Since the minimum packet size for data packet is 32 bits, two identical 10-bit numbers are sent for each transmission.

At the other end of the communication link, the same Nordic nRF2401A transceiver module is used but it is configured as a receiver. The data from the nRF2401A is sampled using the Saleae logic analyzer that is connected to a laptop through the USB port.
5.4 Experimental Results

The wirelessly-powered temperature system is initially constructed with discrete components on individual circuit boards with a patch antenna and tested off-body as shown in Figure 5.2. In the later section, the final system will be constructed on a single piece of circuit board with all the electronics soldered with the CP textile antenna bent on a human arm as a wearable wirelessly-powered temperature system.

![Figure 5.2: The complete wirelessly-powered system with a 2.45GHz patch antenna, the rectifier, 12LF1822 µC, DS600 temperature sensor, Seiko charge-pump IC and Nordic RF transmitter module.](image-url)

5.4.1 Time Interval between Received Data

Once the transponder has received enough energy to charge up the storage capacitor to 2.4 V, it will turn on the µC to sense the analogue signal from the temperature sensor. Then the signal is converted to a 10-bit number. Finally, it will be shifted into the transmitter serially.

The measured required minimum input power levels for the single diode rectifier with SMS-7630 rectifier is -15 dBm which compares well with the results reported [27, 123]. This means
that the voltage sensitivity is about 40 mV for a 50 Ω system and only 32 µW is required to turn on the whole wirelessly-powered system.

For the system using the low power temperature sensor (MAX6607), two 22 µF capacitors (total 44 µF measured) are used as a power storage device for the charge-pump IC. On the other hand, for the high power temperature sensor (DS600), three 22 µF capacitors (total 67 µF measured) are used to deliver more power to the load at the expense of the charging time. The minimum time intervals between samples received at the base station are recorded, summarized in Table 5-1 and plotted in Figure 5.3.

<table>
<thead>
<tr>
<th>Input Power Level (dBm)</th>
<th>Minimum time interval between samples received with a 44 µF capacitor (seconds)</th>
<th>Minimum time interval between samples received with a 67 µF capacitor (seconds)</th>
</tr>
</thead>
<tbody>
<tr>
<td>-4</td>
<td>1.7</td>
<td>2.5</td>
</tr>
<tr>
<td>-5</td>
<td>2.1</td>
<td>3.1</td>
</tr>
<tr>
<td>-6</td>
<td>2.5</td>
<td>3.7</td>
</tr>
<tr>
<td>-7</td>
<td>3.1</td>
<td>4.6</td>
</tr>
<tr>
<td>-8</td>
<td>3.9</td>
<td>5.8</td>
</tr>
<tr>
<td>-9</td>
<td>4.9</td>
<td>7.4</td>
</tr>
<tr>
<td>-10</td>
<td>6.4</td>
<td>10</td>
</tr>
<tr>
<td>-11</td>
<td>8.6</td>
<td>13.3</td>
</tr>
<tr>
<td>-12</td>
<td>12.1</td>
<td>18.8</td>
</tr>
<tr>
<td>-13</td>
<td>18.2</td>
<td>29.2</td>
</tr>
<tr>
<td>-14</td>
<td>30.6</td>
<td>50.7</td>
</tr>
<tr>
<td>-15</td>
<td>70.3</td>
<td>135.8</td>
</tr>
</tbody>
</table>

Table 5-1: The minimum time intervals between samples received at the base station.

As shown in Figure 5.3, at an input power level of -4 dBm, the charge time for both capacitors are reasonably fast, although the difference in sample interval is about 47%. At an input power level of -10 dBm, the sample interval is 6.4 second and 10 second for 44 µF and 67
µF storage capacitors respectively. This means it takes approximately 56% longer to charge up the larger capacitor to drive a bigger load. At an input power level of -15 dBm, the charge time is increased dramatically to 135s for a 67 µF storage capacitor which is almost 2 times longer than for the 44 µF counterpart.

![Figure 5.3: The minimum time interval versus input power for received samples from the transponder.](image)

Note that Table 5-1 shows the minimum time interval between samples received, but there is a chance that the packet is corrupted by signals from the environment such as WiFi signals or background noise. In practice, the interval between samples can be longer than shown in the table due to packet corruption.

