OPTIMIZATION OF A DISCONTINUOUS CONDUCTION MODE FLYBACK FOR ACOUSTICAL ENERGY HARVESTING

By

ROBERT JAMES TAYLOR

A THESIS PRESENTED TO THE GRADUATE SCHOOL OF THE UNIVERSITY OF FLORIDA IN PARTIAL FULFILLMENT OF THE REQUIREMENTS FOR THE DEGREE OF MASTER OF SCIENCE

UNIVERSITY OF FLORIDA

2004
This document is dedicated to my friends and family.
ACKNOWLEDGMENTS

Partial financial support for this project is provided by NASA Langley Research Center (Grant #NAG-1-2261) and NASA KSC (Grant #NAG-10-316).

I would like to thank my advisors, Dr. Toshikazu Nishida, Dr. Khai Ngo, Dr. Louis Cattafesta, for all of their support and continually pushing me to excel. The research experience gained with the help of them will benefit me for the rest of my life. I would also like to thank Dr. Mark Sheplak for his help and guidance in my research endeavors.

I am indebted for life to my friends and coworkers in IMG. Without all of the support and help my research would have never been completed. Thanks go especially to Karthik Kadirvel, Dave Martin and Jian Liu for being great office mates and preventing me from losing my mind. I would also like to thank Steve Horowitz and Fei Liu for all of their help with my research.

Finally, I want to thank you to my parents for all of their love and support, my friends for all of the positive encouragements and good times, and last, but not least, I thank Rosemarie Arzaga for all of her love, support, and for always pushing me to do my best.
TABLE OF CONTENTS

| ACKNOWLEDGMENTS ................................................................................................................ iv |
| LIST OF TABLES .................................................................................................................... vii |
| LIST OF FIGURES .................................................................................................................. viii |
| ABSTRACT ................................................................................................................................... xii |

CHAPTER

1 INTRODUCTION .................................................................................................................. 1

   Energy Harvesting .................................................................................................................... 1
   Mechanical Vibrations ........................................................................................................... 1
   Thermal Gradients ................................................................................................................... 2
   Solar Power ............................................................................................................................. 3
   Acoustical Sound Pressure ...................................................................................................... 4

   Energy Harvester Circuits ........................................................................................................ 5
   Research Objectives ................................................................................................................ 5
   Thesis Organization ................................................................................................................ 6

2 THEORETICAL BACKGROUND .......................................................................................... 7

   Power Converter Topologies ................................................................................................ 7
   Buck Converter ....................................................................................................................... 7
   Boost Converter ..................................................................................................................... 9
   Buck-Boost Converter ........................................................................................................... 10
   Discontinuous Conduction Mode ......................................................................................... 12
   Buck-Boost in DCM ............................................................................................................... 13
   Flyback Circuit ...................................................................................................................... 15

3 DESIGN OF THE FLYBACK CIRCUIT ................................................................................. 18

   System Partitioning ................................................................................................................. 18
   Bridge Rectifier ...................................................................................................................... 18
   Control Circuit ....................................................................................................................... 22
   Power Stage ............................................................................................................................ 23
   Additional Circuits .................................................................................................................. 25
<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Energy Harvester Circuit</td>
<td>27</td>
</tr>
<tr>
<td>Experimental Verification</td>
<td>29</td>
</tr>
<tr>
<td>4 PARAMETRIC STUDY AND OPTIMIZATION OF THE FLYBACK CIRCUIT</td>
<td>34</td>
</tr>
<tr>
<td>Parametric Study Constraints</td>
<td>34</td>
</tr>
<tr>
<td>Flyback Converter Efficiency</td>
<td>35</td>
</tr>
<tr>
<td>Loss Estimations</td>
<td>37</td>
</tr>
<tr>
<td>Input Impedance</td>
<td>41</td>
</tr>
<tr>
<td>Optimal Designs</td>
<td>43</td>
</tr>
<tr>
<td>5 APPLICATION OF A DCM FLYBACK CONVERTER TO ACOUSTIC ENERGY HARVESTING</td>
<td>47</td>
</tr>
<tr>
<td>Helmholtz Resonator</td>
<td>47</td>
</tr>
<tr>
<td>Lumped Element Model</td>
<td>48</td>
</tr>
<tr>
<td>Frequency Response</td>
<td>50</td>
</tr>
<tr>
<td>Maximum Power Transfer Theory</td>
<td>52</td>
</tr>
<tr>
<td>Helmholtz Resonator at Maximum Power Transfer</td>
<td>56</td>
</tr>
<tr>
<td>Acoustic Energy Harvester Simulations</td>
<td>57</td>
</tr>
<tr>
<td>Experimental Verification</td>
<td>58</td>
</tr>
<tr>
<td>Helmholtz Resonator Frequency Response</td>
<td>59</td>
</tr>
<tr>
<td>Helmholtz Resonator Optimal Load</td>
<td>61</td>
</tr>
<tr>
<td>Acoustic Energy Harvester</td>
<td>65</td>
</tr>
<tr>
<td>6 CONCLUSIONS AND FUTURE WORK</td>
<td>73</td>
</tr>
<tr>
<td>Summary of Results</td>
<td>73</td>
</tr>
<tr>
<td>Future Work</td>
<td>74</td>
</tr>
<tr>
<td>APPENDIX: BILL OF MATERIALS</td>
<td>76</td>
</tr>
<tr>
<td>LIST OF REFERENCES</td>
<td>79</td>
</tr>
<tr>
<td>BIOGRAPHICAL SKETCH</td>
<td>83</td>
</tr>
</tbody>
</table>
# LIST OF TABLES

<table>
<thead>
<tr>
<th>Table</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1-1.</td>
<td>Summary of power levels achieved for vibrational energy harvesting</td>
<td>3</td>
</tr>
<tr>
<td>2-1.</td>
<td>DCM criteria and steady-state voltage transfer characteristics.</td>
<td>14</td>
</tr>
<tr>
<td>2-2.</td>
<td>CCM and DCM characteristics of a flyback converter.</td>
<td>16</td>
</tr>
<tr>
<td>4-1.</td>
<td>Inductor specifications for the parametric study.</td>
<td>35</td>
</tr>
<tr>
<td>4-2.</td>
<td>Optimal energy harvester designs.</td>
<td>43</td>
</tr>
<tr>
<td>4-3.</td>
<td>Summary of dimensions for three optimal designs.</td>
<td>45</td>
</tr>
<tr>
<td>5-1.</td>
<td>Description of the lumped element parameters used in the Helmholtz resonator</td>
<td>49</td>
</tr>
<tr>
<td></td>
<td>equivalent circuit.</td>
<td></td>
</tr>
<tr>
<td>5-2.</td>
<td>Physical dimensions of the Helmholtz resonator.</td>
<td>50</td>
</tr>
<tr>
<td>5-3.</td>
<td>Estimates for the lumped element parameters of the Helmholtz resonator.</td>
<td>50</td>
</tr>
<tr>
<td>A-1.</td>
<td>Bill of materials for 1.5mH design.</td>
<td>76</td>
</tr>
<tr>
<td>A-2.</td>
<td>Bill of materials for 4.7mH design.</td>
<td>77</td>
</tr>
<tr>
<td>A-3.</td>
<td>Bill of materials for 14.4mH design.</td>
<td>78</td>
</tr>
</tbody>
</table>
# LIST OF FIGURES

<table>
<thead>
<tr>
<th>Figure</th>
<th>Description</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1-1</td>
<td>System overview of the acoustic energy harvester.</td>
<td>4</td>
</tr>
<tr>
<td>2-1</td>
<td>Schematic of an ideal step-down buck converter.</td>
<td>8</td>
</tr>
<tr>
<td>2-2</td>
<td>Typical switching operation for a buck converter.</td>
<td>9</td>
</tr>
<tr>
<td>2-3</td>
<td>Schematic for a boost converter.</td>
<td>9</td>
</tr>
<tr>
<td>2-4</td>
<td>Switching operation of a boost converter.</td>
<td>10</td>
</tr>
<tr>
<td>2-5</td>
<td>Schematic of an ideal buck-boost converter.</td>
<td>11</td>
</tr>
<tr>
<td>2-6</td>
<td>Typical waveforms of the ideal buck-boost converter.</td>
<td>11</td>
</tr>
<tr>
<td>2-7</td>
<td>PWM switch model for DCM converters.</td>
<td>12</td>
</tr>
<tr>
<td>2-8</td>
<td>Steady-state buck-boost converter with PWM switch model.</td>
<td>13</td>
</tr>
<tr>
<td>2-9</td>
<td>The derivation of the flyback circuit from the buck-boost.</td>
<td>15</td>
</tr>
<tr>
<td>2-10</td>
<td>The ideal switch waveforms for a DCM flyback converter.</td>
<td>17</td>
</tr>
<tr>
<td>3-1</td>
<td>The main sub-circuits of the flyback converter.</td>
<td>18</td>
</tr>
<tr>
<td>3-2</td>
<td>Passive full wave rectifier bridge using diodes.</td>
<td>19</td>
</tr>
<tr>
<td>3-3</td>
<td>Actively controlled rectifier bridge using MOSFETs.</td>
<td>20</td>
</tr>
<tr>
<td>3-4</td>
<td>Signals that are required to drive the MOSFETs in the active rectifier bridge.</td>
<td>20</td>
</tr>
<tr>
<td>3-5</td>
<td>Circuit used to create the control signals for the active bridge rectifier.</td>
<td>21</td>
</tr>
<tr>
<td>3-6</td>
<td>Circuit used to control the MOSFET switch in the power stage.</td>
<td>22</td>
</tr>
<tr>
<td>3-7</td>
<td>Power stage of the flyback converter.</td>
<td>24</td>
</tr>
<tr>
<td>3-8</td>
<td>Schematic of the TPS71501 voltage regulator.</td>
<td>26</td>
</tr>
<tr>
<td>3-9</td>
<td>Simple circuit used to measure small AC currents.</td>
<td>27</td>
</tr>
</tbody>
</table>
5-6: Simplified model of the Helmholtz resonator. ..............................................................52
5-7: Helmholtz resonator circuit with a rectifier bridge, filter capacitor and load. ............53
5-8: Norton equivalent circuit of the Helmholtz resonator connected to the bridge rectifier. ..........................................................................................................................53
5-9: Theoretical waveforms of the Norton equivalent circuit..............................................54
5-10: The expected output power of the Helmholtz resonator vs. load resistance. ..........57
5-11: Acoustic setup for testing the Helmholtz resonator. ..................................................59
5-12: Complete experimental setup for testing the acoustic energy harvester system. ...60
5-13: Magnitude response of the Helmholtz resonator.......................................................61
5-14: Output power of the Helmholtz resonator for varying resistive loads. .................62
5-15: Incident acoustic power of the Helmholtz resonator for varying loads resistances........................................................................................................................................63
5-16: Efficiency of the Helmholtz resonator versus load resistance...............................64
5-17: Helmholtz resonator output voltage versus incident pressure..................................64
5-18: Three possible configurations of connecting the Helmholtz resonator to a rechargeable battery ................................................................................................................66
5-19: Output power of the direct charging method versus incident pressure. ...............67
5-20: Output power of the 4.7mH optimized design and the direct charging method versus incident pressure. ........................................................................................................68
5-21: Output power for a linear regulator, 4.7mH design and direct charging versus incident pressure.................................................................................................................68
5-22: Output power of the 14.4mH optimized design and the direct charging method versus incident pressure. .................................................................................................................69
5-23: Output power of the 1.5mH optimized design and the direct charging method versus incident pressure. .................................................................................................................69
5-24: Output power of the 3.3V linear regulator and the direct charging method versus incident pressure.................................................................................................................70
5-25: The Helmholtz resonator efficiency versus incident sound pressure level for the 5.0V linear regulator, 4.7mH design, 14.4mH design and 1.5mH design. ............71
5-26:  Converter efficiency versus sound pressure level for the 5.0V linear regulator, 4.7mH design, 14.4mH design and 1.5mH design.................................................................72

5-27:  System efficiency of the direct charging method, 5.0V linear regulator, 4.7mH design, 14.4mH design and 1.5mH design versus incident pressure. ..............................72
Abstract of Thesis Presented to the Graduate School of the University of Florida in Partial Fulfillment of the Requirements for the Degree of Master of Science

OPTIMIZATION OF A DISCONTINUOUS CONDUCTION MODE FLYBACK FOR ACOUSTICAL ENERGY HARVESTING

By

Robert James Taylor

August 2004

Chair: Dr. Toshikazu Nishida
Cochair: Dr. Khai Ngo
Major Department: Electrical and Computer Engineering

The need for wireless remote sensors is an increasing trend in engineering. The sensors need to survive without maintenance for long periods of time or indefinitely. Energy harvesting is a possible solution to this problem. A device that is capable of absorbing energy from its ambient surroundings would not need an external power source.

