Energy scavenging for autonomous system

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**ENERGY SCAVENGING FOR AUTONOMOUS
SYSTEM**

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2017

Energy Scavenging for Autonomous System

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the degree of
Doctor of Philosophy

2017

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research and has not been submitted for a higher degree to any other University or Institution.

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Abstract

With huge advancements in CMOS technology, portable electronics can now be operated at a lower voltage, hence lowering the stress index on energy storage elements. A relevant application will be the ubiquitous smart sensors deployed for health, structural and environmental monitoring. With regards to the research topic, the most important component in a smart sensor is its power source. The power source should be relatively compact and have a near-infinite energy capacity for an uninterrupted sensing operation throughout its lifetime. For that reason, energy harvesting techniques are being looked upon as a replacement or supplement to improve the reliability of the power source. The processing circuits of the energy harvesting system must be simple and highly efficient in order to be popular with the consumer market.

Harvesting energy from ambient sources such as light, heat, motion and radio-frequency is an attractive alternative to the finite capacity primary batteries. Among these available ambient energy, light energy has the highest power density and is readily available under indoor and outdoor conditions. Hence, light energy harvesting techniques are being looked upon as a complement to improve the reliability on the power source of the miniaturised wireless sensor. In order to extract the maximum power from a photovoltaic (PV) panel, maximum-power-point-tracking (MPPT) circuitries are required. The MPPT circuitry includes a DC-DC converter and a controller to extract power from the PV panel by operating the PV panel at its maximum power points which varies with irradiance and temperature.

This thesis introduces two topologies for light energy harvesting system with MPPT control. The first design is an autonomous wireless sensor node with flexible energy harvesting mechanism that is able to conform to the body's contours, is proposed for biometric monitoring. To fulfill the requirements of

sustainability and compact size, the proposed system is equipped with an ultra-low power management circuit specially designed on a flexible printed circuit board. The flexible power management circuit is able to transfer near maximum electrical power (88.5 %) from the flexible solar panel to store in the supercapacitor for powering the wireless sensor node. The power consumption of the flexible power management circuit only takes up 32.86 μW . Experimental results show that under indoor conditions with typical average lighting intensity of 320 lux, the wearable sensor node is able to continuously monitor, read and transmit the temperature of the wearer back to the base node without the need of any batteries. In addition, the designed FEH sensor node aligned to the body's contours at an angle of 30° generated 56 μW of electrical power, which was sufficient to sustain its operation for more than fifteen hours.

The second design deals with the improvement of the interface circuitry used for maximum power point tracking of a solar panel. A new maximum power tracking technique for the solar panel is introduced. The new MPPT technique is implemented with 0.18 μm CMOS process. The proposed circuitry is equipped with leakage reduction technique and the power consumption of this circuitry is only 12.113 nW while maintaining a tracking efficiency of 98.6 % for an input power of 268 μW .

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Abbreviations

4TBB	Four Transistor Non-inverting Buck-Boost
ADC	Analog-to-Digital Controller
BAN	Body Area Network
CCM	Continuous Conduction Mode
CMP	Comparator
D	Duty cycle ratio
DCM	Discontinuous Conduction Mode
DTC	Dead Time Controller
EH	Energy Harvesting
ESR	Equivalence Series Resistance
FEH	Flexible Energy Harvesting
FPMC	Flexible Power Management Board
FOCV	Fractional Open-Circuit Voltage
f_s	Switching frequency
i_{pv}	Current at output terminals of PV module
IoT	Internet of Things
K_{oc}	FOCV Factor
INC	Incremental Conductance
L	Inductance
LDO	Low Dropout
MI	Multiple Input
MO	Multiple Output
MPPT	Maximum Power Point Tracking
MPP	Maximum Power Point
M_v	Voltage transfer function
PWM	Pulse Width Modulation
PV	Photovoltaic
PMU	Power Management Unit

P&O	Perturb and Observe
P_{PV}	Output Power of PV module
RF	Radio Frequency
R_s	Series Resistance of PV module
R_{SH}	Shunt Resistance of PV module
SIMIMO	Single Inductor Multiple Input Multiple Output
TEG	Thermoelectric Generator
TFT	Thin-film transistor
V_{PV}	Output Voltage of PV module
V_{MP}	Voltage of PV module at MPP
V_{OC}	Open-Circuit Voltage of PV module
V_{HOLD}	Holding Voltage of S/H Circuit
ZCD	Zero Crossing Detector

Chapter 1

Introduction

This chapter provides an insight into the need for ultra-low power energy harvesting and the motivation for this research. Various renewable energy used for an energy harvesting system are examined. The work undertaken aims to design a highly efficient interface circuit for harvesting energy from indoor lightings. The energy processing circuits must maintain high efficiency, while remaining as an attractive solution from the energy, size and cost perspective. A detailed list of original contributions of this thesis are presented at the end of the chapter.

1.1 Background

“Whether one is engaging in a health promoting activity such as exercise or is living with a chronic disease such as asthma, he or she is responsible for day-to-day management” [1]. With the world’s aging population (people aged 65 or older) been projected to increase to nearly 1.5 billion (16 % of world’s population) by the year 2050 [2], the rapid increase in demand for more sophisticated health care services is an inevitability. Fortunately, the advancement of technology has made it possible for medical support to be integrated into our daily lives to ease this global burden.

Remote monitoring of one’s physiological activities, such as sleep and cardiac patterns can be performed through the use of Body Area Networks (BANs). As shown in Figure 1.1, the wireless BAN, made up of several sensor nodes possessing sensing, processing and communicating abilities, is responsible for monitoring of bodily functions of the user.

Figure 1.1 Illustration on Internet of Things for BSN in healthcare system adapted from [20]

Typically, the data sensed by the miniaturized sensors is wirelessly transmitted to a remote data terminal for post-processing. Although the remote biometric monitoring sounds promising, several major issues must first be addressed before a full implementation of the concept is possible. Among these concerns are security protection, maintainability and connectivity of the health monitoring system [3].

This research will focus on addressing the maintainability of the bio-monitoring system. In order to ensure the maintainability of the biometric sensor, an ultra-low power and highly-efficient power management circuit design is required for the delivery of energy from the energy reservoir to the sensor nodes.

The energy reservoir chosen for this application would be the ambient energy. Employing renewable energy to power the bio-monitoring system will ensure a self-sustainability of the sensor nodes throughout its lifetime.

For centuries, human beings have taken advantage of renewable energy to drive mechanical and electrical loads. Some of the widespread applications that exploit renewable energy are the windmill, watermill and solar farm. These large-scale deployments convert light and kinetic energy into useful energy that possess the similar benefits brought about by the usage of non-renewable resources. Also as a bonus, renewable energy will not be depleted and does not contribute to global warming. Hence, extending renewable energy to drive more applications will contribute to the Earth's sustainability.

For the purpose of this research, however, the large-scale energy harvesting systems will not be covered, instead, a detailed examination of micro-energy harvesting for low-powered autonomous systems will be elaborated. Question is, do we require a complex micro-energy harvesting circuit to supply power to the sensor node when a simple battery can effectively perform the same task?

This leads back to the application of the sensor node mentioned in the introduction. The sensor node, intended for health monitoring can be applied in different areas such as military and sports industries. At this point, it is worthwhile to mention that the sensor node monitoring systems can also be extended to applications such as structural monitoring [4] and environment monitoring [5]. However, most of the above examples require the sensor to have a long lasting uninterrupted operation, which is often not the case.

Take for example, a primary coin-battery (Panasonic, 23mm x 3mm) having a capacity of 255mAh supplying power to a temperature sensor node (Texas Instruments, eZ430-RF2500) which is designed to transmit the sensed data at every second (average current consumption is 40.457 μ A). Under ideal operating conditions, the coin-battery can only sustain the operation for approximately 8.6 months. A failure of the sensor due to the depletion of a finite energy supply (battery) might not be immediately recognized by the user and this could lead to disastrous consequences in emergencies. In addition, when the elderly are the users of the sensor applications, they might not possess the technical knowledge or physical ability to repair and maintain the functionality of the biometric system. Thus, energy harvesting from ambient energy is a promising long-term solution for an infinite energy supply to power the sensor nodes with minimum effort. The self-powered sensor will remain functional, providing decades of monitoring operation as long as the ambient energy is available.

1.2 Energy Scavenging Solution for Wireless Biometric Sensor Node

The limited lifespan of the primary batteries as a source of energy supply is the bottleneck to an autonomous operational period of a wireless sensor. These primary batteries are discarded when they can no longer provide sufficient

electrical energy for the connected load. In addition, the primary batteries pose a great challenge in the aspect of waste disposal [19].

The energy storage mediums are typically classified into two distinct groups, namely: (1) primary batteries and (2) secondary/ rechargeable batteries. The secondary batteries convert the chemical energy contained in it into electrical energy through a reversible electrochemical process. Hence, employing a rechargeable energy storage medium is more favorable and desirable as compared to a primary battery with finite energy supply. The different types of energy storage technologies and their power and energy densities, as well as characteristics are shown in Figure 1.1.

Figure 1.2 Ragone plot showing the power density vs. energy density of different energy storage technologies adapted from [17]

Let us compare the pros and cons of each medium. Referring to Figure 1.2, it can be observed that the supercapacitors have a higher peak power density as compared to the primary and secondary batteries. This implies that the

supercapacitors are able to deliver more electrical power to the connected load within a short period of time.

This corresponds to the high power consumption required by the wireless sensor during its burst mode transmission. In addition, the supercapacitors are shown to be able to maintain their capacities after 1,000,000 charge/ discharge cycles as compared to a lithium-ion battery which has a much lower average of 300 charge/ discharge cycles [18].

However, batteries have a higher energy density as compared to the supercapacitors. This implies that the batteries can deliver electrical power to its connected load for a far longer period of time as compared to the supercapacitors. In conclusion, the supercapacitors are able to meet the power requirement of the wireless sensors during its burst transmission mode but not able to maintain the operational lifetime of the wireless sensor due to its low energy density. Hence, energy scavenging from the ambient environment to replenish the energy in the supercapacitor is an ideal complement for it to sustain the operational lifetime of the wireless sensor nodes.

Energy scavenging refers to a technique that captures and converts the unused ambient energy into usable electrical power. The electrical power can be used to recharge a battery/ supercapacitor and power wearable biometric sensors or wireless sensors. The end application typically consume power in the range of sub-milliwatts to microwatts. Table 1.1 shows a comparison between the commonly found ambient energy.

Table 1.1 Characteristics of various energy sources available in the ambient adapted from [14]

Source	Source Power	Harvested Power
Ambient Light (DC)		
Indoor	0.1 mW/cm ²	10 μ W/ cm ²
Outdoor	100 mW/ cm ²	10 mW/ cm ²
Vibrational/ Motion (AC)		
Human	0.5 m/s ² @ 1 Hz 1 m/s ² @ 50 Hz	4 μ W/cm ²
Industrial	1 m/s ² @ 5 Hz 10 m/s ² @ 1 kHz	100 μ W/cm ²
Thermal (DC)		
Human	20 mW/ cm ²	30 μ W/cm ²
Industrial	100 mW/ cm ²	1-10 mW/cm ²
RF (AC)		
Cell Phone	0.3 μ W/ cm ²	0.1 μ W/cm ²

From Table 1.1, it can be clearly observed that there is no single interface circuitry able to harvest the ambient energies from all the different environment conditions. In addition, from Table 1.1, out of the four types of ambient source mentioned, the energy obtained from light and heat has the highest power density, in outdoor and industrial conditions respectively. However, conditions fluctuate.

During a cloudy or rainy day, the outdoor light energy harvested by the solar panels will drop tremendously. Harvesting high power density through the “Seebeck Effect” of thermal transducer requires a high temperature gradient and this method is not suitable for the context of this thesis.

Hence, the next ideal choice will be harvesting energy from illuminated indoor environments to power wearable biometric sensors. Indoor light energy is abundant and has a relative high power density compared with the other similar sources. In addition, employing energy harvesting for portable applications can

provide numerous benefits to users [15, 16]. The following are the three main benefits that can be brought about:

1. Reduction in form factor of the biometric sensor. With technology scaling in CMOS process, the power consumption of wireless sensors is reduced greatly. Hence, there is a reduction in the dependency for a large energy reservoir battery and energy harvesting may be a sufficient power source for the wireless sensors.
2. Reduction in maintenance cost. Energy harvesting can supplement the battery to provide power for the wireless sensors. These sensor will then be able to be left unattended after deployment and still able to remain functional for a long period of time.
3. Reduction in environment impact. Energy harvesting can eliminate the need of frequent replacement of primary batteries which are non-recyclable.

Over the years, multiple researchers have worked on designing energy harvesting systems that are capable of converting sufficient ambient energy into usable electrical energy for powering micro-power portable applications. Some of the major research interests highlighted are: (1) a simple and efficient maximum power tracking or extracting algorithm [6, 7]; (2) an effective hybrid energy harvesting system to overcome the intermittent ambient energy [8]; (3) power management technique to improve the efficiency of energy harvesting system [9, 10], and (4) cold-start of the complete energy harvesting system [11-13].

However, research gaps between the literatures are identified and they will be described in the following chapter. There is a need to solve the identified research gaps to improve the reliability of the energy harvesting systems for the market to

adopt energy harvesting technique as a primary power source or as a supplement with the battery in portable applications.

1.3 Contributions of the Thesis

Although solar energy is in abundance, it is imperative that solar energy can be harvested in the most efficient method for ultra-low power applications such as the miniaturised medical sensors used in wireless body area networks (WBANs) for ubiquitous healthcare. The main contribution of this thesis is summarized as:

- i. Proposed an indoor light energy harvesting system with optimization on the use of a flexible PV panel. The electrical characteristics of the indoor flexible PV panel is modeled and operated with MPPT capability to improve the overall system's power efficiency.
- ii. A circuit implementation for autonomous wearable sensor nodes with flexible energy harvesting capability under low illumination condition. The proposed design is implemented on a flexible PCB with low-power energy processing circuit to provide a degree of flexibility to the user.
- iii. Proposed a MPPT circuit with hill-climbing algorithm for solar energy harvesting system. The proposed circuit employs the use of sample and hold circuit with a dynamic comparator to achieve optimal MPPT accuracy. A charge leakage reduction technique employing recycled charge is implemented for the MPPT circuit to cater for a wide MPPT cycle.

1.4 Organizations of the Thesis

As described in the literature survey, indoor solar energy harvesting appears to be a promising solution to supplement the primary batteries employed in wearable biometric sensors. The work undertaken as part of this Ph.D. program aims to develop a practical energy harvesting system with a highly efficient power delivery circuitry while the controller maintains a highly low-power consumption.

Chapter 1 has introduced the background and motivation for the thesis work. The dependence on battery as the sole power source for the low-power wireless sensor is impeding the realization of a fully autonomous system. With advancement in CMOS technology, the power consumption of wireless sensors now is comparable with the ambient energy found in indoor conditions. Hence, indoor light energy harvesting can be used to power a wearable biometric sensor.

Chapter 2 discusses the working mechanisms and electrical characteristics of a solar cell. With an in-depth understanding of the solar cell, selection of solar cells for practical deployment can be made with ease and accuracy. Different MPPT algorithms and DC-DC converters are analyzed and insights are given on how the power converters emulate the resistance of a solar cell to achieve maximum power harvesting. The analysis done in this chapter is used for the implementation of a flexible energy harvesting system in Chapter 3.

To determine the feasibility of a flexible indoor light harvesting system for wearable biometric sensor, an off-the-shelf flexible solar panel has been selected to conform to the human body for harvesting of fluorescence light. In Chapter 3, the operating characteristics of the flexible solar module has been modelled and verified with the experimental data.

An interface circuitry capable of harvesting maximum power from the flexible solar module is discussed and designed. The controller for harvesting energy from the flexible solar module is also aimed to possess a degree of flexibility while consuming a low power to maintain its efficiency.

The result for the proposed flexible energy harvesting system is presented and discussed in Chapter 4. An ultra-low power and flexible power management circuit board with maximum power tracking capability has been illustrated and implemented. With the proposed method, the designed flexible energy harvesting system is able to conform to the human contour at an angle of 30° and sustain the operation of the biometric sensor for more than 15 hours. The maximum power tracking circuitry is able to achieve an efficiency of 88.5 % for an input power of $77.65 \mu\text{W}$.

Chapter 5 begins with the discussion on the drawbacks of the employed fractional open-circuit voltage (FOCV) method used for maximum power point tracking of a solar cell. Although the interface circuit designed for FOCV can achieve an efficiency of 88.5 %, the tracking accuracy appears to be lower as compared to the modern MPPT circuitry.

A new technique is then presented that allows maximum power point tracking of solar cell to be easily implemented across different platforms while maintaining an efficiency comparable to the current MPPT circuitries. This technique is elaborated with actual circuit design to realize the theoretical approach of achieving MPPT with four transistor buck boost power converter.

