Radio Frequency
Energy Harvesting for
Embedded Sensor Networks in
the
Natural Environment

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Abstract

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Radio Frequency Energy Harvesting for
Embedded Sensor Networks in the Natural Environment

The agricultural sector is an emerging application area for Wireless Sensor Networks (WSNs). This requires sensor nodes to be deployed in the outdoor environment so as to monitor pertinent natural features, such as soil condition or pest infestation. Limited energy supply and subsequent battery replacement are common issues for these agricultural sensor nodes. One possible solution is to use energy harvesting, where the ambient energy is extracted and converted into usable electrical form to energise the wireless sensors.

The work presented in this thesis investigates the feasibility of using Radio Frequency (RF) energy harvesting for a specific application; that is powering a generic class of wireless ground-level, agricultural sensor networks operating in an outdoor environment. The investigation was primarily undertaken through a literature study of the subject.

The first part of the thesis examines several energy harvesting/wireless energy transfer techniques, which may be applicable to power the targeted agricultural WSN nodes. The key advantages and limitations of each technique are identified, and the rationale is being given for selecting far-field RF energy harvesting as the investigated technique. It is then followed by a theoretical-based system analysis, which seeks to identify all relevant design parameters, and to quantify their impact on the system performance. An RF link budget analysis was also included to examine the feasibility of using RF energy harvesting to power an exemplar WSN node – Zyrox2 Bait Station.

The second part of the thesis focuses on the design of two energy harvesting antennas. The first design is an air-substrate-based folded shorted patch antenna (FSPA) with a solid ground plane, while the second design is a similar FSPA structure with four pairs of slot embedded into its ground plane. Both antennas were simulated, fabricated and tested inside an anechoic chamber, and in their actual operating environment – an outdoor field. In addition, a power harvester circuit, built using the commercially available off-the-shelf components, was tested in the laboratory using an RF signal generator source. The results from both the laboratory and field trial were analysed. The measurement techniques used were reviewed, along with some comments on how to improve them.

Further work on the RF energy harvester, particularly on the improvement of the antenna design must be carried out before the feasibility and viable implementations for this application can be definitively ascertained.
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Publications

1) Conference/ Journal Papers:


2) Poster Presentation:

1 Introduction

1.1 Introduction

The work presented in this thesis is in the area of Radio Frequency (RF) energy harvesting. The main aim of this work is to investigate the feasibility of using RF energy harvesting to power a generic class of wireless ground-level sensor network deployed in an outdoor environment.

In recent years, there has been a growing industrial and academic interest in the deployment of wireless sensor networks (WSNs). These network systems, consisting of spatially distributed sensor nodes, are used in various civilian and industrial applications, such as structural monitoring [1, 2], habitat monitoring [3], smart healthcare [4, 5] and inventory tracking [6]. One emerging WSN application is in agriculture sector where sensor nodes are deployed to monitor soil conditions [7], such as moisture, mineral content and temperature. The data collected from these agriculture sensors can be used to manage irrigation, predict crop yield, and to control fertilisation systems, thereby improving crop quality.

A number of WSN applications for agriculture have been implemented worldwide. In the United States, a group of researchers from Iowa State University [7] developed a prototype wireless soil sensor to collect and send data about soil moisture within a field. The sensors were designed to be buried about a foot deep (30 cm) in a grid pattern 80 to 160 feet apart. By assessing the data gathered from the sensors, researchers could understand precisely how water moves through a field and develop better models to predict crop growth and yield. The sensors could also help farmers to manage their nutrient and water resources, enabling them to maximise yields and profits.

Accenture Technology Lab [8] constructed a wireless network covering a 30 acre vineyard in Sonoma Valley, California. By monitoring soil moisture, leaf moisture, and air temperature, irrigation of the vineyard was controlled in a more efficient and economic way. The system was also used to give warning of possible grape frost damage. Another WSN for agriculture was constructed by LOFAR-Agro Consortium from the Netherlands [9], in which wireless sensor nodes measured the microclimate in potato crops. Information gathered from the wireless sensors was analysed to detect the onset of phytophthora - a common fungal disease found in potato crops.
1.2 Motivation

Energy supply has always been a key limiting factor to the lifetime of the agricultural WSNs. Current commercial WSN nodes are typically powered by onboard batteries, as shown in Figure 1.1, which have fixed energy rating and limited lifespan. To maintain the sensor operability, these batteries need to be replaced in due time. The maintenance cost of WSN nodes can be very high, especially if the network is deployed in hard-to-service locations. For example, a buried wireless soil sensor node must be unearthed before the exhausted batteries could be replaced. This would increase the labour cost. Depending on the deployment location, the cost of changing batteries could range from $10 per node for easily accessible nodes up to $100 per node [10]. This replacement cost would be even higher if the network consists of thousands of nodes distributed across a large field. ON World, one of the leaders in wireless research, predicted that the labour cost of changing batteries in the wireless sensors sector could be $1.1 billion over the next several years [11].

In addition, there are growing concerns over the disposal of batteries. Most wireless sensor batteries contain heavy metals such as lead, cadmium and mercury, which can pollute the environment if they are improperly disposed of. In the UK, it is reported that more than 600 million household batteries are disposed to landfill sites every year [12]. In May 2009, The Waste Batteries and Accumulators Regulations [13] were introduced in the
U.K. with the aim of reducing the amount of waste batteries going to landfill, as well as increasing battery recycling rates.

The motivation behind this research is to overcome the drawbacks of limited battery life and subsequent replacement, by exploring possible alternatives to conventional, disposable batteries in powering the agricultural wireless sensor nodes, thus making the WSN indefinitely self-sustaining. One possible solution is to use energy harvesting, in which ambient energy is captured, converted into electrical energy, and stored. Several energy harvesting techniques using different ambient sources, such as light radiation, temperature difference, electromagnetic field, human power and vibrations, have been reported in the literature [14-16]. Depending on the intended application, selection of a feasible energy harvesting scheme is mainly determined by two criteria; these are the environmental conditions under which the sensors will operate, and their power requirements.

1.3 Exemplar Sensor Node

As mentioned in Section 1.1, the application envisaged of this research is a generic class of wireless ground-level sensor networks deployed in an outdoor setting. Although some of these agricultural WSN nodes [17, 18] are often amenable to energy harvesting techniques based on photovoltaic (notably from sunlight) and kinetic energy based generators (such as those utilising wind), these techniques are not ubiquitously applicable. There is a significant class of potential applications where individual sensor node may be buried within a structure or soil, and not subject to daylight for all or appreciable lengths of time. Such duties may include WSNs for sub-soil irrigation management and detection of pest infestation.

In this research, we will focus on this particular type of application, where the sensor nodes are distributed across an outdoor field within a vicinity of domestic, agricultural or commercial buildings, as shown in Figure 1.2. A typical deployment consists of about 20 to 30 nodes. Each node is placed within a range up to 5 metres from the periphery of the buildings. All nodes are static at all time and buried at a depth of soil (hence the given name of ‘wireless ground-level sensor node’) with their transceiver slightly raised above the ground to enable wireless communication between the nodes and the base station. Due to a wide dynamic range of outdoor environments, these wireless ground-level sensor nodes could be located in an open field, under the shade of trees, or surrounded by growing vegetations, such as shrubs and crop canopies.
Figure 1.2: Deployment architecture of the wireless ground-level sensor nodes.

A wireless ground-level sensor node reported in the literature: Zyrox2 Bait Station [19] will be used as the exemplar WSN node for this feasibility study. The Zyrox2 Bait Station, as shown in Figure 1.3 (a) contains a sensor exploiting the change in optical transmission through a sheet of bait material to detect the presence of termite activity. The bait material is sandwiched between two light guides, with one of them angles a light source normal to the bait material and other directs any light passed through the bait material back to a detector. In the absence of termites, the bait material absorbs majority of the incident light, and the detector outputs low. When some fractions of the bait material are consumed by the termites, additional light passes through to the detector and a sensor hit is flagged [19]. A photograph of the prototype sensing unit is shown in Figure 1.3 (b).

A typical WSN node usually consumes more energy during communication. For instance, the Zyrox2 Bait Station consumes the most energy during its ‘data transmission’ mode, i.e. 25.92 J (24 mA at 3.0 V for 6 minutes). However, this is not always the case as the energy consumption of the sensing operation can sometimes be higher depending on the sensors being employed in the node. For example, the GAIA Soil-Mote [20], a wireless ground-level sensor node deployed to monitor the soil condition, consumes more energy when it operates in ‘sensor data acquisition’ mode than in ‘communication module transmitting/receiving’ mode.

The average power consumption of a WSN node can be calculated based on its duty cycle, i.e. the amount of time that a node spends in an active state to perform sensing, transmitting and receiving of data. In the worst case scenario (i.e. its highest possible duty cycle), the Zyrox2 Bait Station have an average power consumption of 450 µW. Further
details on its operational duty cycle and energy consumption at different functional states are provided in Chapter 3.

Figure 1.3: (a) Schematic of optical cell; (b) Photograph of prototype unit of Zyrox2 Bait Station [19].

1.4 Thesis Outline

This thesis is divided into five chapters, organized as follows. Chapter 2 provides the background on energy harvesting principles, supplemented by a broad review of several energy harvesting methods, which may be applicable for the intended wireless ground-level sensor network. Strengths and weaknesses of each method are evaluated, followed by a discussion on selecting RF energy harvesting as the main subject of this investigation. A further review on four different wireless energy transfer techniques, using magnetic resonant induction, radio/microwaves, laser and ultrasonic waves, are also presented.

Chapter 3 performs a theoretical-based analysis on the investigated system: an outdoor RF energy harvesting system. The analysis seeks to identify all relevant design parameters and quantifies their impact on the overall system performance. In addition, an RF link budget analysis is also included to examine the feasibility of using RF energy harvesting to power the exemplar WSN node - Zyrox2 Bait Station.
Chapter 4 presents the design and testing of the antennas and power harvester circuit which could be potentially used for the investigated system. Based on a range of system requirements, two air-substrate-based, folded, shorted patch antennas (FSPAs) are proposed. The performance of both antennas is simulated, and measurement results are presented for the fabricated antennas tested in an anechoic chamber, and in an outdoor field. A parametric simulation study is also conducted to investigate the effect of different antenna parameters on the performance of the FSPA with slotted ground configuration. In addition, a power harvester circuit is built using commercially available off-the-shelf components, and it is tested with an RF signal generator source in the laboratory. The measurement techniques used are also reviewed, along with some comments on how to improve them.

Chapter 5 concludes the work presented in the thesis and identifies key research areas for future investigations.
2 Literature Review

2.1 Introduction

This literature review contains two major parts. The first part begins with an overview of the energy harvesting principles, and is followed by a broad survey of several energy harvesting techniques, which may be applicable for agricultural WSNs. Four potential energy harvesting techniques, based on solar, thermal, wind and radio frequency (RF) energy sources, are studied. The techniques of interest are discussed in terms of their operating principles, as well as the reported systems and performance from the published literature. The second part of this chapter focuses on the investigated technique: RF energy harvesting. Various wireless energy transfer technologies using magnetic resonant induction, radio or microwaves, lasers and ultrasonic-waves transmission are reviewed. The rationale is then given for selecting far-field RF energy harvesting as the preferred method of this research.

2.2 Energy Harvesting Principles

Energy harvesting, also known as power harvesting or energy scavenging, is defined as the process of extracting energy from the surroundings of a system, and converting it into usable electrical energy [15]. Although ‘energy harvesting’ and ‘energy scavenging’ are often used in the literature, there exists a difference between these two terms. Energy scavenging is usually referred to environments where the ambient sources are unknown or highly irregular, while energy harvesting is more suitable for situations where the ambient energy sources are well characterised and regular [21]. Table 2.1 provides some examples of this difference.

<table>
<thead>
<tr>
<th>Energy Sources</th>
<th>Scavenging</th>
<th>Harvesting</th>
</tr>
</thead>
<tbody>
<tr>
<td>Photonic</td>
<td>Random lightings</td>
<td>Daily solar cycles</td>
</tr>
<tr>
<td>Thermal</td>
<td>Forest fires</td>
<td>Furnace covers</td>
</tr>
<tr>
<td>Kinetic flow</td>
<td>Winds</td>
<td>Air-conditioning ducts</td>
</tr>
<tr>
<td>Electromagnetic</td>
<td>GSM stations/ WLANs</td>
<td>Dedicated transmitters</td>
</tr>
</tbody>
</table>

Table 2.1: Comparison of energy scavenging and energy harvesting [21].
In general, an energy harvesting system consists of the following components:

- An energy source;
- A transducer - converting ambient energy into electrical energy;
- An energy storage component - storing the harvested energy and powering the sensor system; typically a rechargeable battery or a capacitor;
- Power electronics - conditioning the harvested electrical energy, maximising the amount of stored energy, and managing the power flow to the load system for which it is connected.

Several energy harvesting techniques using natural or man-made energy sources have been reported in the literature [14, 16, 22]. These sources can be classified into one of the four categories: photonic, thermal, kinetic and electromagnetic energy. Photonic energy sources include solar radiation and artificial indoor lighting. Thermal energy sources include the objects heated by sun, ambient temperature, exhaust pipes or automobile engines. Kinetic energy sources include flowing fluids, wind or vibrational motions caused by moving structures or humans. Electromagnetic energy exists in the form of alternating magnetic fields surrounding AC power lines, or radio waves emitted by nearby transmitters.

Figure 2.1 shows various ambient energy sources with their respective energy conversion devices. As seen from the Figure, the source energy is first converted into electricity through a transducer element, e.g. photovoltaic cell, thermoelectric element, piezoelectric transducer, antenna, etc. The electrical energy is then conditioned and used to charge a battery, which stores and supplies the energy to a load, i.e. a WSN node. Each energy source has its own unique characteristics in terms of controllability, predictability and magnitude; hence all these factors will need to be considered when choosing the most suitable source for a specific application.
2.3 Potential Energy Harvesting Techniques

The purpose of this section is to provide a broad review on several energy harvesting techniques which maybe applicable for the intended agricultural WSN application. It should be noted that, this section does not intend to provide an exhaustive literature survey on all existing techniques. A significant amount of literature has been published on energy harvesting in recent years, and they have been discussed extensively by various authors in several useful review articles [15, 16, 23-25]. Therefore, in an effort to limit the scope of this literature survey, it will only focus on the energy sources which are more likely to be found in the actual operating environment of the targeted WSN nodes: an outdoor field.

Four potential energy harvesting techniques based on solar, thermal, wind and radio frequency (RF) energy sources are identified, and they will be discussed in the following sections. Another widely reported technique: piezoelectric energy harvesting based on mechanical vibrations or human motions was not considered in this research due to the static nature of the wireless ground-level sensor nodes.
2.3.1 Solar Energy Harvesting

Operating Principles

Solar energy can be converted into electrical energy using a solar cell. The solar cell is basically a semiconductor diode consisting of a large-area p-n junction. When the cell is illuminated with light (photons) having energy greater than the bandgap energy of the semiconductor, electron-hole pairs are generated due to the absorption of photons. These electron-hole pairs are separated by a built-in electric field created by the cell junction, with the electrons swept towards the n-side and holes towards the p-side. The electrons and holes are then collected by the contacts of each side of the cell, thus forming an electrical potential. If an external load is connected to the cell, the current will flow and thereby generating electrical power. This process is known as the photovoltaic effect. Figure 2.2 shows the current versus voltage (I-V) characteristic of a typical solar cell, with (illuminated) and without (dark) incident radiation.

![I-V characteristics of a typical solar cell](image)

As seen from the Figure, the electrical performance of a solar cell can be characterised by two parameters: short circuit current, \( I_{sc} \) (at zero voltage) and open circuit voltage, \( V_{oc} \) (at zero current). For a given level of light intensity, an optimal operating point exists on the I-V curve (somewhere between the two shaded regions in Figure 2.2), which the cell can produce maximum output power, \( P_{mpp} \) [26]:

\[
P_{mpp} = I_{mpp} V_{mpp} = F F I_{sc} V_{oc}
\]

(2.1)

where \( I_{mpp} \) and \( V_{mpp} \) are the cell current and voltage at maximum power point, respectively. \( FF \) is the fill factor, which is a measure of the cell quality ranging from 0 (poor) to 1.
Another important property is the solar conversion efficiency $\eta_{\text{solar}}$, which is given by the ratio of the generated electrical power to the incident power on the solar cell, $P_{\text{solar}}$ [26]:

$$
\eta_{\text{solar}} = \frac{V_{\text{mpp}} I_{\text{mpp}}}{P_{\text{solar}}} = \frac{FF I_{\text{sc}}}{P_{\text{solar}}} \quad (2.2)
$$

The efficiency of the commercial solar cells ranges from a low of approximately 8% to state-of-art values of 20%, with some experimental technologies reaching as high as 35% [16]. The most common materials used for the cells are crystalline silicon, amorphous silicon, cadmium telluride, copper indium diselenide and gallium arsenide [27]. Crystalline silicon is normally used for outdoor applications since it has high conversion efficiencies (15-20%) under high light conditions. On the other hand, thin film materials, such as amorphous silicon and cadmium telluride, are more appropriate for indoor applications due to the closeness of their spectral response to that of artificial light. A major drawback of the thin films is the low efficiency (10%). Table 2.2 shows the specifications for a number of silicon cells which are available in the market. Note that most of the commercial cells have an open circuit voltage of about 0.6 V. These cells can be placed in series to obtain higher voltages, or placed in parallel to increase the currents.

<table>
<thead>
<tr>
<th>Cell</th>
<th>Dimensions (cm$^2$)</th>
<th>Thickness (µm)</th>
<th>Voc (V)</th>
<th>Isc (A)</th>
<th>FF</th>
<th>Efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Schott® EFG 1030</td>
<td>100</td>
<td>330</td>
<td>0.595</td>
<td>3.15</td>
<td>0.77</td>
<td>14.5</td>
</tr>
<tr>
<td>Photowatt® Af</td>
<td>102</td>
<td>300 ± 50</td>
<td>0.606</td>
<td>3.57</td>
<td>0.73</td>
<td>15.4</td>
</tr>
<tr>
<td>SunPower ®A-300</td>
<td>156</td>
<td>270 ± 40</td>
<td>0.670</td>
<td>5.90</td>
<td>0.78</td>
<td>21.5</td>
</tr>
<tr>
<td>SunPower®Pegasus</td>
<td>21.9</td>
<td>160</td>
<td>0.680</td>
<td>0.88</td>
<td>0.82</td>
<td>22.5</td>
</tr>
</tbody>
</table>

Table 2.2: Specifications for several commercially available solar cells with 1000 W/m$^2$ incident radiation [16].

**Reported Systems and Performance**

Compared to well-developed macro solar-powered systems (i.e. electricity generation for residential houses and business buildings), a micro solar-powered WSN system is more constrained in terms of its energy budget and consumption. A number of research groups from different institutions have investigated the feasibility of using solar energy to power outdoor WSNs, with some of them reaching the point of offering plug-and-play solar energy harvesting modules [25].
Raghunathan et al. [28] developed Heliomote, a solar-powered WSN system built using the Mica2 platform [29]. Their design, in favour of simplicity, employed a single-stage energy storage consisting of NiMH batteries, and a hardware-controlled battery charging circuit. An energy monitoring component was included to measure the instantaneous current provided by solar panels, the battery terminal voltage and the accumulated current over a specified time interval. This information enabled the sensor node to learn its energy environment, and to perform harvesting aware power management adaptation. Since the batteries were directly connected to the solar panel through a diode, they experienced recharge cycles daily. This limits the system’s lifetime to no more than two years, due to a finite number of recharge cycles of the batteries.

Jiang et al. [30] presented Prometheus, built using the TelosB platform [31]. Compared to the architecture of Heliomote, Prometheus used an additional stage of energy storage and a software-based charging control mechanism. A supercapacitor, having nearly infinite recharge cycles, was used as the primary buffer to power the sensor node, and to charge the secondary buffer when excess energy is available. A Li-ion rechargeable battery, acted as the secondary buffer, was only used when the primary buffer energy falling below a certain threshold. By using the supercapacitor as a primary energy source, the access to the battery could be minimised, thereby prolonging the system lifetime. A tradeoff of using a tiered energy storage mechanism (i.e. using a supercapacitor and a battery) is the reduction of harvesting circuit efficiency due to the increased overhead of energy storage management. Jiang and his colleagues predicted that the Prometheus system can operate for 43 years with a duty cycle of 1% (5 hours of light per month). However, their claims were rather too optimistic, since the hardware components of the system, such as solar panels and energy storage components, may degrade over long period of time.

Simjee and Chou [32, 33] from UC Irvine designed a supercapacitor-operated, solar-powered WSN node called Everlast. Unlike Heliomote and Prometheus, Everlast did not use batteries but only a supercapacitor to store energy. To maximise the harvesting efficiency, Everlast employed maximum power point tracking (MPPT). A feed forward, pulse frequency modulated (PFM) regulator was proposed to charge the supercapacitor at near optimal operating points of the solar cells. It was reported that Everlast could achieve higher harvesting efficiency than Heliomote and Prometheus. However, the MPPT algorithm used in the Everlast system would require a microcontroller to stay active all the time, which potentially consumes more power.
Table 2.3 compares the solar energy harvesting systems discussed above, in terms of their solar panel power rating, storage type and storage capacity. As seen from Table 2.3, the amount of solar power harvested by these systems is several hundred milliwatts. This level of power would be more than enough to run most WSN applications. Although the comparison does not show any particular system as the best, it clearly demonstrates the potential of solar energy harvesting in powering outdoor WSN nodes. It should also be noted that these systems were designed and tested under ample of light exposure, i.e. direct sunlight. The researchers in [28, 30, 33] did not evaluate their systems under other possible environmental conditions, such as in shade, as well as seasonal and daily variation of solar radiation.

<table>
<thead>
<tr>
<th>Nodes</th>
<th>Solar Power (mW)</th>
<th>Storage Type</th>
<th>Storage Capacity</th>
<th>Sensor Node</th>
<th>MPP Tracking</th>
</tr>
</thead>
<tbody>
<tr>
<td>Heliomote [28]</td>
<td>210</td>
<td>NiMH</td>
<td>1800 mAh</td>
<td>Mica2</td>
<td>No</td>
</tr>
<tr>
<td>Prometheus [30]</td>
<td>190</td>
<td>Two Supercapacitors and Li-ion</td>
<td>22F 200 mAh</td>
<td>Telos</td>
<td>No</td>
</tr>
<tr>
<td>Everlast [33]</td>
<td>450</td>
<td>Supercapacitor</td>
<td>100 F</td>
<td>Integrated</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Table 2.3: Comparison of the solar energy harvesting systems.