An optimized discontinuous conduction mode flyback for acoustic energy harvesting is presented in this thesis. The system consists of a discontinuous conduction mode flyback converter used in conjunction with a compliant-backplate Helmholtz resonator. A theoretical model is developed for the energy harvester system. Three optimized energy harvester circuits are developed based on three coupled inductors. The circuits are designed to be as small as possible, with the smallest having a surface area just over 0.75 sq inches.
Characterization of the acoustic energy harvesting system for each of the three
designs shows that the circuits outperform direct charging and linear regulator circuits.
The acoustic energy harvester system outperforms the direct charging method by an
average of 260% at input sound pressure levels of 155dB (ref. 20µPa). The typical output
levels of the acoustic energy harvester system for that level of input are on the order of 20
– 30mW.
CHAPTER 1
INTRODUCTION

Harvesting energy from ambient sources is an integral component of replacing non-rechargeable batteries in low power electrical systems. In remote sensors, it is either not feasible or inconvenient to use batteries or other electrical supplies. Replacing batteries in these systems would allow for maintenance-free, self-powered sensors to be developed. Self-powered sensors could be useful in a broad range of applications from structural monitoring of bridges and buildings, to health monitoring for astronauts.

Energy Harvesting

Energy harvesting is not a new concept in electrical system design. Many efforts have been made to reclaim energy from ambient sources [1-27]. The ambient sources used to provide this energy can be, but not limited, to the following: mechanical vibrations, thermal gradients, solar power, gravitational fields, fluid flow and acoustic sound pressure.

Mechanical Vibrations

Mechanical vibrations are present in a number of useful locations where energy harvesting could replace the use of batteries. It has been shown that vibrations due to walking can produce enough energy to power a RFID tag [8]. Other vibration sources that could be considered for energy harvesting include bridges, buildings, and heavy-machinery. Regardless of where the vibrations come from, a transducer is required to convert the vibration energy into electrical energy. Transducers that can be used for this
task include piezoelectric (PZT) patches, PVDF (polyvinylidenefluoride) sheets, electromagnetic coils, and electrostatic devices.

Qualitatively, each of the transduction mechanisms has advantages and disadvantages. Electromagnetic coils produce higher powers but are generally larger in size; electrostatic devices are small but usually require a power source or electret; and PZT devices are simple to fabricate but tend to produce less energy. Electromagnetic coils have been used successfully to harvest vibration energy and produce electrical power. The power produced by the devices ranges from 200 $\mu$W to 0.23 W based on the size and conversion scheme [2, 7-9, 15, 17]. Electrostatic devices usually use a type of variable capacitor to harvest the energy [13, 25]. The power levels produced by these types of generators (10-30 $\mu$W) are significantly lower than electromagnetic devices [13, 25]. PZT and PVDF transduction methods also provide a simple method of turning vibrational energy into electrical energy. Using these methods, power levels of 100 $\mu$W to 30 mW can be achieved [8, 14, 18, 20]. A summary of the achieved power levels for each transduction method is shown in Table 1-1.

**Thermal Gradients**

Thermoelectric generators take advantage of temperature differences to generate electrical power. The main problem in using a thermoelectric generator is that it is difficult to find the necessary temperature gradient in a reasonably sized volume. A temperature gradient of 10 °C could produce a power density of 15 $\mu$W/cm$^3$ [6]. One place in which a large temperature gradient is possible is between the human body, and the ambient. Jung show that 1.6 $\mu$W/cm$^2$ can be produced due to the difference in body to ambient temperature [23]. Another possible location for thermoelectric generators is at
the soil-air interface. This interface provides temperature gradient of about 10 °C, which produces ~50 µW [19].

Table 1-1. Summary of Power Levels Achieved for Vibrational Energy Harvesting

<table>
<thead>
<tr>
<th>Author</th>
<th>Year</th>
<th>Transduction method</th>
<th>Power harvested</th>
<th>Scale/Size</th>
<th>Power density</th>
</tr>
</thead>
<tbody>
<tr>
<td>Williams [2]</td>
<td>1995</td>
<td>Electromagnetic coil</td>
<td>100 µW</td>
<td>5x5x1 mm</td>
<td>4 µW/mm²</td>
</tr>
<tr>
<td>Amirtharajah [7, 9]</td>
<td>1998</td>
<td>Electromagnetic coil</td>
<td>400 µW</td>
<td>Non MEMS</td>
<td>-</td>
</tr>
<tr>
<td>Kymissis [8]</td>
<td>1998</td>
<td>Electromagnetic direct conversion</td>
<td>0.23 W</td>
<td>10x10x5 cm</td>
<td>~0.45 µW/mm³</td>
</tr>
<tr>
<td>Ching [15]</td>
<td>2002</td>
<td>Electromagnetic coil</td>
<td>800 µW</td>
<td>1 cm³</td>
<td>0.8 µW/mm³</td>
</tr>
<tr>
<td>James [17]</td>
<td>2002</td>
<td>Electromagnetic coil</td>
<td>1.25 mW</td>
<td>Non-MEMS</td>
<td>-</td>
</tr>
<tr>
<td>Meninger [13]</td>
<td>2001</td>
<td>Electrostatic capacitor</td>
<td>8 µW</td>
<td>15x5x0.5 mm</td>
<td>0.21 µW/mm³</td>
</tr>
<tr>
<td>Maio [25]</td>
<td>2003</td>
<td>Electrostatic variable capacitor</td>
<td>24 µW</td>
<td>28x28x2 mm</td>
<td>0.015 µW/mm³</td>
</tr>
<tr>
<td>Kymissis [8]</td>
<td>1998</td>
<td>PZT / PVDF</td>
<td>1.8 mW / 1.1 mW</td>
<td>8x10x0.2 cm</td>
<td>0.09 µW/mm³</td>
</tr>
<tr>
<td>Shenck [14]</td>
<td>2001</td>
<td>PZT / PVDF</td>
<td>8.4 mW / 1.3 mW</td>
<td>8x10x0.2 cm</td>
<td>0.42 µW/mm³</td>
</tr>
<tr>
<td>Kasyap [18]</td>
<td>2002</td>
<td>PZT cantilever beam</td>
<td>160 µW</td>
<td>50x50x0.25 mm</td>
<td>0.256 µW/mm³</td>
</tr>
<tr>
<td>Ottman [26]</td>
<td>2003</td>
<td>PZT patch</td>
<td>30.6 mW</td>
<td>46x33x0.25 mm</td>
<td>79.3 µW/mm³</td>
</tr>
</tbody>
</table>

Solar Power

Solar power provides the most familiar form of energy harvesting. Improved efficiencies (~20%) in solar power have increased the amount of power harvested. Solar power has been shown to be able to produce up to 20 mW/cm² on a sunny day. Solar panels however, are bulky and not suitable for locations where light is not available.
**Acoustical Sound Pressure**

Acoustic sound pressure provides a way to harvest energy in locations where vibrations, light, or thermal gradients may not be available. The main drawback to acoustical energy harvesting is that very high sound pressure levels are necessary to achieve suitable power levels. An acoustic energy harvester has been shown to produce 7.4mW power delivered to a resistive load for an incident sound pressure level of 155dB (ref. 20uPa) [16]. The focus of this work will be to improve the amount of energy harvested from a similar device.

The acoustic energy harvester will be comprised of two components, the acoustic energy transducer and the energy harvester circuit. As shown in Figure 1-1, an acoustic energy transducer transforms the acoustic energy (blue) into electrical energy (red). The electrical energy is then conditioned by the energy harvesting circuit and delivered to a load.

![Figure 1-1: A system overview of the acoustic energy harvester.](image)

The conversion efficiency of the system is the output power divided by the input power,

\[ \eta = \frac{P_{\text{out}}}{P_{\text{in}}} \quad (0-1) \]

The efficiency can also be shown to be the product of the individual stage efficiencies,
\[ \eta = \frac{P_{HR} \cdot P_{out}}{P_{in} \cdot P_{HR}}. \]  

By partitioning the system into stages the efficiencies of each stage can be maximized individually.

**Energy Harvester Circuits**

One way to improve upon the efficiency of any energy harvester is to use a circuit that will maximize the amount of power available to a given load. Most of the previously discussed energy harvesters used direct charging of a capacitor or battery for energy storage. There are a number of circuits that could be employed to increase the efficiency of the transducer. Linear regulators have been used to provide a stable output voltage, however, they are inherently less efficient than switching converters [8, 14, 17]. Switching regulators have also been used to optimize the power delivered to a load [7, 9, 14, 20]. A step-down or Buck converter was used by Ottman in order to increase the efficiency of a PZT generator by 325% over direct charging methods [20]. In another approach, Shenck used a forward switching converter to improve the output power of a shoe mounted PZT by a factor greater than two over a linear regulator [14]. A similar approach to these will be used to maximize the power from an acoustical energy harvester.

**Research Objectives**

The main goal of this research is to be able to provide an optimized converter circuit for use with an acoustical energy harvester. The converter will be optimized through a series of parametric studies to maximize the amount of power delivered to a particular load. Once the converter has been optimized it will be used in conjunction with an acoustical energy harvester to demonstrate stand-alone operation.
The parametric study of the converter circuit will focus on circuit variables such as duty cycle, inductor size, switch frequency and size. The parametric study constraints will be to maximize power delivered and minimize the converters size.

The end goal of this research is to be able to demonstrate stand-alone operation of the energy reclamation circuit with an acoustic input. To demonstrate this, the acoustical energy harvester will power the circuit that will in turn store the energy in a battery. It will also be desirable for the circuit to be able to power an electronic load such as a microphone or other sensor.

**Thesis Organization**

The thesis is organized into six chapters. This chapter introduces the background and motivation for energy harvesters. Chapter 2 will provide a theoretical background for the energy harvester circuit. A discussion of discontinuous conduction mode theory will also be covered. Chapter 3 presents the design of the energy harvester circuit. The circuit will be broken into sub-circuits and each will be studied in detail. The sub-circuits will also be experimentally verified. Chapter 4 shows the results of the parametric study and three optimized circuits are presented. In Chapter 5, the background for the acoustic energy harvester will be presented. Chapter 5 also covers the lumped element model, frequency response and maximum power for the acoustic energy harvester. Chapter 6 will give the conclusions and future work.
CHAPTER 2
THEORETICAL BACKGROUND

The theoretical background for the energy harvesting circuit is presented in this chapter. A broad overview of power converter topologies is presented, along with a discussion of discontinuous conduction mode. A system overview for the acoustical energy harvester will also be presented.

Power Converter Topologies

Power converter circuits are important to energy harvesting because they allow the reclaimed power to be conditioned suitable for electronic loads. It has also been shown that the use of power converter circuits can increase the harvested power over direct charging methods [14, 26].

The main purpose of a power converter circuit is to change the input power conditions (voltage and current) into a different set of conditions suitable for the desired load. These converter circuits can change the current and voltage levels of the signal as well as the frequency. The circuits presented here will focus on the DC – DC switching converter type. Three ideal single inductor circuits, buck, boost and buck-boost are shown and analyzed.

Buck Converter

The buck converter is a DC – DC switching converter used to step down voltage levels. The circuit as shown in Figure 2-1 consists of an inductor/output filter, switch, diode, and control electronics. The $+\text{Vin}$ voltage of the buck converter will always be
The inductor volt-second balance method is based on the concept that the current through the inductor cannot change instantaneously [28]. If the current increases during one part of the switching cycle, it must decrease by that same amount before the end of the switch period. The duty ratio $D$, for the buck converter is defined as the time the switch connects the input voltage to the inductor versus the period of the switch. The time that the switch is off, $1-D$ is referred to as $D'$. The voltage $V_x(t)$ under normal switching behavior can be seen in Figure 2-2. The average value of $V_x(t)$ can be found by

\[
< V_x(t) > = \frac{1}{T_s} \int_0^{T_s} V_x(t) dt = DV_{in} \quad (2-1)
\]

The inductor and capacitor in buck converter act as a low pass filter to reject the switching components and harmonics from the voltage $V_x(t)$. The output voltage is simply equal to the average or DC value of $V_x(t)$. The voltage transfer characteristics of the buck converter can be described by the following,

\[
\frac{V_{out}}{V_{in}} = D \quad (2-2)
\]
Figure 2-2: Typical switching operation for a buck converter. (A) shows the D circuit when the switch is on, (B) shows the D’ circuit when the switch is off, and (C) shows the voltage at the switch node Vx.