After understanding the fundamentals of the new MPPT circuitry, the circuit is designed and implemented in Chapter 6. The design tools used were MATLAB/SIMULINK and Cadence-Virtuoso. Initial simulation results showed a degradation in the accuracy of the proposed MPPT circuitry. Improvements to

the circuit was introduced and the post-layout simulations showed that proposed MPPT circuitry can achieve an efficiency of 98.6 % while consuming only 12.113 nW to power its processing circuits.

Chapter 7 provides the conclusions for the work done in this thesis and suggest possible solutions to aid in achieving an autonomous system through energy harvesting.

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Chapter 2

Literature Review

This chapter aims to provide a comprehensive overview of the technology used in solar energy harvesting from past to present. Various interface circuits used for harvesting energy from solar transducers are reviewed, highlighting the advantages and disadvantages of each technique. Finally, the desirable features of an energy harvesting system are examined, as well as addressing some of the drawbacks associated with the state-of-the-art circuits. These observations are particularly useful when conceptualizing an all-new and improved design.

2.1 Overview of an Energy Harvesting (EH) Architecture

Figure 2.1 shows a conventional three-stage energy harvesting architecture. The three stages are: (1) energy harvesting stage; (2) maximum power tracking stage, and (3) voltage regulation stage.

Figure 2.1 Architecture of an energy harvesting system

The first stage is the conversion of ambient energy into electrical energy by the various micro-energy transducers. A compiled list showing the characteristics and power density harvested of the various transducers is given in Table 2.1.

From Table 2.1, it can be seen that each form of ambient energy is unique. Furthermore, the behaviors of the ambient energies are intermittent and fluctuate greatly under different operating and environmental conditions. To increase the reliability of an EH system, there is a need to have multiple energy sources to augment the unavailability and uncertainty of ambient energy sources when in operation.

Table 2.1 Characteristics of an ambient energy for harvesting adapted from [1-2]

Energy Source	Output Power	Remarks
Photovoltaic		
Direct Sunlight	100 mW/cm ²	Efficiency of solar cell is different across different materials*. Common polycrystalline solar cells are 16 %-17 % efficient and standard mono-crystalline cells approach 20 %
Illuminated Office	100 μW/cm ²	
Thermal		
10K gradient **	135μW/cm ²	Typical efficiency of thermoelectric generators are ≤ 1 % for thermal gradient less than 313 K
Vibrational		
Push button	50 μJ/N	Quoted at 3V _{DC} for the MIT Media Lab Device
Ambient wind	1 mW/cm ²	Typical average wind speed of 3 m/s in the ambient
RF		
Ambient radio frequency	<1 μW/cm ²	Unless near a RF transmitter

* Reported efficiency of various solar cell can be found at [3]

** ThermoLife [4]

The second stage consists of a DC-DC/AC-DC converter that is responsible for regulating the output voltage of the energy transducer to its instantaneous maximum power point. To achieve that, a maximum power point tracking (MPPT) algorithm [5] is required to control the duty cycle of the DC-DC converter. This control scheme for maximum power extraction is known as the resistor emulation technique [6-9]. By tweaking the duty ratio of the DC-DC converter with respect to the MPPT algorithm, the input impedance of the power converter can be configured to match the internal impedance of the energy transducer. This allows the instantaneous maximum power harvested by the energy transducer to be transferred to an energy storage element for post-processing. The intermediate storage element is a necessity as ambient energy is usually an intermittent energy source. It acts as a fail-safe when the amount of available ambient energy is

insufficient to meet the instantaneous load requirement, and the energy in storage will ensure the continuous operation of the load.

The third stage usually consists of a step-down DC-DC converter or low-dropout (LDO) regulator that will regulate the energy stored in the battery/ supercapacitor to the respective voltage levels required by the various loads connected to the energy harvesting system. Typically, the connected load will either operate continuously or in an intermittent manner. The manner of operation is determined by balancing the average harvested power with the average power consumption of all the connected loads. When the average harvested power is lower than the required power consumption, the load will then be designed to operate in an intermittent manner.

2.2 Solar Energy Harvesting

From Table 2.1, it can be observed that solar energy yields the highest power density. Although solar energy is intermittent, artificial lighting is readily available in the modern world, thus improving the reliability of a solar energy operated system. Therefore, photovoltaic modules will be chosen as the primary energy transducer in this thesis.

The solar cell is a photonic device that can directly convert light into electric charges, supplying the converted power to the electrical loads connected to its output terminal. A conventional structure of the solar cell is shown in Figure 2.2. The solar cell is manufactured with the p-n junction sandwiched between two metal contacts. The front metal contact is for the collection of the electrons generated by the incident photons from a light source and the base metal contact is responsible for the recombination process of the generated electron-hole pairs. The finger structure of the front metal contact is necessary for maximizing the

light-exposed surface area of the solar cell. The current generated by the solar cell can be collected by the connection of a load across its output terminals.

Figure 2.2 Construction of a solar cell with connected load

2.1.1 Characteristics of Solar Cell under Different Light Sources

The light intensity from artificial lighting in indoor conditions is 10-100 times less than the sunlight in outdoor conditions [10]. It is also known that the indoor light spectrum differs from the solar spectrum. Hence, the authors in [11] conducted a study on the performance of different types of solar cells under different light sources. The types of solar cells investigated were crystalline silicon, cadmium telluride, copper indium gallium selenide, amorphous silicon, gallium arsenide and organic cells with active layer P3HT:PCBM. To measure the external quantum efficiency of the solar cells, the authors made three assumptions: (1) all the light sources were scaled to 500 lux to emulate an average indoor lighting intensity; (2) fill factor of solar is unity, and (3) open circuit voltage of solar cell was approximated as $V_{OC} \approx E_g / q$.

Figure 2.3 The scaled spectral irradiance of solar and artificial lightings verse wavelength adapted from [11]

Figure 2.4 The external quantum efficiency of different solar cells adapted from [11]

Figure 2.5 The relative efficiency of different solar cells in different lighting conditions with reference to crystalline silicon solar cell adapted from [11]

As seen in Figure 2.3, the wavelengths of most artificial lighting are contained within the spectrum of the human eye. This translates to power-efficient lighting as energy is not wasted to generate invisible light. Hence, fluorescent lighting is often preferred for indoor deployment. Referring to Figures 2.4 and 2.5, a-Si and CdTe solar cells are most suitable for use indoors. Their relatively high efficiencies with respect to Si solar cell prove their potential for indoor-light harvesting. The above observations are important as the characteristics of solar cells provided by the manufacturer are normally only tested at AM1.5.

2.1.2 Equivalent Circuit Model of a Solar Cell

In order to understand and simulate the performance of a solar energy harvesting system, the energy transducer must first be examined. The electrical characteristics of the solar module can be simulated through its relatively accurate theoretical model. There exists several relatively accurate mathematical

models of a PV cell in the literature [12-14] to describe the operation of this photovoltaic energy transducer. The most accurate model reported in the literature to define the I-V characteristics of a solar module is the double diode model [90]. The double diode model is highly computational intensive and requires an accurate and extensive characterization of the PV module. Instead a single diode model approach can be adopted. The single diode model is only less accurate than the double diode model when operating under low irradiance condition that is not suitable for energy harvesting. Hence, in this section, an electrical circuit with a single diode is considered as the equivalent photovoltaic model, which consists of n_s number of PV cells in series, as shown in Figure 2.6.

Figure 2.6 A single diode equivalent electrical circuit of solar cell

I_{dc} represents the photocurrent generated by a solar cell under illumination, and diode D_1 represents the diffusion and recombination of the charge carriers in the semiconductor. Series resistance R_s represents the movement of charges carried between the semiconductor and metal contact. Lastly, shunt resistance R_{sh} represents the manufacturing defects in the paths for electron flow.

Figure 2.7 Current versus voltage plot of solar panel at various lux values

Figure 2.7 illustrates the I-V curves of a solar cell at different illuminations. Three identified points in the figure are noted and are described by Equations 2.3, 2.4 and 2.5. The current, i_{pv} , generated at the output terminals of a solar cell can be expressed as:

$$i_{pv} = i_{ph} - i_{ol} * [\exp(\frac{q(V_{out} + i_{pv}R_s)}{n_l k T}) - 1] - \frac{V_{out} + i_{pv}R_s}{R_{sh}} \quad (2.1)$$

where, n_l is the diode ideality factor, k is the Boltzmann constant and T is operating temperature of the solar cell. Solar cells can be connected electrically in series and parallel to increase the open-circuit voltage and photocurrent respectively. When N solar cells are connected in series, Equation (2.1) becomes:

$$i_{pv} = i_{ph} - i_{ol} * [\exp(\frac{q(V_{out} + i_{pv}R_{s,total})}{N n_l k T}) - 1] - \frac{V_{out} + i_{pv}R_{s,total}}{N R_{sh}} \quad (2.2)$$

When an illuminated solar cell is short-circuited, the voltage across the solar cell will be zero, and Equation (2.1) becomes:

$$i_{sc} = i_{ph} - i_{o1} * [\exp(\frac{q(i_{sc}R_s)}{n_1kT}) - 1] - \frac{i_{sc}R_s}{R_{sh}} \quad (2.3)$$

where, i_{sc} is short-circuit current flowing across the output terminals of a solar cell.

Often, assumptions on $i_{ph} \gg i_{d1}$ and $R_{sh} \gg i_{sc}R_s$ are made and the generated photocurrent can be approximated as:

$$i_{sc} \approx i_{ph}$$

As seen in Figure 2.7, i_{sc} increased at stronger illumination levels. This corresponds to the increased generation of electron-hole pairs in the solar cell.

When a solar cell is in an open circuit, Equation (2.1) becomes:

$$0 = i_{ph} - i_{o1} * [\exp(\frac{q(V_{oc})}{nkT}) - 1] - \frac{V_{oc}}{R_{sh}} \quad (2.4)$$

where, V_{oc} is open-circuit voltage derived across the output terminals of a solar cell.

When a matched load is connected to the solar cell at a maintained irradiance level, the solar cell will be operating at its maximum power point (MPP):

$$i_{mpp} = i_{ph} - i_{o1} * [\exp(\frac{q(V_{mpp} + i_{mpp}R_s)}{n_1kT}) - 1] - \frac{V_{mpp} + i_{mpp}R_s}{R_{sh}} \quad (2.5)$$

Operating the solar cell close at MPP will ensure almost all the power harvested by the solar cell to be extracted by the electrical loads. This will improve the overall efficiency of a solar energy harvesting system. However, because of the

non-linear operation of the solar panel, a portion of the harvested power will be allocated for the control circuit to track the maximum power point.

2.2 Maximum Power Point Tracking (MPPT)

In [36], a solar energy harvesting system without a sophisticated MPPT circuitry is proposed. The energy harvesting system employs the SolarWorld 4-4.0-100 solar module as its energy transducer. The electrical operating characteristics of the solar module is shown in Figure 2.8.

Figure 2.8 Operating characteristics of solar module from [36]

From Figure 2.8, it can be identified that the maximum power of the solar module is between 2.7V and 2.9V. Hence, the author clamped the output terminals of the solar module directly to a battery (operating between 2.7V to 3.0V). This forces the solar module to operate at its MPP with respect to the fixed output voltage of the battery without any interface MPPT circuitry. This approach eliminates the need of additional power-hungry circuitry to ensure the solar module is operating at its MPP.

However, this method only applies when battery is the energy storage medium. Capacitive mediums such as supercapacitors cannot be used in the proposed method. This is due the fact that supercapacitors typically have a wider voltage range than batteries. A supercapacitor with a voltage range of 0V to 5V will not be able to ensure the solar module is operating at its MPP. In addition, the MPP voltages of a solar module under different operating environment has a wider operating voltage in comparison. This wide electrical operating characteristics of a solar module under indoor conditions are shown in Chapter 3.

In [37], the authors suggested power-hungry MPPT circuitries are not suitable for solar transducers that harness from a lower illumination intensity. Hence, a semi-MPPT design employing two voltage comparators is adopted as shown in Figure 2.9.

Figure 2.9 Schematic of low-light energy harvesting system from [37]

As opposed to the approach in [36], in [37], the solar module is hard-wired to the supercapacitor as the energy storage element. The two voltage comparators determine the on and off voltages for the wireless sensor node. Once the voltage magnitude in the supercapacitor exceeds the higher threshold voltage of the

comparator, the wireless sensor node starts to perform the sensing and transmitting of data back to the base station. This active mode of the sensor node continues till the voltage in the supercapacitor is pulled down to the lower threshold voltage of the comparator. The sensor node is turned off and the supercapacitor will be recharged by the solar module. Hence, the sensor node is operating within a small window determined by the voltage of the supercapacitor as shown in Figure 2.10. By adjusting the threshold voltages of the two comparators, the solar module can be forced to operate around its maximum power point voltages.

Figure 2.10 Operating frequency of wireless sensor node based on “Voltage Window” semi-MPPT energy harvesting from [37]

However, from Figure 2.10, it can be identified that the output voltage of the energy harvesting circuit is not a regulated voltage. The ‘active time’ and ‘recovery time’ of the supercapacitor determines the operation time and frequency of the wireless sensor node. A smaller supercapacitor capacitor or higher input power of the solar module will result in the wireless sensor operating more frequently. Similarly to the shortcoming in [36], when the energy harvesting system is deployed in a wider operating environment condition, the sensor node will not be able to operate at all when the voltage magnitude of the supercapacitor drops below the lower threshold voltage of voltage comparator.

Hence, to maximize the energy harvesting efficiency of a photovoltaic system, an MPPT circuitry must be installed to ensure that the deployed energy transducer is operating at its respective MPP. In the literature, there are many types of MPPT, which can be grouped into 'Direct' and 'Indirect' methods [15]. The direct methods (true seeking methods) obtain the actual Maximum Power Point (MPP) by measuring the output voltage and current of a solar panel. The measurement and computation are often achieved with the use of energy-hungry microcontroller and analog-to-digital controller (ADC). Conversely, for indirect methods (quasi seeks), MPP is estimated from the output voltage and current of a solar panel, the irradiance, or by mathematical expressions of numerical approximations. Without the use of the energy-hungry devices, indirect methods are relatively simpler to implement and a larger portion of the harvested energy can be delivered to the load. Hence, indirect MPPT methods are often desirable for low power energy harvesting system.

2.2.1 Perturb and Observe (P&O) Method

The P&O method [16-20] is one of the most commonly implemented MPPT algorithms to directly seek the MPP of a deployed PV system. It is an iterative method of locating the MPP of the PV module. It perturbs and measures the PV output characteristics periodically. Then it compares the PV output power (P_{PV}) and voltage (V_{PV}) before and after the perturbation towards the maximum point when $dP_{PV}/dV_{PV} = 0$ is reached. This process is repeated until V_{MP} is located and the system oscillates around this MPP. The algorithm is summarized in Table 2.2. Although the algorithm is commonly implemented with a DSP [17, 20], a simple analog P&O MPPT can be implemented to a DC-DC converter by adding an analog multiplier and some comparators [16, 21].

Table 2.2 Algorithm of perturb and observe

$V_{PV}(t) - V_{PV}(t-1)$	$P_{PV}(t) - P_{PV}(t-1)$	Next V_{PV} perturbation
Decrease	Increase	Decrease
Decrease	Decrease	Increase
Increase	Increase	Increase
Increase	Decrease	Decrease

In [38], a pulse frequency modulation (PFM) power management unit (PMU) with ‘Direct’ MPPT capability solar energy harvesting system is proposed. The block diagram of the proposed approach is shown in Figure 2.11.

Figure 2.11 Block diagram of solar energy harvesting system with PFM controlled buck converter with MPPT from [38]

The proposed energy harvesting system exploits the microcontroller on the sensor node to regulate the frequency and duty cycle of the buck converter to ensure the solar module is operating at the optimum maximum power point. From the figure, it can be identified that a current sensor is employed to sense the output current of the solar module. Subsequently the sensed current and voltage of the solar module is processed by the MCU to determine the P&O algorithm. The MCU will constantly monitor the output of the solar module and control the PFM

controller to ensure the Buck converter is emulating the optimum impedance for the solar module. The proposed system is able to achieve a 80 % efficiency at 20mW input power. The MPPT circuitry consisting of the current sensor, MCU and PFM controller, requires 5mW to be operational. This complex and power-hungry MPPT circuitry is not suitable for indoor light energy harvesting, as the output power of a solar module under indoor condition is $100\mu\text{W}/\text{cm}^2$ (Table 2.1).

In addition, the main disadvantages of the P&O method are the speed of MPP tracking and steady-state oscillations. With fixed perturbation values, multiple steps are required for the system to locate the MPP with the changing environmental conditions. Larger step values will allow the system to locate the MPP at a faster rate but also cause higher steady-state oscillations, which will result in unnecessary power loss. To achieve a more efficient MPP tracking speed with minimal oscillation, a variable step P&O method [22] can be implemented with a more complex algorithm and additional hardware.