Discussion

Solar energy is one of the most abundant and readily accessible energy sources in a typical outdoor environment. On a bright sunny day, approximately 1000 W/m² of solar power impinges on the Earth’s surface [26]. For a solar cell with 15% conversion efficiency, this would produce an output power density of 15 mW/cm² [34]. Compared to other techniques, solar energy harvesting can generate the highest output power density, making it an ideal choice to power the WSNs deployed outdoors.

However, the output power of a solar cell is highly dependent on the environmental conditions, i.e. varying light intensity. The first and most obvious is that, a solar cell could only operate during daylight hours. As a result, a secondary storage component (e.g. rechargeable battery) would be required to store the energy, so that it can be used to operate the sensor nodes at night. Other environmental factors, such as shadowing and cloud cover, could also reduce the level of light intensity upon the cells significantly. Sonnenenergie [35] reported that a core shadow cast by nearby object (e.g. trees or buildings) reduces the incident solar energy on the cells by approximately 60 to 80 percent, while a partial shade could cause a reduction up to 50 percent. On a cloudy day, the
output power density of a solar cell (with 15% of conversion efficiency) is expected to drop to 150 μW/cm² [34].

The incident power on the solar cells also decreases with the cosine of the angle of incidence of the Sun’s rays [25]. To maximise the output power, the angle of incidence must remain to zero at all times, i.e. the rays are perpendicular to the cell surface. Solar tracking are commonly used in the macro solar-powered systems to make sure the cells are always facing towards the Sun. Thomas et al. [16] carried out an analysis to examine the effects of solar tracking on the cell’s performance using four different configurations: horizontal flat (parallel to ground), fixed tilt on some optimum angle for the given location, one-axis tracking and two axes tracking. As expected, the tracking yields better energy harvesting performance, in which the one-axis and two-axes tracking yielded 50% and 54% more energy than the horizontal flat configuration. However, this comes with the expense of added weight, complexity, and cost of tracking control equipment. Therefore, an analysis must be carried out to evaluate the trade-offs between the desired performance and overall cost before adding a solar tracking mechanism to a micro solar-powered system.

2.3.2 Thermoelectric Energy Harvesting

Operating Principles

Thermoelectric energy harvesting capitalises on the presence of naturally occurring temperature variations in the environment. A thermoelectric generator (TEG) can be used to convert the thermal gradients into electrical energy based on the Seebeck effect. A basic structure of the semiconductor-based thermocouple is depicted in Figure 2.3. The thermocouple comprises of p-type and n-type semiconductor elements placed between a hot and a cold junction. When the two junctions are held at different temperatures, an open-circuit voltage (known as Seebeck voltage) is developed.
To obtain a practical output voltage and power at low temperature difference, numerous thermocouples can be arranged electrically in series and thermally in parallel, forming a TEG device. The Seebeck voltage, $V_G$ of a TEG with $N$ number of thermocouples can be expressed as [36]:

$$V_G = \Delta V_{pn} = N\alpha_{pn} (\Delta T_{TEG}) = N\alpha_{pn} (T_{HJ} - T_{CJ})$$

(2.3)

where $\Delta T_{TEG}$ is the temperature difference across the hot and cold junctions, and $T_{HJ} - T_{CJ} = \beta (T_H - T_C)$ with $\beta$ being the coefficient taking into account of the thermal losses of the ceramic plates. Meanwhile, $\alpha_{pn} = \alpha_p - \alpha_n$ is the Seebeck coefficient between the p-type and n-type semiconductor materials. When the thermocouples are connected to a load with resistance $R_L$, a current will flow. This current, $I_L$ is given by [36]:

$$I_L = \frac{V_G}{R_{in} + R_L} = \frac{N\alpha_{pn} (T_{HJ} - T_{CJ})}{R_{in} + R_L}$$

(2.4)

$$R_{in} = \frac{N2\rho h}{A_{leg}}$$

(2.5)

where $R_{in}$ is the internal electrical resistance of the thermocouples, $\rho$ is the electrical resistivity of the thermocouple material, with $h$ and $A_{leg}$ being the height and the area of a
single thermocouple leg, respectively. The output power of a TEG is simply the product of voltage and current across the load [36]:

\[
P_{\text{TEG}} = I_L V_L = N^2 \alpha_{pn}^2 \Delta T_{\text{TEG}}^2 \frac{R_L}{(R_{in} + R_L)^2}
\]  

(2.6)

As shown in Equation (2.6), a higher number of thermocouples, larger temperature difference between the hot and cold junctions, together with higher absolute Seebeck coefficient, and lower internal resistance are desirable for a TEG to generate higher output power. The TEG can generate maximum output power on match load conditions, i.e. when the load resistance \( R_L \) is equal to its internal electrical resistance \( R_{in} \).

For thermoelectric generation, the following efficiency factor is normally used [37]:

\[
\eta_{\text{TE}} = \frac{\eta_{\text{carnot}}}{\sqrt{1 + ZT_m} - 1}
\]

(2.7)

The Carnot efficiency, \( \eta_{\text{carnot}} \) is the maximum theoretical efficiency of a heat engine, while \( ZT_m \) is a dimensionless figure of merit quantity for the thermocouples used in the TEG [37]:

\[
ZT_m = \frac{\sigma}{\kappa} \left( \frac{\alpha_p - \alpha_n}{2} \right) \left( \frac{T_H + T_C}{2} \right)
\]

(2.8)

where \( \alpha_p \) and \( \alpha_n \) are the absolute Seebeck coefficients, \( \sigma \) is the electrical conductivity, and \( \kappa \) is the thermal conductivity. Equation (2.8) shows that higher thermoelectric conversion efficiency can be obtained by using materials with larger values of \( ZT_m \), i.e. higher electrical conductivity (to increase the output current), higher Seebeck coefficients (to produce higher output voltage), and lower thermal conductivity (to retain the heat at the junctions) [38]. Most commonly used thermoelectric materials, such as bismuth-telluride and antimony-telluride, have a \( ZT_m \) value of about 1 when operating close to their maximum temperature limit [16]. For more information on the recent advancements in thermoelectric material research, Hudak and Amatucci [39], and Zheng [40] provide good summaries on the topic.

**Reported Systems and Performance**

Macro-scale TEG systems, using waste heat from the industrial processes and exhaust gases of automobiles, have been developed for several decades. These systems [41-43]
can generate large amount of power up to kilowatt scale at very high temperatures of several hundreds of °C. On the other hand, a micro-scale TEG system generates much less power in the order of milliwatts or lower from ambient heat sources with lower temperatures. In this section, a few micro-scale thermoelectric energy harvesting systems deployed outdoors are examined.

As pointed out in Section 2.3.1, solar energy is abundant in outdoor environments. Instead of using conventional photovoltaic conversion, one could employ TEGs to convert solar energy indirectly into electricity through thermoelectric energy harvesting. Sodano et al. [44] investigated the use of TEGs to recharge a NiMH battery from solar radiation. The TEGs consisted of eight commercially available thermoelectric modules, each with a dimension of 30 x 34 x 3.2 mm. The overall system comprised of the TEGs, the NiMH battery, and a diode. The diode was used to force the current flowing only in one direction, and to prevent the TEGs from drawing power off the battery at lower temperature difference, i.e. when the generated voltage was insufficient to charge the battery. A schematic showing the system layout is depicted in Figure 2.4. A greenhouse was created where a Plexiglas window was placed above a heat sink to trap the thermal energy, allowing it to accumulate and to increase the thermal gradient over the TEGs. To further increase the amount of thermal energy stored, a solar concentrator was used to focus the sunlight onto the hot side of the heat sink. The heat sink was painted in black so that it did not reflect much of the visible light but rather absorb the Sun’s ray. The system was built and experimentally tested under direct sunlight condition. It was capable of recharging an 80 mAh and a 300 mAh NiMH battery in 3.3 minutes and 17.3 minutes, respectively. The researchers, however, did not provide clear information on the actual temperature gradient between the hot and cold sides of the TEGs when performing their measurements.

![Figure 2.4: Schematic of thermal energy harvesting device using solar radiation [44].](image-url)
Another potential ambient heat source is the soil and air boundary. Lawrence and Snyder [45] proposed a TEG system exploiting the natural temperature difference between air and soil to generate small amount of electrical power. Figure 2.5 illustrates the proposed concept where two heat exchangers were used - one exposed to the air at ground level and another to the soil. The temperature difference between the two heat exchangers was applied to a TEG. During day time, the air heat exchanger heated up and a 30-cm long water-based heat pipe was used to carry the heat across the TEG to the underground exchanger for heat dissipation, as shown in Figure 2.5. An advantage to this system is that, even at night when the air cools down, the generator can still generate power since the temperature profile is reversed. A prototype was built without the TEG, and the heat flow across a brass dummy element was measured to estimate the amount of harvestable power. Lawrence and his colleague predicted that, if a prototype were equipped with a TEG, daily temperature variations would cause its output power to fluctuate between 0 to 100 µW, peaking up to 350 µW in the early afternoon [45]. The average output power was estimated at 50 µW. A major drawback of their system is the low conversion efficiency, mainly due to the small temperature gradients (less than 1 K) across the TEG. According to the researchers, most of the temperature drops in the system were across the heat pipe. Hence, they suggested the use of a different heat pipe design or the complete removal of heat pipe to optimise the heat flow across the TEG.

Figure 2.5: Thermoelectric harvesting device proposed in [45]. ‘HX’ denotes ‘Heat Exchanger’.
Using the same concept proposed in [45], Meydbray et al. [46] experimentally investigated the effects of TEG module’s surface area. The TEG used in their experiment was a commercially available Marlow DT12-2.5, and it was thermally coupled to a square copper plate, with another side attached to a ceramic plate. The copper plate was buried within the soil so that it can reach the same temperature as the ground. However, the actual depth of the copper plate placement was not stated. The experiment was performed outdoors in a field away from all objects which could cast a shadow on the module. Three experimental setups with different surface areas for TEG modules were evaluated for 110 hours under various weather conditions: clear, overcast, cloudy and misty. The maximum temperature drop across the TEG module was recorded at ± 4ºC. The experimental results are shown in Table 2.4. The results indicate a strong dependence of output power on the surface area. Besides, the experiment data reported in [46] also shows a significant drop (by about 75%) in the output power during times of overcast.

<table>
<thead>
<tr>
<th>Setup</th>
<th>Total Harvested Energy (J)</th>
<th>Average Power (mW)</th>
<th>Surface Area (cm²)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>9.03</td>
<td>0.0228</td>
<td>9</td>
</tr>
<tr>
<td>2</td>
<td>28.12</td>
<td>0.071</td>
<td>33</td>
</tr>
<tr>
<td>3</td>
<td>227.70</td>
<td>0.575</td>
<td>131</td>
</tr>
</tbody>
</table>

Table 2.4: Output power from TEG modules using air-soil temperature gradient [46].

**Discussion**

A TEG is a robust and highly reliable energy converter. Similar to solar cells, its output is a DC signal. Therefore, it does not require any form of rectifying, which is usually the main energy loss in an RF or a wind energy harvesting system. However, the relatively low thermoelectric conversion efficiency - typically less than 5% for most of the commercial TEGs when operating at their maximum temperature limit [16, 47] - has limited its use to low power application. As seen from Equation (2.7), the thermoelectric conversion efficiency of a TEG is constrained by the fundamental Carnot limit, which is given by [16]:

\[
\eta_{\text{carnot}} = \frac{T_H - T_C}{T_H}
\]

where \( T_H \) and \( T_C \) are the high and low temperatures of the thermodynamic cycle, respectively. Figure 2.6 shows the plot of Carnot efficiency for a heat engine, and the thermoelectric conversion efficiency for a thermoelectric material having \( ZT_m \sim 1 \), as a
function of temperature with one of the thermal reservoirs held constant at room temperature (25ºC).

As seen from the Figure, the Carnot efficiency is very low for small to modest temperature differences. For instance, when a heat source is 10 ºC above room temperature, the Carnot efficiency would be only 3.2%. Even if the heat source is increased to 100 ºC, this efficiency would still be low at 20%. Note that the Carnot efficiency limit is the maximum theoretically possible efficiency. For real world conversion devices, their thermoelectric conversion efficiency would be much lower than the Carnot efficiency. Most of the current commercial TEGs operate below 40% of the Carnot efficiency [25]. Because of the low conversion efficiency, a large and sustained thermal gradient with huge amount of heat flows across the two surfaces of the TEG are necessary to generate practical voltages and powers. In addition, it was also observed that the outdoor thermoelectric energy harvesting systems reported in [44-46] are heavily relied on solar radiation. In order to obtain a large thermal gradient, the hot surface of the TEGs must be exposed to direct sunlight.
2.3.3 Wind Energy Harvesting

Operating Principles

Wind energy harvesting can be realised using a wind turbine. The turbine converts the wind flow into a shaft rotation using a rotor consisted of one or more airfoil blades. The shaft is attached to a generator which contains strong magnets and coils inside. As the magnets rotate around the coils, it creates changing magnetic fields, thereby inducing an electrical current. The available theoretical power available from wind flow is equal to the rate of flow of kinetic energy per second [16]:

\[
P_w = \frac{1}{2} \frac{dm}{dt} V^2 = \frac{\rho AV^3}{2}
\]

(2.10)

where \( \rho \) is the air density in kg/m\(^3\), \( A \) is the cross sectional area of flow in m\(^2\) (i.e. the area swept by the rotors) and \( V \) is the upstream wind velocity in m/s. A wind turbine cannot extract all of this power. When the wind stream passes the turbine, part of its kinetic energy is transferred to the rotor, and the air leaving the turbine carries the rest away. The power which can be extracted by a practical wind turbine is given by [16, 17]:

\[
P_m = C_p \frac{\rho AV^3}{2} = C_p P_w
\]

(2.11)

\[
C_p = \left[ \frac{1 + \frac{V_{down}}{V}}{2} - \left( 1 - \left( \frac{V_{down}}{V} \right)^2 \right) \right]
\]

(2.12)

where \( V \) is the upstream wind velocity at the entrance of the rotor blades, and \( V_{down} \) is the downstream wind velocity at the exit of the rotor blades. \( C_p \) is the rotor power coefficient governed by the Betz’s limit [17] (analogous to Carnot’s efficiency), and it has a theoretical maximum value of 0.593 when \( V_{down}/V \) is equal to one-third.

Large scale wind turbines can be highly efficient, with \( C_p \) greater than 0.5 being achievable, while smaller scale wind turbines are expected to have lower \( C_p \), about 0.1 [48]. Assuming the wind turbine is coupled to a transmission with efficiency \( \eta_m \) driving a generator having efficiency \( \eta_g \), then the generated electrical power will be [17]:

\[
P_e = \eta_m \eta_g C_p P_w
\]

(2.13)
In general, the wind turbines can be divided into two primary types based on the axis of rotation: horizontal and vertical. Horizontal axis wind turbines (HAWTs) have the rotor shaft parallel to ground and they must be oriented into the direction of wind flow to operate efficiently. HAWTs are normally used for large-scale windmills. On the other hand, vertical axis wind turbines (VAWTs) have the rotor shaft arranged vertically. The most common VAWTs include Darrieus designs which rotate the rotor blades based on aerodynamic lift forces, and Savonius designs that use drag to produce rotation [16]. One of the main advantages of VAWTs is that they can operate for all horizontal wind flow direction, allowing them to be used in the areas with high variable wind direction. In addition, the VAWTs’ generator and gearbox can be located close to the ground, making them suitable for the design of small-scale wind energy harvesters.

Reported Systems and Performance

While most studies on wind energy harvesting focused on macro-scale generation, only a few work have demonstrated the small-scale airflow harvesters for outdoor WSNs. Weimer et al. [49] proposed a wind energy harvesting system based on anemometer solution. Their system utilised the motion of the anemometer shaft to drive a compact axial flux alternator to generate electrical power for the sensor nodes. The harvested energy was rectified via a full wave rectifier built using Schottky diodes, and then being converted to a battery potential via a buck-boost converter. It was found that by maintaining a constant input resistance at the input port of the converter, the alternator could be biased to operate at its peak power point over a wide range of wind speeds. As expected, the output power of their proposed system increased nonlinearly with wind speed, due to the increased energy availability at the anemometer and the reduced impact of the power loss in the converter control circuitry [49]. The available output power was 650 µW at high wind speed (8 m/s), whereas 5 to 80 µW was obtained at lower wind speeds (2 to 3.5 m/s). Nonetheless, the system’s actual size was not reported in this paper. Furthermore, no clear information was given on the anemometer design, i.e. number of anemometer cups used and their respective diameters.

Carli et al. [50] developed a small-scale wind energy harvester based on a four-bladed, horizontal-axis wind turbine with a diameter of 6 cm. To characterise their wind generator, the researchers performed the measurements at the rectifier’s output port using different load resistances at three selected air flow speeds, shown in Table 2.5. The data shows that, for a fixed air flow speed, there is an optimum load value in which the wind generator could generate maximum output power. By using the same resistor emulation approach
proposed in [49], Carli and his colleagues developed a buck-boost converter-based maximum power point (MPP) circuit. The converter was designed such that a constant input resistance was maintained at its input port, close to the optimum load values $R_{L,\text{opt}}$ on a wide range of operating conditions. In addition, an effective power-saving scheme was also proposed to control the harvesting circuit, via an ultra low-power comparator, which turns off the energy harvester during wind calms. Other relevant work on small-scale airflow harvesters based on the wind turbine principle can be found in [51, 52].

<table>
<thead>
<tr>
<th>Air Flow Speed (m/s)</th>
<th>Optimum Load, $R_{L,\text{opt}}$ (Ω)</th>
<th>Voltage (V)</th>
<th>Maximum Power (mW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.4</td>
<td>715</td>
<td>2.40</td>
<td>2.02</td>
</tr>
<tr>
<td>4.2</td>
<td>559</td>
<td>4.21</td>
<td>7.93</td>
</tr>
<tr>
<td>4.7</td>
<td>549</td>
<td>4.68</td>
<td>9.95</td>
</tr>
</tbody>
</table>

Table 2.5: Characterisation data of the wind generator in [50].

Relative performances of several small-scale wind harvesters reported in the literature are compared in Figure 2.7. The amount of power generated from these harvesters depends on the airflow speed, total swept area, as well as the efficiency of the generator and the power converter used. In addition to developing more efficient centimeter-scale wind turbines, ongoing work must also address the low efficiency issue of the harvesters when operating at much lower flow rates [48], i.e. less than 1 m/s.

![Figure 2.7: Expected output power density as a function of flow speed for small airflow harvester in [50-52].](image-url)
Discussion

Similar to solar energy, wind energy is abundant and easily accessible in outdoor environment. This makes it an attractive solution to power WSN nodes deployed in locations which are subjected to constant wind, i.e. an open field or a large farm area. However, since the wind is a variable energy source (i.e. wind speed and direction change from year to year, season to season, and with time of day), a practical wind energy harvesting system would require the information of the average wind speed in the area where the nodes are located. It should be noted that most of the wind maps are usually quoted for altitude of 10 metres or above. Therefore, these data cannot be applied for wireless ground-level sensor nodes. Additional measurement must be performed to collect more data on the average wind speed which is available at a specific height of the intended application. Furthermore, the wind speed also varies with height of ground. At lower ground, it could be significantly lower due to the surface roughness of the surrounding area of nearby obstacles, such as trees or buildings, and by the contours of the local terrain [53]. The average wind speed $V_h$, at height $h$ above a ‘rough’ surface can be estimated by the following equation [16, 26]:

$$V_h = V_{ref} \left( \frac{h}{h_{ref}} \right)^{\alpha}$$  \hspace{1cm} (2.14)

where $V_{ref}$ is the wind speed at height $h_{ref}$, and the exponent $\alpha$ is the friction coefficient of different terrains. Table 2.6 shows the empirical values of $\alpha$ for various terrains. As expected, the friction coefficient is generally higher in small town and city area with tall buildings, compared to those in countryside. The wind speed is nominally zero at ground level.

<table>
<thead>
<tr>
<th>Terrain Type</th>
<th>Friction Coefficient, $\alpha$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lake, ocean, and smooth, hard ground</td>
<td>0.10</td>
</tr>
<tr>
<td>Foot-high grass on level ground</td>
<td>0.15</td>
</tr>
<tr>
<td>Tall crops, hedges, and shrubs</td>
<td>0.20</td>
</tr>
<tr>
<td>Wooded country with many trees</td>
<td>0.25</td>
</tr>
<tr>
<td>Small town with some trees and shrubs</td>
<td>0.30</td>
</tr>
<tr>
<td>City area with tall buildings</td>
<td>0.40</td>
</tr>
</tbody>
</table>

Table 2.6: Friction Coefficient $\alpha$ of Various Terrains [26].
2.3.4 Radio Frequency (RF) Energy Harvesting

Operating Principles

Another possible alternative to power outdoor WSN nodes is by RF energy harvesting. In cities or very populated areas, the background RF radiation, emitted by broadcast transmitters, cell phone towers or wireless local area networks, could serve as a potential energy harvesting source. In this context, ‘radio frequency’ refers to electromagnetic waves with frequencies in the range from 3 kHz to 300 GHz. RF energy harvesting converts RF waves into DC power using a rectifying antenna, which is a combination of antenna and rectifying circuit. The amount of harvestable power depends on the incident power density of the location, efficiency of power conversion, and size constraints of the energy collection device, i.e. antenna or induction coil.

RF radiation is normally quantified in terms of electric field strength, which is in V/m. In the far-field region, the electric field strength can be converted into incident power density using the following equation:

\[
S = \frac{E^2}{Z_0}
\]  

(2.15)

where \( S \) is the incident power density (in W/m²) and \( Z_0 \) is the characteristic impedance of free space (approximated by 377 ohms). The incident power density is also a function of distance between the source and the receivers. The possible power to be harvested in the far-field region can be estimated using the Friss’ Transmission Equation, given by:

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

(2.16)

where \( P_r \) is the received power (watts), \( P_t \) is the transmitted power (watts), \( G_t \) and \( G_r \) are the gain of the transmitting and receiving antenna (with respect to an isotropic antenna), \( \lambda \) is the wavelength of the operating frequency (metres), and \( d \) is the distance between the source and the harvester (metres). Equation (2.16) assumes there is an unobstructed line-of-sight (LOS) path between the transmitting and receiving antennas.
Reported Systems and Performance

In RF energy harvesting, two different approaches can be employed. The first approach is by exploiting existing electromagnetic radiation in the surroundings, such as Global System for Mobile Communications (GSM) or Wireless Local Area Network (WLAN) signals. When harvesting energy from these signals, one has to deal with very low power density levels. Alternatively, another approach is by deliberately broadcasting a signal at a specific frequency using a dedicated transmitter. Radio frequency identification (RFID) is one of the examples of the second approach where passive tags are wirelessly powered by a reader from a distance of several metres. However, it is debatable as to whether this approach can still be considered as energy harvesting as the power is directly supplied by the user. This section primarily focuses on the first approach: RF energy harvesting using the ambient sources.