**Boost Converter**

The boost DC-DC switching converter is used to step up voltage levels. The circuit shown in Figure 2-3 is a schematic for an ideal boost converter.

Figure 2-3: Schematic for a boost converter.

The volt-second product method will be used to analyze the boost converter. The same conventions will be followed for the boost converter as for the buck converter. Figure 2-4 shows the inductor voltage of the boost converter under normal switching operations.
Figure 2-4: Switching operation of a boost converter. (A) shows the D circuit, when the switch is on, (B) shows the D’ circuit when the switch is off, and (C) shows the voltage across the inductor.

The total volt-second product applied to the inductor over one switching period is given by the following,

\[ \int_{0}^{T_s} V_L(t) dt = (V_{in}) DT_s + (V_{in} - V_{out}) D'T_s. \]  

(2-3)

By setting this equation equal to zero the voltage characteristics of a boost converter can be described by the following,

\[ \frac{V_{out}}{V_{in}} = \frac{1}{D} = \frac{1}{1 - D}. \]  

(2-4)

**Buck-Boost Converter**

The buck-boost converter is different from the buck and boost converters because it can step the voltage either up or down. The buck-boost converter is also unique because it causes the output voltage to be inverted. Figure 2-5 shows a schematic of an ideal buck-boost converter.
Figure 2-5: Schematic of an ideal buck-boost converter.

Again the inductor volt-second balance method will be used to analyze this converter.

The switch waveform of the inductor voltage can be seen in Figure 2-6.

The following shows the volt-second balance of the inductor,

\[
\int_0^{T_s} V_L(t)dt = (V_{in})DT_s + (V_{out})D'T_s. \tag{2-5}
\]

Equating this expression to zero yields the following voltage transfer characteristics for the ideal buck-boost converter,
The equations and models presented here are for ideal converters operating in continuous conduction mode. Another mode of operation that can occur when using switching power converters with unidirectional semiconductor devices (i.e. diodes) is discontinuous conduction mode (DCM).

**Discontinuous Conduction Mode**

The characteristics of power converter circuits change dramatically when DCM occurs. The peak current to rms current ratio is much greater when the circuit operates in DCM. The voltage conversion equations, output impedance, and input impedance also change [28]. These changes in certain situations can be used to the benefit of the designer.

Many models have been developed for studying steady state operation of switching power converters [28-35]. The model type that will be used in this work is the pulse width modulation (PWM) switch model for DCM operation; Figure 2-7 shows the c-PWM switch model.

\[
\frac{V_{out}}{V_{in}} = -\frac{D}{D'} = -\frac{D}{1-D}. \tag{2-6}
\]

The value \(u\), shown in the switch model is given by Vorperian [29].
\[ u = \frac{d^2 v_{cp}}{X i_a} = \frac{d^2 v_{ac}}{X i_p}. \tag{2-7} \]

Where \( X \) is proportional to the switching frequency \( f_s \) and the inductance \( L \),

\[ X = 2 L f_s. \tag{2-8} \]

The PWM switch model shown above, will be used to analyze the buck-boost converter operating in DCM.

**Buck-Boost in DCM**

The buck-boost converter is forced into DCM operation when the circuit has a large load \( R \), a small \( L \), or low \( f_s \). The c-PWM switch is natural for the buck-boost converter because the c terminal is grounded. Figure 2-8 shows the buck-boost circuit with the PWM switch model.

![Buck-Boost in DCM](image)

Figure 2-8: Steady-state buck-boost converter with PWM switch model.

As shown in the figure above \( V_{ac} = V_{in} \) and \( i_p = V_{out}/R \), combining these with Equation (2-7), the following is obtained [32],

\[ u = \frac{D^2 V_{ac}}{XI_p} = \frac{D^2 V_{ac}}{XV_{out}} \frac{R}{D} = \frac{D^2}{XD}. \tag{2-9} \]

Equation (2-9) can also be used to shown the boundary between DCM/CCM for the buck-boost converter,

\[ D^2 R > X \text{ for } DCM \]
\[ D^2 R < X \text{ for } CCM. \tag{2-10} \]
The voltage transfer characteristics of the buck-boost in DCM can be found by inspection using Equation (2-9) and Figure 2-8,

\[
\frac{V_{\text{out}}}{V_{\text{in}}} = u = \frac{D^2 V_{\text{in}}}{X V_{\text{out}}} = R = D \sqrt{\frac{R}{X}}.
\]  
(2-11)

The input impedance of the buck-boost converter can also be found by inspection using Equation (2-7) and Figure 2-8,

\[
R_{\text{in}} = \frac{V_{\text{sw}}}{i_u} = \frac{X}{D^2} = \frac{2 f_s L}{D^2}.
\]  
(2-12)

As seen in Equation (2-11) the voltage transfer characteristics scale linearly with the duty ratio D. The input resistance is linear with the reactance of the inductor L and controllable by the duty ratio D. The input resistance is also independent of the load resistance R. By being able to control the duty ratio, this provides an excellent tool for impedance matching and maximum power transfer.

The buck and boost circuits can be analyzed using the PWM switch model in the same manner as the buck-boost. The voltage transfer characteristics and DCM criteria for each circuit are shown in Table 2-1.

### Table 2-1. DCM Criteria and Steady-State Voltage Transfer Characteristics.

<table>
<thead>
<tr>
<th>Converter Type</th>
<th>DCM Criteria</th>
<th>Voltage Gain ((\frac{V_{\text{out}}}{V_{\text{in}}}))</th>
</tr>
</thead>
<tbody>
<tr>
<td>Buck-Boost</td>
<td>(D^2 R &gt; X)</td>
<td>(\frac{D \sqrt{\frac{R}{X}}}{2})</td>
</tr>
<tr>
<td>Buck</td>
<td>(D' R &gt; X)</td>
<td>(1 + \sqrt{1 + \frac{4X}{D^2 R}})</td>
</tr>
<tr>
<td>Boost</td>
<td>(D'^2 D R &gt; X)</td>
<td>(1 + \sqrt{1 + \frac{4D^2 R}{X}})</td>
</tr>
</tbody>
</table>
As shown in the table, the voltage gain for each DCM converter is dependent on the load resistance. It can also be seen that the voltage gains for the buck and boost converters are more complicated in DCM as opposed to CCM.

The buck-boost circuit operating in DCM has characteristics that can be very useful in energy harvesting circuits. The control of the input resistance with the duty cycle, and the ability for the output voltage to be either higher or lower than the input voltage, make this circuit a good choice for energy harvesting. A buck-boost derived circuit operating in DCM will be used as the converter circuit for the acoustical energy harvester.

**Flyback Circuit**

The flyback circuit can be derived from the buck-boost converter topology. Figure 2-9 shows how the circuit is derived.

![Figure 2-9: The derivation of the flyback circuit from the buck-boost. (A) Shows the basic buck-boost circuit; (B) the inductor is replaced with a coupled inductor; (C) the windings of the inductor are split; (D) the circuit is adjusted to have positive output voltage.]

The governing equations for the flyback converter can be derived in the same manner as the previous circuits. The flyback converter behaves the same as the buck-boost in many respects. Table 2-2 shows the characteristics of the flyback converter for CCM and DCM operation.
Table 2-2. CCM and DCM characteristics of a Flyback Converter.

<table>
<thead>
<tr>
<th></th>
<th>CCM</th>
<th>DCM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Voltage Transfer Characteristics (Vout/Vin)</td>
<td>$M = \frac{V_{out}}{Vin} = \frac{D}{D'}$</td>
<td>$M = \frac{V_{out}}{Vin} = D \sqrt{\frac{R}{X}}$</td>
</tr>
<tr>
<td>Input Impedance</td>
<td>$Z_{in} = \frac{D^2 R}{D^2}$</td>
<td>$Z_{in} = \frac{2f_s L}{D^2}$</td>
</tr>
<tr>
<td>Mode of Operation Criterion</td>
<td>$D^2 R &lt; X$</td>
<td>$D^2 R &gt; X$</td>
</tr>
</tbody>
</table>

The DCM characteristics of the flyback converter are very attractive for impedance matching as with the buck-boost converter. The ideal switch waveforms of the flyback converter are shown in Figure 2-10. While the MOSFET switch is on ($D_1 T_s$), the current in the coupled inductor begins to ramp up. Once the switch is turned off, the energy stored in the inductor is transferred to the secondary ($D_2 T_s$), the current in the secondary ramps down to zero. When the current reaches zero, it tries to reverse directions. However, the diode prevents this and the current remains at zero. The flyback circuit operated in DCM will be used as the main component of the energy harvester circuit.
Figure 2-10: The ideal switch waveforms for a DCM flyback converter. (A) Shows the ideal flyback circuit; (B) shows the $V_{DS}$ of the MOSFET switch; (C) shows the inductor current on the primary side of the transformer; (D) shows inductor current on the secondary side of the transformer.
CHAPTER 3
DESIGN OF THE FLYBACK CIRCUIT

The design and fabrication of the flyback circuit will be presented in this chapter. This will include the circuit design, component selection, printed circuit board layout and tests of the operation.

System Partitioning

The flyback circuit is divided into sub-circuits based on function. This provides a straightforward way to design and analyze the circuit. The main sub-circuits of the flyback converter are the bridge rectifier, control circuit, and power stage. Each of these subsystems will be discussed in detail. The partitioned system is shown in Figure 3-1.

![Diagram of Flyback Converter Sub-circuits](image)

Figure 3-1: This diagram shows the main sub-circuits of the flyback converter.

Other circuits which are necessary for operation of the flyback converter and for measurement purposes will also be covered.

Bridge Rectifier

The purpose of the bridge circuit is to full wave rectify the ac signal coming from the acoustic energy transducer. This full wave rectified signal is then filtered by a large
capacitor to give an approximate DC input to the power stage of the flyback circuit. Two
types of bridge circuits can be used to rectify the ac signal. The first is a simple diode
bridge, and the second is a more complex active bridge. Figure 3-2 shows the diode
bridge circuit.

![Diode Bridge Circuit](image)

**Figure 3-2:** A passive full wave rectifier bridge using diodes.

The diode rectifier bridge provides a compact and simple solution to rectify the ac
signal. The bridge needs to be able to handle voltages up to $50 \, V_{pk-pk}$, the expected
output of the acoustic energy harvester. This means that the diodes will need to be able
to sustain reverse voltages up to 25 V each. The forward drop of the diodes must also be
as low as possible to reduce the amount of power loss. Based on these specifications, the
diodes that are chosen for the bridge are Fairchild BAT54s [36].

The bulk input capacitor must be able to handle voltages up to 25 V. The capacitor
also needs to be large enough that when the power stage draws current, the voltage does
not dip. The capacitor that will be used is a 220 $\mu$F, 25 V, aluminum electrolytic
capacitor.

When the expected voltages are low, a more complicated and efficient bridge can
be used to reduce the wasted power. The more efficient bridge uses MOSFETs to replace
the diodes. The main problem with using this setup is that control signals must be
generated to switch the MOSFETs. Generating these control signals will also consume
power. Figure 3-3 shows the actively controlled rectifier bridge.

![Image: An actively controlled rectifier bridge using MOSFETs.]

Figure 3-3: An actively controlled rectifier bridge using MOSFETs.

The actively controlled rectifier bridge uses NMOS transistors for M2 and M4, and
PMOS transistors for M1 and M3. The NMOS transistors are Fairchild BSS138s [37].
The PMOS transistors are BSS84s [38]. The actively controlled bridge requires two
control signals to turn on the appropriate MOSFETs. These two control signals will be
referred to as A high B low (AHBL) and B high A low (BHAL). The control signals are
displayed in Figure 3-4 relative to input signal.

![Image: Signals that are required to drive the MOSFETs in the active rectifier bridge.]