2.2.2 Incremental Conductance (INC) Method

The INC method [23-28] is based on differentiating of the P_{PV} with respect to V_{PV} and setting this result to zero. This chain differentiation is expressed as Equation (2.6).

$$\frac{1}{V_{PV}} \frac{dP_{PV}}{dV_{PV}} = \frac{I_{PV}}{V_{PV}} + \frac{dI_{PV}}{dV_{PV}} \cong \frac{I_{PV}}{V_{PV}} + \frac{\Delta I_{PV}}{\Delta V_{PV}} = 0 \quad (2.6)$$

hence,

$$\frac{I_{PV}}{V_{PV}} = - \frac{\Delta I_{PV}}{\Delta V_{PV}} \quad (2.7)$$

From Equation (2.7), the location of the MPP can be determined by comparing

the value between the instantaneous conductance $\frac{I_{PV}}{V_{PV}}$ and the incremental conductance $\frac{\Delta I_{PV}}{\Delta V_{PV}}$. The algorithm is summarized in Table 2.3.

Table 2.3 Algorithm of incremental conductance

$V_{PV}(t-1)$	Location of V_{MP}	Next V_{PV} perturbation
$\frac{\Delta I_{PV}}{\Delta V_{PV}} > -\frac{I_{PV}}{V_{PV}}$	$V_{PV} < V_{MP}$	Decrease
$\frac{\Delta I_{PV}}{\Delta V_{PV}} < -\frac{I_{PV}}{V_{PV}}$	$V_{PV} > V_{MP}$	Increase
$\frac{\Delta I_{PV}}{\Delta V_{PV}} = -\frac{I_{PV}}{V_{PV}}$	$V_{PV} = V_{MP}$	Increase

Theoretically, the INC method is able to locate the exact location of the MPP with no oscillations. However, practically it is not possible for the PV system to determine the exact MPP with progressive steps. Hence, similar to P&O method, the INC-method PV system will oscillate around the MPP, suffering the same drawbacks. Furthermore, with the involvement of more complex computations and variables, this method will also require even more hardware than the former.

2.2.2 Fractional Open-Circuit Voltage (FOCV) Method

The *FOCV* [29-34] is an indirect method that exploits the approximately linear relationship between the PV module output voltage at maximum power point (V_{MP}) and its open-circuit voltage, V_{OC} . This relationship is expressed as Equation (2.8).

$$V_{MP} = k_{OC} V_{OC} \quad (2.8)$$

The voltage proportional constant, k_{OC} , is mainly dependent on the PV fill factor, the cell technology and environmental conditions. Hence, this proportional constant is unique to each individual PV module that is being deployed and must be determined beforehand. Studies have shown that the value of k_{OC} ranges from 0.7 to 0.8.

To determine V_{MP} , the PV module is periodically disconnected from the power converter to measure its open-circuit voltage. Subsequently, the MPP is tracked and maintained based on the relationship given in Equation (2.8). This method can be easily employed with minimal hardware requirements. Hence, FOCV approach is often employed for harvesting indoor light energy. However, as the *FOCV* method is only an approximation, the accuracy of this method is generally lower than the other tracking methods. Furthermore, the periodic disconnection of the PV module from the power converter will cause a temporary loss of input power to the system.

2.3 DC-DC Converter Topologies

As mentioned in Section 2.1, the DC-DC converter is used as an interface circuit to perform impedance matching between the PV module and the electrical load. Depending on the designed MPPT approach, the duty cycle of the power converter will be programmed accordingly to ensure its input resistance is matched with the MPP of the solar module. The power converter is also used to regulate the voltage for the connected electrical loads. In the discussion of this section, the primary input power of the power converter is the ambient energy.

The intermittent nature of ambient energy led to a decrease in the reliability of the power system when deploying a specific renewable energy as the sole input power source. These ambient energies should either be combined together

or be used to supplement conventional power supplies to effectively provide an uninterruptable power supply for the electrical system.

Hence, multiple-input (MI) converters can be used to achieve the combination of multiple input power source. The most straightforward method is to employ multiple single-input converter to regulate the different input power supply and have the output of the power converters be either connected in series [39], parallel [40] or through the use of magnetic coupling of an inductor [41]. The three different types of connections for the output of the converter are shown in Figure 2.12. Regardless of the connection topology, it can clearly be seen that the shortcoming of this approach is the increase in the number of components. For every additional input power supply, an inductor and two power MOSFET are required to regulate the input voltage to the magnitude usable by the output load. Here, to reduce the additional cost and size for multiple input power supplies for a designed system, a single-inductor multiple-input (SIMI) converter can be employed.

(b)

(c)

Figure 2.12 Multiple-input converters with: (a) series connection adapted from [39] (b) parallel connection adapted from [40] (c) magnetic coupling adapted from [41]

As the name of the SIMI converter suggested, only one inductor is used in the converter to combine the multiple input power source. Similarly, the multiple input of the power converter can be either connected in series [42], parallel [43] or through the use of isolated half-bridge topology [44]. The three different types of connections for the input of the converter are shown in Figure 2.13.

(a)

(b)

Figure 2.13 Multiple-input single-inductor converters with: (a) series connection adapted from [42] (b) parallel connection adapted from [43] (c) half-bridge topology adapted from [44]

From Figure 2.13 (a), the multiple input power sources of the series-connected SIMI converter are isolated from one another through the presence of the diodes. In addition, for the isolation of the multiple inputs to be valid, the magnitude of V_1 must be greater than that of V_2 . Hence, when switch M_1 is switched on, the current from V_1 will be prevented by D_1 to flow to V_2 . Also when switch M_1 is switched off, the current from V_2 will be prevented by the body-diode of M_1 to flow to V_1 . In addition, switches M_1 and M_2 can be switched on simultaneously to allow both the input power sources to energize the inductor of the power converter [45].

Conversely, the switches in parallel connected SIMI converter cannot be switched on simultaneously and the switches have to be forward conduction and bidirectional blocking (FCBB). This is to ensure the multiple input power sources will not short-circuit one another. The half-bridge topology for SIMI converter is not suitable for ultra-low power energy harvesting system, as the introduction of

transformer will increase the size and cost of the circuit design.

With the understanding of requiring multiple input power sources to strengthen the reliability of an energy harvesting system, the designed EH system must also cater for the multiple voltage domains required by the wireless sensor node. In [46], the CMOS image sensor requires a regulated 1.8V power supply for its analog core and a regulated 0.8V power supply for its digital core. Hence, multiple output (MO) converters are required to perform the multiple output voltage regulation. The straightforward method is to deploy a low dropout (LDO) voltage regulator at the respective output voltage domain. However, it is a well-known fact that a large drop out voltage of the LDO will reduce the efficiency of the LDO. In the case of the 1.8V and 0.8V regulated voltages for the designed system, the LDO can only achieve an efficiency of 44 %. This large drop out voltage of the LDO will significantly reduce the overall efficiency of a power system.

Hence, the bang-bang approach [47-49] is proposed for regulating multiple output voltage on a single power converter. The implementation of bang-bang approach with a boost converter in [47] is shown in Figure 2.14. From the figure, it can be identified that each of the output voltage of the power converter is compared against its respective reference voltage. In a single switching cycle, the inductor will transfer the stored energy only to the first output voltage V_{O1} . At the same time, this V_{O1} is compared against its reference voltage. In the same switching cycle, when V_{O1} has reached its reference voltage, switch S_2 for V_{O2} will be switched on and the remaining energy stored in the inductor will be transferred to V_{O2} and V_{O2} is compared against its respective reference voltage.

This sequence is repeated until the last output of the power converter and the output voltage will be feedback to the PWM controller to regulate the duty cycle of the power converter. Using this bang-bang approach, a SIMO converter with

Figure 2.14 Architecture of single-inductor multiple-output converter adapted from [47].

multiple outputs can be achieved. However, it is worthy to note that cross-regulations exist in the multiple outputs of the continuous-conduction-mode (CCM) power converter. To minimize the cross-regulation effect and in an ultra-low power energy harvesting system for light load condition, discontinuous-conduction-mode (DCM) converter is often preferred so as to achieve a higher efficiency in operating the power converter [50-52].

With the intuitive understanding on the requirement of power converter for ultra-low power indoor light energy harvesting system, the following sub-sections will analyses and describe the DC-DC converter topologies that are related in this research. The voltage transfer function and duty ratio of the DC-DC converters will be derive in detail. In addition, the following discussions will assume the DC-DC converters to be operating under Pulse-Width-Modulation (PWM) in Discontinuous-Conduction-Mode (DCM).

2.3.1 Buck Converter

The generic DC-DC buck converter is as shown in Figure 2.15. Under PWM control in DCM operation, the buck converter has three phases of operation. The output voltage of the buck converter is controlled by controlling the duty cycle of the two switches. The output of the buck converter is always lower than the input voltage. The DC voltage transfer function, M_V , between V_{IN} and V_{OUT} with respect D is expressed as Equation (2.9). The duty cycle of the first phase of operation with respect to the input resistance is expressed as Equation (2.10).

$$M_V = \frac{V_{OUT}}{V_{IN}}$$

$$= \frac{D_1}{D_1 + D_2} \quad (2.9)$$

$$D_1 = \sqrt{\frac{2f_s L}{R_{IN}(1 - M_V)}} \quad (2.10)$$

Figure 2.15 Schematic of a buck converter

2.3.2 Boost Converter

The generic DC-DC boost converter is as shown in Figure 2.16. Under PWM

control in DCM operation, the boost converter has three phases of operation. The output voltage of the boost converter is controlled by controlling the duty cycle of the two switches. The output of the boost converter is always greater than the input voltage. The DC voltage transfer function, M_V , between V_{IN} and V_{OUT} with respect D is expressed as Equation (2.11). The duty cycle of the first phase of operation with respect to the input resistance is expressed as Equation (2.12).

$$M_V = \frac{V_{OUT}}{V_{IN}}$$

$$= \frac{D_1 + D_2}{D_2} \quad (2.11)$$

$$D_1 = \sqrt{\frac{2f_s L (M_V - 1)}{M_V R_{IN}}} \quad (2.12)$$

Figure 2.16 Schematic of a boost converter

2.3.3 Inverting Buck-Boost Converter

The generic DC-DC inverting buck-boost converter is as shown in Figure 2.17. During DCM operation, the converter will have an output voltage, V_{OUT} ,

magnitude that can be either greater than or less than the input voltage, V_{IN} . However, the polarity of V_{OUT} is always negative with respect to the grounding of the converter. The negative output of an inverting buck-boost converter cannot be directly used to supply power to modern electronic circuits. This voltage magnitude is controlled by adjusting the duty cycle, D of the switches, M_1 and M_2 . The relationship between V_{IN} and V_{OUT} with respect D is expressed as Equation (2.13).

$$V_{IN} = \frac{1-D}{D} V_{OUT} \quad (2.13)$$

Figure 2.17 Schematic of an inverting buck-boost converter

2.3.4 Four Transistor Non-inverting Buck-Boost (4TBB) Converter

The non-inverting buck-boost converter consisting of four switching transistor is shown in Figure 2.18. In this section, the continuous conduction mode (CCM) and discontinuous conduction mode (DCM) of the non-inverting buck-boost 4TBB converter will be discussed. The continuous conduction mode is typically employed for DC-DC converters that are handling heavy loads, while discontinuous conduction mode is configured for light load DC-DC converters.

Figure 2.18 Schematic of a 4TBB converter

Under CCM and pulse-width modulation control, the converter has two phases of operation. Referring to Figure 2.19, during the first phase, where $0 < t \leq DT$, M_1 and M_2 are turned on, while M_3 and M_4 are turned off. The input voltage, V_{IN} is applied across the inductor, L and charges the inductor linearly with a slope of $\frac{V_{IN}}{L}$. With M_3 and M_4 turned off, the load is supplied by the output capacitor, C_{OUT} . Hence, the peak-to-peak value of the inductor ripple current, Δi_{L+} at $0 < t \leq DT$ is expressed as Equation (2.14).

$$\Delta i_{L+} = \frac{V_{IN}}{L} DT \quad (2.14)$$

Referring to Figure 2.20, during the second phase, where $DT < t \leq T$, M_3 and M_4 are turned on, while M_1 and M_2 are turned off. With M_3 and M_4 are turned on, L is connected with the load. The energy stored in L is supplied to the load and charging C_{OUT} at the same time. Hence, the peak-to-peak value of the inductor ripple current, Δi_{L-} at $DT < t \leq T$ is expressed as Equation (2.15).

$$\Delta i_{L-} = -\frac{V_{OUT}}{L} (1-D)T \quad (2.15)$$

Therefore, using the inductor volt-second balance, the DC voltage transfer function of the four transistor non-inverting buck-boost converter under CCM is expressed as Equation (2.16).

$$\Delta i_{L+} + \Delta i_{L-} = 0$$

$$V_{IN} = \frac{1-D}{D} V_{OUT} \quad (2.16)$$

Figure 2.19 Phase 1 operation of a 4TBB converter

Figure 2.20 Phase 2 operation of a 4TBB converter

Under DCM and pulse-width modulation control, the converter has a third phase of operation in addition to the two phases discussed above. The third phase

operation is shown in Figure 2.21. During the third phase, where $D_2T < t \leq T$, M_1, M_2, M_3 and M_4 are all turned off. The load is supplied by the output capacitor, C_{OUT} .

Figure 2.21 Phase 3 operation of a 4TB converter

The Δi_{L+} at $0 < t \leq D_1T$ is expressed as Equation (2.17).

$$\Delta i_{L+} = \frac{V_{IN}}{L} D_1 T \quad (2.17)$$

The Δi_{L-} at $D_1T < t \leq D_2T$ is expressed as Equation (2.18).

$$\Delta i_{L-} = -\frac{V_{OUT}}{L} D_2 T \quad (2.18)$$

Therefore, using the inductor volt-second balance, the DC voltage transfer function of the four transistor non-inverting buck-boost converter under DCM is expressed as Equation (2.19). The duty cycle of the first phase of operation with respect to the input resistance is expressed as Equation (2.20).

$$\Delta i_{L+} + \Delta i_{L-} = 0$$

$$M_V = \frac{V_{OUT}}{V_{IN}} = \frac{D_1}{D_2} \quad (2.19)$$

$$D_1 = \sqrt{\frac{2f_s L}{R_{IN}}} \quad (2.20)$$

In addition, from the 4TBB converter, aside from the buck-boost mode operation, the 4TBB converter can also be operated in its individual buck or boost mode. Hence, there is a need to discuss the efficiency of a 4TBB converter operating under these three modes.

Using the electrical characteristics of an indoor solar module and the derivations from Equation (2.10), (2.12) and (2.20), a plot showing the voltage gain of the 4TBB against its respective duty cycle is shown in Figure 2.22. It is known that larger duty cycles will lead to higher conduction power losses. From Figure 2.22, it can be identified that operating the 4TBB converter in boost mode is more efficient than operating the converter in buck-boost mode when the voltage gain of the converter is designed to be more than 1. Similarly, operating the 4TBB converter in buck mode is able to achieve a larger voltage gain of the power converter as compared to operating in buck-boost mode.

Figure 2.22 Plot of voltage gain vs. duty cycle of 4TBB converter operating in DCM

Figure 2.23 Plot of output voltage vs. duty cycle of 4TBB converter operating in DCM

As discussed in this section, most electronic products require multiple voltage domains for operation. Hence, a plot of varying output voltage of the 4TBB converter against the corresponding duty cycle is shown in Figure 2.23. From the figure, it can be observed that the varying output voltage has no effect on the duty cycle when the 4TBB converter is operating in buck-boost mode. This corresponds to the derivation of Equation (2.20). When the converter is operating in buck-boost mode, only the resistance of the input power supply will affect the duty cycle of the converter. This implies that a 4TBB power converter can effectively provide the optimum impedance matching capability to a solar module while being able to regulate multiple output voltage required by the connected load.

2.3.5 Single Inductor Multiple Input Multiple Output (SIMIMO) Converter

Renewable energy is clean and abundant, hence using this energy source to

replace conventional batteries to power wireless micro-sensor networks is an ideal solution. However, their availability and energy densities vary greatly with the environment. This intermittent nature shows renewable energy to be an unreliable power source for the sensors when deployed alone. An effective EH system should combine both renewable and conventional energy sources to guarantee an undisrupted power supply. Such a system will require a power converter with multiple inputs and multiple outputs. A generic SIMIMO converter [35, 36] is shown in Figure 2.24.

Figure 2.24 Generic schmeatic of a SIMIMO DC-DC Converter

This converter only requires one inductor to perform regulated power transfer between the energy transducers, battery and RF load. With proper design of a SIMIMO in a EH system, the converter is able to: (1) adapt the optimum electrical operating point of the energy transducer to the ambient environment; (2) supply power directly from the energy transducer to the RF load; (3) supply

power from the energy transducer to recharge the battery, and (4) transfer power from the power to supply the RF load when the energy transducer fails to generate sufficient power.