Due to health concerns with human exposure to electromagnetic radiation, safety guidelines have been proposed by the International Commission on Non-ionising Radiation Protection (ICNIRP) [54] and IEEE [55] on the maximum permissible exposure (MPE) for the general public. This has led to a number of studies on the incident power densities at various locations, which can be used as a benchmark for RF energy harvesting using ambient sources. Visser et al. [56] assessed the power density levels produced from the GSM base stations by studying the measurement data gathered in several European countries [57, 58]. Their data was based on single frequency spot-measurements at 935–960 MHz and 1805–1880 MHz, which are the downlink frequency band for GSM-900 and GSM-1800 base stations, respectively. The measurements were performed in multiple locations (e.g. city, industrial area, small town, rural area, and countryside) with different site characteristics (e.g. indoor, outdoor, ground level, and on roof or balcony). At the distances between 25 to 100 metres from a GSM base station, a peak power density level ranging from 0.01 to 1 µW/cm^2 were measured. Another study was carried out by Burch et al. [59] to quantify the RF exposure levels in a residential area near Denver, Colorado. The investigated area contained 15 radio and television broadcast transmitters, operating at frequencies from 55 to 687 MHz. Measurements of RF power densities were performed inside and outside of 161 houses located within 3 km of the antenna cluster. It was found that the average indoor and outdoor RF power densities were 0.8 and 2.6 µW/cm^2. These findings [56, 59] suggested that the GSM base stations or broadcast transmitters are unlikely to produce sufficient ambient RF energy to wirelessly power a sensor node, unless an antenna with large effective area is used for harvesting.
Sample and Smith [60] from Intel Research attempted to harvest the RF power from a TV transmission tower, broadcasting 960kW ERP at 674-680 MHz, to power a commercially available thermometer/hygrometer. Their system, as depicted in Figure 2.8, comprised of a broadband log periodic antenna (with 5 dBi) and a power harvesting circuit using a four-stage charge pump, placed outside a balcony at a distance of 4.1 kilometres from the TV tower. With the antenna manually pointed towards the tower, a maximum power of 60 µW (0.7V across an 8K Ohm load) was measured. As expected, the measurements were somehow lower than the theoretical predictions using Friss’ Transmission Equation. Despite a low level of harvested power, it is still sufficient to run the thermometer/hygrometer device which only consumes an average power of 37.5 µW. When the antenna was oriented directly towards the TV tower, the device functioned normally. However, it was found that the power dropped drastically as the antenna was directed away from the tower. This could be attributed to the high directivity characteristic of the log periodic antenna used in the system. Although the researchers succeeded in capturing certain amount of energy, they have yet to demonstrate the feasibility of using ambient RF energy harvesting for outdoor WSN applications.

Discussion

While a small number of systems have demonstrated their capability of harvesting energy from ambient RF sources, the amount of captured energy is still low. Given the typical ambient RF power densities and the size limitation on the energy harvesting antenna used for WSN applications, the harvested power levels would be in the range of a few
microwatts, which is insufficient to run most of the sensor nodes. This suggests that the RF energy harvesting using ambient sources is an approach with very limited applicability. The use of ambient RF power for WSN applications requires either very close proximity to the radiating sources, or the use of a large collection area (antenna) which is much larger than the actual size of the sensor node itself [39]. Another drawback is that the ambient RF energy sources are random and uncontrollable. A number of studies [56, 59] showed that the power density levels produced from these ambient sources could vary depending on the locations, the proximity between the antennas, and the transmit power of the nearest base station. In addition, these power levels could also change with time and traffic density [61, 62], e.g. the peak hours and off-peak hours of GSM base stations.

On the contrary, intentional sources such as dedicated RF transmitters can provide a more predictable energy compared to ambient sources. Based on the requirements of the intended application, the amount of power and transmission time (i.e. duty cycle) can be manually controlled for the transmitter to operate either in continuous mode or on a scheduled basis. A network with multiple transmitters can also be used to improve the coverage over a wider area. However, strict regulations have been imposed by the governing bodies around the world, such as European Telecommunications Standards Institute (ETSI) and Federal Communications Commission (FCC), on the permissible power level radiated by dedicated transmitters. This restriction, in turns limits the amount of harvestable RF energy from the intentional sources.

### 2.3.5 Selection of Energy Harvesting Method

Energy harvesting appears to be an attractive alternative as it serves as a long term solution to replace the conventional primary batteries in powering the WSN nodes. However, each deployment environment would have different types of ambient energy sources. Therefore, there will be no ‘one size fits all’ solution for all WSN applications. The output power of an energy harvesting system depends on the availability of the ambient energy sources, the size/ weight of the energy collection elements, the efficiency of the collection, as well as the conversion efficiency to electrical energy [16].

Table 2.7 summarises all the discussed techniques. Note that a direct comparison of these techniques is difficult, since energy harvesting is extremely source dependent. For example, comparing the efficiency of a solar cell to that of a TEG or an antenna will not be useful due to the variation in source characteristic. Instead, they will be compared in terms of the expected power density. The power density values with (*) are the numbers
obtained based on the experimental work rather than the theoretical optimal values. Clearly, photovoltaic is the most attractive solution as it has the highest power density compared to other techniques. Most of the agricultural WSN nodes would be easily powered by a solar energy harvesting system provided that they are deployed in a well-lit location. In contrast, RF energy harvesting based on ambient sources has the lowest power density of less than $1 \mu W/cm^2$, making it the most ineffective energy harvesting technique in powering the outdoor WSN nodes.

<table>
<thead>
<tr>
<th>Energy Source (Transducer)</th>
<th>Power Density (Condition)</th>
<th>Commercially Available</th>
<th>Source of Information</th>
</tr>
</thead>
<tbody>
<tr>
<td>Solar (Photovoltaic Cell)</td>
<td>15 mW/cm$^2$ (direct sun)</td>
<td>Yes</td>
<td>Roundy et al.[34]</td>
</tr>
<tr>
<td></td>
<td>150 $\mu W/cm^2$ (cloudy day)</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>10 $\mu W/cm^2$ (indoor)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Thermal (Thermoelectric Generator)</td>
<td>5 $\mu W/cm^2$ (air-to-soil temperature variation at $\pm 4^\circ C$)</td>
<td>Yes</td>
<td>Meydbray et al.[46]</td>
</tr>
<tr>
<td></td>
<td>15 $\mu W/cm^2$ ($\Delta T = 10^\circ C$)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Radio Frequency (Antenna)</td>
<td>1 $\mu W/cm^2$ (Ambient sources)</td>
<td>Yes</td>
<td>Burch et al. [59]</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Visser et al. [56]</td>
</tr>
<tr>
<td>Wind (Micro Wind Turbine)</td>
<td>0.2 mW/cm$^2$ (wind speed at 4 m/s)</td>
<td>No</td>
<td>Federspiel et al. [51]</td>
</tr>
</tbody>
</table>

Table 2.7: Comparison of various ambient energy sources.

In this work, RF energy harvesting was chosen as the main subject of the investigation primarily due to the nature of the operating environment expected for the wireless ground-level sensor nodes. As stated in Chapter 1, the nodes are mainly buried within the soil. Although their transceiver parts are slightly raised above the ground, they may not be subjected to sunlight for all or appreciable length of time due to the obstruction of surrounding buildings, foliages or growing vegetations. Under these conditions, solar energy harvesting may not be an appropriate solution. Same constraint applies for thermoelectric energy harvesting which exploits temperature difference between air and soil. As clearly demonstrated in [45, 46], it is desirable to expose the hot side of the TEG to
direct sunlight so that it can be heated up to maintain a large temperature gradient across the TEG. Another discussed technique - wind energy harvesting using small-scale wind turbine - was also not considered in this research. Since the WSN nodes are located close to the ground, the wind speed will be nominally zero.

It has already been shown in Table 2.7 that, the power density of RF energy harvesting using ambient sources is extremely low (in the range of 1 μW/cm²). By considering the exemplar sensor node (Zyrox2 Baits Station) which has an average power consumption of 450 μW, this would require a huge collection area in order to harvest useful amount of energy to power the sensor device. A possible solution is to use a dedicated transmitter; however its maximum radiated power level must conform to the radio regulations in different countries. In addition, the performance of an RF energy harvesting system could also be affected by other environment factors, such as weather conditions and foliage effects. All of these factors will be examined later in the next chapter.

2.4 Further Review on RF Energy Harvesting

In recent years, RF energy harvesting has gathered increasing interests from both the industrial and academic communities due to widespread usage of RFID systems and wireless portable devices. At present, passive RFID tags are the only large-scale implementation of this technology. Powering embedded systems such as WSNs, however, remains challenging as they require much higher operating power than a simple passive RFID tag. In this section, a further literature review on the state-of-the-art in RF energy harvesting is presented. Note that, the terms ‘RF energy harvesting’, ‘wireless energy transfer’ and ‘wireless power transmission’ are often used interchangeably throughout much of the literature, to describe the same process of transmission of electrical energy from a power source to an electrical load without the connection of wires. The first part of this review provides a theoretical background on antenna field regions, i.e. near- and far-field. It is then followed by an assessment of several wireless energy transfer techniques, which include magnetic resonant induction, radio/ microwave, laser and ultrasonic-wave-based wireless energy transfer. Strengths and weaknesses of each technique are evaluated, along with a further discussion on choosing an appropriate wireless energy transfer method to power the intended agricultural WSN application.

2.4.1 Antenna Field Regions

Before assessing different wireless energy transfer techniques, it is useful to examine the antenna field regions. According to Stutzman [64], the space surrounding an antenna can
be divided into three regions, as illustrated in Figure 2.9: the reactive near-field, the radiating near-field and the far-field region. The boundaries separating these field regions are not well-defined, and the location of the transition region between near- and far-field depends on the dimension of the antenna.

The reactive near-field region is the closest region to the antenna in which the reactive fields predominates. In this region, the energy is stored but not radiated. For \( D \gg \lambda \), this region extends to a distance up to \( r = 0.62 \sqrt{\frac{D^3}{\lambda}} \) from the antenna [64], where \( D \) is the largest dimension of the antenna and \( \lambda \) is the wavelength of the operating frequency. For electrically small antennas, such as an ideal dipole (\( D \ll \lambda \)), the reactive near-field region exists within \( r = \frac{\lambda}{2\pi} \), the distance for which the maximum radiated and reactive powers are equal for an ideal dipole [64]. This region is sometimes known as the radian sphere.

![Figure 2.9: Field regions of an antenna [65].](image)

Meanwhile, the radiating near-field region lies between the reactive near-field and radiating far-field regions. In this region, the radiative fields predominate. The fields exhibit non-plane wave behavior and their angular field distribution is dependent upon the distance \( r \) from the antenna. If the antenna has a maximum dimension \( D \) which is smaller compared to the wavelength \( \lambda \), the radiating near-field region may not exist [65]. For antennas with \( D \gg \lambda \), the boundaries of the radiating near-field region can be defined as \( 0.62 \sqrt{\frac{D^3}{\lambda}} < r < 2 \frac{D^2}{\lambda} \) [64].
In the far-field region, the angular field distribution (e.g. radiation pattern of an antenna) is independent of the distance from the antenna. The electric and magnetic field components are perpendicular to each other, and also to the propagation direction. The fields exhibit plane wave behavior and have $1/r$ magnitude dependence. For an observation point to be in the far-field region, three conditions must be satisfied simultaneously [64]: $r > 2D^2/\lambda$, $r >> D$ and $r >> \lambda$. These three conditions are based on certain criteria for maximum phase difference and amplitude errors (parallel ray approximation), as well as the reduction of higher-order terms ($1/r^2$ and $1/r^3$) of the electric and magnetic fields of an antenna.

In [64], it is suggested that the criteria for far-field distance could be more specifically described by: $r > 2D^2/\lambda$, $r > 5D$ and $r > 1.6\lambda$. These criteria were chosen based the desired phase and amplitude errors, and the reduction in relative magnitudes of the higher-order components of the antenna fields [66]. The authors in [66, 67] plotted the three inequalities with the distance and aperture size normalised to wavelength. To determine the regions of validity, all three inequalities must be satisfied simultaneously. Figure 2.10 shows that the first criterion $r > 2D^2/\lambda$ is valid for antenna size $D > 2.5\lambda$. The second criterion $r > 5D$ applies for $1/3\lambda < D < 2.5\lambda$, whereas the third criterion $r > 1.6\lambda$ sets the far-field behavior when the antenna is very small, for which $D < 1/3\lambda$. It should be noted that, the regions of validity will change if the right-hand side of these criteria is multiplied by a different factor.

![Figure 2.10: Plot of the three alternative inequalities to determine far-field distance versus antenna aperture size, with both normalised to wavelength [66].](image)
2.4.2 Different Methods of Wireless Energy Transfer

Wireless energy transfer or wireless power transmission is a process where electrical energy is transmitted from a power source to an electrical load without the connection of wires. This technology is suitable for applications where instantaneous or continuous energy is required but wires are impractical (e.g. for the systems which are deployed in hostile or inconvenient environments). In general, wireless energy transfer systems can be divided into two categories: near- and far-field systems. The near-field systems operate based on the principle of inductive coupling, whereas the far-field systems using propagating electromagnetic waves to transfer energy from its source to the receivers. In the following sections, four different techniques of wireless energy transfer are reviewed in terms of their underlying power transfer mechanism, range, application and efficiency.

2.4.2.1 Magnetic Resonant Induction

Inductive coupling is a mature and well-understood method of wireless energy transfer. The basic principle of an inductive power transfer (IPT) system is illustrated in Figure 2.11. It consists of an AC power source, a primary (transmitting) coil, a secondary (receiving) coil and an ac-to-dc rectifying circuit in the receiver. When an alternating current drives the primary coil, a changing magnetic field is generated. Energy transfer takes place when the secondary coil is located within that field, which induces a voltage across the coil terminals. By connecting a load to the secondary coil, the induced current will flow, thereby generating power.

![Figure 2.11: Schematic overview of a typical IPT system [68].](image-url)
An electrical transformer is the simplest application of this type of wireless energy transfer, in which the magnetic field is usually confined to a core with high permeability. Other applications based on this mechanism include rechargeable toothbrushes, inductive cookers and recently proliferating ‘charging surface’ which require the devices to be placed directly on top, or very close to the charging base or pad. These techniques can be very efficient; however their transfer distances are limited to a few centimetres, with the efficiency rapidly dropped outside the range.

Recently, there has been substantial interest in exploiting wireless energy transfer using magnetic resonant induction. This technique is capable of solving two main problems associated with non-resonant induction methods: relatively short transfer distance and inefficiency over greater distances. Karalis et al. [69] presented a theoretical analysis of using resonant objects coupled through their near-fields to achieve efficient, non-radiative, mid-range wireless energy transfer. In this context, mid-range can be referred as transfer distance comparable to, or a few times the characteristic size of the transfer device. Intuitively, two objects with the same resonant frequency tend to exchange energy efficiently, while having weak interactions with other off-resonant objects. In systems of coupled resonance (e.g. acoustic or electromagnetic), there is often a general ‘strongly coupled’ regime of operation, in which the coupling rate is much faster than the intrinsic loss rate [69]. The authors pointed out that, if one could operate in that regime for a given system, the energy transfer is expected to be very efficient.

Based on the concepts and theoretical groundwork laid out in [69], Kurs et al. [70] experimentally demonstrated the feasibility of using magnetic resonant coupling for non-radiative, mid-range wireless energy transfer. To prove the concept, they conducted an experiment (shown in Figure 2.12) using two identical copper coils of 60 cm in diameter, with one inductively coupled to a power source, and another inductively coupled to a light bulb. Both coils were aligned coaxially, and designed to resonate at the same frequency of 10 MHz. A power of 60 watts was successfully transferred to the light bulb at a distance of two metres with an efficiency of about 40 percent. Although the preliminary results looked promising, there are a few practical issues remained to be addressed. First, the researchers observed that the amount of power transmitted to the load dropped sharply if either one of the coils was detuned from resonance, i.e. a shift in resonant frequency due to the effects of external objects placed within a few centimetres from the coils. Even though the coils can be manually tuned with moderate efforts, a practical system will require a feedback mechanism, e.g. a control circuitry which can perform the tuning
automatically. Secondly, their system has yet to conform to the safety regulations recommended by the IEEE or ICNIRP guidelines. When transferring 60 watts over 2 metres, the researchers found that the electric and magnetic fields measured at the halfway point between the transmitting and receiving coils exceeded the ICNIRP limits by a factor of 7.5 and 14, respectively. Thirdly, the coils used in the proposed system are too large for most of the wireless applications, such as mobile phones, PDAs and WSN nodes. The size of these coils must be further reduced so that they can be fitted into these portable devices.

Figure 2.12: Experiment conducted by Kurs et al.[70].

One of main advantages of magnetic resonant induction is that it enables wireless energy transfer at mid-range distance with high efficiency and less interaction with extraneous objects in the surrounding. According to Kurs et al. [70], magnetic resonances are suitable for everyday applications since the biological tissue and other common materials do not interact strongly with the magnetic fields, making the system safer and more efficient. In addition, most of the energy not picked up by the receiving coils will remain bound to the vicinity of the transmitting coils, instead of being radiated into the surroundings and lost. This ensures minimal wastage of power. The systems reported in [70, 71] have successfully demonstrated the potential of strongly coupled, magnetic resonant induction to deliver power with higher efficiency than far-field approaches, and at larger distances than conventional inductive coupling methods.

However, there are also some drawbacks associated with this technique as well. As discussed earlier, the magnetic resonant induction method is based on non-radiative,
near-field coupling. For that reason, the separation distance between the two coils must be kept within the range of the effect, which is about 0.16 wavelength of the operating frequency \((r < \lambda/2\pi)\). As the operating frequency increases, the distance over which the near-field coupling can be achieved decreases [72]. Another main limitation is that, the energy available for induction is a function of distance from the transmitting coil. The magnetic field produced from the coil drops off at a factor of \(1/r^3\), where \(r\) is the separation between the transmitting and receiving coils, along a center axis perpendicular to the coil’s plane. Waffenschmidt and Staring [73] performed an investigation on the achievable efficiency for resonant inductive transmission systems with varying distance and size ratios. They found that the power transfer efficiency of these systems started to drop off rapidly when the distance between the two coils increases beyond its optimal operating point (normally starts from a distance of approximately a coil’s diameter). Their analysis also showed that an optimum efficiency can only be achieved when both the coils having an identical size.

### 2.4.2.2 Radio/ Microwaves

Radiowaves are part of the electromagnetic spectrum with frequencies ranging from 3 Hz to 300 GHz. Microwaves occupy a subsection of this range between 300 MHz to 300 GHz. According to Brown [74], wireless power transmission using radio/ microwaves is a three-step process: conversion of a DC power source to RF power, transmission of RF power through space to some distant point, and collection and conversion of RF power into DC power with a rectenna. The rectenna is a combination of an antenna absorbing the incoming radiation, and a rectifying circuit (normally a diode) converting the alternating current (AC) signal across the antenna into usable DC output to power to an external load.

The concept of radio/ microwave power transmission has been in existence for more than a century since the early experiments of Heinrich Hertz and Nicola Tesla in the 1890s. Following their early work, very few attempts on wireless power transmission using radio waves were made during the first half of the 20th century. The interest in microwave power transmission (MPT) revived in 1960s when William C. Brown successfully demonstrated the powering of a small helicopter at an attitude of 50 feet using a microwave beam at a frequency of 2.45 GHz [75]. Since then, MPT has been proposed and researched in the context of high-power beaming. Applications for this type of power transfer include solar-powered satellite-to-ground transmission [76], intersatellite power transmission, and power transmission to areas which are inaccessible to the power grid [77], such as remote islands or mountain locations. A good review on the history and technology of such
wireless power transmission is given in [74]. In general, these MPT systems used large collection areas (antennas) so that watts or kilowatts of power could be transmitted over large distances.

In recent years, increasing attentions have been given to much lower power applications. Due to the rise of consumer electronics and increasing demand for longer battery life, research efforts have looked into practical ways to deliver power wirelessly to these portable devices using radio/microwaves. Shipley [78] attempted to supply wireless power to a household lamp through microwaves. His prototype successfully lit a 2.1V LED at a distance of 0.9 metres. Although the system is functional as a whole, the receiving antenna was relatively large. A more compact antenna design is required so that it can be fitted into household applications. Furthermore, the power level radiated from the transmitter used in his experiments did not comply with FCC regulations.

Sample et al. [79] from Intel Research developed Wireless Identification and Sensing Platform (WISP), a platform for sensing and computation in which wireless power is supplied by a commercial off-the-shelf UHF RFID reader operating at 915 MHz. The WISP comprises a 2 dBi dipole antenna, impedance matching components, RF power harvester (built using four-stage charge pump), demodulator to extract reader-to-WISP data, backscatter modulator for WISP-to-reader data, voltage regulator, programmable microcontroller and optional external sensors [60]. Each WISP consumes 2µW to 2mW and could be operated at distances of up to three meters from the readers that transmits 4W EIRP. Unlike conventional RFID design as an IC, the WISP was developed using PCB design due to its faster design iteration time, lower cost, and the ease of adding or removing sensors to the WISP platform. To provide lower power consumption and smaller form factor, the researchers suggested an IC implementation for future research.