Figure 3-4: Signals that are required to drive the MOSFETs in the active rectifier bridge.
The circuit to produce the control signals for the rectifier bridge uses an op-amp and two
comparators, the schematic is shown in Figure 3-5.

![Circuit Schematic](image)

**Figure 3-5:** Circuit used to create the control signals for the active bridge rectifier.

The comparators used for this circuit are TLV3492 (U2, U3), and the op-amp is an
OPA2349 (U1) [39, 40]. The resistors, R1 and R2, are used to create a voltage slightly
higher than ground to trip the comparators. The components of this circuit require a
stable supply voltage to operate properly (VCC). This voltage will be provided by the
bulk input capacitor of the bridge after it has charged to a sufficient voltage. The active
bridge will act as a passive diode bridge until that voltage is stable.

There are tradeoffs between the passive rectifier bridge and the active rectifier
bridge. The passive bridge is quite simple and provides good performance at high input
voltages. The active bridge, while more complex, provides superior performance at low
input voltages. The active bridge does, however, require supply voltage and control
electronics. In most cases, the passive diode bridge is a reasonable choice, and its
simplicity makes it favorable over the active bridge.
Control Circuit

The control circuit of the flyback converter is used to turn the MOSFET switch on and off. This circuit is used to control the switching frequency \( f_s \) and duty ratio \( D \).

Figure 3-6 shows a schematic of the control circuit.

![Control Circuit Diagram](image)

**Figure 3-6:** Circuit used to control the MOSFET switch in the power stage.

The control circuit consists of two cascaded stages; the first contains comparator U4 and resistors R3-R6, and the second contains comparator U5, and resistors R7 and R8. The purpose of the first stage is to create a triangle wave at the designated frequency. This is accomplished by the following:

1. R3 and R4 setup a reference voltage equal to approximately 0.5 the VCC value.
2. This causes the output of comparator U4 to try and go to the VCC value.
3. Capacitor C2 begins to charge through resistor R6; the values of these components determine the frequency.
4. Once C2 reaches the value of the reference voltage the output of the comparator U4 begins to ramp down toward zero.
5. This causes C2 to begin to discharge until the value is below the reference.
6. Resistor R5 creates a hysteretic loop that allows the triangle wave to have a larger amplitude.
The second stage of the control circuit creates a square wave of the desired duty cycle at the same frequency as the triangle wave. Resistors R7 and R8 setup a DC reference voltage. Whenever that voltage is greater than the incoming triangle wave, the output of comparator U5 is high when the reference voltage is smaller the output is low.

The comparators need to be able to operate at very low input voltages and not consume very much power. The comparators also need to be able to operate from rail to rail in order to be able to produce a signal suitable for switching the MOSFET. The parts that were chosen are Texas Instruments TLV3492s. They are nanopower push-pull output comparators. The TLV3492 accepts an input voltage from 1.8 – 5.5 V and has an output range that goes rail to rail [40].

The resistors used are all surface mount 0805 thick film chip resistors. R3 and R4 are chosen to be as large as possible to reduce power loss; the value chosen is 5 MΩ. R5 also needs to be large to reduce power loss and is selected to be 2 MΩ. The capacitor C2 is a 16 V, 1000 pF ceramic capacitor. The values of R6 – R8 are used to set the frequency and duty cycle. Specific values will be discussed in the next chapter. However, it should be noted that the resistance values should be kept above 1 MΩ between VCC and GND to reduce the power losses.

**Power Stage**

The power stage of the flyback converter is where the actual voltage and current conditioning takes place. This sub-circuit contains the power MOSFET switch, diode, coupled inductor and filter capacitors. Figure 3-7 shows the power stage of the flyback circuit.
The components for the power stage must be chosen with care. The components must be able to handle the expected voltage and current levels without breakdown or failure. The expected input voltage range that the power stage will see is 0 – 25 V, with a maximum peak current of less than 100 mA. The output voltage will be clamped to 5 V or less, with maximum currents not to exceed 100 mA.

Based on the specifications, the power MOSFET M5 that is chosen is a Fairchild BSS138 [37]. The drain to source breakdown voltage of this device is 50 V, and it can handle up to 220 mA of continuous current. The device also has a relatively low $R_{\text{DSon}}$ of 1 Ω, with a $V_{\text{GS}}$ of 4.5 V. The breakdown levels of BSS138 exceed the expected levels of the converter, and the low $R_{\text{DSon}}$ will keep power losses to a minimum.

The diode D5 used for the power stage is a schottky barrier diode Fairchild BAT54 [36]. This device supports currents up to 200 mA, and a reverse breakdown voltage of 30 V. The primary reason for using a schottky barrier diode, is the forward drop is much lower than for a standard diode. The forward drop of the diode is approximately 0.25 V for the current range of interest.
The coupled inductor needs to be able to handle peak currents of up to 100 mA without saturation. Each of the inductors has a turns ratio of 1:1. Three different inductors are chosen based on size and inductance. The inductance values are 14.4 mH, 4.7 mH, and 1 mH. The 14.4 mH is a custom wound inductor and the other two are commercial products. All of the inductors used exceed the circuit requirements. These inductors will be used as a part of the parametric study in the next chapter.

The final component of the power stage is the output capacitor C3. The value and size of the output capacitor depend on the load electronics and maximum ripple voltage. In general, the output capacitor that will be used is a 100 uF, 16 V, aluminum electrolytic capacitor. A smaller 1 uF ceramic capacitor can also be placed in parallel to reduce ripple due to the ESR of the capacitor.

**Additional Circuits**

There is a need for additional circuits to help with the operation of the flyback converter and for measurement. Two circuits will be discussed in detail, the first is a voltage regulator for the control circuit, and the second is a circuit used to measure small ac currents.

The voltage regulator circuit is needed because the comparators in the control circuit can handle a maximum supply voltage of 5.5 V. The VCC voltage from the bridge rectifier can go as high as 25 VDC. The voltage regulator needs to be able to supply a voltage within the range of the comparators and also be able to operate with input voltages as high as 25 V. A voltage regulator that meets these specifications is the Texas Instruments TPS71501 [41]. The device is rated for 24 V input, however it was tested and was able to handle up to 30 V with no problems. A schematic of the circuit is shown in Figure 3-8.
The capacitors, C4 and C5, are 25V, 1μF, ceramic chip capacitors. The capacitor, C5, is used to stabilize the internal control loop of the TPS71501. The resistors, R9 and R10, and an internal reference voltage are used to set the output voltage. The output voltage is determined by the following equation,

\[
V_{out} = V_{\text{ref}} \left(1 + \frac{R9}{R10}\right) = 1.205V \left(1 + \frac{R9}{R10}\right). 
\]  

(3.1)

The desired supply voltage for the control circuit is approximately 2.5V. Therefore, the values for R9 and R10 are chosen to be 1MΩ.

Measuring very small ac currents accurately is not an easy task. A simple circuit using an instrumentation amplifier can be used to achieve this goal. This circuit is not necessary for the flyback converter to operate. However, it simplifies the measurements of necessary currents. The circuit must be placed in series with the current that is to be measured. Figure 3-9 shows the schematic of the circuit. The instrumentation amplifier used in this circuit is a Texas Instruments INA114BP [42]. The INA114 is a precision, low drift instrumentation amplifier with external gain control.
Figure 3-9: A simple circuit used to measure small AC currents.

Resistor R12 is used to set the gain according to the following formula,

$$V_{out} = (V_{in+} - V_{in-}) \left(1 + \frac{50k\Omega}{R12}\right).$$

(3.2)

The current measurement is then proportional to this output voltage. The resistors, R11 and R12, can be chosen based on the expected input current and desired output voltage. In general, R11 should be kept small to keep the impedance of the current shunt small, however it must be large enough to produce a voltage greater than the offset of the amplifier, 50 uV [42].

**Energy Harvester Circuit**

The energy harvester circuit incorporates the three sub-circuits; the power stage, control circuit, and bridge rectifier, with the voltage regulator to form the complete system. Figure 3-10 shows a schematic of the energy harvester circuit. The dotted lines in the diagram illustrate connections between the sub-circuits. The illustrated energy harvester circuit uses the passive diode bridge to rectify the AC signal. The circuit also includes the voltage regulator to provide the supply for the control circuit.
Figure 3-10: A schematic of the complete energy harvester system.
Experimental Verification

To ensure that the sub-circuits of the energy harvester all work as expected, each one is tested individually. The experimental verification was performed using the following equipment:

1. HP/Agilent 33120A waveform generator.
2. HP/Agilent 3630A triple output DC power supply.
3. HP/Agilent DMM 3457.
4. Tektronix TDS224 four channel, 1MHz, digital oscilloscope.
5. Wavestar digital oscilloscope software.

The digital oscilloscope is connected to a PC via GPIB. The Wavestar software is then used to capture the data. Once the verification for each sub-circuit is complete, the entire energy harvester circuit will be tested.

The first block to be tested is the bridge rectifier. The input to the bridge is a 2.5kHz, 10V_{pk-pk}, sine wave. The input signal is generated using the Agilent 33120 function generator and passed through an isolation transformer to prevent grounding issues. Figure 3-11 shows the bridge rectifier waveforms.

![Bridge Rectifier Waveforms](image)

Figure 3-11: The waveforms for the bridge rectifier.
Shown in the figure are the input sine wave, the rectified sine wave with no capacitor, and the rectified sine wave with a 220μF aluminum electrolytic capacitor. The difference in the rectified sine wave and the input sine wave is due to the forward voltage drop of the diodes.

To test the control circuit block, an Agilent 3630A DC power supply is used to supply the 1.8V VCC. Two potentiometers are used to control the frequency (R6) and duty cycle (R7 and R8). Figure 3-12 shows the control circuit operating at a frequency of ~6kHz and duty cycle of 10%.

![Control Circuit Waveforms](image)

Figure 3-12: Waveforms showing the operation of the control circuit.

The control circuit behaves as expected. Using the potentiometers, the gate drive can be set to frequencies between 1kHz and 30kHz. The duty cycle can be set from 0-100%.

The voltage regulator is tested using the Agilent 33120 function generator to provide a ramp voltage from 0 – 10V. The regulator output is set to 2.4V. Figure 3-13 shows the regulator in operation. The regulator begins working at an input voltage of approximately 1.5V. It then tracks the input voltage until it reaches the desired output.
voltage of 2.4V. The waveforms only show the regulator being tested to 10V. This is due to a limitation of the function generator. The regulator was tested with a DC power supply to a voltage of 30V.

![Voltage Regulator Waveforms](image)

**Figure 3-13:** 10V ramp verifies the operation of the TPS 71501 voltage regulator.

The individual stages of the flyback converter perform as expected. The subcircuits are connected as shown in Figure 3-10. The Agilent 33120 function generator is connected to the bridge rectifier input through an isolation transformer. The input voltage is set to $10V_{\text{pk-pk}}$ at a frequency of 2.5kHz. No other power is provided to the circuit. The output is connected to a Panasonic 3V rechargeable lithium battery model number VL1220. The nominal capacity of battery 7mAh. The battery is initially charged to a voltage of ~3V, this clamps the output to this voltage. Figure 3-14 shows two periods of the flyback in operation. Figure 3-15 zooms in on the time when the switch is on. The inductor current shown is for the primary side. The MOSFET drain to source voltage goes to zero when the switch is on. Once the switch is off, the voltage flies back or goes to the value of the input voltage plus the output voltage. The voltage remains at
this level until the current in the inductor on the secondary side reaches zero. The voltage then begins to oscillate or ring due to parasitic inductance of the board, leakage inductance of the inductor and parasitic capacitances.

Flyback Waveforms

Figure 3-14: Waveforms for two switching cycles of the flyback converter.

Flyback Waveforms

Figure 3-15: A zoomed in waveform showing the time when the MOSFET switch is on.
The input waveforms to the energy harvester circuit are also captured. Figure 3-16 shows the input sine wave, input current, and filtered rectified signal.