### 5.4.2 Power Efficiency of the System

The efficiencies of the overall system at different power levels with a load resistance of 180 Ω are measured. The 180 Ω load is chosen because the µC with the RF transmitter has similar input resistance when it is transmitting the sensor signal back to a base station. The storage capacitor recharge interval represents the idle time of the µC because the charge-pump IC takes time to charge up the 44 µF storage capacitor to 2.4 V. At the lowest operational input power
level of -15 dBm, it takes the charge-pump IC about 70 seconds to charge up the capacitor to 2.4 V to turn on the μC. However, at an input power level of -10 dBm, it only takes about 6.4 seconds to charge up the storage capacitor. The measurable efficiencies within the harvester are defined in Chapter 2.

The power efficiencies of different sections of the system based on Table 5-2 are plotted in Figure 5.4. The lowest operational input power level of the single diode rectifier with SMS-7630 is -15 dBm and the total system efficiency is 2.6%. The charge-pump IC efficiency is decreasing from 52% to 32% as the power level is increasing except when the power level is at -15 dBm due to the fact that it is barely turned on. The rectifier efficiency is increasing as the power level is increasing at a decreasing rate up to maximum of 27%. The system efficiency increases with a decreasing rate as the power level increases beyond -9 dB as the efficiency of the charge-pump IC decreases faster than the efficiency of the rectifier increases. The total system efficiency reaches 9.2% when the input power is -9 dBm.

<table>
<thead>
<tr>
<th>Input Power Level (dBm)</th>
<th>Rectifier Efficiency (%)</th>
<th>Charge-pump IC Efficiency (%)</th>
<th>System Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>-4</td>
<td>26.5</td>
<td>31.8</td>
<td>8.42</td>
</tr>
<tr>
<td>-5</td>
<td>26.7</td>
<td>32.1</td>
<td>8.58</td>
</tr>
<tr>
<td>-6</td>
<td>27.1</td>
<td>33.5</td>
<td>9.08</td>
</tr>
<tr>
<td>-7</td>
<td>25.1</td>
<td>36.8</td>
<td>9.21</td>
</tr>
<tr>
<td>-8</td>
<td>23.3</td>
<td>39.6</td>
<td>9.22</td>
</tr>
<tr>
<td>-9</td>
<td>23.2</td>
<td>39.8</td>
<td>9.24</td>
</tr>
<tr>
<td>-10</td>
<td>21</td>
<td>42.3</td>
<td>8.91</td>
</tr>
<tr>
<td>-11</td>
<td>17.7</td>
<td>47.1</td>
<td>8.34</td>
</tr>
<tr>
<td>-12</td>
<td>15.7</td>
<td>47.6</td>
<td>7.47</td>
</tr>
<tr>
<td>-13</td>
<td>13.1</td>
<td>47.7</td>
<td>6.25</td>
</tr>
<tr>
<td>-14</td>
<td>9</td>
<td>51.7</td>
<td>4.68</td>
</tr>
<tr>
<td>-15</td>
<td>8</td>
<td>32.2</td>
<td>2.56</td>
</tr>
</tbody>
</table>

Table 5-2: The power efficiency of different input power levels.
Figure 5.4: Measured efficiencies vs. input power level of the rectifier, the charge-pump IC and the overall system.

5.4.3 Resolutions of the Temperature Sensors

Although the temperature error for a factory-calibrated temperature sensor DS600 is ±0.5 °C [122], there are quantization errors from the 10-bit ADC and a maximum 7% internal temperature reference error from the µC that will degrade the resolution of the temperature reading, so the expected worst case resolution which can be achieved is ±0.7 °C. In case of a lower power MAX6607, the expected worst case maximum resolution will be ±2.1 °C within the +20 °C to +50 °C temperature range due to larger sensor error according to the data sheet.
Figure 5.5: Measured quantized sensor reading vs. temperature change.

As shown in Figure 5.5, the 10-bit quantized sensor voltages are measured against the temperature for both systems. The internal constant 2.048 V voltage reference inside the µC is utilized to increase the sensor resolution and minimize the effect of the supply voltage variation. The measured results show that the DS600 and MAX6607 has the linearity of $R^2 = 0.9997$ and $0.9999$, respectively. The resolution of the DS600 and MAX6607 is 0.3 °C and 0.2 °C, respectively from -10 °C to 80 °C. The reason why the lower cost MAX6607 has better performance than DS600 is because it has wider voltage variation against temperature change and its linearity is better than the worst case in the datasheet. The DS600 may have a less absolute temperature offset due to the factory calibration, but this is not important for in this case because the calibration can be done in the laboratory and it can be compensated within the program of the µC [124, 125]. The temperature resolution cannot be lower than 0.2 °C due to the limitation of the 10-bit quantizer in the µC.
5.4.4 Performance of the Wirelessly-Powered Temperature Sensor System