2.4 Effects of Sub-Optimal Dead-Time

A traditional asynchronous PWM buck DC-DC converter circuit is shown as an example for effects of sub-optimal dead-time control in Figure 2.25 (a). The converter consists of a power MOSFET used as a controllable switch M_P , a diode D_N , an inductor L , and a output filter capacitor C_{OUT} . The typical voltage drop across the diode D_N is 0.7V which results in a higher conduction power loss as compared to the use of a power MOSFET. Hence, this inefficient diode rectifier is commonly replaced by a synchronous rectifier as shown in Figure 2.25 (b). The voltage drop across the controllable power MOSFET M_N is significantly lower than D_N , resulting in a lower conduction power loss. However, M_N has to be controlled by a programmable pulse with optimal dead-time, t_{DT1} and t_{DT2} inserted between the ON time of the high-side PMOS and the low-side NMOS. This can be identified in Figure 2.26 (c). Failure to provide the optimum dead-time control, t_{D-OPT} to the low-side power MOSFET will result in additional power loss through effects such as shoot-through current loss, body-diode conduction loss and C_{PAR} discharging loss.

Figure 2.25 Circuit diagram of DC-DC buck converter: (a) asynchronous buck converter (b) synchronous buck converter (c) gate timing diagram

2.4.1 Shoot-Through Current Loss

The shoot-through current loss occurs when $t_{DT1} < 0$, switch M_N is switched on before switching off M_P completely. This phenomenon is illustrated in Figure 2.26. When both of the switches are switched on, a large shoot-through current will flow from the input voltage supply to the ground through the low ON resistance of the switches. The large shoot-through current will not only introduce a large power loss to the converter; it may also permanently damage the functionality of the DC-DC converter. Hence, the low-side power MOSFET must never be switched on before switching off the high-side power MOSFET.

Figure 2.26 Shoot-through current ($t_{DT1} < 0$)

2.4.2 Body-Diode Conduction Loss

The body-diode conduction loss phenomenon is illustrated in Figure 2.27. This loss occurs when $t_{DT1} > t_{D-OPT}$, switch M_N is not switched on near the optimal dead-time and the inductor current continues to discharge to the output capacitor. This phenomenon causes the node voltage V_X to be pulled below ground by about 0.7V, implying the conduction of the body-diode, D_N of switch M_N .

Figure 2.27 Body-diode conduction ($t_{DT1} > t_{D-OPT}$)

The conduction of the body-diode of M_N will lead to the conduction loss and degrades the overall efficiency of the DC-DC converter.

2.4.3 C_{PAR} Discharging Loss

When switch M_P is switched off but switch M_N is yet to be switched on, the charges stored at the parasitic capacitance of node V_X are being discharged to the output capacitor by the inductor current through the body diode of M_N . However, when $0 < t_{DT1} < t_{D-OPT}$, the above-mentioned charge transfer phenomenon is skipped as switch M_N will pull node V_X to ground as illustrated in Figure 2.28. The charges in the parasitic capacitance of node V_X will be fully discharged through the low resistive path of switch M_N instead of being transferred to the output capacitor. This will introduce C_{PAR} discharging loss and degrades the overall efficiency of the power converter.

Figure 2.28 C_{PAR} discharging loss ($0 < t_{DT1} < t_{D-OPT}$)

2.5 Dead-Time Controller (DTC)

The above section demonstrates that inaccurate dead-time for the low-side switch of a buck DC-DC converter will degrade the overall performance and efficiency of the converter. Hence the literature review to mitigate the effects of sub-optimal dead-time with the use of a dead-time controller can be classified into three categories [53], namely, (1) fixed DTC, (2) adaptive DTC and (3) predictive DTC. Traditionally, the DTC employs long fixed dead-time approach to cater to the worst case scenarios for the operating DC-DC converter. Hence, for most of the operation of the DC-DC converter, the body-diode of the power MOSFET is conducting in every switching cycle. This will result in a sub-optimal design, as the long period of body-diode conduction will degrade the efficiency of the power converter.

Subsequently, the design of DTC begins to move towards the employing of adaptive approach [54-56], where the dead-time for the power MOSFET can be adjusted in accordance to the variations of the output load current. The current mode adaptive DTC [57] is able to perform calibration of the dead-time during the soft-start of the power converter. However, an integrated load current sensor is required to accurately optimize the dead-time and the current sensor has to be process, voltage and temperature (PVT) independent for the DTC to be robust.

Other than the current-mode adaptive DTC, shown in Figure 2.29, employing cross-coupled topology on the gate voltages of the power MOSFET is able to eliminate the possibility of shoot-through current. However, this is not the solution to achieve an optimal dead-time control. To fully achieve an optimum dead-time control, high speed comparators that detects the node voltage V_X is proposed [58-60]. However, the drawbacks of employing high speed comparators will be introducing high quiescent power during steady state condition to the PMU of the design and the subsequent mismatch of the comparator may induce

Figure 2.29 Cross-coupled topology for buck converter

offset voltage that will lead to an inaccurate sensing of zero-crossing at V_X .

Hence, integrating a discrete time fully differential switched capacitor comparator [61] with auto-zeroing is used to improve the accuracy. The tradeoff is that the discrete time switched capacitor requires a high clock frequency for operation. Similarly, this power consuming approach will add burden to the power budget for the PMU. Hence, sensorless dead-time control approach is proposed [62-64]. The dead-times of the power converter are optimized through information obtainable from the converter control loops. The approach used the obtained information to accurately model and calculate the optimal dead-time required by the power MOSFET. The optimization of the dead-time is achieved by either minimizing the input current [65, 66] or the duty ratio of the control loop of the power converter. The tradeoff for this approach is that the processing of the algorithm is time-consuming and hence this approach is not suitable for applications that deal with fast changing load transitions.

When the synchronous power converter is deployed in a varying operating environment, fast optimization of an optimal dead-time is required. Predictive

dead-time approach [67-71] can be employed for such power converters. The predictive approach detects and processes the occurrence of the body-diode conduction obtained from the previous switching cycle to determine the optimum dead-time required for the next switching cycle. Avoiding the usage of analog comparators, in [72, 73], a NOR gate to detect the body-diode conduction and a “One-Step” method to quantize the error signal obtained from current dead-time delay in the time domain. This proposed circuit is shown in Figure 2.30. After which, this signal is being subtracted from the current time delay to obtain the optimized dead-time required by the power converter. The shortcoming of this approach is that the NOR gate cannot prevent cross-conduction and shoot-through current that will degrade or damage the functionality of the power converter.

Figure 2.30 Predictive ONE-step approach for buck converter

2.5.1 Inverter-Based Dead-Time Controller (DTC)

For an indoor light energy harvesting system, the available input power is low and hence the power consumption of the PMU must be maintained as low as possible to increase the efficiency of the power converter. Replacing the power-hungry comparators with inverters to optimize the dead-time required by the

converter will increase the overall efficiency of the converter. The block diagram of the inverter-based DTC employed in a BC is shown in Figure 2.31. Without using high-speed comparators, the two inverters, U_1 and U_2 can be used to detect the zero-crossing of V_X and V_{TH} -crossing of V_{GN} . By sizing the transistors of the inverters appropriately, the threshold voltage of the inverters can be used to detect the above-mentioned signals accurately and be less susceptible to noise at the switching nodes [53, 73-74]. Referring to inverter U_2 , which is used for detecting the V_{TH} -crossing of V_{GN} , the width of the NMOS of U_2 is sized to be larger than its counterpart PMOS to bring the threshold voltage of U_2 to be close to the threshold voltage of the NMOS. This will ensure the V_{TH} -crossing of V_{GN} can be detected accurately. The time difference between the two crossings is detected by a phase detector.

Figure 2.31 Inverter-based DTC with digital delay generator for buck converter

The phase detector will only be activated when the control signal to the low-side switch goes high. When the voltage of node V_X crosses zero before V_{GN} becomes larger than the V_{TH} of U_2 , signal X will arrive before G and DN will be set. The DN signal will decrease the n-bit delay signal that controls the DPWM generator to reduce the dead-time. Similarly, when V_{GN} becomes larger than the V_{TH} of U_2 before the voltage of node V_X crosses zero, signal G will arrive before X and UP will be set. The UP signal will increase the n-bit delay signal that controls the

DPWM generator to increase the dead-time. Although this approach has a very slow response, the dynamic digital controlled DTC does not require quiescent power and hence this approach is suitable for implementing in ultra-low power energy harvesting system.

2.6 Integrated Current Sensor

In Section 2.2, direct MPPT approach requires the information on the output current of the solar module to perform the tracking of the maximum power point of the solar module. Hence, current sensors are often integrated into the design of a solar energy harvesting system.

Conventional current sensing approach will insert a designed sense resistor, R_{SEN} that is in series with the inductor of the power converter. This is illustrated in Figure 2.32. By measuring the voltage drop across the sense resistor, the magnitude of the input current can be known. This method is fairly simple and straight forward but the introduced sense resistor will induce additional conduction loss to the power converter and lower the overall efficiency of the power converter.

Figure 2.32 Series-sense resistor for buck converter

To avoid introducing additional losses to the power converter, lossless current sensing methods are proposed in the literature. The lossless current sensing methods exploit the loss that are already physically present in the circuit system instead of introducing additional losses to the system. In [75], the voltage drop ($V_{DS(on)}$) across the power MOSFET, M_1 can be used to determine the input current to the power converter. The circuit diagram of this method is shown in Figure 2.33. This is due to the fact that when M_1 operates in the deep triode region, it behaves similarly to a resistor that has an equivalent resistance that can be expressed as Equation (2.21).

Figure 2.33 Power MOSFET R_{DS} sensing for buck converter

$$R_{DS} \approx - \frac{1}{\mu_p C_{OX} \left(\frac{W}{L} \right) (|V_{GS}| - |V_{TH}|)} \quad (2.21)$$

, where μ is the mobility, C_{OX} is the oxide capacitance per unit area and V_{TH} is the threshold voltage. However, this current sensing approach is inaccurate as the value of R_{DS} has to be known in order to determine the magnitude of the input current flowing into the power converter. Furthermore, R_{DS} varies greatly with the operating point, process and temperature [76].

The next current sensing technique uses the first-order low pass filter network to

filter the voltage across the inductor of the power converter. The input current to the power converter is then sensed through the equivalent series resistance (ESR) of the inductor. The circuit of this method is shown in Figure 2.34. The voltage across the inductor is given as Equation (2.22).

$$V_L = I_L(R_L + sL) \quad (2.22)$$

Figure 2.34 Current sensing through filtering of voltage across inductor for buck converter

Hence, the sense voltage, V_{SEN} across the capacitor of the low-pass filter can be expressed as Equation (2.23).

$$V_{SEN} = I_L R_L \frac{1 + s(\frac{L}{R_L})}{1 + s(R_L C_1)} \quad (2.23)$$

From Equation (2.23), by equating $\frac{L}{R_L}$ to $R_L C_1$, V_{SEN} can be simplified to

Equation (2.24).

$$V_{SEN} = I_L R_L \quad (2.24)$$

This sensing approach is popular for discrete circuit design as the values of L and R_L can be obtained easily. However in the case for integration circuit design, the RC filter is practically too large for integration.

An observer approach shown in Figure 2.35 [77, 78], uses the inductor voltage to measure the current that is flowing through the inductor. This is based on induced voltage across the inductor as given in Equation (2.25).

$$V_L = L \frac{di}{dt} \quad (2.25)$$

The inductor current can be obtained by integrating the voltage across the inductor over time. It is worthy to note that in this approach, the error present in the integration accumulates over time and will eventually reduce the accuracy of the approximated inductor current.

Figure 2.35 Observer current sensing technique for buck converter

Hence, a more accurate method illustrated in Figure 2.36 that is suitable for integrated circuit design is proposed in [79-81]. This current sensing approach for monolithic converter uses a SENSEFET (M_{SEN}) to copy the current flowing through the power MOSFET, M_I . The W/L ratio ($\frac{W_{SEN}}{L_{SEN}}$) of the M_{SEN} is N times smaller than M_I ($\frac{W_1}{L_1}$). This will reduce the power loss due to the introduction of a SENSEFET into the system. By forcing the node voltages of M_{SEN} to be equal to M_I , the current flowing through M_I can be copied to M_{SEN} at a ratio that is N times ($\frac{W_{SEN}/W_1}{L_{SEN}/L_1}$) smaller. The sense current, I_{SEN} will then flow through a designed resistor for the detection of V_{SEN} . Proper layout techniques are required to ensure matching between the SENSEFET and the power MOSFET to achieve an accurate sensing of the current flowing through the power MOSFET.

Figure 2.36 SENSEFET for buck converter

2.7 Flexible Energy Harvesting System (FEH)

In the recent years, wearable health sensors that are able to sense heart rate, blood

pressure, blood oxygenation and body temperature have been implemented on flexible devices [82]. These wearable flexible biomedical monitoring are lightweight, unobtrusive, and are dependent on primary batteries for power supply. The power is utilized by the biomedical sensor to sense and analyses the data collected, and display or transmit the result to the user or medical authorities. The power required by this sophisticated operation of the biomedical sensor range from the order of microwatts to a few milliwatts [82, 83]. To avoid the replacement of battery for the sensor or plugging wires to recharge the battery of the sensor, the on-board energy storage of the biomedical sensor must have enough capacity to sustain the sensor to operate throughout its lifetime. A battery with larger energy capacity will translate to an increase in the form-factor of the sensor and may eventually be obtrusive to comfort and mobility of the wearer. Hence, employing energy harvesting for the power supply poses as a promising solution to ensure the biomedical sensor is able to operate throughout its lifetime.

Photovoltaics transducer can be a favorable device to supplement the on-board energy storage of the biomedical sensor. As discussed in Section 2.1, the power density of light energy harvesting under indoor and outdoor conditions is relatively larger than other ambient sources. In addition, the recent development of flexible thin-film solar cells and organic photovoltaics (OPV) [84, 85] are well suited to be employed in wearable applications. These lightweight and flexible energy transducer can be integrated with the wearable biomedical sensors to achieve an autonomous energy harvesting system.

The PMU of a flexible energy harvesting system can be realized by attaching surface-mount technology (SMT) components or flexible components onto a flexible substrate. The flexible printed circuit board is an established technology that provides highly conductive copper interconnects for the components mounted on it. The bulky inductor for the power converter can be incorporated into the flexible energy harvesting system through screen printing [86] or

laminating layers spiral inductor windings onto a flexible substrate [87].

In [88], a bendable energy harvesting system shown in Figure 2.37 is proposed. The bendable EH system comprise of a thin-film OPV module, a thin-film NiMH battery and indium gallium zinc oxide TFT-based capacitive DC-DC converter. The proposed bendable system is able to charge a 6V NiMH battery to full capacity under 1 sun condition. However, the EH system does not possess any MPPT capability and only has an efficiency of 4 %.

Figure 2.37 Schematic diagram and photograph of the bendable energy harvesting system adapted from [88]

In [89], a flexible energy harvesting system shown in Figure 2.38 is proposed. The proposed EH system comprises of a flexible amorphous silicon solar module, Li-ion thin-film batteries and amorphous silicon based thin-film transistor PMU. The system is able to harness indoor light energy and powered up the system designed on the thin-film sheet and to off-sheet load wirelessly. A power of 5mW can be delivered to the on-sheet load at an efficiency of 60 % and 8mW can be delivered to the off-sheet load wirelessly (1mm separation) at an efficiency of 21 %. Similarly, the proposed system uses only TFT technology and hence the circuit design was kept to minimal without the addition of any power-hungry MPPT circuitry.

(a)

(b)

Figure 2.38 A fully thin-film solar energy harvesting system adapted from [89]: (a) system concept and assembly of PMU (b) system prototype

2.8 Conclusion on Literature Review

This chapter has covered the architecture of an energy harvesting system. The operation and electrical characteristics of a solar energy transducer is described in detail. An efficient solar energy harvesting system should be able to adapt the electrical operating point of the solar panel to its *MPP* while operating under any amount of illumination. With knowledge on the equivalent electrical circuit of the solar cell, the performance and functionality of a PV energy harvesting system can be modeled and simulated.

The preferred method to achieve an optimum *MPPT* should be fairly simple to implement while requiring minimal processing energy. This is especially true for indoor light energy harvesting system. Under indoor conditions, the output power density of the solar transducer is approximated to be $100 \mu\text{W}/\text{cm}^2$. FOCV approach that avoid the use of power-hungry MCU or ADC, can be employed for such ultra-low power energy harvesting system. Here, the FOCV method will be employed for the experiment that will be discussed in Chapter 3.

An overview of the inductive-based DC-DC converter is discussed. SIMOMO converters are required to improve the cost, size and efficiency of an energy harvesting system. In addition, it is worthy to note that duty ratio of the buck-boost operation in a 4TBB converter is independent of the variations in the output voltage. Hence, this operation mode is suitable for emulating the optimal resistance of the solar module for maximum power point tracking. The power losses in a synchronous buck converter is also discussed. It is worthy to note that the discussed analysis can be applied for other type of converters. Proper DTC has to be designed to cater for these losses. In an ultra-low power energy harvesting system, DTC with quiescent power consumption will limit the power budget for the PMU. Hence, a dynamic inverter-based DTC can be applied to increase the overall efficiency of the power converter.