![Energy Harvester Input Waveforms](image)

**Figure 3-16:** Input waveforms for the energy harvester circuit.

The input current displayed in Figure 3-16 is measured after the diode bridge and before the bulk input capacitor. The current is measured using the circuit shown in Figure 3-9. The diodes conduct when the voltage of the input sine wave is larger than the voltage of the bulk input capacitor.
CHAPTER 4
PARAMETRIC STUDY AND OPTIMIZATION OF THE FLYBACK CIRCUIT

The energy harvester circuit has many parameters that can be adjusted to improve the performance. These parameters include, but are not limited to, the switching frequency, inductor size and duty cycle. In addition to these parameters, there is also a concern about the size of the converter. This chapter will present a study of the effects of changing these parameters on the efficiency of the converter, as well as the overall converter size. Three optimized circuits will be presented with schematics, component values and PCB layouts.

Parametric Study Constraints

Some basic constraints are used to put a limit on the possible converter configurations. The three components that will be varied are the switching frequency, duty cycle and inductance. The three factors can be controlled independently. However, they must ensure that the energy harvester maintains the same input impedance throughout the tests. The input impedance for the DCM flyback converter, as previously discussed, is given by

\[ Z_{in} = \frac{2 f_c L}{D^2}. \] (4.1)

For this study, the input impedance will be kept at a value of 20kΩ. This value is an estimate of optimal load for the acoustic energy transducer (this will be discussed in more detail in Chapter 5).
Three inductors of varying physical size and inductance values are chosen for the study. Two of the inductors are commercially available, and the third is a custom wound inductor. Table 4-1 shows the specifications for the inductors.

<table>
<thead>
<tr>
<th>Part Number</th>
<th>MFG.</th>
<th>Inductance</th>
<th>Series R</th>
<th>Saturation I</th>
</tr>
</thead>
<tbody>
<tr>
<td>DW3316-475</td>
<td>Coilcraft</td>
<td>4.7mH</td>
<td>26Ω</td>
<td>160mA</td>
</tr>
<tr>
<td>DRQ74-152</td>
<td>Coiltronics</td>
<td>1.5mH</td>
<td>7.8Ω</td>
<td>160mA</td>
</tr>
<tr>
<td>NA</td>
<td>Robert Taylor</td>
<td>14.4mH</td>
<td>0.987Ω</td>
<td>&gt;100mA</td>
</tr>
</tbody>
</table>

The switching frequency will be varied from 350Hz to 20kHz. The duty cycle will be calculated for each setup (frequency and inductor) to ensure the proper input resistance.

The experimental setup will use the same equipment used previously to test the functionality of the energy harvester circuit. The Agilent 33120 function generator connected through an isolation transformer will provide the input to the circuit. The input voltage will be 10 Vpk – pk at 2.5 kHz, and will not change throughout the experiments. The output of the energy harvester will be connected to a 3 V rechargeable lithium battery. The current and voltages at the output are monitored using the HP3457 digital multimeter. The input voltage and input current will be monitored and recorded using the Tektronix TDS224 oscilloscope and Wavestar software.

**Flyback Converter Efficiency**

The main purpose of the parametric study is to maximize the efficiency of the energy harvester circuit. The efficiency of the circuit is defined as the output power divided by the input power. The output power is the DC voltage of the battery times the DC current flowing into the battery. The input power is defined to be the voltage across the bulk input capacitor times the current measured to be flowing out of the diode bridge. Figure 4-1 shows efficiency vs. switching frequency for each of the three inductors. The
frequency was varied from 350Hz – 20kHz. The data was taken twice to show repeatability of the measurements. Figure 4-2 also shows the efficiency vs. switching frequency. However, this time, the switching frequency was limited between 1kHz and 10kHz.

![Efficiency vs. Switching Frequency](image)

Figure 4-1: The efficiency vs. switching frequency for 3 coupled inductors, with the switching frequency varying from 350Hz – 20kHz for two measurements, #1 and #2.

It is immediately evident from the plots that the converter is more efficient for a larger inductor size. For a given frequency, the smaller the inductor size the smaller the duty cycle must also be to keep the input impedance constant. For example, at 5kHz, the duty cycle for the 14.4mH inductor is 8.5%; for the 1.5mH inductor it is 2.74%. If the duty cycle is shorter, that means the on time for the MOSFET is also shorter. The average current for both cases must be the same if the impedance is the same. Since the on time for the 1.5mH inductor is so much shorter, the peak currents must be much larger. The larger peak currents lead to resistive conduction losses in the MOSFET, diodes, inductor and lower converter efficiency.
Efficiency vs. Switching Frequency

![Efficiency vs. Switching Frequency Graph]

Figure 4-2: The efficiency vs. switching frequency, for frequencies 1kHz – 10kHz for two measurements, #1 and #2.

It can also be observed from the graphs that the smaller the inductor size, the higher the optimal point for the switching frequency. This is also related to the peak currents being larger for the smaller inductors. In general, when the switching frequency is increased, the switching losses increase and the conduction losses decrease. The switching losses increase because the MOSFET is turning off and on faster; the parasitic capacitances of this and other devices consume power. The conduction losses are decreased because the peak currents are reduced. The maximum efficiency for the converter occurs when the switching losses and conduction losses for a given configuration are balanced.

**Loss Estimations**

The losses for each component of the flyback converter are calculated and compared to experimental results. For all of the experiments and calculations the input conditions are the same as before. The following are the losses that are considered;
conduction losses of inductor, MOSFET, and diode, control circuit losses and switching losses. The conduction losses for a given parasitic resistance \( R_{\text{par}} \) and current \( i \) are given by

\[
P_{\text{cond}} = \frac{1}{T_s} \int_0^{T_s} i \cdot i \cdot R.
\]  

(4.2)

The peak current in the inductor, MOSFET, and diode can be calculated from the following,

\[
V_{\text{input}} = L \frac{di}{dt}.
\]  

(4.3)

The control circuit losses are estimated from two components; first the power consumption of the comparators with no load, and second the estimated power consumption of the comparators as gate drivers. The first component is determined experimentally; the second \( P_{\text{comp}} \) is determined by the following,

\[
P_{\text{comp}} = f_s \cdot V_{\text{input}} \cdot Q_T.
\]  

(4.4)

Where \( Q_T \) is the total gate charge of the MOSFET. The switching loss for the MOSFET with a turnoff time given by \( T_{\text{off}} \) is the following,

\[
P_{\text{switch}} = \frac{1}{4} \left( V_{\text{input}} + V_{\text{output}} \right) \cdot I_{\text{peak}} \cdot T_{\text{off}} \cdot f_s.
\]  

(4.5)

Using these equations, the losses for each of the three designs are calculated. Included in these losses are the conduction losses of the MOSFET, diode and Inductor. Also included are the MOSFET switching losses and the control circuit losses. Figure 4-3 shows the total losses for the 14.4mH case, Figure 4-4 shows the breakdown of the calculated losses. The calculated losses are shown to be less than measured losses. This is mainly due to inaccuracies in the calculations and underestimates. Figure 4-5 shows
the total losses for the 4.7mH case, Figure 4-6 shows the loss breakdown of the calculated losses. Figure 4-7 shows the total losses for the 1.5mH case, Figure 4-8 shows the loss breakdown of the individual components. The loss estimations are less accurate for the smaller inductor sizes. The main reason for this is that during the experiments it is very hard to control the duty cycle with precision. Small fluctuations in the duty cycle change the peak currents and thus make the estimations inaccurate.

**Figure 4-3:** Losses for the 14.4mH flyback vs. switching frequency.

**Figure 4-4:** Loss breakdown for the 14.4mH components vs. switching frequency.
Losses vs. Switching Frequency

![Graph: Losses vs. Switching Frequency]

**Figure 4-5:** Losses for the 4.7mH flyback vs. switching frequency

4.7mH Calculated Individual Component Losses vs. Switching Frequency

![Graph: 4.7mH Calculated Individual Component Losses vs. Switching Frequency]

**Figure 4-6:** Loss breakdown for the 4.7mH components vs. switching frequency
Figure 4-7: Losses for the 1.5mH flyback vs. switching frequency.

Figure 4-8: Loss breakdown for the 1.5mH components vs. switching frequency.

Input Impedance

Ideally, the input impedance of the energy harvester circuit should be equal to the input impedance of the power stage to maximize power flow. However, this is not the case. Since the control circuit is also connected to the same point as the power stage,
some current flows into that circuit and it presents its own impedance. The two impedances are in parallel. If the impedance presented by the control circuit is much larger than the power stage impedance it can be neglected. The impedance of the control circuit is directly related to switching frequency, and for sufficiently high switching frequencies, the impedance is not negligible. Figure 4-9 shows the resistance vs. switching frequency for the three different inductors. These values were calculated by measuring the input voltage and current simultaneously, the values are averaged over one period and divided to get the input resistance.

**Input Resistance vs. Switching Frequency**

![Graph showing input impedance vs. switching frequency for different inductors.]

Figure 4-9: Input impedance to the energy harvester circuit vs. switching frequency.

The impedance variations are fairly similar for the three different inductors. As expected, the impedance drops with switching frequency. The impedance of the system varies from approximately 25kΩ to 16kΩ.
Optimal Designs

Based on the efficiency data presented earlier in the chapter, three designs are obtained, one for each inductor. Table 4-2 shows the designs, along with the necessary component values.

| Design #1   | 14.4mH | 3000 | 6.57% | 241kΩ | 604 kΩ | 342kΩ | 1.77 |
| Design #2   | 4.7mH  | 6000 | 5.31% | 117kΩ | 610 kΩ | 336 kΩ | 1.82 |
| Design #3   | 1.5mH  | 7000 | 3.24% | 98.5kΩ | 618kΩ | 328 kΩ | 1.88 |

The component values are based on measurements taken while the converter is in one of the desired setups. The last column of the table refers to the ratio of R7 to R8. Specific values are not important as long as this ratio is met. If space is not a concern, then potentiometers should be used so that tweaking is possible. A complete bill of materials for each design is presented in Appendix.

PCB layouts are completed for each of the optimal designs. PCAD 2001 is the software used to create the PCB layouts. The layouts are done to try and minimize the overall size. Also, to minimize size, components are placed on both sides of the board. Figure 4-10 shows the PCB layout for the 1.5mH, optimized design. The dimensions of the PCB layout are shown on the top and right, and are given in inches. Figure 4-11 shows the layout for the 4.7mH design, and Figure 4-12 shows the layout for the 14.4mH design. It is easy to see that the size of the PCB increases greatly as the inductor size increases. The surface area of each design can be seen from the layouts. However, also of importance, is the height estimates for each design. The height for each design is as follows: 14.4mH – 1 inch, 4.7mH – 0.4 inches, and 1.5mH – 0.4 inches.
Figure 4-10: Top and bottom views of the PCB layout for the 1.5mH design. Dimensions are given in inches.

Figure 4-11: Top and bottom views of the PCB layout for the 4.7mH design. Dimensions are given in inches.
Figure 4-12: Top and bottom views of the PCB layout for the 14.4mH design. Dimensions are given in inches.

The layouts as shown in Figures 4-10, 4-11, and 4-12 are not shown to scale. To provide a better comparison of the size of each design, Figure 4-13 shows outlines of all three designs drawn to scale. A summary of all the dimensions, areas, and volumes is provided in Table 4-3.

Table 4-3. Summary of Dimensions for Three Optimal Designs.

<table>
<thead>
<tr>
<th>Inductor</th>
<th>Length</th>
<th>Width</th>
<th>Height</th>
<th>Area</th>
<th>Volume</th>
</tr>
</thead>
<tbody>
<tr>
<td>Design #1</td>
<td>14.4mH</td>
<td>2.02 in</td>
<td>1.57 in</td>
<td>1.00 in</td>
<td>3.17 in²</td>
</tr>
<tr>
<td>Design #2</td>
<td>4.7mH</td>
<td>1.15 in</td>
<td>0.85 in</td>
<td>0.40 in</td>
<td>0.98 in²</td>
</tr>
<tr>
<td>Design #3</td>
<td>1.5mH</td>
<td>0.91 in</td>
<td>0.85 in</td>
<td>0.40 in</td>
<td>0.77 in²</td>
</tr>
</tbody>
</table>
Figure 4-13: Outlines of the three optimized designs drawn to scale.
The main focus of this chapter is to discuss an application of the DCM flyback converter. More specifically the chapter will cover the role of the converter in acoustic energy harvesting. First the acoustic energy transducer will be discussed and second experimental data will be provided to verify the need for the DCM flyback converter.