The comparison of our work to the other monolithic temperature sensors is shown in Table 5-3. Due to the high threshold voltage of 0.25 μm CMOS technology, the power sensitivity of the monolithic solutions is only up to -12.5 dBm. However, the RF sensitivity of our power harvester is -15 dBm and it is best in Table 5-3 due to the combination of a highly efficient Schottky diode and an ultra-low input voltage Seiko charge-pump IC. The 900 MHz and 450 MHz monolithic solutions provide a better power consumption due to the integration of the sensor and logic and lower transmission frequency. The major current consumption of this system is due to the Nordic transceiver. Although the peak current consumption is 13 mA according to the datasheet, it only happens over a very short period of time according to the ShockBurst™ technology [118]. Therefore, the measured total system current consumption whilst active is only 1.38 mA with the DS600 and μC running at 500 kHz. However, the current consumption can be reduced by 13% by using a newer Nordic transceiver or by 19% using a lower transmission power with the expense of higher data error rate [118, 126]. Note that extra power has been consumed to compute the 16-bit CRC checksum for data integrity and a total of 56 bits are sent in each packet for each 32-bit data. Lowering the RF frequency can lead to a longer operation range according to the Friis free space equation. However, the size of a 450 MHz or 900 MHz antenna is much bigger than its 2.45 GHz counterpart.
<table>
<thead>
<tr>
<th>References</th>
<th>This work [127]</th>
<th>[25]</th>
<th>[26]</th>
</tr>
</thead>
<tbody>
<tr>
<td>RF Power Frequency</td>
<td>2.45 GHz</td>
<td>900 MHz</td>
<td>450 MHz</td>
</tr>
<tr>
<td>Sensitivity (dBm)</td>
<td>-15</td>
<td>-12.3</td>
<td>-12.5</td>
</tr>
<tr>
<td>Maximum Output Voltage</td>
<td>2.4</td>
<td>2</td>
<td>2.5</td>
</tr>
<tr>
<td>Temperature Resolution (°C)</td>
<td>0.2 or 0.3 (-10 °C to 80 °C)</td>
<td>0.8 (-10 °C to 80 °C)</td>
<td>&gt;0.8 (-14 °C to 40 °C)</td>
</tr>
<tr>
<td>Data Transmission Method</td>
<td>32-bit data with 16-bit CRC checksum + 8-bit header</td>
<td>Backscattering</td>
<td>frequency modulation at 2.3 GHz</td>
</tr>
<tr>
<td>Current Consumption during Active (mA)</td>
<td>1.4</td>
<td>1.1</td>
<td>1.1</td>
</tr>
<tr>
<td>Fabrication Process</td>
<td>discrete components on FR-4</td>
<td>0.25µm CMOS + Schottky diodes</td>
<td>0.25µm CMOS</td>
</tr>
</tbody>
</table>

Table 5-3: The performance comparisons of wirelessly-powered temperature sensors.

### 5.4.1 Range Test of the Wirelessly-Powered Temperature Sensor System

The single diode rectifier with an SMS-7630 is incorporated to make a full rectenna system using a 12 dBi gain 4x4 patch antenna array attached to a Rhode & Schwarz signal generator with a 5 dBi gain single patch antenna attached to the transponder as shown in Figure 5.3. With an RF power amplifier attached to the signal generator, it can radiate a total of 16.6 dBm (0.046 W) after power losses in the cabling to the anechoic chamber. The RF receiver is attached to a logic analyzer for sample recording and the setup is shown in Figure 5.6.
Figure 5.6: The 2.45 GHz patch antenna array (on the left of the figure) is powering up the transponder inside an anechoic chamber with a laptop receiving the temperature samples.

The longest measured operational distance is 2 metres. The measured power loss between the antenna pair including the connectors is 30 dB with the maximum distance. Therefore the received input RF power to the harvester is approximately -14 dBm. Additional power losses are due to polarization and beam pattern mismatch between the two antennas. Based on these measurements, according to the Friis transmission equation stated earlier, 1 W of transmitted power can easily power the harvester over a few metres.

5.4.2 Range Test of the Wearable Wirelessly-Powered Temperature Sensor System with the Textile Antenna

Lastly, the final system is assembled on a single PCB for a better mobility as a wearable sensor system as shown in Figure 5.7a. The textile antenna is bent and attached to the
transponder on a human body as shown in Figure 5.7b. A RHCP helix antenna with 12.5 dBic gain is attached to a Rhode & Schwarz signal generator and it can radiate a total of 16.6 dBm (0.046 W) after power losses in the cabling to the anechoic chamber are included. The RF receiver is attached to a logic analyzer for sample recording to the laptop as before. The longest measured operational distance is 1.74 meters and it is comparable to the system with the linearly polarized patch antennas on FR-4 off-body.