In 2009, Powercast Inc. [80], a company based in Pennsylvania, introduced the very first commercial RF energy harvesters based on radio frequency broadcasts. Several receiver modules named Powerharvester™ have been designed for charging batteries, other energy storage devices, and for direct power applications. These harvesters can operate at frequency range from 850 to 950 MHz and they are intended to be used with an external transmitter. According to Powercast, their technology was capable of delivering 1 mW wirelessly to any receiver circuits placed within a meter radius of an antenna radiating 1 W inside a room. Powercast is currently examining the use of beam-steering and patch antennas to power a sensor located 14 metres away from the transmitter [81].
The advantage of using radio/microwaves as the medium for wireless energy transfer is that, it offers the possibility of low as well as high power transfer over significant distances (perhaps even kilometres). The received power is proportional to the effective aperture of the antenna. In order to maximise the collection of radiated energy at longer distances, an antenna with larger aperture is usually required. One possible way to improve the efficiency of radio/microwave power transmission is by using highly-directional antennas to confine and preferentially direct the radiated energy towards the receiving devices. Nonetheless, it will require an uninterrupted line-of-sight or complicated beam-steering mechanisms. Like other electromagnetic waves, the radio/microwaves propagating in outdoor environment may experience attenuation, reflection or scattering due to intervening hydrometeors and other environmental factors, such as trees, buildings and other obstructions along the propagation path. All these effects must be taken into consideration when designing an RF energy transfer system for the outdoor WSN applications. Another limitation of this technique is that the transmitter output power is constrained by the government regulations. This restriction, in turns, limits the amount of power which is transferrable to the receivers to low milliwatt and microwatt range. Current research efforts in microwave power transmission focus on the antenna design (material, geometry, etc) for optimal energy collection, and the improvement on RF-DC conversion efficiency of the rectifying circuits.

2.4.2.3 Laser Radiation

Wireless energy transfer based on laser radiation is often called "power beaming" [82] as the power is beamed at a receiver, and then converted into usable electrical energy. A laser device produces a highly collimated, monochromatic and coherent beam of light through a process of optical amplification based on stimulated emission of photons [83]. In principle, laser energy transmission systems are similar to the microwave energy transmission technology. A power source (i.e. electricity) is converted into a laser beam that is directed to a receiver comprising photovoltaic cells. The cells absorbed the energy and transform it back into DC power. A major difference between microwave and laser-based systems lies on the wavelength used, where the first uses microwave frequencies (e.g. 2.45 or 5.8 GHz with wavelength of 0.12 – 0.05m), whereas the latter relies on much higher frequencies in the visible or near infrared spectrum (with wavelengths range from hundreds of nanometers up to a few tens of micrometers). The laser power delivered to the cells is dependent upon the aperture of the beam director, laser wavelength, and the degree of laser beam coherence [84].
Over the years, a number of laser-based wireless power transmission experiments and applications have been proposed and described. A group of researchers from NASA [85] demonstrated a beam-powering system for a small-scale, light weighted aircraft using an invisible, ground based laser. The laser tracks the aircraft in flight, with its energy beam directed towards the photovoltaic cells carried onboard in the aircraft. The cells convert the laser energy into electricity powering a small electric motor which spins the plane’s propeller. Their work was perceived as a significant step towards the capabilities of beam-powering an aeroplane. Nonetheless, their demonstrations were only taken place in an indoor environment, which is inside the building. Further investigations are required to examine the outdoor factors such as wind and weather condition on the test-flight performance.

In [86], Afzal et al. proposed a recharge mechanism model for WSNs based on variable infrared laser through the incorporation of corner cube retroreflector (CCR) and its variant, thin-film corner cube retroreflector (TCCR) mounted on the sensor node. The CCR forms a wireless optical communication link between a node and the base station. It consists of three mutually orthogonal reflective mirrors forming a concave corner. Two of the mirrors are fixed, while the third mirror is intentionally misaligned and can be tilted using electrostatic force to modulate the retroreflected beam. The TCCR has almost the same structure as CCR except that two of the mirrors are partially reflecting mirrors coated with thin film. The light transmitted in these mirrors is absorbed, and is used to recharge the thin-film battery of the node. The practicability of their proposal remains to be validated as no actual prototype has yet to been reported in the public domain.
PowerBeam Inc. [87], a company based in the United States, also explored the use of laser-beaming technology in charging consumer devices, such as wireless keyboard, mobile phones and speakers. Like similar laser-beam applications discussed above, the electricity is first turned into optical energy using an invisible, infrared laser and directed to a photovoltaic detector attached to the targeted device. The received energy is then converted into electricity. News release indicated that their technology is capable of generating 1.5 W of power at 10 metres [87] with an uninterrupted line-of-sight (LOS). An innovative part of the PowerBeam solution is a patented-pending safety system [87], which terminates the power transfer instantaneously whenever a person or animal interrupts the beam. PowerBeam Inc. has recently announced a collaboration with a major original equipment manufacturers, and they will be launching their first optical, wireless power consumer product by end of 2012.

On the whole, wireless energy transfer based on laser radiation is still a relatively new technology. Further research and development work in this area are necessary before its implementation can become feasible in powering the WSN applications. Due to its inherent properties such as high directivity and narrow beam, laser-beaming can deliver higher power to the targeted device at greater distances than microwaves. Another important advantage of laser-beaming is the reduction size of the transmitting and receiving devices. Since the laser beams have shorter wavelengths (about 5 orders of magnitude) compared to microwaves, the receiver can be designed at much smaller size, which could possibly lead to lower manufacturing costs. However, the laser-based energy wireless transfer solution must require a direct LOS path, making it difficult to be integrated into portable devices. A possible way to overcome the LOS issue is to use the redirecting mirrors controlled through a computer programmed pan or tilt mechanism [87], but this could only be realised with the expense of a more complex and expensive design. Besides, laser beams are also very susceptible to weather-induced attenuation. The attenuation due to scattering is the highest when the wavelength is comparable with the particle size in the atmosphere, such as rain or snow [88]. Another main concern about laser-beaming technique is safety. Since the power flux density in lasers is much higher than those in microwaves, the beams could potentially be hazardous as they can damage the retina of eyes, or even the skin. As a precaution, all laser systems must operate such that they are conformed to the regulatory limits on eye and skin exposure. Besides, it is desired to have a safety system, such as the one proposed in [87], to monitor any undesired intrusion into the beam, and shutting down the power transfer immediately upon detection.
2.4.2.4 Ultrasonic Waves

Another alternative for wireless energy transfer is by using ultrasonic-waves, which are generated from cyclic sound pressure. The ultrasonic-waves consist of frequencies greater than 20 kHz and can be in excess of 25 MHz. Like laser-beaming, the concept of energy transfer using ultrasounds is relatively new. At present, this technique is mainly used for energy transmission through human tissues in powering medical implanted devices [89, 90], and through metal walls [91, 92] for wireless sensing in pressure vessels or ships. Their transfer distance is typically within the range of a few centimetres. Ishiyama et al. [93] presented a wireless energy transmission system based on ultrasonic air transducers. Their developed system consisted of a transmitter unit (a pulse generator, a power amplifier and a piezoelectric air transducer) and a receiver unit (a receiving transducer and a rectifying circuit). A plastic horn structure, as shown in Figure 2.14, was attached to the front of the transmitter in order to focus the ultrasonic waves towards the receiver, thereby improving the overall energy transfer performance.

![Figure 2.14: Proposed ultrasonic wireless transfer system in [93].](image)

Their system was operated at 28 kHz, and it was capable of transmitting 0.8 mW over a distance of 0.3 metres (when the transducer was supplied with an input voltage of 120V peak-to-peak). Although the experimental results in [93] showed some promises in using ultrasonic-waves as a possible wireless powering solution, its suitability for outdoor applications have yet to be fully explored. Similar to other wireless energy transfer systems, its performance could be affected by the surrounding environmental conditions since the attenuation of ultrasonic-waves is a function of temperature, humidity, frequency and air pressure [94]. For that reason, further investigations are needed to examine these effects, as well as on how the distance and mutual orientation of the transducers may affect the...
overall power transmission performance. Up to date, no work in the literature has successfully demonstrated the capability of ultrasonic wireless power transmission over large distances, i.e. more than a metre.

2.4.3 Selection of Far-field RF Energy Harvesting

As discussed in previous sections, wireless energy transfer can be achieved using magnetic resonant induction, radio/ microwaves, lasers or ultrasonic waves. Selection of a proper technique depends upon two main criteria: the required transfer distance and the allowable size of the transmitting/ receiving device. For the intended WSN application, the system is required to power the WSN nodes up to 5 metres from the source. In addition, it is desirable that the size of the antenna/ coil structure to be practically small so that it can be conveniently attached to the targeted sensor node for harvesting the RF energy. Based on these two considerations, a far-field RF energy harvesting system, using radio/ microwaves as the energy transfer medium, has been chosen as the subject of the investigation. Although the magnetic resonant induction has higher transfer efficiency compared to other far-field approaches, its transfer distance is well limited to near-field region. One could possibly extend the near-field range by using the coils with lower self-resonant frequency. However, this would normally require a larger coil size (with bigger diameter and higher number of turns), thus making it impractical for the intended application. In contrast, wireless energy transfer based on radio or microwaves offers the possibility of low as well as high power transfer over large distances. However, in the case of microwaves, a direct LOS is needed. Furthermore, the required aperture size of the antenna increases with the transfer distance. Laser radiation is another possible solution which enables wireless energy transfer over long distances. Nonetheless, this technology is still relatively new, and it requires further development, especially on resolving the concerns and issues related to laser safety. For this reason, the laser technique was not considered in this research.
3 System Analysis

3.1 Introduction

This chapter presents a theoretical-based analysis on a far-field, outdoor RF energy harvesting system in powering the wireless ground-level sensor nodes. It seeks to identify all relevant design parameters, and to quantify their impact on the system’s overall performance. An RF link budget analysis is also included to investigate the feasibility of using RF energy harvesting to power the exemplar node: Zyrox2 Baits Station.

3.2 Overview of Outdoor RF Energy Harvesting System

In general, an outdoor RF energy harvesting system consists of the following components: RF energy source(s), external environment, receiving antenna, power harvester circuit, energy storage and load. A block diagram of the complete system is depicted in Figure 3.1. As seen from the Figure, a radiating source (e.g. a dedicated transmitter or a GSM base station) is used to broadcast a high-intensity RF signal wirelessly through the air. The transmitted RF signal is collected by a receiving antenna attached to the WSN node, and is being converted into usable DC voltage via a power harvester circuit. The DC voltage is then stored in an energy storage element before being made available for the operation of the load, which is the microcontroller unit (MCU) and the sensor devices. In the following sections, each component of the system will be discussed in detail.

![Figure 3.1: Block diagram of an outdoor RF energy harvesting system.](image-url)
3.3 RF Energy Source

As mentioned in Chapter 2, the sources for RF energy harvesting can be divided into two categories: ambient sources and intentional sources. In order to supply a more predictable and controllable power to the wireless ground-level sensor nodes, it was decided to employ a dedicated transmitter as the energy source of the system. This approach offers the possibility of higher power densities, and allows the system to be designed based on prior knowledge of the transmitted RF signal, i.e. orientation and polarisation.

3.3.1 Frequency of Operation

The selection of a proper operating frequency band for the RF energy harvesting system is crucial as it can affect other related parameters, such as the size of the transmitting/receiving antennas, the operating range of the system, as well as the design of the power harvester circuit. An Industrial, Scientific and Medical (ISM) frequency band was chosen as the operating band of the investigated system owing to its license free radio, large spectrum allocation and global availability [6]. Two Ultra High Frequency (UHF) ISM bands: 867 MHz and 2.45 GHz, were evaluated in terms of their path losses.

Free space path loss (FSPL) is the loss in signal strength when an electromagnetic wave propagates over a line-of-sight path in free space, with no obstruction between the transmitter and the receiver. It can be expressed in decibel form as [95]:

\[
FSPL = 20 \log_{10} \left( \frac{4 \pi df}{c} \right)
\]  

(3.1)

where \(d\) is the distance between the transmitter and receiver (in metres), \(f\) is the operating frequency (in Hertz), and \(c\) is the speed of light, which has a value of \(3 \times 10^8\) ms\(^{-1}\). The FSPL equation above often leads to a false belief that free space attenuates the propagating wave according to its frequency. However, this is not the case, as there is no physical mechanism causing it to happen. The FSPL expression actually contains two effects. The first effect is the spreading out of energy over the increasing surface area of a sphere, which can be described by the inverse square law, and it is not a frequency-dependent effect. The second effect results from the receiving antenna’s aperture, i.e. how well an antenna can pick up the power from an incoming wave. This term is dependent on the wavelength, and this is where the frequency-dependent effect of the path loss equation
arises. Figure 3.2 shows the FSPL of the two evaluated ISM bands: 867 MHz and 2.45 GHz.

As seen from the Figure above, the FSPL at 2.45 GHz is about 9 dB higher than that at 867 MHz. This indicates that the received power of an antenna operating at 867 MHz could be almost 8 times more than that at 2.45 GHz if they are located in the same location from a radiating source. Due to its much lower free space attenuation, 867 MHz was chosen as the operating frequency of the system. Table 3.1 summarises the frequency allocation as well as the radiated power limit of the two evaluated ISM bands in various countries [96, 97]. Note that these frequency bands are also allocated for the use in RFID applications [98].

<table>
<thead>
<tr>
<th>Countries</th>
<th>Frequency Bands</th>
<th>Radiated Power Limit</th>
</tr>
</thead>
<tbody>
<tr>
<td>United States</td>
<td>902 – 928 MHz</td>
<td>4 W EIRP</td>
</tr>
<tr>
<td></td>
<td>2400 – 2500 MHz</td>
<td>4 W EIRP</td>
</tr>
<tr>
<td></td>
<td>865.6 – 867.6 MHz</td>
<td>2 W ERP</td>
</tr>
<tr>
<td>United Kingdom</td>
<td>2446 – 2454 MHz</td>
<td>500 mW EIRP (outdoor)</td>
</tr>
<tr>
<td></td>
<td></td>
<td>4 W EIRP (indoor)</td>
</tr>
<tr>
<td>Japan</td>
<td>952 – 954 MHz</td>
<td>4 W EIRP</td>
</tr>
<tr>
<td></td>
<td>2450 MHz</td>
<td>1 W EIRP</td>
</tr>
</tbody>
</table>

Table 3.1: Frequency allocation and radiated power limit for the selected UHF bands [96, 97, 99].
3.3.2 Transmitter Unit

The transmitter unit of the investigated system is constructed using a commercially-available equipment. A TM8110 Mobile Radio [100] manufactured by Tait Electronics Ltd, will be used as the radiating source of the system. This radio can be programmed to operate between 762 to 870 MHz with a maximum power of 35 W, as shown in the specifications listed in Table 3.2. In order to use this equipment for this research, a non-operational license was acquired from the Office of Communications (Ofcom).

<table>
<thead>
<tr>
<th>Power Supply</th>
<th>10.8 - 16 VDC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output Power</td>
<td>2W, 5W, 15W, 35W</td>
</tr>
<tr>
<td>Operating Frequency</td>
<td>762 - 870 MHz</td>
</tr>
<tr>
<td>Connector Type</td>
<td>BNC- Female</td>
</tr>
<tr>
<td>Output Impedance</td>
<td>50 Ω</td>
</tr>
</tbody>
</table>

Table 3.2: Specifications of TM8110 mobile radio [100].

For the transmitting device, two types of antenna can be employed: omni-directional and directional antenna. An omni-directional antenna radiates or receives energy equally in all directions in one plane, where the radiated or received power decreases with elevation angle above or below that plane. Since the omnidirectional antenna is designed to provide a donut-shaped, 360 degrees horizontal radiation pattern, it is ideal for point-to-multipoint links when coverage in all directions from the antenna is required. The major drawback of using the omnidirectional antenna is that its energy is greatly diffused when broadcasting in 360 degrees. This eventually limits the operating range and signal strength at longer distance. On the other hand, a directional antenna radiates or receives greater energy in one and more directions. As the gain of the directional antenna increases, the angle of radiation usually decreases [101]. This provides a longer coverage distance, but at the expenses of reducing the coverage angle. Directional antennas are primarily used for point-to-point wireless links.

In this research, the targeted sensor nodes are non-mobile, and they are deployed in a fixed location. For this reason, the directional transmitting antenna appears to be a better choice as it can direct most of the energy towards the stationary receivers. A directional dual-band patch antenna, ID850-SF00 [102], from Laird Technologies, was chosen as the transmitting antenna of the investigated system. Table 3.3 shows the main specifications of the selected patch antenna, provided by the antenna manufacturer.
<table>
<thead>
<tr>
<th>Frequency Range</th>
<th>800 – 960 MHz / 1710 – 2500 MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain</td>
<td>7 dBi / 10 dBi</td>
</tr>
<tr>
<td>Beamwidth Horizontal Plane</td>
<td>70° / 60°</td>
</tr>
<tr>
<td>Beamwidth Vertical Plane</td>
<td>50° / 45°</td>
</tr>
<tr>
<td>Polarisation</td>
<td>Vertical</td>
</tr>
<tr>
<td>Impedance</td>
<td>50 Ω</td>
</tr>
<tr>
<td>VSWR</td>
<td>&lt; 1.5</td>
</tr>
<tr>
<td>Dimension</td>
<td>21 × 18 × 4.39 cm</td>
</tr>
</tbody>
</table>

Table 3.3: Specifications for the dual band patch antenna [102].

3.4 Receiving Antenna

In addition to the transmitting antenna, a receiving antenna is another key component of an RF energy harvesting system as it is responsible to extract the incident RF energy from the radiating sources. The performance of an antenna is mainly determined by some basic parameters, such as radiation pattern, gain, efficiency and polarisation. In order to characterise the antenna performance, definition of these parameters is necessary. In this section, a few of the critical parameters are discussed.

3.4.1 Radiation Pattern

The radiation pattern of an antenna is defined as “a mathematical function or a graphical representation of the radiation properties of the antenna as a function of space coordinates. In most cases, the radiation pattern is determined in the far field region, and it is represented as a function of the directional coordinates. Radiation properties include power flux density, radiation intensity, field strength, directivity, phase or polarisation” [65]. An example of the radiation pattern plot of a generic directional antenna is shown in Figure 3.3. As seen from the Figure, the radiation pattern consists of several radiation lobes. Definitions of each lobe are described below:

- **Main Lobe**: A radiation lobe containing the direction of maximum radiation.
- **Minor Lobe**: Any lobes other than the main lobe. Minor lobes usually represent radiation in undesired directions. A good antenna design should minimize the minor lobes.
- **Side Lobe**: Minor lobes adjacent to the main lobe. Side lobes are normally the largest among the minor lobes.
• Back Lobe: A radiation lobe which occupies the hemisphere in a direction opposite to that of the main lobe.

• Nulls: A zone where the effective radiated power is at a minimum. It is useful for several purposes, such as suppressing interfering signals at a given direction.

Figure 3.3: Radiation lobes and beamwidth of an antenna pattern, adapted from [65].

3.4.2 Directivity

The directivity of an antenna is defined as “the ratio of the radiation intensity in a given direction from the antenna to the radiation intensity averaged over all directions. The average radiation intensity \( U_{ave} \) is equal to the total power radiated by the antenna divided by \( 4\pi \)” [65]. In its mathematical form, the directivity can be written as:

\[
D = \frac{U}{U_{ave}} = \frac{4\pi U}{P_{rad}}
\]

(3.2)

If the direction is not specified, the direction of maximum radiation intensity is normally implied. In this case, the maximum directivity is given by:

\[
D_{max} = \frac{U_{max}}{U_{ave}} = \frac{4\pi U_{max}}{P_{rad}}
\]

(3.3)

where

\( D \) is the directivity of the antenna (dimensionless)

\( D_{max} \) is the maximum directivity of the antenna (dimensionless)

\( U \) is the radiation intensity of the antenna (W/ unit solid angle)

\( U_{max} \) is the maximum radiation intensity of the antenna (W/ unit solid angle)
$U_0$ is the radiation intensity of an isotropic source (W/ unit solid angle)
$P_{rad}$ is the total radiated power (W)

3.4.3 Efficiency

Antenna efficiency is the measure of an antenna’s ability to convert its input power into radiation [103]. A high efficiency antenna has most of the delivered power to the antenna’s input radiated away, whereas a low efficiency antenna has most of the power absorbed as ohmic losses within the antenna, or reflection losses due to mismatch between the transmission line and the antenna. The ohmic losses within the antenna are typically dielectric and conduction losses. In general, there are a few efficiency terms associated with an antenna. The antenna radiation efficiency ($e_{cd}$) is defined as the ratio between the radiated powers to the input power delivered to the antenna [65]. It can also be expressed as a product of the conduction efficiency ($e_c$) and the dielectric efficiency ($e_d$).

$$e_{cd} = \frac{P_r}{P_{in}} = \frac{P_r}{P_r + P_{ohm}} = e_c e_d \quad (3.4)$$

Another commonly used term is the antenna’s total efficiency, where it includes the losses within the antenna structure, as well as the reflection losses due to mismatch at the antenna terminals. It can be expressed as:

$$e_o = e_{cd} e_r = e_{cd} (1 - |\Gamma|^2) \quad (3.5)$$

where $e_o =$ total efficiency (dimensionless)
$e_r =$ reflection/ mismatch efficiency (dimensionless)
$e_c =$ conduction efficiency (dimensionless)
$e_d =$ dielectric efficiency (dimensionless)
$\Gamma =$ voltage reflection coefficient at the antenna’s input terminals (dimensionless)

3.4.4 Gain

Gain is perhaps one of the most important design parameters in an RF energy harvesting system. An antenna designer must take into consideration of the antenna’s application before determining the gain for the antennas. This is normally done through an RF link budget (see Section 3.8). The gain of an antenna is closely related to its directivity and radiation efficiency, and it can be written as [65]:

$$G(\theta,\phi) = e_{cd} D(\theta,\phi) \quad (3.6)$$
Note that if the antenna radiation efficiency is 100%, then the antenna gain would be equal to its directivity. Occasionally, the antennas are expressed in absolute (realised) gain term. This term takes into account the reflection or mismatch losses when the antenna element is connected to a transmission line.

\[ G_{\text{abs}}(\theta, \phi) = e_r G(\theta, \phi) \]  

(3.7)

The antenna gain is a passive phenomenon, in which the power is not added by the antenna but being redistributed to provide more radiated power in a certain direction than it would be transmitted by an isotropic antenna [103]. Published figures for the antenna gain are usually expressed in decibel form (dB), in which the unit of dBi and dBD are used when describing the antenna gain relative to an ideal isotropic radiator and a half-wave dipole, respectively. The antennas with high gain have the advantage of longer transmission range but they must be pointed carefully in a particular direction. Meanwhile, the antennas with lower gain have shorter range, but their orientation’s effect on the overall system performance is less significant.