**Helmholtz Resonator**

The main component of the acoustic energy harvester is a Helmholtz resonator with a compliant piezoceramic composite backplate. A diagram of the Helmholtz resonator is shown in Figure 5-2. The Helmholtz resonator, as shown above, is a multi energy domain system. The system contains elements from the acoustical, mechanical and electrical domain. The compliant back-plate of the Helmholtz resonator is used to convert the acoustic energy into mechanical energy. The piezoelectric element then uses the mechanical energy to generate electrical energy. An effective way to understand and analyze the system is through lumped element modeling.
Lumped Element Model

Lumped element modeling is an extremely useful tool for the design and analysis of systems containing multiple energy domains. The lumped element model for systems, directly relates design parameters to the frequency response of the system. In order for the lumped element model to be valid, it is important that certain assumptions about the system are met. Most importantly, the wavelength of interest must be significantly larger than the characteristic length of the system.

Using lumped element modeling, an equivalent circuit has been developed for the compliant-backplate Helmholtz resonator [43, 44]. The equivalent circuit is shown in Figure 5-3. The elements shown are in the acoustic and electrical domains.

Figure 5-3: Lumped element model equivalent circuit for the compliant-backplate Helmholtz resonator.
Table 5-1 describes the elements shown in the equivalent circuit. The elements are grouped according to the structure that from which they originate. The equations relating the geometry of the structures to values in the model are shown in the work done by Liu et al. [43] and Horowitz et al. [44]. The first subscript of the symbol gives the domain ("a" for acoustic, "e" for electrical), the second subscript of the symbol gives the location ("N" for neck, "C" for cavity).

Table 5-1. Description of the Lumped Element Parameters used in the Helmholtz Resonator Equivalent Circuit.

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>Neck</td>
<td></td>
</tr>
<tr>
<td>MaN</td>
<td>Acoustic mass of the neck</td>
</tr>
<tr>
<td>RaN</td>
<td>Acoustic flow resistance of the neck</td>
</tr>
<tr>
<td>Cavity</td>
<td></td>
</tr>
<tr>
<td>Cac</td>
<td>Acoustic compliance of the cavity</td>
</tr>
<tr>
<td>Cad</td>
<td>Short circuit acoustic compliance of the piezoelectric composite diaphragm</td>
</tr>
<tr>
<td>Diaphragm</td>
<td></td>
</tr>
<tr>
<td>Radr</td>
<td>Acoustic resistance of the diaphragm</td>
</tr>
<tr>
<td>Mad</td>
<td>Acoustic mass of the piezoelectric composite diaphragm</td>
</tr>
<tr>
<td>Madrad</td>
<td>Acoustic radiation mass of the diaphragm</td>
</tr>
<tr>
<td>Electrical</td>
<td></td>
</tr>
<tr>
<td>φ</td>
<td>Impedance transformation factor</td>
</tr>
<tr>
<td>CEB</td>
<td>Blocked electrical capacitance of the piezoelectric composite diaphragm</td>
</tr>
</tbody>
</table>

The frequency response of the Helmholtz resonator contains two separated resonant peaks. The first resonant peak is due to Helmholtz frequency. The second resonant peak is caused by the natural frequency of the diaphragm. The value of this resonance is roughly given by the following,

$$\omega = \sqrt{\frac{C_{EB} + \phi^2 C_{ad}}{C_{ad} C_{EB} (M_{ad} + M_{adrad})}}.$$  \hspace{1cm} (5.1)

It is important for the acoustical energy harvester to be operated at the resonance of the electromechanical diaphragm to ensure the maximum amount of energy is available.
Frequency Response

The frequency response of the Helmholtz resonator is dependent on the geometry and material properties of the device. The dimensions for the compliant backplate Helmholtz resonator used in this study are shown in Table 5-2. These dimensions are shown in millimeters and are used to estimate the values for the lumped element model.

Table 5-2 shows the estimates for each element of the model.

<table>
<thead>
<tr>
<th>Table 5-2: Physical Dimensions of the Helmholtz Resonator.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Section</td>
</tr>
<tr>
<td>------------------</td>
</tr>
<tr>
<td>Neck</td>
</tr>
<tr>
<td></td>
</tr>
<tr>
<td></td>
</tr>
<tr>
<td>Cavity</td>
</tr>
<tr>
<td></td>
</tr>
<tr>
<td>Diaphragm</td>
</tr>
<tr>
<td></td>
</tr>
<tr>
<td></td>
</tr>
<tr>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Table 5-3: Estimates for the Lumped Element Parameters of the Helmholtz Resonator.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Description</td>
</tr>
<tr>
<td>--------------------------------------------------</td>
</tr>
<tr>
<td>Acoustic mass of the neck</td>
</tr>
<tr>
<td>Acoustic flow resistance of the neck</td>
</tr>
<tr>
<td>Acoustic compliance of the cavity</td>
</tr>
<tr>
<td>Short circuit acoustic compliance of the diaphragm</td>
</tr>
<tr>
<td>Acoustic resistance of the diaphragm</td>
</tr>
<tr>
<td>Acoustic mass of the diaphragm + Acoustic radiation mass of the diaphragm</td>
</tr>
<tr>
<td>Impedance transformation factor</td>
</tr>
<tr>
<td>Blocked electrical capacitance diaphragm</td>
</tr>
</tbody>
</table>
The estimates are given in the acoustic domain, as well as the electrical domain. To convert the values from the acoustic domain to the electrical domain and vice versa, the transformer turns ration ($\phi$) is used to reflect the values into the desired domain. These values can then be used with a simulation tool to calculate the frequency response of the system. The transfer function for the lumped element model shown is found using basic circuit analysis. Once the transfer function is obtained, it is entered into MATLAB to obtain the frequency response of the system.

The magnitude of the frequency response is shown in Figure 5-4; the frequency range is from 1000 – 3000Hz, the transfer function and units are given by

$$\frac{V_{HR\text{out}}}{P_{in}} = \frac{Volts}{Pascals}.$$  \hspace{1cm} (5.2)

The graph shows two separated resonant peaks, as expected. The resonance of the diaphragm is at approximately 2500Hz, and the resonance due to the cavity is at approximately 1750Hz. The phase response of the system is shown in Figure 5-5.

![Magnitude Response of the Helmholtz Resonator](image)

Figure 5-4: Graph showing the magnitude frequency response of the Helmholtz resonator.
Figure 5-5: Graph showing the phase response of the Helmholtz resonator.

**Maximum Power Transfer Theory**

The concept of maximum power transfer is important to energy harvesting to ensure that the energy harvesting system is able to take full advantage of the available power. In Figure 5-6, the model of the Helmholtz resonator is simplified by reflecting all acoustic parameters across the transformer to the electrical domain, thus eliminating the transformer.

Figure 5-6: Simplified model of the Helmholtz resonator.

This simplified model will be used to explore the concept of maximum power transfer for the Helmholtz resonator.
The output of the Helmholtz resonator is an ac voltage signal. The DCM theory that provides the impedance matching is based on the converter input being a constant voltage. Since the ac voltage is not desirable for the input to the converter circuit, the signal must be rectified. Typically after a rectifier bridge a large bulk capacitor is used to filter the ac components and generate a DC voltage. Using a bulk capacitor after the rectifier bridge is the approach taken by Ottman et al. [20, 26].

Figure 5-7 shows a model of the Helmholtz resonator connected to a rectifier bridge with a large bulk capacitor, $C_{\text{rect}}$. This model of the Helmholtz resonator can be simplified to a Norton equivalent. The Norton equivalent circuit makes analysis of the maximum power transfer easier to handle. Figure 5-8 shows the Norton equivalent connected to the bridge rectifier.

![Figure 5-7: Helmholtz resonator circuit with a rectifier bridge, filter capacitor and load.](image)

![Figure 5-8: Norton equivalent circuit of the Helmholtz resonator connected to the bridge rectifier.](image)
The maximum power for a circuit of this form can be found by following the approach used by Ottman [20, 26]. In the work done by Ottman, they do not consider a loss element (Rp) in the model.

The circuit shown in Figure 5-8 can be broken into two time intervals per period. The first interval will be referred to as the commutation interval. This is when the diodes are all off. In the second interval, the diodes conduct and the load is connected to the source. Figure 5-9 shows the theoretical waveforms.

![Theoretical waveforms of the Norton equivalent circuit](image)

Figure 5-9: Theoretical waveforms of the Norton equivalent circuit

The output current $i_o(t)$ can be defined by the following if $C_{rect}$ is much larger than $C_p$,

$$i_o(t) = \begin{cases} 0 & 0 < \omega t \leq u \\ i_p(t) & u < \omega t \leq \pi. \end{cases}$$

To find the average output current, the following is used

$$<i_o(t)> = \frac{1}{\pi} \left( \int_0^{\pi} I \sin(\omega t) \, d(\omega t) \right),$$

$$<i_o(t)> = \frac{I}{\pi} \left( 1 + \cos(u) \right).$$

During the commutation interval, the following relation is used
\[ I \sin(\omega t) \frac{v_p(t)}{R_p} = C \frac{\partial v}{\partial t}. \quad (5.6) \]

Integrating this equation over the interval 0 to \( u \), gives the following

\[ \int_0^u I \sin(\omega t) \hat{d}(\omega t) = C \left( v_p(u) - v_p(0) \right) + \frac{V_{\text{rect}} * u}{R_p}, \quad (5.7) \]

\[ \frac{I}{w} \left( \cos(\omega t) \right)_{0}^{u} = 2V_{\text{rect}} C \frac{\omega}{p} \frac{V_{\text{rect}} * u}{R_p}. \quad (5.8) \]

The equation above can be reduced to

\[ \cos(u) = 1 - \frac{2V_{\text{rect}} \omega C_p}{I} - \frac{V_{\text{rect}} * \omega * u}{IR_p}. \quad (5.9) \]

Inserting Equation (5.9) into Equation (5.5) gives the result for the average load current

\[ <i_o(t)> = \frac{I}{\pi} \left\{ 2 - \frac{2V_{\text{rect}} \omega C_p}{I} - \frac{V_{\text{rect}} * \omega * u}{IR_p} \right\}. \quad (5.10) \]

The average value of the output voltage is

\[ <v_o(t)> = V_{\text{rect}}. \quad (5.11) \]

The average output power is the product of the output voltage and output current given by the following

\[ <P(t)> = \frac{2V_{\text{rect}}}{\pi} \left\{ I - V_{\text{rect}} \omega C_p - \frac{V_{\text{rect}} * \omega * u}{2R_p} \right\}. \quad (5.12) \]

Using Equation (5.9) \( V_{\text{rect}} \) can be found in terms of \( u \)
Using Equation (5.12), Equation (5.13), and given the values for $C_p$, $R_p$, and $\omega$, the maximum average power and optimal load resistance can be found for any system that can be represented by the circuit in Figure 5-8.

**Helmholtz Resonator at Maximum Power Transfer**

The Helmholtz resonator can be modeled as the system in Figure 5-8. The values for $C_p$ and $R_p$ at the resonant frequency are found using Matlab. Table 5-4 shows the values that were used to calculate the maximum power for the Helmholtz resonator.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Symbol</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Resonant Frequency</td>
<td>$f$</td>
<td>2500Hz</td>
</tr>
<tr>
<td>Device Capacitance</td>
<td>$C_p$</td>
<td>3.8nF</td>
</tr>
<tr>
<td>Device Resistance</td>
<td>$R_p$</td>
<td>50kΩ</td>
</tr>
<tr>
<td>Bulk Capacitor</td>
<td>$C_{\text{rect}}$</td>
<td>100uF</td>
</tr>
<tr>
<td>Input Current</td>
<td>$I$</td>
<td>0.75mA</td>
</tr>
<tr>
<td>Load Resistance</td>
<td>$R_{\text{load}}$</td>
<td>1kΩ-100kΩ</td>
</tr>
</tbody>
</table>

The value for the bulk capacitor, $C_{\text{rect}}$, was simply chosen such that it was much larger than $C_p$. If the value of $C_{\text{rect}}$ is not large enough then it begins to interact with $C_p$ reducing the output power. The value for the input current was chosen to simulate an input sound pressure level of approximately 134 dB referenced to 20uPa. The other values were determined using Matlab. Figure 5-10 shows the expected power vs. load resistance for the Helmholtz resonator.
Acoustic Energy Harvester Simulations

An advantage of using a lumped element model for the Helmholtz resonator is that the system can easily be simulated using a circuit simulation tool. This reduces the amount of time required to test multiple designs. The simulator that is used for this particular case is Saber. The complete lumped element model for the Helmholtz resonator is used, as well as the bridge rectifier and power stage of the flyback circuit. Because of the complexity, the control circuit is not used in order to reduce simulation times. Instead of the control circuit, a controllable square wave source is used.