Lastly, energy harvesting for an autonomous wearable biometric sensor is discussed. For the wearable biometric sensor to be autonomous, energy harvesting must be adopted to complement the energy storage element in the wireless sensor node. The literature showed that past works has made use of thin-film transistors to achieve a truly flexible energy harvesting circuit. However, MPPT circuitry is not included in the design and the overall efficiency of the system is low. Hence, in Chapter 3, a flexible energy harvesting system for biometric monitoring sensor will be proposed.

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Chapter 3

Flexible Energy Harvesting for Autonomous Wearable Biometric Monitoring

The main aim of this experiment is to investigate the feasibility of a wearable indoor energy harvesting system for powering a biometric monitoring sensor. The selection of renewable power source, designing of power management circuit, selection of energy storage medium and performance of the RF load are described in detail.

3.1 Flexible Energy Harvesting System (FEH)

With the advancement in the technological and medical fields, attention has been placed on integrating medical support to our daily lives. An area of interest is monitoring applications, where wearable biometric wireless monitoring sensors are deployed in a network for health and athlete performance monitoring systems [1-3]. These distributed sensors, typically powered by batteries, would sense and monitor bodily functions then transmit the collected data to a remote data terminal for processing. The main concerns for wearable monitoring systems are that they have to be lightweight, unobtrusive, and be able to operate long durations with minimal power consumption [4]. Various textile-based systems such as wristwatches have appeared in the literature [5-7]. However, these devices often have a bulky black box, containing the electronics and batteries. Although the authors in [8] had come up with an ultra-thin and flexible physiological monitoring system to overcome the constraints, the battery's limited capacity and operating duration [1, 9] remain a challenge due to frequent maintenance and high operational cost.

The introduction of an energy harvesting source is identified as a feasible way to increase the sensor's operating duration. Various works utilizing solar harvesting techniques demonstrated the suitability of indoor PV in supplying energy to small, low-power consuming devices [10, 11]. Nonetheless, these energy harvesting solutions are far from being wearable on human subjects, because they are either too bulky and rigid to be worn [12]. In this chapter, a solution to overcome the above drawbacks is proposed - a self-sustainable wearable body sensor, incorporating indoor PV energy harvesting without requiring any primary batteries for the physiological monitoring of human bodies. This prototype possesses some degree of flexibility when worn, making it more comfortable for users.

The challenge in designing this wearable monitoring system is to ensure that the whole system is able to: (1) conform to the contours of the human body, and (2) harvest sufficient power from the ambient energy to be autonomous.

Figure 3.1 Deployment of developed FEH wireless body sensor prototype on the arm for distributed biometric monitoring

Referring to Figure 3.1, a practical deployment of the developed FEH system with wireless body sensor is shown. It can be seen that certain amount of flexibility of the monitoring system is required for it to conform to the contours of the human body. It is crucial to ensure that upon bending of the proposed FEH system, the harvested ambient power is sufficient for normal operation of the RF load.

3.1.1 Selection of Ambient Energy

The objective of the indoor energy harvesting is to ensure that the operation of the biometric monitoring sensor will not be impeded by battery failure. The indoor ambient energy should be continuous and provide sufficient power to sustain the sensor's operation and recharge the energy storage. Referring to Table

3.1, both light and thermal energy meet the above mentioned requirements. However, a TEG with a rigid and bulky heat sink is rather invasive to the user. On the contrary, a flexible solar energy transducer is able to provide the ability to conform to the body's contours. Hence an off-the-shelf PowerFilm MPT3.6-75 flexible solar panel, made of amorphous silicon on a polymer substrate, with physical dimensions of about 75 mm x 72 mm x 0.5 mm (effective area of solar cells is 60 mm x 72 mm) is chosen as the energy transducer for this experiment.

Table 3.1 Available ambient energy for wearable indoor energy harvesting system

Energy Source	Remarks
Light	Continuous source when the artificial lightings are switched on
Thermal	Continuous source generated from the human body
Vibration	Only available when the human body is moving
RF	Only effective at close proximity to the wireless router

3.1.2 Characterization of Solar Panel

The characterization of the Sundance Solar MPT3.6-75 (60 mm × 72 mm) flexible PV panel operating in an indoor environment is plotted in Figure 3.2. As shown, under the average indoor lighting intensity level of 320 lux, the peak power harvested by the flexible PV panel is 77 μ W.

Figure 3.2 Plotted Power against Voltage (V_{PV}) plot at various lux values

To study the impact of bending, a test setup to measure the effect of the angle of incident G has on the efficiency of PV panel [19] is built and shown in Figure 3.3. As shown, the bent PV follows the arc of a semi-circle. The PV is attached to the body of a circular object to simulate the bending angle. The bending angle is measured by taking the angle between the tangent line at the end of the effective area of the PV and the straight line joining both ends of the PV. This tangent line is perpendicular to the radius of the circular object. The test was conducted in a single-tube fluorescent lighting and air-conditioned laboratory. The illumination intensity was measured using a light meter, Testo545, and four assumptions were made:

1. The PV panel is bent only to a circular shape.
2. Only the incident light to the center of the PV panel, Q1, is measured for the experiment.
3. The temperature throughout the tests remains consistent.
4. The PV is only bent between 0° to 180° .

Figure 3.3 Illustration of solar panel bent at 30°

To understand the relationship between the electrical power harvested from the light source and the bending angle exerted on the PV module, a set of experiments were conducted. Figure 3.4 shows the harvested power from the flexible PV panel with 30° of bending, dropped to about 4.85 % less of a flat PV panel (0.882 V, 76.646 μ W to 0.871V, 72.9 μ W). The bent PV panel shows a decrease in power produced as less surface area of the PV panel is exposed perpendicular to the ceiling lights. However, this power percentage drop is not that large since the laboratory is well-lit.

Figure 3.4 Plotted data of PV panel when subjected to lighting intensity of 320 lux

3.1.3 MPPT Algorithm

Maximum Power Point Tracking (MPPT) is a frequently used technique to vary the optimum electrical operating point of the PV panel. Various MPPT techniques are grouped into 'Direct' or 'Indirect' methods [17]. The direct methods ("true seeking methods") obtain the actual Maximum Power Point (MPP) from the measures of the PV generator's voltage and current. The processing circuitry in the direct methods is normally very energy-consuming.

Conversely, for indirect methods ("quasi seeks"), MPP is estimated from the measures of the PV generator's voltage or current, the irradiance, or by mathematical expressions of numerical approximations. Such indirect methods are relatively simpler and less energy-consuming than the direct methods. One of the indirect method is the constant voltage (CV) method. However, from Figure 3.2, it is observed that there is no single optimum voltage that aligns the maximum power point of the flexible solar panel, Hence, another indirect method, fractional open circuit voltage (FOCV) approach is investigated in this design for suboptimal MPPT of the flexible solar panel.

The maximum power point voltage, $V_{MPP} = K_{oc} * V_{oc}$, where V_{oc} is the open circuit voltage of the PV and K_{oc} is the voltage factor [18]. Referring to Figure 3.5, the measured peak power of the PV panel, without bending in a well-lit and air-conditioned room, is found to be near K_{oc} constant values of ≈ 0.6 . As shown in Figure 3.6, the maximum power points are also approximately aligned along a K_{oc} value of 0.6, at different light intensity and bending angles on the flexible PV panel. Hence, FOCV method will be adopted in the design of this wearable flexible energy harvesting system.

Figure 3.5 Plotted Power against K_{OC} plot at various lux values when panel is flat

Figure 3.6 Plotted power against K_{OC} with the PV panel bent at bending angle of $\approx 30^\circ$ at different light intensity

Figure 3.7 shows a plot of the V_{MPP} against its respective V_{OC} at lighting intensity ranging from 200 lux to 1000 lux and at a bending angle of 30° . These two factors appeared to have a linear correlation and they can be represented by a constant, $K_{OC} = V_{MPP}/V_{OC}$. Hence, the K_{OC} constant of 0.6 is used in the design of the flexible power management circuit.

Figure 3.7 Plotted data of various V_{MPP} against V_{OC}

3.1.4 Transmission Profile of Sensor Node

The wireless biometric sensor node (TI eZ430-RF2500) used in this research is used on the wearer to measure the body temperature and send the information wirelessly back to the base station for post processing and monitoring. The sensor node has an operating voltage range of 1.9 V to 3.6 V, RF range of 2400 MHz to 2483.5 MHz, 2.4GHz chip antenna and physical dimension of 35 mm x 20 mm x 3.5 mm. With the default settings, the sensor node is able to communicate with the access point at a line-of-sight range of 50 meters. The wireless transmission pulse profile of the sensor node is illustrated in Figure 3.8. The description of the transmission pulse is explained in Table 3.2.

Figure 3.8 Transmission profile of a wireless sensor node

Table 3.2 Description of transmission profile

Mode	Description
A	Wake up
B	Temperature Sample
C	Processing
D	PLL (Phase Lock Loops)
E	Rx
F	Rx Switching to Tx
G	Tx
H	Low Power

3.1.5 Design of Power Management Circuit Board

To reduce traditional wearable monitoring systems, which are bulky, rigid and battery dependent, a flexible power management circuit (FPMC) is proposed. A flexible, single sided coppered epoxy base sheet (AN10, C.I.F) was used as the substrate for circuit fabrication. A positive photoresist (AZ9260, Sigma Aldrich)

was spun onto the substrate at 4000 rpm for 30 seconds. Following, soft bake and lithography (P300MT, Fusion UV System) was performed at 105°C for 6 minutes and 10seconds, respectively. Hereafter, post bake at 120°C and develop using photoresist stripper 1:3 (AZ developer, Sigma Aldrich). The various components were then soldered manually using a lead solder. As shown in Figure 3.9, two regions have been identified for the soldering of the surface mount devices.

Figure 3.9 Flexible power management circuit

3.1.6 System Implementation and Integration

The overall circuit diagram of the FEH wireless body sensor for wearable biometric monitoring is depicted in Figure 3.11. As shown in the figure, there are two main parts in the FEH system: (1) power train and (2) sensing, control and drive signals.

The power train carries the harvested electrical power from the solar panel to the supercapacitor and the wireless sensor node (WSN), via a DC-DC boost converter with MPPT capability. The MPPT control provides a simple means of

impedance matching for maximum power extraction. When the solar panel voltage, V_{PV} , falls below the MPPT target reference voltage, V_{REF} , value, the duty cycle of the DC-DC converter will be adjusted accordingly by the gain controller to track V_{PV} , to V_{REF} .

The sample and hold circuit for FOCV algorithm comprises of two N-type MOSFETs (NMOS1 and NMOS2). The open circuit voltage of the solar panel, V_{OC} , is obtained by switching off NMOS1 to isolate the solar panel from rest of the circuit. NMOS2 is switched on to connect V_{OC} to the resistor divider. The ratio of the resistor divider is determined by the K_{OC} constant. The voltage obtained from the resistor divider corresponds to the required V_{MPP} and this voltage is stored in a capacitor. The gate signals are controlled by the preprogrammed timer signal from onboard microcontroller, MSP430, of the sensor node.

Once the open circuit voltage sensing cycle is completed, NMOS2 is switched off to disconnect the sample and hold circuit from the solar panel. NMOS1 is switched on for the MPP tracking operation. During the tracking operation, V_{PV} is compared with the reference voltage, V_{REF} , at the differential amplifier. When V_{PV} is lower than V_{REF} , a positive control voltage signal generated by the differential amplifier will be sent to the voltage comparator (LMC7215).

This positive control signal is compared against a high frequency saw tooth-like waveform generated by a voltage-controlled oscillator (LTC6906) and a RC-filter. The resultant duty cycle with switching frequency of 10 kHz is fed to the NMOS3 switch of the DC-DC boost converter. Hence, the output voltage of the solar panel will always be regulated to V_{MPP} for maximum power extraction.

The flexible solar-based wireless body sensor with its designed fractional open circuit voltage MPPT control circuitry is integrated together into the final FEH

system prototype as shown in Figure 3.10. The power train, sensing, control and drive circuits are fabricated onto a flexible PCB substrate. This allows the energy transducer and PMC to possess a certain degree of flexibility when being worn by a human. The overall system architecture is shown in Figure 3.11.

A supercapacitor, 100 mF, is selected as the energy storage to handle the energy required by the burst transmission mode of the wireless sensor node and MPPT control circuitry. The supercapacitor with millions of charging/discharging cycles, is able to sustain the sensor node operation for a reliable period of time.

Figure 3.10 Final integrated FEH system prototype

Figure 3.11 Flexible energy harvesting wireless body sensor with its designed fractional open circuit voltage MPPT control circuitry

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Chapter 4

Results and Discussion on FEH System

The proposed design of flexible energy harvesting for autonomous wearable biometric monitoring is discussed and designed in Chapter 3. PCB prototypes are fabricated for the purpose of testing and verifying the design. In this chapter, performance results obtained by testing the prototypes with photovoltaic transducers are presented and discussed in detail.

4.1 Analysis on Bending Effects of Flexible Solar Panel

Referring to the experimental results shown in Figure 4.1, it can be seen that the short-circuit current of the PV, I_{SC} , is approximately linear to the bending angle at a gradient, β . An additional parameter, θ , proportional to the incidence light angle is added function of I_{SC} , which is defines as:

$$I_{SC}(\theta) = -\theta\beta + I_{SC} \quad (4.1)$$

Using Least Square method, the β obtained from the various lighting intensities ranged from 0.1304 to 0.139. Hence, taking the average of these values, β was found to be 0.1349. Applying the averaged β to Equation (4.1), the percentage error between the measured and approximated I_{SC} is tabulated in Table.1.

Figure 4.1 Plotted data of PV panel when subjected to different bending angles at various lighting intensity

Referring to Table 4.1, it is observed that the percentage error calculated is lesser than 5 %. The calculated error difference is mainly due to the uneven distribution of light intensity at Q-points of the bent panel. At these Q-points, the average light intensity is lower than at Q1, and this non-linearity will affect the total

Table 4.1 Measured error between true I_{SC} and approximated I_{SC} obtained when $|\beta|=0.1349$

Lux Bending°	300	350	390	500
0	0.00%	0.00%	0.00%	0.00%
66	-4.19%	-2.47%	-3.27%	-2.89%
70	-3.82%	-2.21%	-2.38%	-2.32%
80	-3.40%	-2.03%	-2.76%	-2.48%
114	-1.71%	-0.95%	-0.66%	-0.26%
140	-4.14%	-0.90%	-1.28%	-0.74%
210	-4.23%	0.16%	-0.87%	-1.21%

current generated by the flexible panel. Assuming that $I_L \approx I_{SC}$, the general I-V characteristic equation of a PV module can be written as follows:

$$I_{PV} = I_{SC}(\theta) - I_o \left[\exp\left(\frac{V_{PV} + I_{PV}R_s}{n_s n_i V_t}\right) - 1 \right] - \frac{V_{PV} + I_{PV}R_s}{n_s R_{sh}} \quad (4.2)$$

To validate Equation 4.2, the PV panel was modeled and the simulated power curves were compared with the measured power curves and plotted in Figure 4.2.

Figure 4.2 Measured and simulated power against K_{OC} curves of PV panel bent at different bending angles at 550 lux

In the simulation, the modeled parameter, I_{SC} was obtained from Equation 4.1 and the other parameters (I_0 , R_{SH} , R_s , n_s , n_1 and V_T) were kept constant with respect to the different bending angles. The simulated power curve of the flat PV panel shown in Figure 4.2 was fairly accurate as compared to the measured power curve. It can also be seen that at a bending angle at 80° , the percentage error of the maximum power obtained between the simulated and measured waveform is greater than 7.5 %. The error may be due to the partial shading of the PV panel at large bending angles. However, in the actual deployment of the FEH system in a well-lit environment, the impact of the partial shading should be reduced. In addition, it is impractical to deploy the flexible PV panel at a large bending angle as maintaining a maximum exposure of the effective area of the PV panel to the incident light would imply a higher photon to electric current conversion efficiency. Hence, the proposed equation is suitable to estimate the available power to be harvested from the flexible PV panel with respect to different degrees of bending.

4.2 Flexible Power Management Board

For a higher degree of flexibility, the placement of components on the flexible PCB should be fairly spaced out. Figure 3.8 shows that the placement of components at Region A, which is the axis of symmetry of the board, should be avoided. The solder threads in Region A will experience a higher strain when the FPMC is conformed to the body contour. Larger components (MAX4471, SOIC-8L) should be soldered in Region B to minimize the risk of loosening the soldered components when the PCB is flexed.

4.3 Analysis on Current Consumption of Sensor Node

The measured current consumed by the sensor node at each transition stage of a wireless transmission is tabulated in Table.4.2.