Figure 5.7: a) The complete wirelessly-powered system on a single circuit board, b) Testing the system on a human arm inside the anechoic chamber.

5.4.3 Conclusion

The design of a complete wirelessly-powered sensor system with wideband circularly polarized textile on-body antenna is presented. The textile antenna is bent on a human arm with the wirelessly-powered sensor system proved to be operational over a 1.7-meters range with
under 0.05 W power transmitted from a base station. This demonstrates that a very low-cost and ultra low-power battery-less µC sensor platform can be implemented and integrated into clothes for other sensing applications in the future [14, 19, 128-130].
Chapter 6: Conclusion and Future Work

6.1 Conclusion

It has been demonstrated that a highly power sensitive rectifier at 2.45 GHz ISM band can be constructed using off-the-shelf components. By combining a simple input impedance matching, a 1-stage rectifier and a charge-pump IC together, the power sensitivity level of -15 dBm is achieved and it is very competitive with a standard CMOS solution.

A simple circularly-polarized antenna geometry has been investigated and it is found that the single-sided wide-slot antenna not only features the wide impedance bandwidth, but it also has the wide axial ratio bandwidth by using the simple tapered L-shaped feed line. The CPW feed enables the whole antenna structure to be contained in a single layer of metal without an extra layer for a ground plane. This design leads to a robust, compact and cost effective textile antenna to be worn on a human body.

The circularly-polarized textile antenna is then optimized on single conductive layer with proximity to a human body. Due to its wideband characteristics, it is able to accommodate the shifting in central frequency due to effect of a human body and its wide impedance band is able to cover the whole 2.45 GHz ISM band with a good purity of circular polarization. It is shown that the performance of this CP textile antenna is significantly better than any other on-body CP textile patch antennas have been reported.

Lastly, a complete wirelessly-powered temperature sensor system with the textile antenna is constructed and worn on a human arm. It is shown that with just 50 mW of microwave power,
the system is able to operate at 1.74 metres and send the temperature data back to the laptop at a regular time interval.

Therefore, this work presented in this thesis leads to a power sensitive and robust wearable wirelessly-powered sensor system and it is ready to be applied to any other sensing applications [129].

6.2 Future Work

This thesis concentrates on the antenna design for only on-body power transmission; however, the antenna design to transmit energy to an implanted sensor at mid-field is still under development in the community [131]. A wearable low frequency coil array for a near-field energy transfer to band-aid sensor has been investigated, but optimized design for a such array at microwave frequency to power up a mid-field implanted cardiac sensor is still not developed [131]. Recently, it has been reported that the advantages using the 1.7 GHz over 200 MHz in terms of the compactness and power efficiency in simulation as shown in Figure 6.1.

![Figure 6.1: Wireless power transmission using 1.7 GHz (right) and 200 MHz (left). Red indicates more power transmitted as shown in the right [131].](image)
As the orientation of the antenna inside the body is unknown and changing, the powering antenna outside the body should be CP so that the power transmission efficiency is unchanged and independent to the orientations [132, 133]. Therefore, the CP textile antenna developed in this thesis may be a possible candidate for the above application.

![Wirelessly-powered ingestible camera from the Chinese University of Hong Kong](image)

**Figure 6.2: Wirelessly-powered ingestible camera from the Chinese University of Hong Kong [132].**

Although the single feed CP antennas in Chapter 3 and 4 are very simple to implement and have very wide impedance and axial ratio bandwidths; however the radiation with good axial ratio only covers limited solid angle close to the centre of the antenna. A CP wide-slot antenna with multiple feed lines has never been reported and it may significantly improve the solid angle coverage with good axial ratio if the similar feeding structure of the wideband CP patch antenna can be applied as shown in Figure 3.4 [74].
APPENDIX

//For RF2401A transmitter
/

  config_setup word 16 bits

  23: 0 Payloads have an 8 bit address

  22: 0

  21: 1

  20: 0

  19: 0

  18: 0

  17: 1 16-Bit CRC

  16: 1 CRC Enabled

  15: 0 One channel receive

  14: 1 ShockBurst Mode

  13: 1 1Mbps Transmission Rate

  12: 0

  11: 1

  10: 1

    9: 1 RF Output Power

    8: 0 RF Output Power

    7: 0 Channel select (channel 2)

    6: 0
#include <htc.h>
#define _XTAL_FREQ 2000000 // oscillator frequency for _delay()