The Saber simulations are used estimate the power delivered to a battery. Two cases are simulated to show the expected improvement of using the energy harvester circuit. The first case uses the 4.7mH, optimized design to charge the battery. In the second case the battery is connected directly across the large capacitor of the rectifier bridge. These two cases will be compared to experimental data later in the chapter.
Experimental Verification

Experimental results are obtained for the acoustic energy transducer and for the energy harvester circuit. The experimental data includes the frequency response, maximum output power of the Helmholtz resonator, and the output power of the acoustic energy harvester system. The experimental results will be compared to theoretical and simulation results presented in the previous chapter.

A simple experimental setup will be used to characterize the acoustic energy harvester. All of the experimental results will be obtained using small variations of this setup. The Helmholtz resonator is flush mounted to the end of a normal incidence plane wave tube. This tube is used to provide the acoustic excitation for the resonator. The tube consists of a 38-inch long, 1 square inch duct that permits characterization in a known acoustic field up to 6.7kHz. Four Bruel and Kjaer (B&K) type 4138 microphones are used to monitor the acoustic field, cavity pressure, and incident pressure simultaneously. Figure 5-11 shows a diagram of the acoustic setup. The four microphones are connected to a B&K PULSE Multi-Analyzer System Type 3560. The PULSE system is used to supply a bias voltage for the microphones. In addition, it is also used as a data acquisition system, and to generate the source waveforms. The output of the PULSE system is fed into a Techron 7540 power amplifier that is connected to a BMS 4590P compression driver. The compression driver is capable of producing acoustic waves from 200Hz to 22kHz, and is connected to a transition piece and mounted to the end of the PWT.
Figure 5-11: Acoustic setup for testing the Helmholtz resonator.

The electrical terminals of the compliant-backplate Helmholtz resonator are connected to a 1:1 isolation transformer to prevent ground loops. The various electrical signals are monitored using a Tektronix TDS224 digital oscilloscope. The DC voltages and currents are measured using an HP 3457 digital multimeter. Figure 5-12 shows a diagram of the complete setup.

**Helmholtz Resonator Frequency Response**

The frequency response of the Helmholtz resonator is necessary to determine the natural frequency of the device. The experimental setup builds upon the basic setup discussed. The output of the Helmholtz resonator is connected to the PULSE system. The incident microphone acts as a reference and is compared to the output of the resonator.

A pseudo random waveform is generated by the PULSE system to test the frequency response. The signal is filtered by the PULSE system to only produce frequencies between 1kHz – 6.4kHz. A 3200 bin FFT analyzer with 1000 averages was used on the incoming signals.
Figure 5-12: Complete experimental setup for testing the acoustic energy harvester system.

The frequency span of the FFT was from 0 – 6.4kHz, giving a bin width of 2Hz. The signal on the incident microphone is taken to be the same as the input to the acoustic energy harvester. The frequency response is then taken to be the electrical output of the Helmholtz resonator divided by the incident microphones signal, the transfer function and units are shown by,

\[ \frac{V_{HRout}}{P_{in}} = \frac{Volts}{Pascals}. \]  

(5.14)

Figure 5-13 shows the magnitude of the frequency response for the Helmholtz resonator. The results of the experimental frequency response compare well with results from the lumped element model results presented in the last chapter. The peak of the diaphragm resonance is close to 2.5kHz and the resonance due to the cavity is approximately
1.75kHz. The resonant frequency of the diaphragm will be used for the rest of the experiments.

![Magnitude Response of the Helmholtz Resonator](image)

Figure 5-13: Magnitude response of the Helmholtz resonator.

**Helmholtz Resonator Optimal Load**

The load presented to the Helmholtz resonator affects the efficiency and output power. In order to extract the maximum amount of power, the optimal load needs to be used. The Helmholtz resonator is connected through an isolation transformer to a rectifier bridge and large capacitor. A variable load resistor is connected in parallel with capacitor. The resistor is swept from 1kΩ - 100kΩ in order to find the optimal load.

The PULSE system is setup to generate a single tone sine wave at the resonant frequency of the diaphragm. The incident microphone is used to measure the pressure input to the Helmholtz resonator, the pressure is 134 dB referenced to 20 µPa. The microphones, mic 1 and mic 2, are flush mounted in a rotating plug to the sidewall of the
plane wave tube. The microphones need to be rotated to cancel any phase mismatch. The two microphones are used to calculate the incident acoustic input power.

The power delivered to the load is monitored using the TDS224 digital oscilloscope. The voltage is monitored directly, and the current is measured using the instrumentation amplifier circuit discussed earlier. These signals are captured using the Wavestar software, and transferred to a PC. Simultaneously, the acoustic information is also captured using the PULSE system. The FFT analyzer is set to a frequency span of 6.4kHz, with 3200 spectral lines. The analyzer is set to take 100 averages. The results of the load test are compared to the theoretical results and are plotted in Figure 5-14.

![Output Power vs. Load Resistance](image)

Figure 5-14: Output power of the Helmholtz resonator for varying resistive loads.

The experimental data agrees well with the theoretical results. The experimental data points are all within 5% of the theoretical results. The optimal load according to the experimental results is approximately 20kΩ.
The incident acoustic power of the Helmholtz resonator is also monitored for varying load conditions. The incident acoustic power is found by using the two-microphone method for sound intensity [45]. The electrical load of the resonator changes the acoustic impedance and thus, also varies the incident acoustic power. Figure 5-15 shows the acoustic power for load resistances from 1kΩ to 100kΩ.

![Incident Acoustic Power vs. Load Resistance](image)

Figure 5-15: Incident acoustic power of the Helmholtz resonator for varying loads resistances.

The general trend in Figure 5-15 is for the acoustic power to decrease with increasing load resistances. The efficiency of the acoustic energy transducer can also be calculated using this data. Figure 5-16 shows the efficiency versus load resistance. The efficiency curve approximately follows the output power curve. The peak efficiency is at 20 kΩ and is approximately 17.5%.

In addition to the efficiency it is also of importance to note the output voltage of the Helmholtz resonator for increasing incident sound pressure levels. The optimal load of
20 kΩ is connected to the output and the voltage is monitored using a digital multimeter.

Figure 5-17 shows the voltage across the load for increasing sound pressure levels.

Figure 5-16: Efficiency of the Helmholtz resonator versus load resistance.

Figure 5-17: Helmholtz resonator output voltage versus incident pressure.
Acoustic Energy Harvester

The design of the acoustic energy harvester, simulations and previous experiments culminate in the final verification of the acoustic energy harvester. The resonant frequency and optimal load of the Helmholtz resonator have been determined. The energy harvester circuit has been designed and optimized for operation with the Helmholtz resonator. The energy harvester circuit will be connected to the Helmholtz resonator and used to charge a battery.

Three configurations of the energy harvester circuit will be compared with two linear regulators and a direct charging method. The three configurations of the energy harvesting circuit will be the 14.4mH circuit, 4.7mH circuit and 1.5mH circuit. The two linear regulators utilize the TPS71501; one design set the output voltage to 3.3V and the other design sets the output voltage to 5.0V. The direct charging method connects the battery directly across the output of the rectifier bridge. Figure 5-18 illustrates the three possible connections of the Helmholtz resonator to the battery. All three configurations use the same passive diode bridge. The bridge has a small, current sense resistor and circuit to monitor the current flowing into the particular setup. The power will be monitored at three locations: the acoustic input power, the converter input power and power delivered to the battery. The input acoustic power is measured using the two-microphone method. The converter input power and power delivered to the battery are both measured by recording the voltage and current simultaneously. These power measurements will provide a clear picture of the efficiency of each stage during the energy harvesting process. The three power measurements allow for three efficiencies to be defined. The first efficiency is the Helmholtz resonator efficiency, which is the converter input power divided by the acoustic input power. The second efficiency is the
converter efficiency, which is the power delivered to the battery divided by the converter input power. The final efficiency is the total efficiency, or the power delivered to the battery divided by the acoustic input power.

Figure 5-18: Three possible configurations of connecting the Helmholtz resonator to a rechargeable battery, also shown are three points at which the power will be monitored.

The experimental setup is similar to the setup discussed previously. Four microphones will be used to capture the acoustic information. The digital oscilloscope and digital multimeters will be used to acquire the electrical data. The PULSE system is setup in same configuration as the load test. For each case, the incident acoustic pressure is varied from approximately 125 dB – 150 dB.

The first case to be tested is the direct charging method. This setup will be used as a baseline to compare all other data. Figure 5-19 shows a comparison of the experimental data, and the data from the Saber simulations. The figure shows good agreement between the simulation data and the experimental data. The power delivered to the battery for a 150 dB acoustic input is 9.35 mW.
The second case to be tested is the 4.7 mH, DCM flyback converter design. The experimental data is compared to the Saber simulation data and the data from the direct charging method. Figure 5-20 shows the simulation and experimental data for the 4.7 mH design and the direct charging method. The 4.7 mH experimental data agrees well with the simulation data. The simulation data predicts a slightly higher output power. This is due to the fact that the control circuit is not properly accounted for during the simulations. The experimental data for the 4.7 mH design and the direct charging method show the same trend as the simulation data. There is a 267% increase in the amount of power delivered to the battery when the energy harvester circuit is used.

The third case to be tested is the linear regulator. The output voltage of the linear regulator is designed to be 5.0 V. The experimental data is compared to the direct charging method and 4.7 mH design. Figure 5-21 shows the power delivered to the battery for each of the test cases.
Figure 5-20: Output power of the 4.7mH optimized design and the direct charging method versus incident pressure.

Figure 5-21: Output power for a linear regulator, 4.7mH design and direct charging versus incident pressure.

It is obvious from Figure 5-21 that the energy harvester circuit is superior to the linear regulator and the direct charging method at high sound pressure levels.
The remaining cases (14.4mH design, 1.5mH design, and 3.3V linear regulator) are each tested and compared to the direct charging method. Figure 5-22 shows the experimental data for the 14.4mH; Figure 5-23 shows the data for the 1.5mH design; and Figure 5-24 shows the data for the 3.3V linear regulator.

**Figure 5-22:** Output power of the 14.4mH optimized design and the direct charging method versus incident pressure.

**Figure 5-23:** Output power of the 1.5mH optimized design and the direct charging method versus incident pressure.
The 14.4mH and 1.5mH versions of the energy harvester circuit behave similar to the 4.7mH case. The estimated increase in power delivered for the 14.4mH case is 277% at an incident pressure of 148dB and for the 1.5mH case it is 235% at an incident pressure of 148dB. The main reason for the difference is the increase in efficiency of the converter with increasing inductor size. The 3.3V linear regulator does not perform very well compared to the direct charging method. The main reason for the decrease in performance is because the output voltage of the regulator is very close to the nominal voltage of the battery. When charging the battery, the voltage needs to be able to go above the nominal voltage and the regulator prevents this from happening. These results are difficult to compare directly to the results of others in the field (such as Ottman[20, 26]) because of the difference in power levels.

The experimental data shows that the three configurations of energy harvester circuit outperform direct charging of the battery, as well as the two versions of linear
regulators. This is true for sufficiently high incident sound pressure levels. At lower sound pressure levels, the energy harvester circuit is less efficient than the other methods of power conversion.

The Helmholtz resonator efficiency is shown in Figure 5-25 for the 5.0V linear regulator, 14.4mH design, 4.7mH design and 1.5mH design.

![HR Efficiency vs. Incident Pressure](image)

Figure 5-25: The Helmholtz resonator efficiency versus incident sound pressure level for the 5.0V linear regulator, 4.7mH design, 14.4mH design and 1.5mH design.

The Helmholtz resonator efficiency stays approximately constant for the three-flyback designs, while the efficiency of the linear regulator drops drastically. The converter efficiency is show in Figure 5-26 for the 5.0V linear regulator, 14.4mH design, 4.7mH design and 1.5mH design. Figure 5-27 shows the total efficiency of the direct charging method, 5.0V linear regulator, 4.7mH design, 1.5mH design and 14.4mH design. The total efficiency is defined as the amount of power stored in the battery divided by the acoustic input power. The data shown in Figure 5-25 indicates that for sound pressure levels above 137 dB, it is advantageous to use the energy harvesting system. Below the
level of 137 dB, the direct charging and linear regulator prove to be more efficient. The data also shows the 14.4 mH design to be more efficient than the 4.7 mH design, as expected.