Table 4.2 Current consumed by sensor node during wireless transmission

Mode	Description	Measured current consumption ($\mu\text{A*s}$)
A	Wake up	0.9
B	Temperature Sample	2.496
C	Processing	2.17
D	PLL (Phase Lock Loops)	6.08
E	Rx	9.72
F	Rx Switching to Tx	0.09
G	Tx	17.55
H	Low Power	0.156
Total		39.162 $\mu\text{A*s}$

In every wireless transmission, the average current consumption (I_{ave}) of the sensor node per second transmission period (T_{Tx}) can be expressed as follows:

$$I_{ave} = (I_{sleep} + I_{Tx} \text{ Total}) / T_{Tx} \quad (4.3)$$

$$I_{sleep} = (I_{idle, MSP430} + I_{idle, CC2500}) * (T_{Tx} - T_{app_exe}) \quad (4.4)$$

If the target board transmits less frequently, the average current consumed will be reduced accordingly. Hence, without crippling the efficiency of the biometric monitoring sensor, the minimal data transmission rate required in this paper is set to be once every 60 seconds. The measured current consumed by the modified RF transceiver settings is tabulated in Table 4.3.

Table 4.3 Current consumption across different time transmission periods

Current	Value
I_{Tx} Total	39.162 $\mu A \cdot s$
I_{idle} , MSP430 + I_{idle} , CC25000	1.29993 μA
T_{app_exe}	3.91 ms
I_{sleep} for 1 sec T_x period	1.295 $\mu A \cdot s$
I_{ave} for 1 sec T_x period	40.457 μA
I_{sleep} for 60 sec T_x period	77.996 $\mu A \cdot s$
I_{ave} for 60 sec T_x period	1.953 μA

Referring to Table 4.3, the modified RF transceiver consumes an average current of 20 times less than the original transmission rate. With the modified RF transceiver set to operate at 2 V, the average power consumption of the transceiver is approximated at 3.91 μW . The modification is acceptable as the body temperature does not rise instantly. During burst transmission mode, the transceiver draws energy mainly from the supercapacitor, which is able to manage the sudden current surge. Subsequently, the transceiver goes into sleep mode and the supercapacitor gets recharged by the solar panel.

4.4 Performance of MPPT Control Circuit and FEH System

The power line diagram of the FEH system at 320 lux and with no bending is shown in Figure 4.3. The characterization of the solar panel at 320 lux reflects a true output power of 76.65 μW at MPP. The characterization also showed the true MPP was obtained at $V_{PV} = 0.882$ V. With the implementation of the proposed FEH system, the circuit was able to harvest 67.86 μW from the 320 lux at $V_{PV} = 0.875$ V (measured in Figure 4.3), which is a difference of 11.5 % from the true maximum power.

Figure 4.3 Line power diagram consisting of power harvested and power conversion efficiency of proposed flexible energy harvesting wireless body sensor

This power loss, as a result of the FOCV approach, is justifiable for use in this flexible energy harvesting system. The reason is that it do away with the need of an ADC and a microcontroller as opposed to the conventional Perturb and Observe (Hill-Climbing) approach.

Figure 4.4 shows the performance of the MPPT under intentional shading. As seen, under the intentional shading, the voltage, V_{HOLD} , in the holding capacitor will drop according to the ratio of the K_{OC} . The V_{PV} gradually returns to the MPP after the shading was removed.

As seen in Figure 4.5, when the PV panel is bent at 30°, the harvested power obtained from the FOCV, drops by as much as 20 %, to around 56 μW. Hence, an energy efficient management strategy is needed to harness and use whatever little power available to maintain the operation of the end device.

To achieve a self-sufficient and autonomous flexible energy harvesting wireless body sensor, it is crucial to ensure that the onboard power management circuitry consumes a minimal portion of the total harvested power. A power reduction approach utilized in this proposed FEH system is to operate the system at its minimum DC voltage level of 2 V. By balancing the power exchange between the input harvested power from the solar panel and the output power consumed

Figure 4.4 Flexible solar panel voltage with MPPT control under a lighting intensity of 320 lux

Figure 4.5 Plotted verse Measured Output of flexible PV panel

by the electrical loads including the wireless sensor node and the supercapacitor, the operating DC voltage can be maintained at the desired value.

The power budget of the designed ultra-low power management circuit with closed loop MPPT control to extract maximum available power from the solar panel is tabulated in Table 4.4. It can be observed that 36.39 μW of electrical

Table 4.4 Power consumption of MPPT control circuit

Description	Current (μA)	Voltage (V)	Power (μW)
LTC6906	7.79	2	15.58
LMC7215	4.75	2	9.5
MAX4471	3.7	2	7.4
Total	16.24	2	32.48

power is shared among the electrical loads that connected to the output of the boost converter and $7.71 \mu\text{W}$ of electrical power is leftover to be stored in the supercapacitor.

A sustainability test of the FEH prototype is conducted in a simulated indoor environment with a lighting intensity of 320 lux. The supercapacitor was set to 2.0 V and the FEH system with RF load was bent to a bending angle of 30° . The discharging voltage waveform of the supercapacitor is shown in Figure 4.6. With an ambient indoor lighting condition of 320 lux as the only input power source, it is observed that the supercapacitor voltage maintained in the range of 2 V to 2.1 V for more than 10 hours. Thus, this positive observation validates that the prototype is able to be self-sustainable at any indoor light intensity more than 320 lux.

Referring to Figure 4.6, it was observed that there were some fluctuations in the supercapacitor's voltage waveform. This is because as the supercapacitor gets charged up; its equivalence series resistance (ESR) changes, causing impedance mismatch between the solar panel and the electrical load. FOCV is an approximation technique and the power extracted from the solar panel might not always be at its maximum in relation to the new load. Therefore, power consumed may be more than the power produced, leading to a drop in voltage across the supercapacitor. As the voltage drops, the ESR changes, reaching a point where the power flowing is more than the power consumed, resulting a positive net power to charge up the supercapacitor.

Figure 4.6 Charging voltage waveform of supercapacitor under a constant light source of 320 lux when solar panel is bent at 30° with MPPT control

Another similar experiment was conducted with the solar panel connected directly to the supercapacitor and RF load via a diode. The supercapacitor was set to 2.0 V. As seen in Figure 4.7, the solar panel is unable to sustain sufficient power in the supercapacitor to supply the RF load. The experimental results verified that the proposed FEH system with MPPT is able to sustain the practical deployment of a wireless body sensor node, thus replacing the need of using a battery cell as power source.

Comparing the PV panel with a thermoelectric generator (TEG) [1], direct-current energy transducer counterpart, the PV panel is shown to have a higher power density under indoor environment. Note that the FOCV approach can also be employed to achieve maximum power tracking for the TEG. Light harvesting will be a more suitable energy harvesting choice over thermal harvesting when the wearable monitoring system is to be extended to outdoor environments where the irradiance is 100-1000 times higher.

Figure 4.7 Discharge voltage waveform of supercapacitor under a constant light source of 320 lux when solar panel is bent at 30° without MPPT control

4.5 Conclusion on FEH System

A self-sustainable, flexible energy harvesting (FEH) prototype incorporated with maximum power point tracking (MPPT) and ultra-low power circuitry has been introduced. Self-sustainability using flat, as well as bent PV panels were investigated and experimentally verified. Under constant light intensity of 320 lux in a controlled environment, the prototype is able to operate in a self-sustaining mode, where the supercapacitor voltage maintains at around 2 V–2.1 V, and with the flexible PV panel bending at an angle of 30°.

The comparison of this work with the state-of-the-art solar energy harvesters is shown in Table 4.5. It can be seen that this work has integrated flexibility attributes into the EH system, facilitating wearable and autonomous operation of biometric sensors for the users with a low input power of 77.65 μW . Although this work has reported a lower MPPT efficiency, the proposed system is able to perform optimum maximum power tracking when the energy transducer is bent to different bending angles. The optimized design in the proposed flexible EH system is able to achieve an overall system efficiency of 69.5%. In addition, the

flexible design can be further modified to incorporate other harvesting techniques, to ensure a more robust and reliable EH system for the wireless sensor nodes.

Table 4.5 Comparison with existing flexible energy harvesting system

Parameters	[2]	[3]	[4]	This work
Technology	Discrete	CMOS TFT	Discrete	Discrete
Flexibility	No	Yes	Flexible PV panel	Flexible PV panel + PMU
Topology	Inductive buck	Diode coupled + charge pump	Inductive buck-boost	Inductive boost
Input Power	200 mW	78 mW	94.4 mW	77.65 μ W
MPPT efficiency	89% (FOCV)	N.A	93.7 % (Load current sensing)	88.5% (FOCV + PV panel bend to 30°)
End-to-end efficiency	~79 %	4 %	66.2 %	69.5 %

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Chapter 5

Ultra-Low Power MPPT for SIMIMO Power Converter

The main aim of this experiment is to investigate the feasibility of a ultra-low power MPPT circuitry designed for indoor energy harvesting system for powering a biometric monitoring sensor. The theoretical approach for the proposed MPPT circuitry, design and implementation of the circuitry are described in detail.

5.1 Introduction

The Internet of Things (IoT) has been the new technological paradigm for various research fields [1-2]. In future aspects of medical and healthcare, patients and professionals will be connected seamlessly with IoT applications [3-4]. However, a large number of these autonomous sensors still rely on primary batteries for operation which when depleted, will require replacements. Extra resources will be needed to maintain power. The key requirement for a sensor to be autonomous is to have an uninterrupted power supply.

Energy harvesting is a promising solution for providing power to the miniaturised medical sensors used in wireless body area networks (WBANs) for ubiquitous healthcare [5]. The miniaturised sensors can either be powered by light, mechanical, thermal energy, or a combination of the energies available in the ambient environment [6-10]. However, as discussed in the Chapter 2, light energy has a higher power density among the other ambient energies and is readily available in both outdoor and indoor environments.

Hence, solar energy harvesting will be use to supplement the battery in the sensors to achieve an uninterrupted power supply. This form of harvesting energy from the light source can be accomplished by employing photovoltaic modules. A maximum power point tracking circuitry is customarily required to harvest light energy through PV modules effectively.

There exists numerous techniques to realise a MPPT circuitry [11-14]. The most established techniques are the Perturb and Observe method and Fractional Open-Circuit Voltage method. The FOCV method has already been discussed and implemented in Chapter 3, with two drawbacks observed.

The first drawback is the periodic disconnection of the PV module from the RF load. This momentary disconnection results in a temporary loss of input power to the system. The second shortcoming is that FOCV method only achieves a MPPT efficiency of 88.5 %. This translates to a loss of 11.5% of input power. This loss is a critical flaw for ultra-low power applications.

Hence, there is a need to adopt a new MPPT method that has a tracking efficiency higher than 90 %, and the MPPT circuitry will consume less than 1 % of the input power to maintain its functionality. A continuous operation of the PV panel and a high tracking efficiency can be achieved by employing closed-loop MPPT controller. This close-loop controller is often based on P&O or hill-climbing algorithm. These algorithms require the measurement of the output voltage and current of the PV panel and calculation of the output power of the PV panel to perform the MPPT operation. However, these circuitries often employ analog to digital converters (ADCs) or power inefficient current sensors, increasing the system complexity, degrading the overall power efficiency of the harvesting system and are not suitable to be employed for low-power PV modules. In this chapter, a new, accurate and ultra-low power consumption hill-climbing based MPPT of a digitally-controlled SIMIMO converter for ultra-low power application is presented

An overview of the MPPT circuit with DC-DC power converter for solar energy harvesting is shown in Figure 5.1. From Figure 5.1, the SIMIMO converter is employed as the interface circuit for maximum power extract from the PV panel and output voltage regulation. Using only a single inductor, the energy harvesting system is designed to operate with two input voltages and three output voltages.

Under normal operation, the converter operates in buck-boost mode and discontinuous conduction mode (DCM) to deliver the harvested energy from the PV module to 1V supply, 1.8V supply or the battery. When the harvested energy

is low, the 1V and 1.8V supply rails are no longer regulated to the designed magnitude and hence, they are insufficient to regulate the power requirements for the control circuitries. The converter will operate in buck mode and draw power from the battery to regulate the two supply rails.

Figure 5.1 Overview of solar energy harvesting system with MPPT

5.2 Selection of MPPT Algorithm

A ‘Direct’ MPPT algorithm [15], hill climbing method will be adopted in this experiment to overcome the drawbacks discussed in Section 5.1. Figure 5.2 shows the flow diagram of the hill climbing method. From the flowchart, the instantaneous voltage, V_{PV} and power, V_{PWR} of the PV module are necessary information for the tracking of the optimal MPP. From the literature [16, 21-24], ADCs and/ or power inefficient current sensors are employed to estimate V_{PWR} . Hence, a new technique to estimate V_{PWR} will be discussed in Chapter 5.

Figure 5.2 Flowchart of hill climbing based MPPT system

5.3 Theoretical Approach to Estimate Instantaneous Power of PV Module

A true MPPT algorithm will measure the generated power of the PV module and adapt the electrical operating point of the PV module to the MPP under any light condition. The following sub-sections will discuss a different approach to estimate the instantaneous power of the PV module in detail.

5.3.1 Characteristics of the PV Module

The solar module used in this design is primarily for harvesting light energy from fluorescent lamps under indoor conditions. In comparison to outdoor conditions, the availability of light energy under indoor condition to be harvested is 1000

times smaller and there is a lower variation in the ambient temperature [20]. Using a solar simulator, the power curves of the solar module in relationship with the V_{PV} and load resistance is shown in Figure 5.3 and Figure 5.4 respectively. From the figures, the maximum power operating points of the solar module is non-linear with respect to the illuminance. Hence, there is a need for a MPPT circuitry to continuously adapt the electrical operating point of the solar module to ensure the solar module is delivering its maximum power under a given illuminous intensity.

Figure 5.3 P-V Curves of solar module at different lux. A curve joining the MPPs is also shown.

Figure 5.4 P-R Curves of solar module at different lux. A curve joining the MPPs is also shown.

5.3.2 Emulated Resistance of Power Converter

A SIMIMO power converter will be used to perform the adaptation of the electrical operating point of the PV module. This power converter will be operating in the discontinuous conduction mode to facilitate low-power operation. Figure 5.5 illustrates the first phase DCM operation a single stage SIMIMO converter.

Figure 5.5 Phase 1 operation of a single stage SIMIMO converter with solar panel

During the first phase, where $0 < t \leq D_1T$, M_1 and M_2 are turned on, while M_3 and M_4 are turned off. The voltage of the solar panel, V_{PV} is applied across the inductor, L and charges the inductor linearly with a slope of $\frac{V_{PV}}{L}$. The average input current $\overline{I_{IN}}$ of the converter can be expressed as Equation (5.1).

$$\overline{I_{IN}} = \frac{V_{IN}}{2L} D^2 T \quad (5.1)$$

The input resistance of the power converter, which is also known as the emulated resistance [17-19], R_{EM} of the PV module can be viewed as the ratio of the average input voltage to the average input current. This ratio can be expressed as Equation (5.2).

$$R_{EM} = \frac{\overline{V_{IN}}}{\overline{I_{IN}}} = \frac{2f_s L}{D_1^2} \quad (5.2)$$

From Equation (5.2), it can be seen that the emulated resistance of the PV module is only dependant on the duty cycle of the power transistor, switching frequency of the power transistor and the inductor. The only changing variable with respect to time is D_I . Hence, the effective load resistance of the power converter can be controlled by the duty cycle to emulate the source resistance of the PV module to achieve impedance matching under different illuminance intensity. Thereby, the harvested power from the PV module is always at its optimum MPP.

5.3.3 Relationship between Maximum Power of Solar Module and Inductor Current

From Equation (5.2), input impedance of the power converter is only dependant on three variables. Out of the three variables, only the duty cycle, D_I of transistors, M_I and M_2 is varying with time. Hence, there is a need to investigate on how D_I will affect the converter in emulating the resistance of the solar module. D_I will directly affect the charging current of the inductor. The instantaneous current of the inductor across $0 < t \leq D_I T$ is given as Equation (5.3)

$$i_L(t) = \frac{V_{PV}(t)}{L} D_I T \quad (5.3)$$

From Figure 5.4, using Kirchhoff's first law, the instantaneous current of the inductor across $0 < t \leq D_I T$ can also be given as Equation (5.4).

$$i_L(t) = i_{PV}(t) + i_C(t) \quad (5.4)$$

The voltage waveform at Node A in Figure 5.5 is shown in Figure 5.6.

Figure 5.6 Illustration of the voltage waveform at node A of figure 5.4

At Node A, the capacitor, C_{IN} is charging and discharging periodically. Interval T_C corresponds to the third operation phase of power converter, where $D_2T < t \leq T$, M_1 , M_2 , M_3 and M_4 are all turned off. The PV module is not connected to the power converter, and is charging C_{IN} . This is expressed as Equation (5.5), for $D_2T < t \leq T$.

$$i_{PV}(t) = C_{IN} \frac{dV_{PV}}{dt} \quad (5.5)$$

In addition, the output power, P_{PV} of the PV panel is equal to the incoming power at C_{IN} . This is expressed as Equation (5.6).

$$P_{PV}(t) = V_{PV}(t)i_{PV}(t) = V_{PV}(t)C_{IN} \frac{dV_{PV}}{dt} \quad (5.5)$$

At interval T_D , switches M_1 and M_2 are turn on with a duty ratio of D_1 . The solar module voltage, V_{PV} is applied across the inductor. Assuming the power transistor

are ideal switches, the current across the switches and inductor is given as Equation (5.3).