### 3.4.5 Return Loss

Return loss measures the effectiveness of power delivery from a transmission line to a load such as an antenna [104]. It represents the reduction in amplitude of the reflected wave in comparison to the incident one. Expressed in dB, the return loss is given by:

\[ RL(\text{dB}) = 10 \log_{10} \left( \frac{P_{\text{in}}}{P_{\text{ref}}} \right) \]  

(3.8)

where \( P_{\text{in}} \) is the incident power sent towards the antenna-under-test (AUT), and \( P_{\text{ref}} \) is the power reflected back to the source. The power can also be expressed in terms of voltage (or equivalently as field strength) in a transmission line or waveguide:

\[ RL(\text{dB}) = 10 \log_{10} \left( \left| \frac{1}{\Gamma} \right| \right) = -20 \log_{10} |\Gamma| \]  

(3.9)

where \( \Gamma \) is the complex reflection coefficient at the input of the AUT. Table 3.4 shows the relationship between the reflection coefficient, the return loss and the reflected power for a perfect match, an ideal open/short circuit, and three cases within the extremes.
<table>
<thead>
<tr>
<th>Reflection Coefficient, ( \Gamma )</th>
<th>Return Loss (dB)</th>
<th>Reflected Power (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>( \infty )</td>
<td>0 (( Z_L = Z_0 ))</td>
</tr>
<tr>
<td>0.03</td>
<td>30.0</td>
<td>0.1</td>
</tr>
<tr>
<td>0.1</td>
<td>20.0</td>
<td>1.0</td>
</tr>
<tr>
<td>0.316</td>
<td>10.0</td>
<td>10.0</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>100.0 (( Z_L = ) open, short)</td>
</tr>
</tbody>
</table>

Table 3.4: Relations between the mismatch parameters.

In order to achieve maximum power transference between the receiving antenna and the power harvester circuit, the impedance of the antenna (\( Z_{ANT} \)) must be matched to that of the circuit (\( Z_L \)). As shown in Table 3.4, if the antenna is perfectly matched to the load (i.e. \( Z_{ANT} = Z_L \)), the return loss would be infinite, which indicates no reflected power from the load. For most wireless applications, a higher value of return loss (more than 10 dB) is desirable.

### 3.4.6 Polarisation

The polarisation of an antenna is the orientation of the electric field vector of the radiated wave with respect to the Earth’s surface and is determined by the physical structure and orientation of the antenna [103]. Most antennas usually radiate either in linear or circular polarisation. An antenna is linearly polarised if it radiates only in one plane containing the direction of propagation. In this case, the electric field vector moves back and forth along a line. Vertical and horizontal polarisations are both considered as linear polarisation. On the other hand, a circularly polarised antenna radiates in both horizontal and vertical planes (including the planes in between). The electric field vector remains constant in length but rotates around a circular path. If the radiated wave is travelling towards the observer and the vector rotates clockwise, it is left-hand circularly polarised. If the rotation is counter clockwise, it is right-hand circularly polarised.

In RF energy harvesting, the antenna polarisation is an important design consideration. In order to transfer maximum power between a transmitting and receiving antenna, both antennas must have the same spatial orientation, polarisation sense and axial ratio. If the antennas are not properly aligned or do not have the same polarisation, the power transfer between the two antennas will be reduced, thereby degrading the overall system efficiency and performance. When the two linearly polarised antennas are rotated from each other by an angle of \( \psi_p \) (which is the difference in alignment angle), the power loss due to the
polarisation mismatch, also known as the Polarisation Loss Factor (PLF), can be determined using the following formula [65]:

\[ PLF = \left| \hat{\rho}_w \cdot \hat{\rho}_a \right|^2 = \cos^2 \psi_p \]  \hspace{1cm} (3.10)

This PLF can also be expressed in dB form, given by:

\[ PLF(dB) = 20 \log(\cos \psi_p) \]  \hspace{1cm} (3.11)

Note that the PLF will be unity if both antennas have the same polarisation, i.e. maximum power transfer occurs. However, if one antenna is vertically polarised and the other is horizontally polarised, the angle between their radiated E-fields will be 90 degrees, and therefore no power will be transferred. In some cases, a circularly polarised antenna might be considered to receive a linearly polarised wave or vice versa if the orientation of the linearly polarisation is unknown. The PLF between linear and circular polarisation would be 3 dB, regardless of the orientation of the linearly polarised wave. In short, the greater the mismatch in polarisation between the two antennas, the greater the power loss will be.

### 3.4.7 Friis Transmission Equation

After reviewing some of the critical antenna parameters, the performance of a typical RF energy harvesting system can now be determined. The power available at the receiving antenna, given a known transmitted power and gain, can be estimated using the Friis Transmission Equation. Due to the importance of this equation, the following section is devoted to its derivation.

Assuming a transmitting antenna fed with power \( P_T \), the power flux density \( P_D \) at a distance \( d \) in a given direction \((\theta, \phi)\) from the antenna can be expressed as [105]:

\[ P_D(\theta, \phi) = \frac{P_T G_T(\theta, \phi)}{4\pi d^2} \text{ (W/m}^2\text{)} \]  \hspace{1cm} (3.12)

where \( G_T \) is the gain of the transmitting antenna. When a receiving antenna with an effective collecting aperture area \( A_e \) is placed in the power flux, a power \( P_R \) will be delivered into a load connected to the antenna terminals (assuming there is no mismatch or feedline losses) [105]:

\[ P_R = \frac{P_T G_r(\theta, \phi) A_e}{4\pi d^2} \text{ (W)} \]  \hspace{1cm} (3.13)
The effective aperture of the antenna can be related to its gain $G_R$ and it is given by [105]:

$$A_e = G_R (\theta, \phi) \frac{\lambda^2}{4\pi} \text{ (m}^2\text{)} \tag{3.14}$$

Suppose the antennas are aligned for maximum directional radiation and reception. Upon substituting Equation (3.14) into Equation (3.13), the received power can then be written in terms of the transmitting and receiving antenna gains $G_T$ and $G_R$, respectively

$$P_R = P_T G_T G_R \left(\frac{\lambda}{4\pi d}\right)^2 \text{ (W)} \tag{3.15}$$

Equation (3.15) above is often known as the Friis Transmission Equation. This equation assumes that the transmitting and receiving antennas are matched to their respective transmission lines and loads, in which the reflection efficiencies are unity. Furthermore, it also assumes that the polarisation of the receiving antenna is matched to that of the incoming wave, such that the polarization loss factor is unity. In a practical system, these two assumptions may not hold true. Therefore, the power $P_R$ delivered to load of the receiving antenna can be more accurately represented by [65]:

$$P_R = P_T G_T G_R \left(\frac{\lambda}{4\pi d}\right)^2 (1-|\Gamma_t|^2) (1-|\Gamma_r|^2) |\rho_t \cdot \rho_r|^2 \text{ (W)} \tag{3.16}$$

where $\Gamma_t$ and $\Gamma_r$ are the reflection coefficients, and $\rho_t$ and $\rho_r$ are the polarisation vectors of the transmitting and receiving antennas, respectively. In order to optimise the overall system performance, the impedance and polarisation mismatch losses must be minimised.

Note that Equations (3.15) and (3.16) can only be used to estimate the upper bound of the received power at the antenna. The actual harvestable power of the system (i.e. the delivered power to the energy storage and the sensor node) could be much lower due to additional losses during the RF-DC conversion process within the power harvester circuit.

### 3.5 Power Harvester Circuit

A power harvester circuit is used to convert the captured RF energy from the receiving antenna into usable DC power. The block diagram of a typical power harvester circuit is depicted in Figure 3.4. As shown from the Figure, a receiving antenna and an impedance matching network precede the front end of the circuit. This is followed by a rectifier (i.e.
voltage multiplier) which converts the captured RF energy into DC voltage. The rectified voltage is regulated through a voltage regulator, and it is then being stored in an energy storage element, such as an ultra-capacitor or a rechargeable battery. Finally, the stored energy is supplied to operate the microcontroller unit (MCU) and the onboard sensors for data acquisition, processing and communication between the WSN nodes and the base station.

![Diagram of a typical power harvester circuit](image)

Figure 3.4: Block diagram of a typical power harvester circuit.

In general, the design complexity of a power harvester circuit depends on the system requirements, as well as its desired functionality. A simple harvester may just provide RF-DC rectification, whereas a more complex harvester may include power management circuitry and other functionality within the harvester component. There are two main design considerations associated with a power harvester circuit. First, the circuit should have high sensitivity so that it can harvest RF energy with low power densities to achieve longer operating distance. As shown in Equation (3.12), the power density in free space drops off at the rate of $1/d^2$. This implies that the available power to a receiver in far-field RF energy harvesting decreases by 6 dB for every doubling of distance from the transmitter. In other words, improving the circuit’s sensitivity allows the RF-DC conversion at greater distances from the radiating source. Secondly, the circuit must have high RF-DC conversion efficiency in order to ensure most of the captured energy from the antenna can be converted into usable DC power. The efficiency level must be maintained over a wide range of operating conditions, such as the variations in input power and output load resistance. In the following sections, each component in the power harvester circuit will be briefly discussed.
3.5.1 Impedance Matching Network

The radiation resistance of the receiving antenna of the investigated system has a typical value of 50Ω at its operating frequency. In contrast, the input impedance of a power harvester circuit normally has a reactance component, owing to the capacitance of rectifying diodes or MOSFETs implemented in the voltage multiplier. Therefore, an impedance matching network is required to transform the input impedance of the circuit to 50Ω. By having an impedance matching network, the power transfer between the receiving antenna and the voltage multiplier circuit can be maximised. In addition, the network can also be used to provide a large voltage gain at the input port, which reduces the power loss at the voltage multiplier. As a result, the overall efficiency of the power harvesting path can be improved [106].

The electrical model of the antenna and power harvester circuit incorporated with an impedance matching network is shown in Figure 3.5. The antenna is modelled as an AC voltage source $V_S$ with a series impedance $Z_{ant} = R_S + j X_{ant}$. $R_S$ is the radiation resistance emulating the power used for transmitting or receiving the electromagnetic wave, while $X_{ant}$ is the reactive part which can be inductive or capacitive depending upon the antenna designs. To achieve maximum power transfer, the input impedance of the power harvester circuit $Z_{in}$ must made equal to the conjugate of the antenna’s impedance, that is $Z_{in} = Z_{ant}^* = R_S - j X_{ant}$. $Z_L$ is the impedance of the voltage multiplier and the ensuing load, where $R_L$ and $X_L$ are the real and imaginary parts of this impedance, respectively.

One of the most commonly used impedance matching circuits for RF energy harvesting and RFID transponders [107, 108] is an L-network. The L-network can be easily constructed using two reactive components (an inductor and a capacitor), one is placed in
series and another one in parallel. There are two possible challenges when designing an efficient matching network for a power harvester circuit. First, the circuit itself is a non-linear device as its input impedance changes as a function of input power and frequency. In order to achieve optimal power transfer for maximum distance, the circuit is normally matched to the receiving antenna at its minimum input power to ensure correct operation of the circuit. Shorter distances lead to higher input power at the circuit, which produces a variation in the circuit’s input impedance. In this case, a mismatching between the antenna and the circuit will occur, but these power losses can be compensated with the increase of the input power [108]. Secondly, it is difficult to control the impedance to a desired value. If the matching circuit is built using off-the-shelf components, it would be hard to obtain the inductors or capacitors having specific inductance or capacitance values. To overcome this problem, a tunable impedance approach is normally employed [79, 109].

3.5.2 RF-DC Rectifier

An RF-DC rectifier converts the captured RF energy from the antenna, and provides a DC output voltage to the rest of the circuit. A variety of rectifier structures suitable for RF power harvester was reported in [110], including a single diode [111] which forms a rectenna together with the antenna, a bridge of diodes [112], and a voltage multiplier [113]. The diode and the diode bridge configurations supply an output DC voltage at the ensuing load, in which its amplitude is lower than that of the incoming signal. The principle and operation of these two circuits are well-described in the standard texts on microelectronics, and thus it will not be covered in this section. A voltage multiplier, shown in Figure 3.6, is more widely used for RF energy harvesting applications. It performs not only the RF to DC conversion, but also produces a DC output voltage which multiplies the peak amplitude of the incoming RF signal.

![Figure 3.6: A single-stage voltage multiplier.](image-url)
The operation of an ideal voltage multiplier can be explained by looking at successive half-cycles of an input signal with amplitude, $V_{IN}$. It is assumed that the diodes ($D_1, D_2$) and the capacitors ($C_1, C_2$) are identical. $V_{TH}$ is the forward conduction voltage of the diodes.

1) During the first negative half-cycle of the input signal (when $V_{IN} < V_{TH}$), $D_1$ is forward-biased and $C_1$ is charged up to a peak voltage, $V_{IN} - V_{TH}$. The voltage across $C_1$ remains at $V_{IN} - V_{TH}$ since there is no discharge path available.

2) For the following positive half-cycle, $D_1$ is reverse-biased and hence it will not conduct any current. Meanwhile, $D_2$ is forward-biased and it conducts current, which charges up $C_2$. This makes the output voltage observed at $C_2$ roughly two times the amplitude of the input signal minus the turn-on voltage of both diodes, $2V_{IN} - 2V_{TH}$.

An attractive feature of this circuit is that it can be arranged in a cascade configuration to achieve higher output voltage. An N-stage voltage multiplier is displayed in Figure 3.7.

Given that $V_{TH}$ is the forward conduction voltage of the diodes, the output voltage of the circuit $V_{OUT}$ can be estimated by:

$$V_{OUT} = 2N(V_{IN} - V_{TH})$$

(3.17)

Figure 3.7: N-stage voltage multiplier [107].
From Equation (3.17), it clearly shows that the rectifying diodes with lower $V_{TH}$ are desirable in a voltage multiplier so that a higher $V_{out}$ can be obtained with minimum $V_{IN}$. Schottky diodes are commonly used due to their low $V_{TH}$ and fast switching capability. The design of a voltage multiplier involves a tradeoff between the power efficiency, useful impedance and desired operating point, i.e. improving one of these may compromise another. It is recommended to use the diodes with large saturation current (leading to lower forward voltage drop), low junction capacitance $C_j$, small series resistance $R_s$, and small parasitic capacitance $C_{sub}$ to minimise the losses within the diode [114]. In addition, the diode area needs to be optimised. Diodes with larger area may have higher saturation current and smaller series resistance; but they also have larger junction and substrate capacitances which will contribute to power losses [115]. Likewise, it is desirable to have the coupling capacitors with small $R_s$ and $C_{sub}$ to minimise the power losses. A method to reduce $C_{sub}$ is by using capacitors with low parasitic capacitances, such as metal-insulator-metal (MIM) capacitors [116].

Equation (3.17) also shows that a higher $V_{OUT}$ can be obtained by increasing the number of rectifier stages, $N$. By multiplying the number of stages, the minimum AC voltage swing required by the rectifier input to produce a certain output can be relaxed. As $N$ increases, the DC output voltage increases until it reaches an optimal point. Adding more stages beyond the optimal point reduces the system $Q$ (due to the linearly increased parasitic capacitance), and it causes a decrease in DC output voltage [117]. Increasing $N$ also affects the input impedance of the voltage multiplier. Therefore, it is important that the number of rectifier stages to be determined through circuit simulation, and to be designed simultaneously with the impedance matching network, in order to maximise the DC output voltage, maintain a high system $Q$, and optimise the power conversion efficiency.

### 3.5.3 Voltage Regulator

Since the WSN nodes are deployed at various distances, i.e. ranging up to 5 metres from the transmitting source, the captured power of the antenna and its resulting RF input power at the power harvester circuit may vary. This will cause voltage fluctuation at the voltage multiplier’s output, thus leading to poor supply voltage stability. For this reason, a voltage regulator can be used to provide a more stable DC supply voltage to charge the energy storage element. Depending on the voltage requirement of the electrical load, two types of regulator can be considered: linear regulators and switching regulators.
A linear regulator has the simplest voltage regulation scheme. It is employed as a variable resistor, wherein it continuously adjusting a voltage divider network to maintain a constant output voltage. The linear regulator provides a regulated output which is smaller than its input voltage \( (V_{\text{OUT}} < V_{\text{IN}}) \), and necessitates a dropout between the input and output, although its value can be very small for the case of low-dropout (LDO) linear regulators [118]. On the other hand, the switching regulators can either step up (boost converter), step down (buck converter) or step-up/down (buck-boost converter) their input voltage. The output voltage is controlled by a current switch with frequency ranging from a few Hz to a few kHz. These switches are usually implemented with transistors and diodes. To maintain a fixed output voltage, duty cycle of the switch is normally adjusted through a feedback control loop. Compared to their conventional linear counterparts, the switching regulators offer better efficiency, i.e. a good design can achieve an efficiency of 90% [119]. In practice, their maximum efficiency is limited by the internal power losses which fall into a variety of categories including switching transistors, clamping diodes, filter capacitors, inductors and controller [120]. The main disadvantages of switching regulators is greater complexity (more external passive components are required in the design) and higher output voltage ripple/ noise.

In most of the RF energy harvesting applications, the power available for extraction is low, normally in the range of micro-watts. As a result, the output of the RF-DC rectifier can vary between high to low DC voltages depending on the distance between the RF energy harvester and the transmitting source. Under this circumstance, a DC-DC converter (buck, boost or buck-boost) is often used to provide a regulated output which charges the energy storage. In [121], Paing et al. presented a boost converter topology based on a resistor emulation technique, which increases the rectified output voltage of a rectenna to charge a thin-film lithium battery of 4.15V. This converter was designed such that it acts as a constant positive resistance at its input port to match the optimal load resistance of the rectenna. Nonetheless, the proposed technique requires an initial characterisation of the rectenna, i.e. to obtain its dc output power curves over a range of incident power densities and load resistances. Based on the curves, the required value of the emulated resistance is determined. The reported system had an overall efficiency of 16.7%, and it was capable of harvesting 8 µW to 420 µW from a 6 cm X 6 cm rectenna with incident RF power ranging from 30µW/ cm² to 70 µW/ cm², respectively.
3.5.4 Energy Storage

In practice, an RF energy harvesting system will not directly power the WSN nodes. This is because the peak current drawn by the nodes for certain operation may go beyond what is achievable using the harvester alone. This leads to a need for energy storage to accumulate and store the harvested energy whenever the WSN nodes are not in their active mode, i.e. sleep mode. If excessive power bursts are required by the sensor nodes, e.g. during data acquisition or transmission, they can be drawn out of this energy buffer.

Selecting the right energy storage for a specific WSN application depends on a variety of factors, including the peak power requirement of the sensor nodes, as well as the cycle life, energy storage capacity, size and cost of the energy storage component. In addition, the leakage current of the device must be taken into consideration. If the leakage current is high, more harvested energy would be required to replace the capacity lost due to leakage, rather than replenishing the storage capacity itself.

There are currently two main technologies for energy storage: supercapacitors and rechargeable batteries. A supercapacitor (also known as electrochemical double-layer capacitor) is formed by two non-reactive porous plates immersed in an electrolyte. The combination of enormous surface area from a porous carbon-based electrode material and extremely small charge separation determined by the size of the ions in the electrolyte enables them to have significant higher capacitance values compared to conventional capacitors, reaching up to several farads (in the size of an AA battery) up to a few hundreds farads (in the size of a D battery)[33]. Since the supercapacitors are based on electrostatic field generation rather than chemical reactions, their energy storage mechanism are highly irreversible, allowing the supercapacitors to be rapidly charged and discharge up to $10^6$ cycles.

In general, the supercapacitors have higher power density but lower energy density (one or two orders of magnitude) than the rechargeable batteries. This means that they cannot store as much energy, but can transfer energy out at a much greater rate than the batteries, which makes them suitable to handle short duration power surges. Lately, such capacitors have been explored for energy harvesting as they are more efficient than batteries and presents higher lifetime in terms of charge-discharge cycles. Nonetheless, they suffers from severe leakage (both intrinsic and due to parasitic paths in the external circuitry), which in turns limits their use for long-term energy storage. If the ambient RF energy is only available for a small part of the day, a supercapacitor with high self-
discharge rate may not be useful for the intended WSN application. One of the solutions is to implement a tiered energy storage mechanism [30, 33] which uses a supercapacitor and a battery by making use of the advantage of both devices. However, the tradeoff is a reduction in efficiency of the power harvester circuit due to the increased overhead of energy storage management.

On the other hand, three types of rechargeable battery can also be used for wireless sensor applications: Nickel Cadmium (NiCd), Nickel-Metal-Hydride (NiMH) and Lithium-Ion (Li-ion). The main characteristic of these three batteries are compared in Table 3.5.

<table>
<thead>
<tr>
<th></th>
<th>NiCd</th>
<th>NiMH</th>
<th>Li-Ion</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal cell voltage (V)</td>
<td>1.25</td>
<td>1.25</td>
<td>3.6</td>
</tr>
<tr>
<td>Cycle life (to 80% of initial capacity)</td>
<td>1500</td>
<td>300-500</td>
<td>500-1000</td>
</tr>
<tr>
<td>Gravimetric energy density (Wh/ kg)</td>
<td>45-80</td>
<td>60-120</td>
<td>110-160</td>
</tr>
<tr>
<td>Volumetric energy density (Wh/ l)</td>
<td>140</td>
<td>180</td>
<td>210</td>
</tr>
<tr>
<td>Self Discharge Rate/ Month (%)</td>
<td>15-20</td>
<td>20-30</td>
<td>5-10</td>
</tr>
<tr>
<td>Operating temperature (˚C)</td>
<td>-40 to 60</td>
<td>-20 to 60</td>
<td>-20 to 60</td>
</tr>
</tbody>
</table>

Table 3.5: Comparison of the most common rechargeable batteries used in energy harvesting system [122, 123].