// Config: ext reset, no code protect, no brownout detect, no watchdog,

// power-up timer enabled, 4MHz int clock
__CONFIG(MCLRE_OFF & CP_OFF & BOREN_OFF & WDTE_OFF & PWRTE_OFF & FOSC_INTOSC);

#define TX_CE LATA1 //Enable transmission
#define TX_CS LATA0 //Enable configuration
#define TX_CLK1 LATA4 //Clocking for N2401
#define TX_DATA LATA5 //Data for N2401
#define SENSOR RA2 //Sensor input

unsigned char data_array[4]; //8-bit data
unsigned int count;
void configure_transmitter(void);
void transmit_data(void);
void main()
{
    unsigned char dcnt; // delay counter
    unsigned char i;
    // Initialisation
    TRISA = 0b000100; // Config RA2 as input
    ANSA2 = 1; // RA2 as analogue sensor input
    IRCF3 = 1; // 1101 4 MHz
    IRCF2 = 1; // 1100 2 MHz
    IRCF1 = 0; // 1011 1 MHz
    IRCF0 = 0; // 0111 500kHz(default)
    // configure ADC
    ADCON1 = 0b01000011;
    //0------- MSB of result in ADRESH<7> (ADFM = 0)
    //-100---- Tad = Fosc/4 = 1 MHz)
    //-------00 Use VCC
    //-------11 Use internal Vref
    ADCON0 = 0b00001001;
    //-11101-- select temp sensor
    //-11111-- FVR output
    //-00001-- AN1
AN2

---1 turn ADC on (ADON = 1)
FVRCON = 0b10000010;

--- enable fixed voltage reference
--- enable temperature sensor
--- enable low range
---10 ADC 2.048V fixed voltage ref
---01 ADC 1.024V fixed voltage ref

// initialize the received data array
data_array[2] = 0x00;
data_array[3] = 0x00;

// Main loop
for (;;) {
    //3ms requirement from powerup to active for n2401
    //But can be smaller than 3 ms
    configure_transmitter();

    GO_nDONE = 1;       // start conversion
    while (GO_nDONE)    // wait until done
        ;
//10 bits result

data_array[0] = ADRESH; //Top 8-bit ADC result

data_array[1] = ADRESL; //Low 2-bit ADC result, MSB justified

data_array[2] = ADRESH; //Top 8-bit ADC result

data_array[3] = ADRESL; //Low 2-bit ADC result, MSB justified

delay_ms(1);
delay_us(500); // Can be smaller than 300us ~250us
transmit_data(); // We send 4 bytes
}

void configure_transmitter(void)
{
    unsigned char i;
    unsigned long config_setup;
    // Config Mode
    TX_CE = 0;
    TX_CS = 1;
    // Delay of 5us from CS to Data is taken care of by the for loop
    // Setup configuration word, we only need upper 24 bits, zero pad the lower 8 bits
config_setup = 0b0010001101111110010000000000;

// ----------------------1111001-----------RF freq
// ---------------------10----------RF power 10 = -5db
// -------------------1-------------------enable 1Mbps
// ---------------1----------------------CRC enable
// ---------------1----------------------16 bit CRC

for(i = 0; i < 24; i++)
{
    TX_DATA = (config_setup & 0x80000000) > 0;    //Display data
    TX_CLK1 = 1;
    TX_CLK1 = 0;
    config_setup <<= 1;    //shift 1 bit to left
}

//Configuration is active on falling edge of CS
    TX_CE = 0;
    TX_CS = 0;
}

//This sends out the data stored in the data_array
//data_array must be setup before calling this function
void transmit_data(void)
{


unsigned char i, j;
unsigned char temp, rf_address; //8-bit RF address
TX_CE = 1;

//Clock in address
rf_address = 0b11100111; //Power-on Default for all units
for(i = 0 ; i < 8 ; i++)
{
    TX_DATA = (rf_address & 0x80) > 0; //Display data
    TX_CLK1 = 1;
    TX_CLK1 = 0;
    rf_address <<= 1; //shift 1 bit to left
}

//Clock in the data_array
for(i = 0 ; i < 4 ; i++) //4 bytes
{
    temp = data_array[i];
    for(j = 0 ; j < 8 ; j++) //One bit at a time
    {
        TX_DATA = (temp & 0x80) > 0; //Display data
        TX_CLK1 = 1;
        TX_CLK1 = 0;
        temp <<= 1; //shift 1 bit to left
    }
TX_CE = 0; //Start transmission
BIBLIOGRAPHY


[130] N. Chaimanonart and D. J. Young, "An adaptively RF-powered wireless batteryless in vivo EKG and core body temperature sensing microsystem for untethered genetically