To reduce the form factor for wearable biometric sensors, the inductance chosen for the power converter will be relatively small. Hence, the PV module alone cannot provide enough current for the inductor during the interval, T_D . The input capacitor, C_{IN} will supplement the PV module for the building up of energy in the inductor. This expression is given as Equation (5.4). The power from the PV module and C_{IN} will be transferred to the inductor. In addition, it is worthy to note that the peak current of the inductor occurs at the end of the interval T_D . Figure 5.7 shows the P-R plot of the PV module and the inductor current waveform plot.

Figure 5.7 P-R and inductor current plots of the solar module

From Figure 5.7, the output power of the PV module is directly correlated to the current flowing through the inductor. The maximum output power of the PV module corresponds to the peak current of the inductor. Therefore, by sensing the inductor current, the output power of the PV module can be approximated.

5.4 Proposed Power Estimator

From Section 5.3, the inductor current is seen to be correlated to the output power of the solar module. Hence, the output power of the solar module can be approximated by sensing Equation (5.3). A solar power sensing circuit is proposed as shown in Figure (5.8). Switches SB_3 , SB_2 , SB_1 and SB_0 are controlled by the digital bits representation of the magnitude of duty cycle D_I in Equation (5.3).

Figure 5.8 Circuit diagram of a PV power estimator

The PV power estimator has three phases of operation. The first phase of operation occurs at the end of the interval T_D , where the current of the inductor is at its peak value. Switch SIN_I will be closed along with switches SB_3 - SB_0 depending on the magnitude of D_I . Hence, the output voltage of the solar module that correspond to the peak inductor current is sampled into the four sampling capacitors. The total charge, Q_{SAMPLE} stored across the sampling capacitors is expressed as Equation (5.6).

$$Q_{SAMPLE} = V_{PV} C (B_3 + \frac{1}{2} B_2 + \frac{1}{4} B_1 + \frac{1}{8} B_0), \text{ where } B_n = '0' / '1' \quad (5.6)$$

Since $(B_3 + \frac{1}{2} B_2 + \frac{1}{4} B_1 + \frac{1}{8} B_0)$ is a constant with respect to the magnitude of D_I , Equation (5.6) can be simplified as Equation (5.7).

$$Q_{SAMPLE} = V_{PV} C \alpha, \text{ where } \alpha \text{ is } 0 < \alpha \leq 1.875 \quad (5.7)$$

In the second phase of operation, Switches SB_3 to SB_0 and SO_{MPP} will be closed. The total charge obtained in phase one will be redistributed among the four sampling capacitors and the holding capacitor, C_P . The resulting holding voltage, V_{HOLD} in C_P is expressed as Equation (5.8).

$$V_{HOLD} = \frac{Q_{SAMPLE}}{\frac{15C}{8} + C_P} \quad (5.8)$$

Assuming $C_P = C$, Equation (5.8) can be rewritten as Equation (5.9).

$$\begin{aligned} V_{HOLD} &= \frac{Q_{SAMPLE}}{\frac{23C}{8}} \\ &= \frac{V_{PV} C \alpha}{\frac{23C}{8}} \\ &= \frac{8V_{PV} \alpha}{23} \end{aligned} \quad (5.9)$$

Comparing Equation (5.3) with (5.9), one can see that V_{HOLD} can be estimated as a function of $i_L(t)$. Hence, the inductor current which correspond to the output power of the solar module can be estimated by the proposed power estimator.

For the third phase of operation of the power estimator, only switch SO_{MPP} will

be closed. The holding voltage in the capacitor, $C_{P(t)}$ which corresponds to the output power of the solar module will be fed to the MPPT circuitry to determine the next perturbation of D_I .

5.5 System Implementation

The overall circuit diagram for the proposed MPPT for solar energy harvesting is depicted in Figure 5.9. The power converter operates in discontinuous conduction mode (DCM) to regulate the solar module with digital pulse width modulation technique (DPWM). The digital controller implementation ensures the MPPT and DC-DC converter to have a high level of programmability and portability across different technologies.

Figure 5.9 Architecture of DC-DC power converter with MPPT control

5.5.1 MPPT Block

The MPPT block consists of a power estimator and a power comparator. The power estimator will perform the sensing of the input voltage and input power of the PV module. Using the proposed algorithm, the sampling and holding circuit in the power estimator will sense and measure the input power of the solar module. The measured magnitude voltage and power will be held in a holding capacitor. The two measured signals will then be fed to the dynamic power comparator for determining the next perturbation of the duty cycle.

5.5.1.1 Power Estimator

The power estimator is shown as Figure 5.10. The MPP tracking operation of the power estimator for $D[0\ 0\ 1\ 1]$ to $D[0\ 1\ 0\ 0]$ is illustrated in Figure 5.11. Each cycle of the power estimator consists of $T_{SAMPLING}$ and $T_{HOLDING}$.

In a new MPPT cycle, during the sampling phase, $T_{SAMPLING(k-1)}$, where the duty cycle of the power converter is at $D[0\ 0\ 1\ 1]$, the output voltage of the solar module will be sensed and processed for MPPT. To conserve charge, the sampling capacitors, C_3 - C_0 will only be discharged if the control bit for the respective capacitor is ‘low’. In this case, only C_3 and C_2 will be discharged to ground. Both the holding capacitors, C_{VI} and C_{PI} will be discharged to ground.

Subsequently, the value of PV_N will be ‘high’ and the voltage from the PV module will be sampled. In this case, only capacitor C_1 and C_0 will charge up to V_{PV} . Base on the operation of the power estimator, capacitors, C_3 , C_2 , C_1 , C_0 and C_{PI} will then be connected. The charges in C_1 and C_0 will be redistributed among these five capacitors. The result is that the holding capacitor, C_{PI} will have a

voltage magnitude correlated to the voltage magnitude of the PV's maximum power.

The power estimator will then sample the input voltage, V_{PV} of the PV module in capacitor C_2 . Then the charges in C_2 will be shared among C_2 and C_{V1} , leaving the sampled input voltage as expressed in Equation (5.10).

$$V_{CV1} = V_{PV} \frac{C_2}{C_2 + C_{V1}} \quad (5.10)$$

The holding capacitors will hold the voltage till the next sampling cycle, $T_{SAMPLING(k)}$.

Assuming the next perturbation of the duty cycle will be $D[0 \ 1 \ 0 \ 0]$. During the sampling phase, $T_{SAMPLING(k)}$, sampling capacitors, C_3 , C_1 and C_0 , and the holding capacitors, C_{V2} and C_{P2} will be discharged to ground. Following, PV_N will be 'high' and the voltage from the PV module will be sampled. In this case, only capacitor C_2 will charge up to V_{PV} . Base on the operation of the power estimator, capacitors, C_3 , C_2 , C_1 , C_0 and C_{P1} will then be connected. The charges in C_2 will be redistributed among these five capacitors. The result is the holding capacitor, C_{P2} will have a voltage magnitude correlated to the voltage magnitude of the PV's maximum power. Similarly, capacitor C_{V1} , will hold a scaled-down input voltage of the PV module. These charges held in the holding capacitors will be fed to the power comparator for the determination of the perturbation of the duty cycle.

Figure 5.10 Schematic of power estimator

Figure 5.11 Timing diagram of power estimator with D[0 0 1 1] changing to D[0 1 0 0].

5.5.1.1 Power Comparator

The power comparator is shown as Figure 5.12. The comparator is implemented with a double-tail latch-type dynamic comparator [25]. Unlike the analog comparator that consumes static power, the dynamic comparator does not require quiescent power. This is essential, as the dynamic comparator will not consume power once the MPPT operation is completed.

Figure 5.12 Block diagram of power comparator

From Section 5.2, the MPPT algorithm requires comparison between the voltage magnitude at cycle k and $k-1$. Hence, the voltage held in the holding capacitors are manipulated by the transmission gates and multiplexer to fulfill the MPPT algorithm before feeding to the inputs of the dynamic comparator. For each T_{MPPT} cycle, the dynamic comparator will perform two switching activities to determine the comparison between the input voltage and input power of the PV module.

Using logic gates and D-flip-flops, the output of the dynamic comparator is processed to determine the direction of perturbation of the duty cycle, D_1 . Following, the DPWM is adjusted accordingly for MMPT operation by the *UP/DOWN* signal from the perturbation block.

5.5.2 DPWM Block

The DPWM block consists of the relaxation oscillator, counters and digital controlled delay generators. The block diagram from the DPWM is illustrated in Figure 5.13. The relaxation oscillator [25] adapted in this design is targeted to provide the energy harvesting system a clock frequency of 10Khz while consuming little overhead power.

The control signals for the MPPT block is controlled by a 7bits counter. In this design, the MPPT block will only activate once every 10ms, thus reducing the overall power required by the MPPT block to perform the voltage and power sensing.

The control signals required by the power converter to emulate the resistance of the PV module is generated by the 3 bits digital control delay generator block. The schematic of the 3 bits up-down counter is illustrated in Figure 5.14.

Figure 5.13 Block diagram of DPWM block

Figure 5.14 Schematic of 3 bits up/ down counter

The most significant bit of the up-down counter is set to ‘high’ during initialisation. This is to ensure that the power converter will not start with $D_I=0$ as the initial duty cycle. The proposed MPPT circuitry will be simulated with MATLAB/ SIMULINK and designed in 0.18 μm CMOS process.

5.6 Summary

In Chapter 2, various state-of-the-art lossless current sensing techniques have been reviewed and discussed. These sensing techniques can only sense the input current and are not suitable for ultra-low power indoor light energy harvesting. In this chapter, a lossless sensing approach to employ hill-climbing MPPT for indoor conditions is described and proposed. The proposed approach is able to sense the inductor current and adapt the sensed signal into the output power of the solar module. Hence, without the use of power-hungry MCU or analog multiplier, the input power and voltage of the solar module can be obtained and processed for ‘direct’ MPPT of indoor light energy.

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Chapter 6

Results and Discussion on MPPT Circuitry

The proposed design of an ultra-low power MPPT enabled energy harvesting for powering a biometric monitoring sensor is discussed and designed in Chapter 5. Simulations are conducted for the purpose of testing and verifying the theory and proposed design. In this chapter, performance results obtained from simulations are presented and discussed in detail.

6.1 Analysis on Proposed MPPT Circuitry

To verify the functionality of the proposed Power Estimator, the simulation set-up as shown in Figure 6.1, is simulated using the MATLAB/SIMULINK software. It is worthy to note that the component names are used across the SIMULINK and CMOS process. The PV module is modelled with the mathematical equations obtained from the electrical characteristics of the photovoltaic cell discussed in Chapter 2. The RF load at V_{OUT} is modelled as a current source. The component values and circuit parameters are tabulated in Table 6.1.

Figure 6.1 Simulation setup for proposed MPPT circuitry with power converter

Table 6.1 Component values and circuit parameters

Switching Frequency	10000 Hz
V_{OUT}	2 V
I_{OUT}	80 μ A
PM_1, PM_4, NM_2, NM_3	100 m Ω
L	22 μ H (ESR = 0.13 Ω)
C_{PV}	0.47 μ F
C_{LOAD}	10 μ F
C_3	1.6 pF
C_2	0.8 pF
C_1	0.4 pF
C_0	0.2 pF
$C_{V1}, C_{V2}, C_{P1}, C_{P2}$	1.0 pF

To begin the simulation test, capacitor, C_{PV} is set to the open-circuit voltage of the PV module and the power converter is operating under DCM with initial $D_1 = D[0 \ 1 \ 1 \ 1]$. The PV array is simulated to perform under a fixed illuminance and consistent room temperature. From here, it is worthy to note that the period for MPP sensing by the power estimator is once every 10 ms.

Figure 6.2 shows the simulated voltage waveform at the input of the power converter and Figure 6.3 shows the corresponding duty cycle fed to the MPPT block. During the initial MPPT cycle, capacitors, C_{V1} and C_{P1} will sample and hold the open-circuit voltage of the PV array, while C_{V2} and C_{P2} discharges to ground.

The MPPT algorithm recognizes the holding voltages in C_{V1} , C_{P1} , C_{V2} and C_{P2} as $V_{PV(k)}$, $V_{PWR(k)}$, $V_{PV(k-1)}$ and $V_{PWR(k-1)}$ respectively. Since $V_{PV(k)} > V_{PV(k-1)}$ and $V_{PWR(k)} > V_{PWR(k-1)}$, the hill-climbing algorithm will determine there is a need to decrease the next perturbation of the duty cycle. This corresponds to the observation in Figure 6.3, where at $t=0$ s, the duty cycle decrease from $D[0 \ 1 \ 1 \ 1]$ to $D[0 \ 1 \ 1 \ 0]$.

Figure 6.2 Output voltages of the PV array with MPPT enabled

Figure 6.3 Switching ratio for the sampling capacitors

In addition, it is also observed that under the influence of the MPPT block, the power converter was able to reach steady state after $t = 0.035\text{s}$.

Knowing the MPPT block is able to perform maximum power tracking of the PV module, there is a need to analyze the efficiency of the proposed MPPT circuitry.

At this point, the accuracy of the MPPT circuitry will be analyzed. The accuracy, η_{ACC} of the MPPT circuitry is expressed as Equation (6.1).

$$\eta_{ACC} = \frac{P_{AVG}}{P_{TRUE}} \quad (6.1)$$

$$= \frac{\frac{1}{T_{MPP}} \int_{T_{MPP}} v_{PV} i_{PV} dt}{\frac{V_{MPP}}{I_{MPP}}}$$

As seen from Equation (6.1), the accuracy of the MPPT circuitry is defined as the ratio of the average power that can be extracted from the PV module at equilibrium to the maximum power that can be extracted from the PV module at its true V_{MPP} .

From Figures 6.2 and 6.3, the voltage of the PV module oscillates around the true V_{MPP} . Hence, the average power, P_{AVG} that can be extracted from the PV module at equilibrium is obtained as an average of the power that can be extracted at intervals, T_{MPP1} and T_{MPP2} . Applying the results obtained from the simulation into Equation (6.1):

$$\eta_{ACC} = \frac{\frac{266.23\mu W + 263.29\mu W}{2}}{268.46\mu W}$$

$$= 0.986$$

Hence, the proposed power estimator can achieve a MPPT accuracy of 98.6 %.

Although the power estimator can achieve a high MPPT accuracy, the primary drawback is that the power estimator is drawing charge from the input capacitor C_{PV} to perform the sensing of input voltage and power of the PV module. To

analyze this drawback, the worst case scenario, where the duty cycle is $D[1 \ 1 \ 1]$ is assumed for the operation of the power estimator.

This will implies that for a MPPT cycle, C_3 , $2 \cdot C_2$, C_1 and C_0 will be turned on to sample the input voltage of the PV module. It is worthy to note that capacitor C_2 is turned on twice for the sampling of the input power and input voltage of the PV module. The percentage power loss, P_{LOSS} from the input power due to the sampling of the power estimator can be expressed as Equation (6.2).

$$P_{LOSS} = \frac{V_{PV}(C_3 + C_2 + C_2 + C_1 + C_0)}{C_{PV}} * 100 \quad (6.2)$$

Substituting the capacitance values from Table 6.1 into Equation (6.2), the percentage power loss obtained is $0.253 \cdot 10^{-3} \%$. Hence, this minute power loss is insignificant to the efficiency degradation of the overall energy harvesting system.

To further examine the robustness of the proposed MPPT circuitry, intentional shading of the PV array was introduced in the simulation. Figure 6.4 shows the simulated voltage waveform at the input of the power converter under intentional shading and Figure 6.5 shows the corresponding duty cycle fed to the MPPT block.

The introduced intentional shading of the PV array in the simulation was achieved by decreasing the short-circuit current of the PV array. As seen, the output voltage of the PV array was able to come back to the steady state after being disturbed.

Figure 6.4 Output voltages of the PV array with MPPT enabled under intentional shading

Figure 6.5 Switching ratio for the sampling capacitors under intentional shading

6.2 Analysis on Proposed MPPT Circuitry in Circuit Design

To verify the functionality of the proposed Power Estimator, the circuit was designed in 0.18 μm CMOS process. The silicon area of the proposed MPPT circuitry with power converter is shown in Figure 6.6. Due to some unforeseen circumstances the chip will be send out for fabrication in September 2017. Hence, the following discussion will be based on post-layout simulations.

Figure 6.6 Silicon chip area of proposed MPPT circuitry with power converter

6.2.1 Power Comparator

To reduce the energy consumption of the processing circuitry, the power comparator used in this design is based on a dynamic comparator. The dynamic

comparator and its corresponding aspect ratio (W/L) are shown in Figure 6.7 and Table 6.2 respectively.