The usage of NiCd batteries becomes less nowadays as it suffers from memory effect. The selection between NiMH and Li-ion batteries involves several tradeoffs. Generally, Li-ion batteries can store more energy compared to NiMH batteries as they offer higher energy density with respect to mass and volume. In other words, they are much lighter than other energy equivalent rechargeable batteries. They also have a higher operating voltage, typically about 3.6 V, which is three times higher than that of NiMH batteries. This makes them suitable in powering most of the WSN nodes which normally operate between 1.8 and 3.6 V. A single Li-ion cell would be sufficient to power the sensors, rather than having multiple NiMH cells. In addition, Li-ion batteries also have a longer cycle lifetime and lower self-discharge rate (5-10% per month) than most NiMH batteries (20-30% per month), with the exception for Low Self-Discharge (LSD) NiMH batteries (1.25% per month). One of the major drawbacks of Li-ion batteries is that they are typically more expensive than NiMH batteries with similar capacity. Furthermore, they present safety issues and commonly require a more complicated charging scheme.
3.6 Load (Sensor Node)

The WSN node is the end consumer of energy in an outdoor RF energy harvesting system. A generic node is comprised of four basic components as illustrated in Figure 3.8: a sensing unit, a processing unit, a transceiver unit and a power unit. The sensing unit usually consists of one or more sensors (depending on the application) and analogue-to-digital converters (ADCs). The sensors generate analogue signals based on the observed physical phenomenon. These signals are then converted into digital signals by the ADCs, and fed into the processing unit. The processing unit controls the sensor node [124], i.e. managing data acquisition, handling communication protocols, as well as scheduling and preparing data packets for transmission, so that it can work together with other nodes to implement the assigned sensing tasks. The transceiver unit connects the sensor node to the network by performing data transmission and reception over a radio channel. The power unit, which is another important component of the sensor node, is supported either by primary batteries or energy harvesters. The load design and optimisation falls outside the scope of this work, hence it will not be covered in this section.

![Figure 3.8: The components of a generic wireless sensor node. [6]](image)

In many WSN applications, it is not necessary for the nodes to be active constantly. To lower its power consumption, energy-efficient techniques such as duty-cycling can be incorporated. In duty-cycling, the node turns its radio on and off periodically (i.e. alternating between active and sleep states), thus conserving energy to extend the lifetime of WSN. A similar technique can also be achieved for the power savings of the sensing device. To size the power supply, the average current consumption \( I_{L,av} \) of the targeted sensor node must be known. This value can be obtained by using the formula below:

\[
I_{L,av} = DI_{active} + (1 - D)I_{sleep}
\]  

(3.18)
where $D$ is the duty cycle defined as $\frac{t_{\text{on}}}{t_{\text{cycle}}}$, where $t_{\text{on}}$ is the time for which the system is active (acquiring, processing, transmitting and receiving data), and $t_{\text{cycle}}$ is the period of one operating cycle ($t_{\text{on}} + t_{\text{off}}$).

In this work, a wireless ground-level sensor node: Zyrox2 Bait Station [19] is used as the exemplar load of the feasibility study. The energy consumption profile of the node at its various operating states is shown in Table 3.6.

<table>
<thead>
<tr>
<th>Operation State</th>
<th>Duration (s)</th>
<th>Current Drawn</th>
<th>Average Current (µA/ %)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wake-on Radio</td>
<td>120 (2 minutes)</td>
<td>6 mA</td>
<td>11.0 7.3</td>
</tr>
<tr>
<td>Sensing (with LED on)</td>
<td>10</td>
<td>35 mA</td>
<td>5.4 3.6</td>
</tr>
<tr>
<td>Transmission (Tx/ Rx)</td>
<td>360 (6 minutes)</td>
<td>24 mA</td>
<td>132.3 88.4</td>
</tr>
<tr>
<td>Sleep</td>
<td>64800 (18 hours)</td>
<td>1 µA</td>
<td>1.0 0.7</td>
</tr>
<tr>
<td>Total</td>
<td>65290</td>
<td>-</td>
<td>149.7 100</td>
</tr>
</tbody>
</table>

Table 3.6: Energy consumption profile of the Zyrox2 Bait Station [125].

As seen from Table 3.6, the Zyrox2 Bait Station has four functional states in one operating cycle: communication module wake-up, sensor data acquisition, data transmission and sleeping states. The sensor node undertakes a full operation of data acquisition, processing and transmission in about 8 minutes, followed by a deep sleep cycle lasting for 18 hours. According to the designer [125], the wireless communication of Zyrox2 Bait Station was devised to take place every 18 hours (instead of 24 hours) to avoid possible interference from the electrical activity which is synchronous with the 24 hour clock. Spreading the time at which the node to communicate (randomly or systematically) could counter the potential risk of being blocked by the same interferer. The current drawn shown in Table 3.6 was measured when the device operating at 3.0 V.

By using Equation (3.18), the value of $I_{L,av}$ can be determined and the result was 149.7 µA. Given that the sensor node is operated at 3.0 V, its average power consumption will be 450 µW. It is interesting to note that, the node consumes the most energy during its communicating state. Therefore, minimising the amount of time for data transmission is essential to conserve energy. It is also possible to reduce the Zyrox2 Bait Station’s
average power consumption by optimising the relative amount of time it spent in low-power sleeping mode. Table 3.7 presents the calculated data for a range of cycle times from 18 hours (worst case scenario) to 96 hours in order to illustrate the impact of the duty cycle-driven operation. As shown, a dramatic reduction in average power consumption of the sensor occurs as the sleeping time is extended from 18 hours to 96 hours.

<table>
<thead>
<tr>
<th>Sleep Time (hours)</th>
<th>Duty Cycle (%)</th>
<th>Total energy per cycle (J)</th>
<th>Average power consumption (µW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>18</td>
<td>0.75</td>
<td>29.33</td>
<td>450</td>
</tr>
<tr>
<td>24</td>
<td>0.56</td>
<td>29.39</td>
<td>340</td>
</tr>
<tr>
<td>48</td>
<td>0.28</td>
<td>29.65</td>
<td>172</td>
</tr>
<tr>
<td>72</td>
<td>0.19</td>
<td>29.90</td>
<td>115</td>
</tr>
<tr>
<td>96</td>
<td>0.14</td>
<td>30.17</td>
<td>87</td>
</tr>
</tbody>
</table>

Table 3.7: Calculated energy budget and average power consumption of Zyrox2 Bait Station for sleep times from 18 hours to 96 hours.

3.7 External Environment

The external environment of an RF energy harvesting system (between the transmitter and the receiving nodes) is another important factor. Depending on the characteristic of the propagation environment, the transmitted waves could be subjected to three basic propagation mechanisms: reflection, scattering or diffraction. A number of propagation models were considered to estimate the path loss of the investigated system. The simplest is the free space path loss (FSPL) model which only considers a clear, unobstructed LOS path between the transmitter and receiver. The FSPL model is rather optimistic and should only be regarded as theoretical maximum. This model poorly reflects on most practical situations where the actual environmental conditions could affect the wave propagation characteristics. As mentioned in Chapter 1, the wireless ground-level sensor nodes envisaged in this work are located in the vicinity of domestic, agricultural or commercial buildings. These buildings could be surrounded by large garden, farm or open field areas with possible presence of light to heavy foliage. Therefore, an additional propagation model which may be applicable to the investigated system has been examined: the vegetation loss model.
3.7.1 Vegetation Loss Model

The presence of randomly distributed shrubs, leaves and crop canopies in the propagation environment can cause attenuation of the radiated waves. Numerous studies have been conducted to characterise and model the effects of vegetation experimentally. They have been reviewed and summarised into a few better-known empirical models, such as Weissberger’s Modified Exponential Decay (MED) model [126], ITU Recommendation (ITU-R) [95] and its Derivative, the COST 235 model [95]. The main advantage of these empirical models lies on the simplicity of their final mathematical expressions describing them. However, these models are only valid for environments similar to those used to derive the expressions, as they are developed from specific measured data sets. On the other hand, analytical models provide a better insight into physical processes involved in the radio wave propagation through vegetation. Nonetheless, they require the use of computation-intensive numerical methods to provide solutions to the mathematical formulations.

To simplify the analysis of the investigated system, the Weissberger’s model [126] was used to provide a rough estimation of the additional loss due to foliages. This model is based on sets of measured data from 230 MHz to 95 GHz, and it is given by:

\[
L(dB) = \begin{cases} 
1.33f^{0.284}d_r^{0.588}, & 14 \leq d_r \leq 400m \\
0.45f^{0.284}d_r, & 0 \leq d_r \leq 14m
\end{cases}
\]  

(3.19)

where \( L \) is the losses in dB, \( f \) is the frequency in GHz, and \( d_r \) is the depth of foliages in metres. It applies in areas where a ray path is blocked by dense, dry, in-leaf trees in temperate climate. In this model, the propagation is assumed to take place through a grove of trees rather than by diffraction over the top of the trees. As seen from Equation (3.19), the amount of foliage loss increases with foliage depth and higher frequency. At 867 MHz, the vegetation loss is estimated at 0.43 dB and 2.16 dB for foliage depths of 1 and 5 metres, respectively. These estimations will be used in the link budget calculation presented in the next section.

Note that, the Weissberger’s model is developed from the measured data of trees ranging in height from 9 to 15 metres. In contrast, the vegetation types expected for the investigated system are mainly shrubs or crop canopies with much shorter height (maximum 2-3 metres). Therefore, this model should only serve as an approximation. To get a better prediction for the actual foliage attenuation, the empirical model must
incorporate other relevant factors, such as foliage types, density of foliages, and moisture content of leaves in the canopy. However, the addition of these factors into such a model cannot be readily achieved as further research supported by measured data is required.

### 3.7.2 Attenuation due to Hydrometeors

The absorption and scattering by hydrometeors, such as rain and snow, can also cause signal attenuation, depending upon the frequency, the link distance, and the geographical location [127]. Studies showed that the specific attenuation (usually expressed in dB/km) caused by falling raindrops starts to become noticeable only at frequencies above 10 GHz [128]. At much lower frequencies, the rainfall and snow attenuation effect is negligible. Given that the investigated system operating around 900 MHz (which is much lower than 10 GHz), the attenuation due to hydrometeors will not be considered in this analysis.

### 3.8 Link Budget Analysis

A link budget analysis can be performed on an outdoor RF energy harvesting system to estimate the power which is available at the receiving nodes. This analysis accounts for the transmitted power, antenna gains, and all the losses in the link prior to the receiver. The path loss is the most significant factor due to its magnitude compared to other terms. It includes free-space loss (FSL) as well as the losses from environmental factors such as foliage and weather conditions. Any waveguide or cable used to connect to the antennas is another loss contributor. Therefore, it is essential to make sure that the waveguides be intact and fitted properly, and that coaxial cables be in good condition. For point-to-point links, directional antennas are typically used to increase the energy directed towards the receivers. In this case, there will be some pointing losses associated with the transmitting and receiving antennas if they are not perfectly aligned. Depending upon the application, one dB or two may be allocated to this pointing loss in a link budget. In addition, the loss due to a mismatch in the transmitting and receiving antenna’s polarisation vector alignment will also reduce the power transfer. This loss is less important for a fixed, point-to-point link, but it must be considered for mobile applications. A standard link budget equation can be expressed in terms of dB as:

\[
P_{RX} = P_{TX} + G_{TX} - L_{TX} - L_{FS} - L_{M} + G_{RX} - L_{RX}
\]  
(3.20)

where \(P_{RX} = \) received power, \(P_{TX} = \) transmitter power, \(G_{TX} = \) transmitter antenna gain, \(L_{TX} = \) transmitter losses (coaxial cables, connectors, etc), \(L_{FS} = \) free space loss,
\( L_M = \text{miscellaneous losses (mismatch loss, pointing error, foliage loss)}, \ G_{RX} = \text{receiver antenna gain and} \ L_{RX} = \text{receiver losses (coaxial cables, connectors, etc)}. \)

### 3.8.1 Link Budget Calculation

In order to investigate the feasibility of the proposed system, a link budget calculation was performed by using the Zyrox2 Bait Station as an exemplar. This calculation was made based on the following conditions:

- The RF energy harvesting system operates at 867 MHz with an EIRP of 3.28 W (\( P_{TX} + G_{TX} = 35.15 \text{ dBm} \)).
- The link distance is between 1 to 5 metres.
- The required \( P_{RX} \) of the intended application is 450 \( \mu \text{W} \) (-3.46 dBm).
- It is assumed that the only loss in the link is the free space loss.

Figure 3.9 shows the required \( G_{RX} \) to form a feasible link in free space (up to 5 metres). As seen from Figure, the receiver has to be designed at a minimum gain of 6.57 dBi in order to obtain a received power of -3.46 dBm. Note that the free space model assumes idealised condition and it did not take into account for other possible losses due to foliages, polarisation mismatch, pointing error, cables/ connectors, etc.

![Figure 3.9: Required \( G_{RX} \) for the link in free space.](image)

For a more practical scenario, a second link budget calculation was performed based on the following agricultural setting:
- The sensor nodes are placed within the shrubberies or crop canopies within a dense foliage depth of 5 metres. The Weissberger’s Model discussed in Section 3.7.1 was used to estimate the foliage loss.
- Directional antennas are used for both transmitters and receivers. A pointing error loss of 0.5 dB and polarisation mismatch loss of 0.2 dB are assumed.
- The receiver and transmitter losses (e.g. connectors, cables) are assumed to 0.5 dB each.

In RF energy harvesting, an additional loss factor – the power loss due to RF-DC conversion of the energy harvesting circuit ($L_{RFDC}$) must also be added into the link budget equation. For this analysis, a commercially-available component (P1110) [129] from Powercast™ is used as the power harvester circuit of the system. Figure 3.10 shows the P1110’s efficiency curve versus input power at 3.0 V, extracted from its datasheet.

As seen from Figure 3.10, when the P1110 operates at 868 MHz, it has an efficiency of about 50% when the input power is 0 dBm (1 mW). This means that the expected

![Figure 3.10: P1110’s efficiency versus frequency [129].](image-url)
harvested power from the circuit will be -3 dBm (500 µW), which is close to the power requirement of the exemplar node (450 µW). Therefore, the $L_{RFDC}$ value for this calculation was chosen as 3 dB. Note that the $L_{RFDC}$ will be different for other power harvester circuit, since the efficiency varies with different input power, frequency and load characteristics [106, 130]. Figure 3.11 shows the required $G_{RX}$ to form a reliable link in powering the Zyrox2 in a typical agricultural setting with the inclusion of all relevant losses.

![Figure 3.11: Required $G_{RX}$ for the link in a typical agricultural surrounding, with $d_f$ = foliage depth.](image)

Clearly, the required $G_{RX}$ at 5 metres has now increased to 13.43 dBi. Achieving such high antenna gain requires large antenna apertures (e.g. parabolic dishes, horn antenna or antenna arrays). In some cases, the expense of such large apertures is excessive, making them almost impractical for the intended WSN application. Therefore, for a more practical implementation (i.e. in terms of size and cost), RF energy harvesting should only be considered for ultra-low power applications, such as sensor-enhanced RFID tags [79, 131] or WSN nodes with extremely low duty cycle. These applications would have much lower power consumption of less than 100 µW. In order to implement an RF energy harvesting system for the Zyrox2 Bait Station, its power consumption level must first be lowered. This could possibly be achieved by operating the node at lower duty cycle (e.g. once a week instead of every 18 hours). For instance, the average power consumption of the Zyrox2 Bait Station can be significantly reduced to 87 µW (see Table 3.7) if it operates once every 96 hours (4 days).
4 Antenna /Circuit Design and Testing

4.1 Introduction

This chapter provides the design of the receiving antenna and the power harvester circuit which could be potentially used in an outdoor RF energy harvesting system. First, a generic design process for the antenna is outlined, followed by a discussion of several key antenna design requirements for the intended WSN application. The basic concept of a microstrip patch antenna and some of its miniaturisation techniques are also discussed. Then, two compact patch antenna designs are presented. The first is an air-substrate-based, folded shorted patch antenna (FSPA) with a small, solid ground plane, whereas the second is a modified FSPA structure with four pairs of slot embedded into its ground plane. The antennas are simulated, fabricated and tested in an anechoic chamber, as well as in an outdoor field. A power harvester circuit, operating in the 860 to 960 MHz industrial, scientific and medical (ISM) frequency range is also built using a commercially manufactured integrated circuit component from Powercast™. The performance of the power harvester circuit is evaluated with an RF signal generator source inside the laboratory.

4.2 Antenna Design Process

A typical design process of the antenna can be visualised with the following flow-chart shown in Figure 4.1. This process begins by identifying the intended application and establishing a set of design requirements for the antenna. Based on the given requirements, an appropriate antenna structure along with the materials which can be used for its construction are determined. The antenna is then modelled, simulated and optimised using CST Microwave Studio until all design requirements are met in the simulations. Finally, the antenna prototype is fabricated and tested, and then compared to the simulation results. The design can only be considered ready if the requirements are met in the testing stage. Otherwise, it has to be modified and optimised until all requirements are satisfied.
4.3 Antenna Design Requirements

As mentioned in Chapter 1, the intended application of this research is a generic class of wireless ground-level sensor network deployed in an outdoor environment. Several key design requirements of the energy harvesting antenna are summarised below:

(i) Frequency of Operation/ Impedance Bandwidth

As stated in Section 3.3.1, a UHF, ISM band (867/ 915/ 953 MHz) was chosen as the operating frequency of the investigated system. Note that the frequency allocation for the selected ISM band varies in different countries. Therefore, for the system to be used worldwide, the antenna must be designed to operate over a wide frequency range from 860 to 960 MHz, with an impedance bandwidth ($S_{11} < -10$ dB) covering the selected frequency band.

(ii) Overall Antenna Size

For this particular application, a specific requirement has been given by the industrial sponsor (Syngenta) on the antenna’s height. The height must not exceed 20 mm above
the soil surface. Although there is no further limitation imposed on the antenna’s lateral dimension (i.e. width and length), it is desirable to keep the overall antenna size as small as possible so that it can be conveniently attached to the wireless ground-level sensor nodes. Note that this overall size must also include the antenna’s ground plane.

(iii) Gain

The link budget analysis in Chapter 3 showed that the exemplar WSN node (Zyrox2 Bait Station) is a power-hungry device. In order to power that device by RF energy harvesting, a high-gain receiving antenna (about 13 dBi) must be used. In this design, a minimum gain of 3 dBi was decided upon for the antenna so that it can still be useful for other wireless ground-level sensor nodes or RFID applications with much lower power consumption.

(iv) Manufacturing Cost

The wireless ground-level sensor network comprises of several sensor nodes distributed across a field. Each sensor node will be embedded with its own energy harvesting antenna. Hence, to make this system practical, the antenna has to be a low-cost device. The antenna structure and the choice of materials for constructing it must be chosen appropriately to maintain a reasonable manufacturing cost.

4.4 Microstrip Patch Antenna

For the investigated system, it was decided to use a microstrip patch antenna (MPA) as the receiving antenna primarily due to its low profile, low cost and ease of fabrication. In the following sections, the basic concepts of a MPA, and its size miniaturisation techniques are discussed.

In its simplest form, a MPA consists of two conducting surfaces: a radiating patch, and a ground plane, separated by a layer of dielectric substrate. A simplified MPA is shown in Figure 4.2. The very thin radiating patch \( t \ll \lambda_o \) is typically made of conducting material such as gold or copper, and it can take any possible shape. Square, rectangular, and circular are the most common shapes due to the ease of analysis, and performance prediction. For a conventional rectangular patch, the length \( L \) of the patch is usually \( \lambda_o/3 < L < \lambda_o/2 \). Radiating elements and feed lines are generally photo-etched on the dielectric substrate. The height of the dielectric substrate is usually \( 0.003 \lambda_o < h < 0.05 \lambda_o \) with a dielectric constant of \( 2.2 < \varepsilon_r < 12 \). To achieve good antenna performance, thicker substrates with lower dielectric constant are desired since they provide better efficiency, larger bandwidth and better radiation characteristics; but at
the expense of larger antenna sizes [65]. Conversely, thin substrates with higher dielectric constant are used to design smaller MPAs. However, they are less efficient and have narrower bandwidths [65].

![Structure of a microstrip patch antenna](image)

**Figure 4.2:** Structure of a microstrip patch antenna [132].

Table 4.1 summarises the major advantages and disadvantages of a MPA, in comparison to conventional antennas.

<table>
<thead>
<tr>
<th>Advantage</th>
<th>Disadvantage</th>
</tr>
</thead>
<tbody>
<tr>
<td>• Low profile, conformable to planar surfaces.</td>
<td>• Low gain, low efficiency.</td>
</tr>
<tr>
<td>• Light weight, mechanically robust when mounted on rigid surface.</td>
<td>• Low power handling capacity.</td>
</tr>
<tr>
<td>• Low fabrication cost, ease of manufacture.</td>
<td>• Narrow frequency bandwidth.</td>
</tr>
<tr>
<td>• Capable of dual or triple frequency band operation.</td>
<td>• Spurious feed radiation.</td>
</tr>
<tr>
<td>• Supports linear and circular polarisation.</td>
<td>• Surface wave excitation within substrates.</td>
</tr>
</tbody>
</table>

**Table 4.1:** Advantages and disadvantages of MPAs.

### 4.5 Antenna Size Miniaturisation Techniques

To ensure the practicability of the system, lower cost material must be used when constructing the energy harvesting antenna. An alternative is to use air as the antenna substrate. Compared to that with other dielectric substrates, air-substrate-based MPAs have better radiation efficiency and enhanced operating bandwidth [133]. Moreover, the cost of manufacturing these antennas is much lower, since they can be easily built using just a conducting metal sheet such as copper or aluminum. Nonetheless, there is a major drawback associated with air-substrate-based antennas. At 860 to 960 MHz, a
conventional, half-wavelength, air-substrate-based MPA has a typical resonant length of about 17 cm. On top of that, the ground plane of the MPA is usually larger than the resonating patch. This would make the overall size of the antenna prohibitively large to be mounted on the sensor nodes. Due to this consideration, the size of a conventional air-substrate-based MPA must be reduced before it can be used for the intended application.

Several miniaturisation techniques for MPAs have been proposed in the literature, including loading the antenna with high-dielectric materials, using shorting elements to ground plane, and incorporating slots or notches on the patch shape. These techniques have been widely used in mobile-communication industries, where the most successful results were achieved by combining several of them for the design of one antenna. However, it should be noted that miniaturising an antenna could affect its radiation characteristics, e.g. gain, bandwidth and efficiency. Hence, a trade-off must be made between size and performance when designing a compact patch antenna.