**Converter Efficiency vs. Incident Pressure**

![Converter Efficiency vs. Incident Pressure](image)

Figure 5-26: Converter efficiency versus sound pressure level for the 5.0V linear regulator, 4.7mH design, 14.4mH design and 1.5mH design.

**Total Efficiency vs. Incident Pressure**

![Total Efficiency vs. Incident Pressure](image)

Figure 5-27: System efficiency of the direct charging method, 5.0V linear regulator, 4.7mH design, 14.4mH design and 1.5mH design versus incident pressure.
CHAPTER 6
CONCLUSIONS AND FUTURE WORK

Summary of Results

A discontinuous conduction mode flyback converter has been designed and optimized for operation with a compliant-backplate Helmholtz resonator. A theoretical model was developed for analyzing the system. The model can also be extended to any similar structure and setup. Three circuits have been developed and optimized based on three different inductors. These circuits vary in size and efficiency. The larger circuits are more efficient but take up much larger space, while the smaller circuit sacrifice efficiency for space. PCB layouts have been completed for each of the circuit designs. The smallest design has a surface area of just larger than 0.75 sq. inches and total volume of 0.3 cubic inches.

Characterization of the optimized circuits using the compliant-backplate Helmholtz resonator show that for sound pressure levels above 143dB referenced to 20uPa, the DCM flyback converter outperforms the direct charging method and linear regulators. Specifically, the energy harvester circuits provide an average improvement of 260% at 156dB in output power delivered to a battery. This number could be increased even more for higher sound pressure levels. The main reason for the drastic improvement in the delivered power is due to the impedance matching. The output power levels of the energy harvester circuit are on the order of 20 – 30mW continuous power for an incident pressure of 156dB. This power level is sufficient for many low power electronics and sensors.
Future Work

The work that has been completed on the acoustic energy harvester focused on impedance matching of the acoustic source to the input of the power converter. The impedance of the power converter is electronically controllable. However, it is affected by the control circuit and bulk input capacitor. Improvements can be made to the control circuit to minimize these effects. The bulk capacitor however, is necessary to provide a stable DC input to the circuit so it cannot be removed. If that capacitor could be removed, more power would be available to the converter circuit. This option was not explored thoroughly for this work because removing the input capacitor provides an unstable voltage to the converter. One solution to this problem is to use multiple converters; some that have large bulk capacitors and some that do not. The converters with large bulk capacitors would then be used to provide a stable bias to the control circuits.

The acoustic energy harvester was tested to an incident sound pressure level of 156dB. This level was not exceeded because the voltage levels would have caused damage to the circuit. The Helmholtz resonator and test structure can handle sound pressure levels in excess of 160dB. If the maximum input range of the energy harvester circuit were increased to handle the voltages at this input level, the circuit performance would increase even more over the direct charging method. Another possible solution would be to change the input impedance of the energy harvester circuit to limit the voltage. This would, however, cause a decrease in total system efficiency.

Another possibility for the acoustic energy harvester is to use MEMS based acoustic generators as a source. The MEMS based acoustic generators are attractive because of batch fabrication, small size, and matched parameters. The energy harvester
circuit could connect to one acoustic generator or to an array of them. The circuit would be able to connect the generators in series or parallel and choose which configuration would provide the most power.
The bill of materials will be presented here for each of the three energy harvester

circuit designs. Table A-1 gives the necessary components for the 1.5mH design, Table
A-2 for the 4.7mH design, and Table A-3 for the 14.4mH design.

Table A-1. Bill of Materials for 1.5mH Design.

<table>
<thead>
<tr>
<th>Count</th>
<th>RefDes</th>
<th>Value</th>
<th>Mfr</th>
<th>Part Number</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>C21</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C22</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C4</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C5</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C11</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C31</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td>1</td>
<td>C2</td>
<td>1000pF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Aluminum, 100-uF, 6.3-V, 20%, FK Series</td>
</tr>
<tr>
<td>1</td>
<td>C3</td>
<td>100uF</td>
<td>Panasonic</td>
<td>EEVFK0J101P</td>
<td>Capacitor, Aluminum, SM, 220-uF, 25-V, 150-milliohms (FK series)</td>
</tr>
<tr>
<td>1</td>
<td>C1</td>
<td>220uF</td>
<td>Panasonic</td>
<td>EEVFK1E221P</td>
<td>Capacitor, Aluminum, SM, 220-uF, 25-V, 150-milliohms (FK series)</td>
</tr>
<tr>
<td>5</td>
<td>D1</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td></td>
<td>D2</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td></td>
<td>D3</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td></td>
<td>D4</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td></td>
<td>D5</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td>1</td>
<td>L1</td>
<td>DRQ74-152</td>
<td>Cooper</td>
<td>DRQ74</td>
<td>Inductor, SMT, 1.5-mH, 0.16-A, 7.8-ohm</td>
</tr>
<tr>
<td>1</td>
<td>U4</td>
<td>TLV3492</td>
<td>Ti</td>
<td>TLV3492</td>
<td>IC, Dual Low Power Comparators</td>
</tr>
<tr>
<td>2</td>
<td>R9</td>
<td>1M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 1M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td></td>
<td>R10</td>
<td>1M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 1M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R5</td>
<td>2M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 2M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>2</td>
<td>R3</td>
<td>5M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 5M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td></td>
<td>R4</td>
<td>5M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 5M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R6</td>
<td>97.6K</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 97.6K-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R8</td>
<td>324K</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 324K-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R7</td>
<td>619K</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 619K-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>U6</td>
<td>TPS71501</td>
<td>Ti</td>
<td>TPS71501DCKR</td>
<td>IC, Regulator, LDO, Micropower, 3.2µA @ 50mA</td>
</tr>
<tr>
<td>1</td>
<td>M5</td>
<td>BSS138</td>
<td>Fairchild</td>
<td>BSS138</td>
<td>MOSFET, Nch, 50V, 200-mA, 3.5 ohms</td>
</tr>
<tr>
<td>Count</td>
<td>RefDes</td>
<td>Value</td>
<td>Mfr</td>
<td>Part Number</td>
<td>Description</td>
</tr>
<tr>
<td>-------</td>
<td>--------</td>
<td>--------</td>
<td>---------</td>
<td>-------------</td>
<td>-----------------------------------------------------------------------------</td>
</tr>
<tr>
<td>6</td>
<td>C21</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C22</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C4</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C5</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C11</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C31</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C1</td>
<td>220uF</td>
<td>Panasonic EEVFK1E221P</td>
<td>Capacitor, Aluminum, SM, 220-uF, 25-V, 150-milliohms (FK series)</td>
<td></td>
</tr>
<tr>
<td></td>
<td>C2</td>
<td>1000pF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1000-pF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C3</td>
<td>100uF</td>
<td>Panasonic EEVK0J101P</td>
<td>Capacitor, Aluminum, 100-uF, 6.3-V, 20%, FK Series</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>D1</td>
<td>BAT54</td>
<td>Fairchild BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>D2</td>
<td>BAT54</td>
<td>Fairchild BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>D3</td>
<td>BAT54</td>
<td>Fairchild BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>D4</td>
<td>BAT54</td>
<td>Fairchild BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>D5</td>
<td>BAT54</td>
<td>Fairchild BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>L2</td>
<td>4.7mH</td>
<td>Coilcraft DW3316-475</td>
<td>Coupled Inductor, SMT, 4.7-mH, 160-mA, 26-ohms</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>U4</td>
<td>TLV3492</td>
<td>TI TLV3492</td>
<td>IC, Dual Low Power Comparators</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>R9</td>
<td>1M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 1M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R10</td>
<td>1M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 1M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>2</td>
<td>R5</td>
<td>2M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 2M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>2</td>
<td>R3</td>
<td>5M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 5M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R4</td>
<td>5M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 5M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R6</td>
<td>118K</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 118K-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R8</td>
<td>357K</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 357K-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R7</td>
<td>649K</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 649K-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>U6</td>
<td>TPS71501</td>
<td>TI TPS71501DCKR</td>
<td>IC, Regulator, LDO, Micropower, 3.2µA @ 50mA.</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>M5</td>
<td>BSS138</td>
<td>Fairchild BSS138</td>
<td>MOSFET, Nch, 50V, 200-mA, 3.5 ohms</td>
<td></td>
</tr>
<tr>
<td>Count</td>
<td>RefDes</td>
<td>Value</td>
<td>Mfr</td>
<td>Part Number</td>
<td>Description</td>
</tr>
<tr>
<td>-------</td>
<td>--------</td>
<td>-------</td>
<td>---------------</td>
<td>---------------</td>
<td>--------------------------------------------------</td>
</tr>
<tr>
<td>6</td>
<td>C21</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C22</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C4</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C5</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C11</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td></td>
<td>C31</td>
<td>1uF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Ceramic, 1-uF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td>1</td>
<td>C2</td>
<td>1000pF</td>
<td>Std</td>
<td>Std</td>
<td>Capacitor, Aluminum, 1000-pF, 25-V, X7R, 10%</td>
</tr>
<tr>
<td>1</td>
<td>C3</td>
<td>100uF</td>
<td>Panasonic</td>
<td>EEVFK0J101P</td>
<td>Capacitor, Aluminum, 100-uF, 6.3-V, 20%, FK Series</td>
</tr>
<tr>
<td></td>
<td>C1</td>
<td>220uF</td>
<td>Panasonic</td>
<td>EEVFK1E221P</td>
<td>Capacitor, Aluminum, SM, 220-uF, 25-V, 150-milliohms (FK series)</td>
</tr>
<tr>
<td>5</td>
<td>D1</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td></td>
<td>D2</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td></td>
<td>D3</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td></td>
<td>D4</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td></td>
<td>D5</td>
<td>BAT54</td>
<td>Fairchild</td>
<td>BAT54</td>
<td>Diode, Schottky, 200-mA, 30-V</td>
</tr>
<tr>
<td>1</td>
<td>U4</td>
<td>TLV3492</td>
<td>TI</td>
<td>TLV3492</td>
<td>IC, Dual Low Power Comparators</td>
</tr>
<tr>
<td>2</td>
<td>R9</td>
<td>1M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 1M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td></td>
<td>R10</td>
<td>1M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 1M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R5</td>
<td>2M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 2M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>2</td>
<td>R3</td>
<td>5M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 5M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td></td>
<td>R4</td>
<td>5M</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 5M-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R6</td>
<td>243K</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 243K-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R8</td>
<td>340K</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 340K-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>R7</td>
<td>604K</td>
<td>Std</td>
<td>Std</td>
<td>Resistor, Chip, 604K-Ohms, 1/10-W, 1%</td>
</tr>
<tr>
<td>1</td>
<td>U6</td>
<td>TPS71501</td>
<td>TI</td>
<td>TPS71501DCKR</td>
<td>IC, Regulator, LDO, Micropower, 3.2µA @ 50mA.</td>
</tr>
<tr>
<td>1</td>
<td>M5</td>
<td>BSS138</td>
<td>Fairchild</td>
<td>BSS138</td>
<td>MOSFET, Nch, 50V, 200-mA, 3.5 ohms</td>
</tr>
<tr>
<td>1</td>
<td>L1</td>
<td>14.4mH</td>
<td>Robert Taylor</td>
<td>NA</td>
<td>Coupled Inductor, Through Hole, 14.4-mH, 100-mA, 0.98-ohms</td>
</tr>
</tbody>
</table>
LIST OF REFERENCES


[45] ASTM-E1050-90, "Impedance and Absorption of Acoustical Materials Using a Tube, Two Microphones, and a Digital Frequency Analysis System."
BIOGRAPHICAL SKETCH

Robert J. Taylor was born June 13th, 1980, in Albion, Michigan. He moved to Jacksonville, Florida, when he was the age of 3. He attended N.B. Forrest High School and graduated as the valedictorian in 1998. In August of 1998, Robert began his studies at the University of Florida. He graduated with honors in May of 2002, with a Bachelor of Science in Electrical Engineering degree. He is currently working towards a Master of Science degree in electrical and computer engineering. After completing his degree, he will work as a power supply design engineer for Texas Instruments.