Figure 6.7 Schematic of proposed power comparator

Table 6.2 Transistor sizing for power comparator

NMOS	(W/L) _N	PMOS	(W/L) _P
M_{N0}	4.28/1.8	M_{P0}	0.9/0.18
M_{N1}	4.28/1.8	M_{P1}	0.9/0.18
M_{N2}	0.88/0.36	M_{P2}	1.06/0.18
M_{N3}	0.88/0.36	M_{P3}	1.06/0.18
M_{N4}	0.88/0.36	M_{P4}	1.06/0.18
M_{N5}	0.88/0.36	M_{N5}	1.06/0.18
M_{N6}	0.88/0.36	M_{N6}	1.06/0.18
M_{N7}	0.88/0.36	M_{N7}	1.06/0.18
M_{N8}	3.52/0.36		
M_{N9}	3.52/0.36		

In Figure 6.7, transistors M_{N0} , M_{N1} , M_{N8} , and M_{N9} are utilized as the input transistors for voltage amplification. Transistors M_{N4} , M_{N5} , M_{P2} , M_{P3} , M_{P4} , and M_{P5} are utilized as the latch circuit. During the reset phase, where clock = 'low', node A is discharged to ground, transistors M_{N8} and M_{N9} are turned off and V_{OUT+} and V_{OUT-} are reset to 'low'. After the reset phase, the tail transistors, M_{P0} , M_{P1} , M_{N2} , M_{N3} are turned on.

At node A, the differential output voltage and the common-mode output voltage increases. When the common-mode output voltage approaches the threshold voltage of transistors, M_{N8} and M_{N9} , these two transistors will turn on. The differential output voltage of this two transistors will increase gradually till the positive feedback loop kicks in and provide a rail-to-rail output, V_{OUT+} .

The proposed power estimator requires a minimum differential voltage of 30 mV between the estimated $V_{PWR(k)}$ and $V_{PWR(k-1)}$ and a minimum differential voltage of 100 mV between the estimated $V_{PV(k)}$ and $V_{PV(k-1)}$. Hence, there is a need to ensure the dynamic comparator has a low offset for comparing the estimate power of the PV module.

From a 100 samples of Monte-Carlo analysis, the efficiency of the power comparator is illustrated in Figure 6.8. Under a 400 kHz switching frequency and common-mode input voltage of 1.1 V, the simulated offset voltage was found to be around 7 mV. Hence, the offset of the dynamic comparator is acceptable in the context of this design. In addition, in this design, the dynamic comparator only consumes 1.132 nW

Figure 6.8 Efficiency of dynamic comparator with different offset voltages

6.2.2 Power Estimator

The output voltage of the PV module is sampled and held in capacitors for MPPT processing. The capacitances of these sampling and holding capacitors are tabulated in Table 6.3. The power estimator circuit was simulated with an input voltage of 2.2 V at D[1 1 0 1]. Figure 6.9 shows the variations in the holding voltage waveform of the estimated input power held in C_{PI} .

Table 6.3 Capacitor sizing and area for sample and hold circuit

Capacitor	Capacitance	Total Area
C_3	6.411 pF	132.54 μm X 132.54 μm
C_2	3.205 pF	
C_1	1.603 pF	
C_0	0.801 pF	
$C_{V1}, C_{V2}, C_{PI}, C_{P2}$	4.004 pF	120.8 μm X 120.8 μm

Figure 6.9 Simulations results for estimated input power, PH_1

From Figure 6.9, it can clearly be seen that the voltage magnitude held in C_{PI} had degraded by 11.5424 mV after an interval of 10 ms. This observation is undesirable as the degradation will reduce the accuracy of the proposed power estimator if the interval for MPP sensing is delayed for a longer period of time. Hence, there is a need to investigate the cause of the degradation in the holding voltage of the holding capacitor. Figure 6.10 depicts the connections of the holding capacitor.

From Figure 6.10, it can be seen that there are three paths connected to the top plate of capacitor, C_{PI} . Hence, the charge leakage from the C_{PI} can be dominated by the junction leakage of the switches in its path. The junction leakage is roughly proportional to the voltage drop across the drain-source and drain-body of the transistor. Hence, by keeping a zero potential for the above voltage drop, a lower leakage for the holding capacitor can be realized.

Figure 6.10 Schematic of the holding capacitor for estimated input power

Figure 6.11 shows the addition of a bias voltage to reduce the junction leakage of C_{P1} . The same bias circuitry with a different time pulse is added to capacitor C_{P2} . From Figure 6.11, it can be seen that the sampled voltage in C_{V1} is recycled and deployed as the bias voltage to reduce the potential difference in the drain-source of the leakage path.

Under the same scenario, the variations in the holding voltage waveform of the estimated input power held in C_{P1} is shown in Figure 6.12. From 6.12, with the aid of the proposed bias circuit, it can clearly be seen that the degradation of voltage magnitude held in C_{P1} had been reduced to 6.7164 mV. This translates to a 41.8 % increase in the holding efficiency of capacitor, C_{P1} . The degradation of V_{BIAS1} , which is the sampled voltage of C_{V1} is not critical, as it will be refreshed to a new sampled voltage in the next MPPT cycle. Hence, lower leakage and longer holding time can be achieved for the holding capacitors, C_{P1} and C_{P2} with minimal power consumption.

In addition, from the simulation, the total power consumed by the control signals of the power estimator is found to be 10.981 nW. Hence, the total power consumed by the MPPT circuitry (power estimator and power comparator) is found to be 12.113 nW.

Figure 6.11 Schematic of the holding capacitor with proposed bias circuit

Figure 6.12 Simulations results for estimated input power, PH_1 with leakage reduction

6.3 Summary

To efficiently harvest solar energy to power the biometric sensor, a novel MPPT circuitry has been introduced. The power converter and MPPT circuitry employs digital switching modulation to minimize the quiescent power of the controller. Hence, the MPPT circuitry can achieve a high tracking efficiency of 98.6 % while consuming only 12.113 nW to power its processing circuits. A charge leakage reduction technique is employed in the holding capacitors of the sample and hold circuit. The introduction of the proposed leakage reduction technique is able to improve the efficiency of the holding capacitor by 41.8 %.

The result of this work is compared with the state-of-art circuits and tabulated in Table 6.4. The intrinsic power loss in the proposed circuit is contributed by the ratio of the on-chip sampling capacitors (pico-farads) to the off-chip input capacitor (micro-farads). This intrinsic power loss in the proposed circuit can be considered negligible when a high-power PV panel is employed as the energy transducer. In addition, the proposed MPPT circuit does not require quiescent power and the dynamic power loss in the MPPT controller is determined by the frequency of MPPT sampling. Hence, the proposed MPPT circuit, comprising of

a sample and hold circuit and a dynamic comparator has high portability across semiconductor technologies and across different power level of solar energy harvesting system.

Table 6.4 Comparison with existing MPPT circuitry

Parameters	[1]	[2]	[3]	[4]	This work
MPPT efficiency (%)	P&O >99	P&O ~70	P&O >99	P&O >99.4	P&O 98.6
Power consumption of MPPT controller	500 μ W @ P_{PV} ≈ 3 W	2.4 μ W @ $P_{PV}=8\mu$ W; 3.5 μ W @ P_{PV} = 1.66mW	120 μ A @ P_{in} = 0.53W	52 μ W @ $P_{PV}=2.7$ mW	12.113nW @ P_{PV} = 268 μ W
Process	0.5 μ m CMOS	0.25 μ m CMOS	0.35 μ m CMOS	Discrete Components	0.18 μ m CMOS

References

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Chapter 7

Conclusion and Future Work

This chapter concludes the important outcomes of this research work and discusses its effectiveness in achieving a low-power processing circuit and a highly efficient power delivery circuitry for a solar energy harvesting system. The chapter ends by suggesting a few plausible improvements to the system as a future work.

7.1 Conclusion

The electrical power of a PV module can be harnessed directly to the connected load through a diode. This direct-coupling solution is not adopted widely as a large solar array is required to ensure the diode is in forward-bias mode. In addition, this approach has a low efficiency in transferring the available power at the solar array to the output load. The energy storage element in such an energy harvesting system will also affect the designing of this approach. A battery with nominal rating ranging between 2.7V to 3.0V can force selected solar modules to operate under its maximum power point. Hence, increasing the overall efficiency of the power system. However, if the energy storage element is a supercapacitor, with a nominal rating ranging between 0V to 5V, the solar module will not always be able to operate at its MPP. In addition, the self-discharge of the supercapacitor will cause failure to the system under low-light conditions.

Hence, an interface circuit is required to achieve a small form factor for the energy harvesting system and to effectively transfer the harnessed power from the energy transducer to the connected load and the energy storage element. The designing of the interface circuit is dependent on which MPPT approach is adopted. High tracking efficiency MPPT method like P&O method will require the output voltage and current of the solar module to be known. Complex and power-hungry current sensors, microcontroller or analog-to-digital converter will be required to perform the algorithm of such an approach. However, less complex MPPT method like the FOCV method can be employed to achieve sub-optimal maximum power point tracking of a solar module. These approaches, albeit with lower tracking efficiency are able to perform MPP tracking using lower energy-processing circuitry as compared to the circuitry employed in a high tracking efficiency MPPT approach. Hence, the lesser tracking efficiency MPPT approach such as the FOCV is conventionally employed for solar energy harvesting system under low-light conditions.

Regardless on the selection of a suitable MPPT approach, a DC-DC converter will always be required to perform the MPPT and to regulate the output voltage of the connected load. An extensive overview of inductive DC-DC converters is discussed in Chapter 2. Among the converters, SIMIMO converters are frequently employed in a complete energy harvesting system. SIMIMO converters have the capability to rectify multiple input power sources, thus enhancing the reliability of the energy harvesting system. The converter will also be able to provide multiple domains of output voltage for the different control blocks in the system. The power switches of a SIMIMO converter can be controlled to perform buck, boost, or buck-boost operation.

The thesis has also demonstrated that operating the converter in the buck-boost mode will achieve a lower conversion efficiency as compared to the individual buck or boost mode operation. However, the lower conversion efficiency of the buck-boost mode can be justified by the fact that the duty cycle of the buck-boost mode operation is independent of the output voltage. This relationship is important for ‘direct’ MPPT of a solar module with multiple output voltage domains.

To perform on-chip integration of the power converter, the converter has to be operated in synchronous mode. Hence, a dead-time controller is crucial to ensure the functionality and efficiency of the power converter. The effects of sub-optimal dead-time control is discussed in the thesis. Literature works on dead-time controllers are also reviewed and discussed. However, in ultra-low power energy harvesting, the power budget for the on-chip power management unit is limited. Hence, an inverter-based DTC is discussed in Chapter 2 of the thesis. Although the digital controlled inverter-based DTC has a slower loop response as compared to analog based DTC, the absence of analog comparator in the inverter-based DTC eliminates the quiescent power of the DTC. This digital controlled inverter-based DTC is adopted in the design in Chapter 5.

The real challenges in employing wearable wireless sensors for various physiological monitoring are: (1) bulky and rigid system design, and (2) limited operational lifespan of batteries with finite energy supply. Written journals have shown the possibility of implementing a flexible power management unit with flexible solar transducer to counter the challenges in the employment of wearable biometric monitoring sensors. Adopting the flexible thin-film solar cells and organic photovoltaics as the energy transducer will extend the operational lifetime of the biometric sensors. These lightweight and flexible energy transducer can be easily integrated with the wearable biometric sensors to achieve an autonomous energy harvesting system for medical applications. The PMU of a flexible energy harvesting system can be achieved by attaching surface-mount technology (SMT) components or flexible components onto a flexible printed circuit board. However, the proposed design in the literature reported low power conversion efficiency and MPPT of the solar module is not implemented.

Hence, Chapter 3 has proposed a flexible energy harvesting system with MPPT capability that is able to overcome the two main constraints for wearable biometric sensors in the above-mentioned. The ability to harness energy from the infinite ambient energy and degree of flexibility existing in the energy transducer and power management circuit will eliminate discomfort and the inconvenience of the older, bulky and battery-dependent wearable wireless sensors. The proposed flexible energy harvesting system is self-sustainable, possesses MPPT capability and is implemented with ultra-low power control circuitry. Simulation and measurement data showed that the proposed MPPT approach suits well for the bendable solar module. In the experiment data, the flexible energy harvesting system is able to maintain operation under a low light illuminance of 320lux while conforming to the human body at an angle of 30°.

Harvesting DC ambient energy to power the ultra-low-power portable electronics is becoming increasingly popular. Traditional interface circuit use in ‘direct’

MPPT approach to extract the maximum available energy harnessed by the energy transducer is inefficient with the power-hungry controllers. In addition, ‘indirect’ MPPT approach has a lower tracking efficiency as compared to ‘direct’ approach and the power consumption of such an approach may still be consider still large in the case of ultra-low power ambient energy. Hence, there is a need to develop a new approach for true ultra-low power energy harvesting. Exploiting the literature review in Chapter 2, the buck-boost mode of a SIMIMO converter is used to develop a new MPPT approach for ultra-low power energy harvesting system.

A new lossless current sensing approach is proposed in Chapter 5. The proposed current sensing for a solar module is accomplished by sensing the peak inductor current. Theoretical calculations and simulation results have showed that the output current of a solar module at its maximum power point corresponds to the peak current flowing through the inductor. Hence, a capacitor network is proposed to approximate the duty cycle of the power converter. By sampling the output voltage of the solar module with respect to the duty cycle, the output current of the solar module can be found. However, the voltage magnitude in the holding capacitor for the sampled voltage and current information will degrade over time. A charge leakage reduction technique is employed in the sample and hold circuit. The proposed MPPT circuitry can achieve a high tracking efficiency of 98.6% while only consuming 12.113 nW to power its processing circuits. Also the charge leakage reduction technique is able to improve the efficiency of the holding capacitor by 41.8%. The proposed MPPT circuitry has high portability across semiconductor technologies and across different power level of solar energy harvesting system.

In conclusion, this thesis has covered works on the design and development of a solar energy harvesting system that is capable of autonomy and flexibility. Most importantly, the proposed energy harvesting system can efficiently deliver the

harnessed power from the solar module to the load while consuming ultra-low processing energy.

7.2 Future Works

The proposed MPPT employs a fixed step size hill-climbing algorithm and is refreshed every 10 ms for the tracking. The transient respond of the MPPT can be improve by employing a variable step size approach. The transient respond can be even further improve by integrating the well-established constant voltage MPPT method with the proposed MPPT circuit. When the output voltage of the PV panel is sensed to fall outside of the window of the constant voltage method, the duty cycle of the DC-DC converter can be adjusted accordingly by the MPPT controller. This allows the PV panel to operate at the maximum power point with a short transient time. With this new architecture, the proposed ultra-low power processing MPPT circuitry will be able to operate over outdoor and indoor conditions while maintaining a high tracking efficiency and short transient time for the PV panel to reach steady state.

To further reduce the quiescent power of the MPPT circuitry, the refresh rate of the MPPT cycle can be reduce. However, as seen in Chapter 6, a longer refresh rate for the MPPT cycle will lead to a reduction in the accuracy of the power estimator circuit. A voltage follower can be included to generate the required bias voltage to reduce the leakage of the holding capacitor. By ensuring minimal potential difference across the drain and source of the transistor for the holding capacitor, the leakage in the holding capacitor can be reduced. This closed-loop mechanism will allow the proposed MPPT circuit to operate under a wider frequency.

The proposed power estimator can be extended to track the MPP of similar DC

ambient energy, such as thermal energy. FOCV method is widely adopted for MPPT in thermal energy harvesting system. With FOCV, the thermal energy harvesting system will suffer the same drawbacks as a solar energy transducer mentioned in Chapter 5. Hence, the proposed MPPT circuit can overcome these drawbacks will maintaining a high efficiency in extracting the maximum power from the thermoelectric energy generator (TEG). In addition, the proposed solar energy SIMIMO energy harvesting system can include the TEG as one of the ambient energy input for the converter. A hybrid ambient energy harvesting system with a single highly effective and efficient MPPT circuitry will further affirm the reliability of a fully autonomous energy harvesting system for the wireless sensors.

List of Publications

W. Y. Toh, Y. K. Tan, W. S. Koh and L. Siek, "Autonomous Wearable Sensor Nodes With Flexible Energy Harvesting," in *IEEE Sensors Journal*, vol. 14, no. 7, pp. 2299-2306, July 2014.

W.Y. Toh, L.Siek and C.H.Vun, "Ultra-Low Power and Highly Accurate Photovoltaic MPP Detector for Micro-Power Energy Harvesting System." – Submitted to IEEE Electronics Letter.

W.Y. Toh, L.Siek and C.H.Vun, "Digitally Controlled H-Bridge DC-DC Converter for Photovoltaic Energy Harvesting with Lossless Current Sensing MPPT" – Expected submission to *IEEE Transactions on Circuits and Systems I* (Waiting on return of chip for measurement results).

W.Y. Toh, L.Siek and C.H.Vun, in preparation for technical disclosure on the proposed ultra-low power MPPT approach for indoor conditions.