4.5.1 Use of Higher Dielectric Constant Substrate

The simplest technique to reduce the size of a MPA is to use higher dielectric constant substrates. Via this technique, the guided wavelength below the patch can be reduced, thereby decreasing the size of the resonating patch [134]. In general, a conventional MPA is a half-wavelength structure and its resonant length in fundamental mode is given by [133]:

\[
L \approx \frac{1}{2} \left( \frac{c}{f \sqrt{\varepsilon_r}} \right)
\]  

(4.1)

where 
\( c = \) speed of light (m/s)  
\( L = \) resonant length of the patch (m)  
\( f = \) frequency (Hz)  
\( \varepsilon_r = \) dielectric constant of the substrate

Equation (4.1) shows that the radiating patch of the MPA has a resonant length \( L \) which is proportional to \( 1/\sqrt{\varepsilon_r} \). Thus, the use of substrate material with higher dielectric constant can result in a smaller physical antenna length at a fixed operating frequency. Although this technique looks fairly simple, it has some disadvantages. As the dielectric constant of the substrate increases, the antenna’s radiation efficiency decreases due to surface wave excitation [135]. Moreover, the antenna’s bandwidth also reduces with the use of higher dielectric constant substrate.
4.5.2 Use of Shorting Elements

Another common technique is to use shorting elements such as shorting walls or shorting pins. At fundamental resonant mode, the electric field is zero (virtual shorting plane) at the centre of a half-wavelength MPA. Hence, a physical shorting wall can be used to reduce the antenna’s physical length by half at a fixed resonant frequency. This type of antenna is normally known as a quarter-wavelength patch antenna. The size of a shorted patch antenna can be further reduced by folding the antenna structure [136, 137]. As shown in Figure 4.3, a quarter wavelength wall-shorted rectangular patch is folded and the antenna length can be reduced up to $\lambda_0/8$ [137].

![Figure 4.3: (a) Conventional half-wavelength antenna; (b) Quarter-wavelength antenna; (c) and (d) Folded quarter-wave length antenna. [137]](image)

Besides the shorting walls, shorting pins or posts can also be used to reduce the size of the MPA. By shifting the null voltage point from the patch centre to their respective edges, the shorted patch can resonate at a much lower frequency. In [138, 139], a shorting pin is incorporated near the feed of a circular microstrip patch antenna. The overall size of this antenna was reduced to one third of a conventional circular patch, but at the expenses of reduced bandwidth and lower gain due to the reduced patch area for radiation.

4.5.3 Use of Slots or Notches on Patch Shape or Ground Plane

The use of slots or notches on the resonating patch can also reduce the antenna size. For the case of a rectangular patch, several narrow slots can be inserted at the non-radiating edges of the patch. An example of the modified MPA leading to a smaller size is illustrated in Figure 4.4. It can be seen that the slots or notches forces the currents to meander, which artificially increases the antenna electrical length for a fixed patch linear dimension. This behaviour results in a lowered fundamental resonant frequency, and thus achieving a
size reduction for the antennas operating at a fixed frequency. Depending upon the shape or size of the insertions, the overall antenna size can be reduced by up to 50% [140, 141].

Figure 4.4: The effect of slots and notches in a microstrip patch antenna for size reduction: (a) conventional microstrip patch antenna (b) microstrip patch antennas with slots and notches [142].

The slot technique can also be applied to the antenna’s ground plane. When the proper slots are incorporated into the ground plane of a MPA, a lowering of the antenna’s fundamental resonant frequency can be achieved [143-145]. In [144], a pair of narrow slots are placed along the center axis of the ground plane perpendicular to the antenna’s resonant direction. Figure 4.5 shows the simulated surface current distributions in the ground plane and radiating patch of the proposed antenna in [144]. As seen, the embedded slots effectively meander the excited surface currents in the ground plane, which in turns causes meandering of the surface currents on the radiating patch and lengthening of the equivalent surface current path.
4.6 Folded Shorted Patch Antenna

In this section, two air-substrate-based folded shorted patch antennas are presented, the first with a solid ground plane (FSPA-GND), and the second with a slotted ground plane (FSPA-SG). Both FSPAs achieve size miniaturisation based on two of the techniques discussed above: use of shorting wall and incorporation of slots into the antenna structure. The two designs are based on a low profile, single-band (1.48 to 1.99 GHz) FSPA which was first proposed in [146] for an indoor base station that is required to service several wireless communication systems by a single compact antenna. This work aims to perform further investigation on the suitability of the proposed FSPA for a different type of application: outdoor RF energy harvesting in powering a wireless ground-level sensor network.

4.6.1 FSPA with Solid Ground Plane

The geometry of the FSPA with solid ground plane (FSPA-GND) is shown in Figure 4.6. A rectangular patch is located in the middle of a square ground plane. Both patch and ground plane are made of copper sheet, which has a thickness of 1 mm. For the patch,
one of its ends is shorted to ground while the other end is folded backwards. The patch structure is separated from the ground plane by an air layer. A coaxial feed and shorting post with the same diameter are positioned along the x-axis, which is the midline of the patch. In [146], the FSPA was designed to cover frequencies between 1.48 and 1.99 GHz. Here, the patch dimension of the FSPA-GND was modified so that it can operate at the desired frequency range from 860 to 960 MHz. Besides, it employed a much smaller ground plane. Assuming $\lambda_0$ is the free space wavelength of the centre frequency at 910 MHz, the FSPA-GND has a ground plane dimension of $160 \text{ mm (0.48} \lambda_0) \times 160 \text{ mm (0.48} \lambda_0)$, which is 70% less area than that of the original antenna proposed in [146]. Detailed dimension of the FSPA-GND is provided in Table 4.2.

Figure 4.6: Geometry of the FSPA-GND.
Table 4.2: Antenna dimension of the FSPA-GND.

<table>
<thead>
<tr>
<th>Dimension</th>
<th>$H_1$</th>
<th>$H_2$</th>
<th>$W$</th>
<th>$L_1$</th>
<th>$L_2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>(mm)</td>
<td>8.0</td>
<td>8.0</td>
<td>125.0</td>
<td>90.0</td>
<td>72.0</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Dimension</th>
<th>$S_{\text{post}}$</th>
<th>$S_{\text{feed}}$</th>
<th>$S_{\text{edge}}$</th>
<th>$d_1$</th>
<th>$D_2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>(mm)</td>
<td>47.5</td>
<td>7.5</td>
<td>35.0</td>
<td>1.3</td>
<td>1.3</td>
</tr>
</tbody>
</table>

4.6.2 FSPA with Slotted Ground Plane

As pointed out in Section 4.3, it is desirable to have a small receiving antenna for the intended application. As shown in Figure 4.6, the overall size of the FSPA-GND is largely dominated by its ground plane dimension. Therefore, one of the techniques discussed in Section 4.5 - embedding slots in the ground plane - was implemented on the FSPA-GND with the aim of reducing its overall size.

The geometry of the FSPA with slotted ground (FSPA-SLG) is almost identical to the one shown in Figure 4.6 with the exception of its slotted ground plane. Four pairs of equally spaced slots were embedded into the ground plane, as illustrated in Figure 4.7. Each slot has the same width ($W_s$) and length ($L_s$). The slot pairs are spaced by the same distance ($S_1$). The FSPA-SLG has a ground plane dimension of $135 \times 135$ mm, which is smaller than that of FSPA-GND. Detailed dimension of the antenna is given in Table 4.3.

Figure 4.7: Geometry of the slotted ground plane of FSPA-SLG.
4.6.3 Results and Discussion

The antennas’ performance was first simulated with the aid of a commercial software package: CST Microwave Studio. In order to validate the simulation results, both antennas were fabricated and then tested in the SMART chamber of the National Physical Laboratory (NPL) at Teddington. The return loss and radiation patterns were measured using a HP 8510 Vector Network Analyser with proper calibration. The Antenna under Test (AUT), which is the FSPA-GND and FSPA-SLG, was illuminated using an EMCO 3147 Log Periodic Antenna as the transmitting antenna at a separation distance of 2.7 metres to guarantee far-field measurements. The AUT was rotated on an Orbit Powercell roll-over-azimuth positioner, and the radiation pattern was measured over the surface of a sphere with roll and azimuth increments of 5 degrees. The positioner, data acquisition and post processing were controlled by the Orbit MIDAS Far Field Antenna Measurement and Analysis software. The setups of the return loss and radiation pattern measurements are shown in Figure 4.8 and Figure 4.9, respectively.
Return Loss

The simulated and measured return loss of the FSPA-GND is shown in Figure 4.10 (a). A 10 dB return loss bandwidths of 20.0% (0.815 to 0.997 GHz) is observed, with a centre frequency of 906 MHz. As seen from the Figure, the antenna has two resonances, a lower resonance at 840 MHz and an upper resonance at 956 MHz. Because these two resonances are close to each other, the antenna obtains a wide impedance bandwidth. The measured and simulated return loss of the second antenna: FSPA-SLG is shown in Figure 4.10 (b), and it has a measured impedance bandwidth of 18.7% (0.834 to 1.006 GHz), with a centre frequency of 920 MHz.
Radiation Pattern and Gain

The co-polar x-z plane radiation patterns of the FSPA-GND and FSPA-SLG at three operating frequencies: 867 MHz, 915 MHz, and 953 MHz, are shown in Figure 4.11 (a)-(c) and Figure 4.12 (a)-(c), respectively. For FSPA-GND, the main beams are tilted at -35° at 867 MHz, -18° at 915 MHz, and -10° at 953 MHz. The minus sign indicates the angle tilted from the broadside direction of the antenna. As for FSPA-SLG, the main beams are tilted at -24° at 867 MHz, -13° at 915 MHz, and -9° at 953 MHz. Note that the tilted angles are reduced with higher frequencies, and the main beams in the x-z plane are squinted to the
left rather than the right since more currents are flowing on the left vertical wall (folded side) than the right one. The co-polar y-z plane radiation patterns of the FSPA-GND and FSPA-SLG are also shown in Figure 4.11 (d)-(f) and Figure 4.12 (d)-(f). In the y-z plane, the main beams at these three frequencies are found to be in the broadside direction of the antenna. The measured peak gains of the FSPA-GND at 867 MHz, 915 MHz, and 953 MHz are 3.9 dBi, 5.9 dBi, and 6.3 dBi, while those of FSPA-SLG are 3.7 dBi, 5.4 dBi, and 5.8 dBi, respectively. These peak gain values lie in the x-z plane of the antennas.
Figure 4.11: (left): Measured and simulated co-polar x-z plane radiation pattern of the FSPA-GND at (a) 867 MHz, (b) 915 MHz, and (c) 953 MHz; (right): co-polar y-z plane radiation pattern of the FSPA-GND at (d) 867 MHz, (e) 915 MHz, and (f) 953 MHz.
Figure 4.12: (left): Measured and simulated co-polar x-z plane radiation patterns of the FSPA-SLG at (a) 867 MHz, (b) 915 MHz, and (c) 953 MHz; (right): co-polar y-z plane radiation pattern of the FSPA-SLG at (d) 867 MHz, (e) 915 MHz, and (f) 953 MHz

On the whole, the measured return loss and radiation patterns have good agreements with the simulations. The slight discrepancies between them are most likely due to the imperfection in the fabrication. By looking at the measurement results, the FSPA-GND and FSPA-SLG are capable of covering the required operating bandwidth range from 860 to 960 MHz. Moreover, they met the minimum gain requirement of 3 dBi as specified in Section 4.3. Both antennas have a comparable impedance bandwidth performance, but with the FSPA-SLG having a smaller patch and ground plane dimension than that of the FSPA-GND. With the slots properly embedded in the ground plane of the FSPA, an overall size reduction of 29% (from 256 cm² for FSPA-GND to 182.25 cm² for FSPA-SLG) can be achieved.

To thoroughly understand the effect of the slots, a simulation was performed on the FSPA-SLG. The slotted ground plane was removed and replaced by a solid ground plane of 135 mm × 135 mm (same area as the slotted one). Figure 4.13(a)-(b) shows the simulated input impedance against frequency curves of the FSPA-SLG with solid and slotted ground plane. The thick black lines in both figures indicate a perfect matching at 50 Ω with zero input reactance. As shown from the Figure, the input reactance of the antenna with solid ground plane, at frequency band of interest (860 to 960 MHz), is dominated by the inductive reactance of the probe feed and shorting pin. By incorporating slot pairs into the ground plane, additional capacitive reactance can be introduced to counteract the inductive reactance, thereby providing better impedance matching across the operating band. Figure 4.14 compares the simulated return losses of the two antenna prototypes. As seen,
the FSPA with ground plane has an impedance bandwidth of 18.4%, while the FSPA with solid ground plane does not even match to 50 Ω.

![Input Impedance Graph](image)

(a) Simulated input impedance of the FSPA-SLG with solid and slotted ground plane: (a) resistance and (b) reactance.

(b) Simulated return losses of FSPA with and without slotted ground plane.

Figure 4.13: Simulated input impedance of the FSPA-SLG with solid and slotted ground plane: (a) resistance and (b) reactance.

Figure 4.14: Simulated return losses of FSPA with and without slotted ground plane.

### 4.6.4 Parametric Analysis of FSPA-SLG

A simulation-based parametric study was performed using CST Microwave Studio to examine the effects of different design parameters on the impedance bandwidth and radiation patterns of the FSPA-SLG. The simulations were carried out by varying the parameter of interest, whilst keeping all other parameters constant. The initial dimensions
of these design parameters are shown in Table 4.3.

First, three parameters characterising the slots of the ground plane - number of slot pairs, slot length, and slot width, were investigated. The effects of slot pairs (labeled from A to E shown in Figure 4.15) on the antenna’s impedance bandwidth are shown in Figure 4.16. As the number of slot pairs increases, the FSPA-SG obtains wider impedance bandwidth. The widest bandwidth is achieved when four pairs of slot (A, B, C and D) are used. However, the inclusion of Pair E does not further improve the antenna’s bandwidth performance. Figure 4.17 and Figure 4.18 illustrate the effect of varying the length (Ls) and width (Ws) of the slot. In most cases, the impedance bandwidth becomes wider with the increments of slot’s length and width. However, no further improvement was found when Ls is more than 58 mm, and Ws is more than 4 mm. In addition, changing the slot’s dimension of the ground plane has no significant impact on the radiation patterns of the FSPA-SLG. Only a small variation in directivity (± 0.2 dB) is observed when varying the slot parameters, i.e. number of slot pairs, Ls, and Ws.

![Figure 4.15: Slot pair A to E in the ground plane.](image-url)
Figure 4.16: Simulated return losses for different number pair of slots (where $L_s = 58$ mm and $W_s = 6$ mm)

Figure 4.17: Simulated return losses for various slot length, $L_s$. 
Three other design parameters related to the patch structure were also investigated: patch width (W), folded patch length (L₂) and height ratio (H₁/H₂). The effects of each parameter on the FSPA-SLG's impedance bandwidth and radiation patterns are displayed in Figure 4.19, Table 4.4 and Table 4.5 respectively. Wider impedance bandwidth can be obtained by increasing W, L₂ and H₁/H₂ up to a certain value, i.e. W = 90 mm, L₂ = 72mm, and H₁/H₂ = 8mm/8mm. The bandwidths start to decrease beyond these values. In addition, it is found that L₂ and H₁/H₂ are the two design parameters which could affect the directivity of the FSPA-SLG. An increase in L₂ or H₁/H₂ results in higher antenna directivity, but with a smaller tilted angle of the main lobe from the broadside direction. Therefore, a tradeoff must be made when selecting these parameters for the FSPA-SLG. Changing W has no significant outcome on the antenna radiation pattern as its maximum variation in directivity is only ±0.1 dB.
Figure 4.19: Simulated return losses for various patch width, $W$.

<table>
<thead>
<tr>
<th>Folded Patch Length, $L_2$</th>
<th>Return Loss</th>
<th>Radiation Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Impedance Bandwidth ($S_{11} &lt; -10$ dB)</td>
<td>Operating Frequency (MHz)</td>
</tr>
<tr>
<td>68 mm</td>
<td>0.888 – 1.040 GHz (15.8%)</td>
<td>867</td>
</tr>
<tr>
<td></td>
<td>915</td>
<td>24°</td>
</tr>
<tr>
<td></td>
<td>953</td>
<td>15°</td>
</tr>
<tr>
<td>70 mm</td>
<td>0.858 – 1.023 GHz (17.5%)</td>
<td>867</td>
</tr>
<tr>
<td></td>
<td>915</td>
<td>17°</td>
</tr>
<tr>
<td></td>
<td>953</td>
<td>12°</td>
</tr>
<tr>
<td>72 mm</td>
<td>0.835 – 1.003 GHz (18.3%)</td>
<td>867</td>
</tr>
<tr>
<td></td>
<td>915</td>
<td>13°</td>
</tr>
<tr>
<td></td>
<td>953</td>
<td>6°</td>
</tr>
<tr>
<td>74 mm</td>
<td>0.816 – 0.980 GHz (18.2%)</td>
<td>867</td>
</tr>
<tr>
<td></td>
<td>915</td>
<td>10°</td>
</tr>
<tr>
<td></td>
<td>953</td>
<td>7°</td>
</tr>
<tr>
<td>76 mm</td>
<td>0.799 – 0.925 GHz (14.6%)</td>
<td>867</td>
</tr>
<tr>
<td></td>
<td>915</td>
<td>8°</td>
</tr>
<tr>
<td></td>
<td>953</td>
<td>6°</td>
</tr>
</tbody>
</table>

Table 4.4: Simulated antenna performance for various $L_2$. 
### Table 4.5: Simulated antenna performance for various H₁/H₂.

<table>
<thead>
<tr>
<th>Antenna Height Ratio (H₁/H₂)</th>
<th>Return Loss</th>
<th>Radiation Pattern</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Impedance Bandwidth (S₁₁ &lt;-10 dB)</td>
<td>Operating Frequency (MHz)</td>
</tr>
<tr>
<td>12 mm/ 4 mm</td>
<td>0.816 – 0.963 GHz (16.5%)</td>
<td>867</td>
</tr>
<tr>
<td></td>
<td></td>
<td>915</td>
</tr>
<tr>
<td></td>
<td></td>
<td>953</td>
</tr>
<tr>
<td>10 mm/ 6 mm</td>
<td>0.835 – 1.003 GHz (18.2%)</td>
<td>867</td>
</tr>
<tr>
<td></td>
<td></td>
<td>915</td>
</tr>
<tr>
<td></td>
<td></td>
<td>953</td>
</tr>
<tr>
<td>8 mm/ 8 mm</td>
<td>0.856 – 1.041 GHz (19.5%)</td>
<td>867</td>
</tr>
<tr>
<td></td>
<td></td>
<td>915</td>
</tr>
<tr>
<td></td>
<td></td>
<td>953</td>
</tr>
<tr>
<td>6 mm/ 10 mm</td>
<td>0.882 – 1.069 GHz (19.2%)</td>
<td>867</td>
</tr>
<tr>
<td></td>
<td></td>
<td>915</td>
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<td></td>
<td></td>
<td>953</td>
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</tbody>
</table>

#### 4.7 Power Harvester Circuit

In this section, a power harvester circuit, operating at the 860 to 960 MHz industrial, scientific, and medical (ISM) frequency range, is presented. The circuit was built using a commercially-available P1110 Powerharvester™ receiver manufactured by Powercast™. The P1110 converts RF energy to DC output, and provides the energy to the attached storage element. This component also has a power management capability where it can prevent the storage element from being overcharged. When an adjustable voltage threshold on the energy storage is reached, the charging will be disabled automatically.

#### 4.7.1 P1110 Powerharvester™ Receiver

The functional block diagram and pin configuration of the P1110 are shown in Figure 4.20. It consists of two major blocks: an RF-DC converter and a voltage monitor. Due to confidentiality reasons, the actual circuit design of these two blocks cannot be obtained from the manufacturer. However, it is believed that the circuit topology used in the RF-DC converter block is similar to the voltage multiplier discussed in Section 3.5.2. The functions of several major pins of the P1110 are described next.
RF<sub>in</sub> is the input pin of the P1110. It can be connected to any standard or custom 50 Ω antenna through a 50 Ω transmission line. According to the datasheet, the P1110 is optimised for 902 to 928 MHz band (U.S.A). However, it can also be used for other frequency bands, such as 865.6-867.6 MHz (United Kingdom) and 952-954 MHz (Japan), but at the expense of lower power conversion efficiency. The actual performance of the P1110 at three operating frequencies of the investigated system (867, 915 and 953 MHz) will be evaluated in the next subsection.

V<sub>OUT</sub> pin is the DC output of the P1110. It is connected to an energy storage element which can be either a rechargeable battery or a capacitor. The maximum output voltage of the P1110 can be adjusted using one of the three pins: LI, ALK or V<sub>SET</sub> pin. The maximum output voltage when using LI and ALK pins are 4.2V and 3.3V, respectively. Both pins are connected directly to ground when used. If a different output voltage is desired, V<sub>SET</sub> pin can be used. This is done by placing an external resistor between the V<sub>SET</sub> pin and ground. The resistor’s value to obtain the desired output voltage can be calculated using the following equation [147]:

\[
R = \frac{12.35 \times 10^6}{V_{\text{out,max}} - 1.235}
\]  

(4.2)

The P1110 chip also has a Received Signal Strength Indication (RSSI) functionality which enables the sampling of received signal to indicate the amount of RF power available for DC conversion. When this functionality is used, both D<sub>SET</sub> and D<sub>OUT</sub> pins are connected to a microprocessor unit (MCU). By setting the D<sub>SET</sub> pin high at 1 V, the D<sub>OUT</sub> pin produces a specific output voltage depending on the received RF input power. Note that when RSSI
functionality is being used, the harvested power will not be stored. Further information of the P1110 component can be found from its datasheet [129].

4.7.2 PCB Layout and Circuit Fabrication

The power harvester circuit was built using a two-sided PCB made of FR4, with both sides having 35μm thick copper. The selected FR4 laminate [148] has a dielectric constant of 4.4, with a tangent loss of 0.022 at the circuit’s operating frequencies. Since the P1110 is internally matched to 50 Ω, no external RF components (e.g. a matching network or stubs) were required in this design. A 50 Ω microstrip transmission line with a width of 3.02 mm (see Appendix A) was used to direct the received energy from the RF source/ antenna to the input pin of the P1110. Figure 4.21 shows the final PCB layout of the power harvester circuit. All components including the P1110 component, external resistor, and rechargeable battery were soldered on the top layer of the PCB. The RF in feed line was made as short as possible to reduce feed line losses. The bottom layer of the PCB is a full ground plane, with the GND pins on the top layer connected to ground by plated through holes. Also, an end-launched SMA female connector was placed at the circuit’s input for measurement purpose. Figure 4.22 shows one of the fabricated prototype circuits.

![Figure 4.21: Final PCB layout of the power harvester circuit.](image)
4.7.3 Power Harvester Circuit Testing

The performance of the power harvester circuit at three different operating frequencies (867 MHz, 915 MHz and 953 MHz) was evaluated in a laboratory environment. The results obtained from this evaluation can provide an insight on the actual power which can be harvested by the circuit when it is integrated into a complete outdoor RF energy harvesting system.

Results and Discussion

First, the charging current flowing into the attached rechargeable battery was measured at varying input power. To carrying out the measurement, three circuits were built. For each circuit, the DC output pin ($V_{OUT}$) was connected to the batteries with different nominal voltages – 1.2 V (NiMH) [149], 3.0 V (Li-Ion)[150] and 3.6 V (NiMH)[151]. For the circuit with a 3.0 V Li-Ion battery, the ALK recharging pin of the P1110 was connected directly to ground for 3.3 V maximum recharging. For other two circuits with NiMH batteries, an external resistor (49.8 MΩ or 4.55 MΩ) was placed between the $V_{SET}$ pin and ground to obtain the desired output voltages (1.48V or 3.95V). The resistors’ value was determined using Equation (4.2).

An Agilent E4438C ESG Vector Signal Generator [152] was used as the source to supply the RF input power to the circuits. The charging current and voltage across the battery during the tests were measured using a MS8221 Multimeter. Figure 4.23 to 4.25 shows the charging current measurement results at three load voltages: 1.25 V, 2.95 V and 3.70V,
respectively. A complete data set presented in the form of tables can be found in Appendix B.

Figure 4.23: Measured charging current as a function of RF input power when $V_{LOAD} = 1.25$ V.

Figure 4.24: Measured charging current as a function of RF input power when $V_{LOAD} = 2.95$ V.
As shown in Figures 4.23, 4.24 and 4.25, the charging current mainly depends on the RF input power fed into the circuit. As the input power increases, the current flowing into the battery increases. This results in a shorter time to fully recharge the battery. Furthermore, it is observed that for a fixed input power, the charging current decreases when higher V\text{LOAD} is used. This could be attributed to the fixed amount of input power available for RF-DC rectification. The circuit exhibits higher sensitivity at lower load voltages. At 867 MHz, the circuit has the sensitivity of -10 dBm for the output voltage of 1.25V with output current of 1.2 µA (output power = 1.5 µW), -4 dBm for the output voltage of 2.95V with output current of 1.0 µA (output power = 2.95 µW), and -2 dBm for the output voltage of 3.70 V with output current of 4.2 µA (output power = 15.54 µW).

Based on the current measurement data, the conversion efficiency of the power harvester circuit (\(\eta_c\)) was also plotted using the following equation [61]:

\[
\eta_c = \frac{P_{\text{DC}}}{P_{\text{RF}}} = \frac{V_{\text{DC}}I_{\text{DC}}}{P_{\text{RF}}}
\]  

(4.3)
where $V_{DC}$ is the DC output voltage, $I_{DC}$ is the output current, and $P_{RF}$ is the RF input power fed to the circuit. Figure 4.26 to 4.28 show the power conversion efficiency of the circuit, with varying input power at three different load voltages.

**Figure 4.26:** Circuit efficiency versus input power, $V_{LOAD} = 1.25V$  

**Figure 4.27:** Circuit efficiency versus input power, $V_{LOAD} = 2.95V$
In general, the efficiency increases with input power, but it saturates as the rate of the increase gradually decreases. The circuit is designed for a wide operating range and has two performance peaks to accommodate the range. A small drop in efficiency at certain input power range is observed for the 867 and 915 MHz curves. This could be possibly due to a higher impedance mismatch between the source and the circuit at that particular range (see Table 4.6 for the cases of 867 MHz and 915 MHz at $V_{LOAD} = 2.95 \, \text{V}$).

The return losses of the power harvester circuit, connected with a 3.0V Li-Ion battery were also measured. The purpose of this measurement is to determine how well the circuit is matched to a 50 $\Omega$ source at its operating frequency. It is desired that the circuit to have an input impedance closely matched to 50 $\Omega$ to ensure maximum power transfer from the source to the circuit. An Agilent's E5071B ENA Series Network Analyser was used to carry out these measurements. The analyser was calibrated with an Agilent's N4431-60003 Electronic Calibration Module before each measurement at different input power. The measured input impedances and return losses of the circuit at three operating frequencies: 867 MHz, 915 MHz and 953 MHz, are shown in Table 4.6. It is observed that the input impedance of the circuit changes at varying input power due to its non-linear characteristic. At $V_{LOAD} = 2.95 \, \text{V}$, the circuit operating at 915 MHz is found to be well-matched ($S_{11} < -10 \, \text{dB}$) in the lower input power range (-3 to 2 dBm) compared to that of 867 MHz (0 to 3 dBm). Meanwhile, the circuit is poorly matched at 953 MHz from -5 to 5 dBm.
<table>
<thead>
<tr>
<th>Input Power (dBm)</th>
<th>Operating Frequency</th>
<th>Impedance</th>
<th>S11 (dB)</th>
</tr>
</thead>
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<td></td>
<td></td>
<td>Real</td>
<td>Imaginary</td>
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<tr>
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<td>867</td>
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<tr>
<td></td>
<td>953</td>
<td>23.16</td>
<td>24.04</td>
</tr>
</tbody>
</table>

Table 4.6: Measured impedance and return loss at 867, 915 and 953 MHz.
It was also realised that the measurement technique used above (dc output and return loss measurements) is inadequate to characterise the performance of the power harvester circuit. The battery voltage will drop when a power load (i.e. MCU or sensor) is drawing a high current from the battery. A better measurement technique is to connect the circuit’s output with a resistive load instead of a battery. Measurements can be made to verify the dc output performance as well as the power which can be delivered to different resistive loads. These different load resistors model different power consumption “workload”. By using the proposed technique, the optimal load resistance values for maximum power transfer (at different input power and operating frequencies) can be determined. In addition, the battery charging capability of the power harvester circuit should also be investigated. This can be done by fully discharging the rechargeable battery and then connected across the output of the circuit. The battery charging profile can then be obtained by measuring the battery voltage and charging current over a period of time.

4.8 Antenna Field Trial

To evaluate the performance of the designed FSPAs, an outdoor field trial was conducted in a farming area located at Huddersfield (UK Grid Reference: SD 046136). The setup of the trial is shown in Figure 4.30. A mobile radio TM8110 [100] was used as the RF energy source of the system. The radio was programmed to transmit at 867.0 MHz with an output power of 33 dBm (2W), and fed to a commercially-available, linearly polarized patch antenna [102], through a 3-dB attenuator and coaxial cable. The radiation pattern and gain of the transmitting antenna was previously measured inside the anechoic chamber. The antenna has a peak gain of 5.2 dBi at 867.0 MHz. As shown in Figure 4.31, the main beam of the transmitting antenna is at its broadside direction with a half-power beamwidth of about 65°.
Figure 4.29: Experimental setup of the field trial.

Figure 4.30: Field trial at Huddersfield.
Since the antenna is not directly mounted to the transmitter, a feeder loss $L_{TX}$ associated with the coaxial cable and the attenuator between the radio and the antenna must be taken into account. The losses of these attenuator and cable were measured at 3.65 dB. Knowing that the radio has an output power $P_{TX}$ of 2W (33 dBm), and the antenna has a gain $G_{TX}$ of 5.2 dBi at 867 MHz, the EIRP of the transmitter system can be calculated. By including the transmitter feeder loss, the maximum radiated power of this system is 34.55 dBm (2.85W EIRP), which is within the power limit given in Table 3.1.

At the receiver side, both FSPAs – FSPA-GND and FSPA-SLG were deployed as the receiving antennas. The received power of each FSPA was measured using an Anritsu S332D Site Master™ [153] with High Accuracy Power Meter connected to a PSN50 sensor [154]. The sensor is then attached to the tested FSPA using RF connectors and coaxial cables. Before carrying out the measurements, the losses of these connectors and cables were determined. The receiver feeder losses $L_{RX}$ were measured at 0.70 dB. The transmitting antenna was placed on a tripod (to simulate the situation where it is mounted on the building’s wall) at three different heights: 0.70, 1.00 and 1.45 metres. As for the receiving antennas, they were located on the turf with their main radiating edge facing towards the transmitter. Both transmitting and receiving antennas were oriented for a polarisation match. The down tilt angle of the transmitting antenna towards the FSPAs was
set at 30°, and the received power of each antenna was recorded at a horizontal distance of 1 to 4 metres from the transmitting source.

**Results and Discussion**

Table 4.7 shows the measured received power (in dBm) of both FSPAs at three different heights: 0.70, 1.00 and 1.45 metres. The predicted received power in free space, using Equation (3.20), are also included for comparison. The predicted values were calculated by taking into account of all the known losses from the attenuator, connectors and cables at both transmitting and receiving ends.

<table>
<thead>
<tr>
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<tr>
<td>4.0</td>
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<table>
<thead>
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<th>Transmitter Height = 1.00 metres</th>
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</thead>
<tbody>
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<td>Horizontal Distance (d)</td>
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<td>3.0</td>
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</table>
Transmitter Height = 0.70 metres

<table>
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<tr>
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<th>Actual Distance (d_a)</th>
<th>Received Power at FSPA-GND (dBm)</th>
<th>Measured</th>
<th>Predicted</th>
<th>Measured</th>
<th>Predicted</th>
</tr>
</thead>
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</tr>
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<td>-8.96</td>
<td>-5.65</td>
<td>-10.59</td>
<td>-5.45</td>
<td></td>
</tr>
</tbody>
</table>

Table 4.7: Measured and predicted received power of FSPA-GND and FSPA-SLG at transmitter different heights.

In general, the received power of the antennas decreases with distance. The FSPA-SLG received less power than the FSPA-GND due to its lower gain at 867 MHz compared to that of the FSPA-GND. As expected, the measurements are less than the predicted values. This could be attributed to the pointing loss between the transmitting and the receiving antenna. The predicted values are calculated based on the peak gain values of both antennas in a free space, with their main beams properly aligned and pointing towards each other. In contrast, the transmitting antenna of the trial is mounted on a tripod at a certain height above the ground with a down tilt angle of 30º. The receiving antennas are placed directly on the ground surface with their main beams slightly tilted from the broadside direction. As a result, part of the radiated energy from the transmitter may not be captured due to the directional nature of the receiving antenna. The power measurement values shown in Table 4.7 include the receiver feeder losses from the connectors and cables. For the real deployment of a full system, the receiving antenna will be directly connected to the power harvester circuit (without the connection of the cable). Therefore, it will be expected that the input power fed into the circuit is 0.7 dB more than the values displayed above.

Further Discussion

Clearly, the number of measurement data presented above is low and it is insufficient to model the actual deployment environment of the wireless ground-level sensor network. In this field trial, only one condition is being considered, where it assumes a direct LOS path between the transmitter and the receiving nodes. In addition, the effect of antenna polarisation on the received power performance of the system was also not investigated.
Therefore, additional outdoor measurements are required to test the antenna performance under a wider range of operating conditions, i.e. placing the antennas at different orientations, locating them within different types of vegetation (e.g. shrubberies or crop canopies), and testing them under different outdoor temperatures and humidity conditions. Also, the measurements data above were spot measurement, i.e. no information of variation of the antenna’s received power versus time is available. Hence, it is recommended to perform the trial for a longer period of time (i.e. a full day) to collect more data. Averaging can then be applied on these collected data to obtain a more sensible received power level expected from the antennas at varying distances and different outdoor scenarios.

The field trial results show that, the FSPA is not a suitable antenna design for the intended application. Depending on the location where the FSPA is placed, a loss of 1 to 3 dB could be incurred as a result of the pointing error between the transmitting antenna (on the wall) and the receiving antennas (on the ground). One may suggest that raising and tilting the FSPAs slightly off the ground can reduce the pointing error, thereby improving the received power performance. However, this would exceed the 2mm maximum height limitation as set forth herein. Therefore, a better antenna option must be explored. Since there is no restriction imposed on the lateral size of the antenna, a phased array antenna with higher gain and beam steering capability [155, 156] could be a possible alternative. Further research is required to examine the suitability of using phase array antenna for this application.
5 Conclusions and Future Work

The work presented in this thesis investigates the feasibility of using RF energy harvesting to power a wireless ground-level sensor network deployed in an outdoor environment. The conclusions from each part of the work are summarised here.

1) The research started with a literature-based study of various potential energy harvesting techniques which may be applicable to power the wireless ground-level sensor network. Four techniques were examined including solar, thermal, wind and RF energy harvesting. The review showed that solar energy harvesting outperforms other techniques in terms of its output power capability (up to 15 mW/cm² in a bright sunny day). In contrast, RF energy harvesting using ambient sources is found to be the most ineffective technique due to its extremely low power density (1 µW/cm²). Due to the nature of the operating environment in which the targeted WSN nodes are deployed, a far-field, RF energy harvesting system based on an intentional source (rather than the ambient sources) was chosen as the subject of the investigation.

2) A theoretical-based analysis was performed on an outdoor RF energy harvesting system. Several design parameters were discussed, including the transmitter unit, the receiving antenna, the power harvester circuit, the energy storage and the load. It was found that the operating distance of the system mainly depends upon the operating frequency, the collection efficiency of the receiving antenna, the accuracy of impedance matching between the antenna and the power harvester circuit, as well as the efficiency of the circuit converting the captured RF energy to DC voltage. A link budget analysis was also carried out to investigate the feasibility of using RF energy harvesting to power the intended WSN application using the Zyrox2 Baits Station as the exemplar. The calculation showed that, in order to power the WSN node in a typical agricultural setting (with foliage depth of 5 metres), a high-gain receiving antenna of at least 13 dBi would be required.

3) The research continued by presenting two antenna designs which could be potentially used in an outdoor RF energy harvesting system. The first design is based on a folded shorted patch antenna with solid ground plane (FSPA-GND) and the second design is a modified FSPA with slotted ground plane configuration (FSPA-SLG). Both antennas were simulated, fabricated and tested in an anechoic
chamber. It was found that by properly embedding slot pairs on the ground plane, the FSPA’s overall size can be reduced by 29% without greatly affecting its impedance bandwidth. Parametric studies were also performed to examine the effects of different dimensional parameters on the performance of a FSPA-SLG. The antenna’s height ratio (H₁/H₂) and its folded patch length (L₂) were found to be the two main parameters which could affect the directivity of the FSPA-SLG.

4) In addition, a power harvester circuit operating at frequency band from 860 to 960 MHz was designed and built using a P1110 component from Powercast™. Although the circuit was optimised for 902-928 MHz band (U.S.A), it could still be used for other frequency bands such as 865.6-867.6 MHz (United Kingdom) and 952-954 MHz (Japan), but with a reduction in efficiency at lower input power range. The measurement results showed that the circuit operating at 867 MHz has a sensitivity of -10 dBm for the output voltage of 1.25V with output current of 1.2 µA (1.5 µW), -4 dBm for the output voltage of 2.95V with output current of 1.0 µA (2.95 µW), and -2 dBm for the output voltage of 3.70 V with output current of 4.2 µA (15.54 µW), respectively. A better measurement technique to characterise the circuit performance (i.e. connecting the output with resistive loads instead of a battery) has been proposed.

5) A field trial was also carried out to evaluate the received power performance of the FSPAs in an outdoor environment. The number of measurement data was found to be insufficient as it only considers one specific condition - a direct LOS path between the transmitter and the receiving nodes. Further experimental work is required to test the antennas under a wider range of environmental situations.

In conclusion, RF energy harvesting is not practical at present to power most of the wireless ground-level sensor nodes due to their high power consumption. The inherent loss along the outdoor RF energy harvesting path (free space loss, foliage loss, RF-DC conversion loss) makes it only suitable for powering sensor-enhanced RFID tags and WSN nodes with extremely low duty cycle. However, with the continued reduction in power consumption of electronic components and increased sensitivity of the power harvester circuit, the practical applications of RF energy harvesting will grow.

The most important follow-on-work is the improvement of the current antenna design. Further research can be carried out to examine the suitability of using phased array
antenna with higher gain and beam steering capability [155, 156] to improve the system performance. Nonetheless, the 20 mm height limitation remains the biggest barrier towards the feasibility of using RF energy harvesting for this particular application, i.e. wireless ground-level sensor network. To make this system more practical, this limitation must be relaxed in the future. Once a better antenna design is identified, a larger scale field trial can be conducted. This trial should consider a wider range of environmental situations as expected by the targeted WSN nodes. Another research area that can be explored is the improvement of the sensitivity, as well as the conversion efficiency of the power harvester circuit. In this work, the designed circuit operates at 867 MHz with a sensitivity of 10 dBm for 1.5 µW output power level. Two possible methods: voltage boosting technique based on an impedance transformation circuit [130] or rectifier circuit using floating gate transistors as rectifying diodes [157], can be employed to achieve better sensitivity and higher conversion efficiency.
References


[54] ICNIRP, "Guidelines for Limiting Exposure to Time-Varying Electric, Magnetic, and Electromagnetic Fields (Up to 300 GHz)."


N. Hiroshi, K. Yoshihiro, and A. Tohru, "Prototype implementation of wireless sensor network using TV broadcast RF energy harvesting," in Proceedings of the 12th ACM international conference adjunct papers on Ubiquitous computing Copenhagen, Denmark: ACM.


[140] L. Gil-Young, K. Yonghoon, L. Jong-Sik, and N. Sangwook, "Size reduction of microstrip-fed slot antenna by inductive and capacitive loading," in *Antennas and


Appendices:

A. Microstrip Transmission Line Calculation

A 50 Ω microstrip transmission line was designed to direct the energy from the RF source to the input pin of the P1110. Figure A.1 shows the cross-sectional view of a microstrip transmission line, which has a strip conductor and a ground plane separated by a dielectric substrate material.

![Cross-sectional view of a microstrip transmission line](image)

Figure A.1: Cross sectional view of a microstrip transmission line.

The characteristic impedance of a microstrip transmission line is determined by the line width (w), thickness of the strip conductor (t), substrate height (h) and the dielectric constant of the substrate used (εr). Assuming the thickness, t of the strip conductor is negligible compared to the substrate height, h (where t/h < 0.005), the following equations can be used to calculate the line impedance [1]:

For a narrow strip line (w/h < 1),

\[
Z_o = \frac{Z_t}{2\pi \sqrt{\varepsilon_{eff}}} \ln \left( \frac{8h}{w} + \frac{w}{4h} \right)
\]

(A.1)

where \( Z_t \) is the wave impedance of free space and it is given as:

\[
Z_t = \sqrt{\frac{\mu_o}{\varepsilon_o}} = 376.8\Omega
\]

(A.2)

with \( \mu_o = \) permeability of free space = 1.256 \times 10^{-6} H/m

\( \varepsilon_o = \) permittivity of free space = 8.854 \times 10^{-12} F/m
\( \varepsilon_{\text{eff}} \) is the effective dielectric constant which represents the dielectric constant of an equivalent homogenous medium enclosing the transmission line (replacing the dielectric substrate and surrounding air). The effective dielectric constant can be expressed as:

\[
\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ \left(1 + \frac{12}{w} \frac{h}{w} \right)^{-1/2} + 0.04 \left(1 - \frac{w}{h} \right)^2 \right]
\]  

(A.3)

For a wide strip line (w/h > 1), a different characteristic line impedance formula is used:

\[
Z_o = \frac{Z_t}{\sqrt{\varepsilon_{\text{eff}} \left(1.393 + \frac{w}{h} + \frac{2}{3} \ln \left( \frac{w}{h} + 1.444 \right) \right)}}
\]

(A.4)

where

\[
\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left(1 + \frac{12}{w} \frac{h}{w} \right)^{-1/2}
\]

(A.5)

The characteristic impedance of a microstrip line as a function of w/h (when \( \varepsilon_r = 4.4 \)) for two cases: narrow strip line (w/h <1) and wide strip line (w/h>1), are plotted in Figure A.2 and Figure A.3. To obtain \( Z_o = 50 \, \Omega \), the w/h ratio is approximately 1.93. Since the substrate thickness, h is known (1.57 mm), the estimated line width can be computed as 3.02 mm.

Figure A.2: Characteristic impedance of microstrip line as a function of w/h for w/h <1 (\( \varepsilon_r = 4.4 \))
Figure A.3: Characteristic impedance of microstrip line as a function of \( w/h \) for \( w/h > 1 \) (\( \varepsilon_r = 4.4 \))

For the proceeding equations, the strip line thickness is assumed to be negligible. In the proposed circuit design, the PCB has a copper thickness, \( t \) of 1 oz (0.035 mm). The ratio \( t/h \) is about 0.02, which is more than 0.005 used for the previous assumption. In this case, the effect of finite copper strip thickness has to be considered. This effect can be estimated as an increase in effective width \( w_{\text{eff}} \) of the strip line due to the occurrence of more fringing fields. The effective width is given by [1]:

\[
21 \ln \left( \frac{w + \frac{t}{2}}{x} \right) = \pi \left( 1 + \frac{2x}{t} \right) = \pi \left( 1 + \ln \left( \frac{2w}{h} \right) \right) = 3.08 \text{ mm}
\]  

where \( t \) is the thickness of the strip line, and \( x = h \) if \( w > h/(2\pi) > 2t \), or \( x = 2\pi w \) if \( h/(2\pi) > w > 2t \). Since \( h/(2\pi) > w > 2t \), the following equation is used to estimate the effective width of the strip line:

\[
w_{\text{eff}} = w + \frac{t}{\pi} \left( 1 + \ln \left( \frac{2h}{t} \right) \right)
\]  

By substituting the line width term, \( w \) in Equations (A.4) and (A.5) with the effective width given in Equation (A.7), the effective dielectric constant and the characteristic impedance of the line can be computed as:
The calculated characteristic line impedance is found to be very close to the target impedance of 50 Ω. By using the obtained results, the microstrip transmission line of the proposed circuit is designed to have a width of 3.08 mm.

Reference

B. Charging Current Measurement of the Power Harvester Circuit

<table>
<thead>
<tr>
<th>Input Power (dBm)</th>
<th>Charging Current (μA), $V_{LOAD} = 1.25$ V</th>
</tr>
</thead>
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<tr>
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<td>-</td>
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<tr>
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Table B.1: Measured charging current as a function of RF input power from -11 to 5 dBm when $V_{LOAD} = 1.25$ V.

<table>
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<th>Input Power (dBm)</th>
<th>Charging Current (μA), $V_{LOAD} = 2.95$ V</th>
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Table B.2: Measured charging current as a function of RF input power from -5 to 10 dBm when $V_{\text{LOAD}} = 2.95$ V.

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<th>Input Power (dBm)</th>
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Table B.3: Measured charging current as a function of RF input power from -5 to 10 dBm when $V_{\text{LOAD}} = 3.70$ V.