Design and Implementation of a Novel Control System for Four Quadrant Operation of a Two-Phase Switched Reluctance Motor

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In

Electrical Engineering

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Dr. Krishnan Ramu, Chair

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In the emergence of switched reluctance motors to the commercial market, two-phase motors have received relatively little attention. Higher power and industrial applications have focused on the use of three and occasionally four phase machines, while low cost applications demanding only modest performance have largely been the domain of single phase machines. By contrast, while two phase systems have been the subject of occasional studies, they have not been widely applied.

Two phase systems represent a compromise between the higher cost but higher performance three phase machines, and the lower cost but lower performance single phase systems. They do not suffer from the same magnitude of peak to peak torque ripple that single phase machines experience due to their wide zero torque arcs. Yet two phase systems keep a relatively low component count in their power-converter designs. The primary drawback to two phase motors is the difficulty of torque production at startup speeds. Although sizably reduced from single phase machines, the zero torque regions in two phase machines can still result in rotor lock unless steps are taken to circumvent them. These steps can include measures such as: placement of permanent magnets or other means to ensure the rotor is positioned outside of these zero torque regions when at rest, mechanically spinning the motor before energizing the phase windings on startup, shaping of the rotor or stator poles to extend the positive torque regions of each phase, or use of the machines mutual inductance with both phases energized to produce enough torque to initiate motion.

This project is intended to develop a variable speed controller for a 4:6 two-phase switched reluctance motor. The motor is to operate in all four quadrants, and is to demonstrate self starting capability. The controller is also supposed to produce signals needed to operate the motor with multiple converter designs. Two different converter designs will be built and tested with the converter. One makes use of a single switch and two diodes per phase, the other has one switch and one diode per phase plus a common switch and common diode shared by all phases.

There are many possible applications of the system being developed in this project. Any application needing four quadrant operation while still being constrained by low cost requirements would be ideal. Some examples include washing machines, power tools, and low power industrial applications.
Acknowledgements

Although it is my name on the cover of this thesis, there is no way in which I could have hoped to have completed either my graduate studies or this thesis project by myself. I would like to offer my sincerest thanks to my advisor, Dr. Krishnan Ramu, for the myriad of opportunities he gave me to work on a wide range of subjects and applications, and for his continuous presence behind the scenes providing me with support and assistance whenever I needed it. I would also like to thank Dr. Jeffrey Reed and Dr. Douglas Lindner for their willingness to act as members of my defense committee.

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<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>gain</td>
</tr>
<tr>
<td>C</td>
<td>capacitance (F)</td>
</tr>
<tr>
<td>$f$</td>
<td>frequency (Hz)</td>
</tr>
<tr>
<td>$f_H$</td>
<td>cutoff frequency of a low-pass filter (Hz)</td>
</tr>
<tr>
<td>$i$</td>
<td>stator current (A)</td>
</tr>
<tr>
<td>I</td>
<td>current (A)</td>
</tr>
<tr>
<td>$I_p$</td>
<td>peak stator current (A)</td>
</tr>
<tr>
<td>$\frac{\partial L}{\partial \theta}$</td>
<td>rate of change of inductance with respect to rotor position</td>
</tr>
<tr>
<td>L</td>
<td>inductance (H)</td>
</tr>
<tr>
<td>$L_a$</td>
<td>inductance when rotor is in the fully aligned position (H)</td>
</tr>
<tr>
<td>$L_u$</td>
<td>Inductance when rotor is in the fully unaligned position (H)</td>
</tr>
<tr>
<td>R</td>
<td>resistance (Ω)</td>
</tr>
<tr>
<td>$R_s$</td>
<td>stator coil resistance (Ω)</td>
</tr>
<tr>
<td>$t_c$</td>
<td>time constant (sec)</td>
</tr>
<tr>
<td>$T_{av}$</td>
<td>average torque (N)</td>
</tr>
<tr>
<td>$T_f$</td>
<td>approximated time for current to fall to zero from its peak value (sec)</td>
</tr>
<tr>
<td>$T_r$</td>
<td>approximated time for current to rise from zero to the peak value assuming 100% duty cycle (sec)</td>
</tr>
<tr>
<td>U</td>
<td>energy (J)</td>
</tr>
<tr>
<td>V</td>
<td>voltage (V)</td>
</tr>
<tr>
<td>$V_{OH}$</td>
<td>high output voltage of an op amp used in a Schmitt-trigger (V)</td>
</tr>
<tr>
<td>$V_{OL}$</td>
<td>low output voltage of an op amp used in a Schmitt-trigger (V)</td>
</tr>
<tr>
<td>$V_{TH}$</td>
<td>high voltage trigger level in a Schmitt-trigger (V)</td>
</tr>
<tr>
<td>$V_{TL}$</td>
<td>low voltage trigger level in a Schmitt-trigger (V)</td>
</tr>
<tr>
<td>$\tau_a$</td>
<td>time constant for the aligned rotor position (sec)</td>
</tr>
<tr>
<td>$\tau_u$</td>
<td>time constant for the unaligned rotor position (sec)</td>
</tr>
<tr>
<td>Symbol</td>
<td>Description</td>
</tr>
<tr>
<td>--------</td>
<td>-------------</td>
</tr>
<tr>
<td>$\theta_a$</td>
<td>fully aligned rotor position (deg or radians)</td>
</tr>
<tr>
<td>$\theta_{adv}$</td>
<td>angle at which stator coils are fired prior to rotor reaching $\theta_i$ so that it will reach $I_p$ at $\theta_i$ (deg or radians)</td>
</tr>
<tr>
<td>$\theta_{comm}$</td>
<td>angle prior to $\theta_f$ at which stator coils begin de-energizing to reach zero current at or before $\theta_f$ (deg or radians)</td>
</tr>
<tr>
<td>$\theta_f$</td>
<td>position at which rotor pole and stator pole completely overlap (deg or radians)</td>
</tr>
<tr>
<td>$\theta_i$</td>
<td>position at which rotor pole and stator pole begin to overlap (deg or radians)</td>
</tr>
<tr>
<td>$\theta_r$</td>
<td>rotor pole arc (deg or radians)</td>
</tr>
<tr>
<td>$\theta_s$</td>
<td>stator pole arc (deg or radians)</td>
</tr>
<tr>
<td>$\theta_u$</td>
<td>fully unaligned rotor position (deg or radians)</td>
</tr>
<tr>
<td>$\omega_r$</td>
<td>rotor speed (rpm or radians/sec)</td>
</tr>
</tbody>
</table>

**List of Special Terminology**

*Firing arc* the angular range over which a given phase is actively switching to conduct current. Typically $[(\theta_i - \theta_{adv})$ to $(\theta_f - \theta_{comm})]$ or $[(\theta_i - \theta_{adv})$ to $(\theta_u - \theta_{comm})]$. 
Chapter 1 – Introduction

1.1 History and Technical Challenges

Switched Reluctance Motors (SRMs) are a relatively new technology in the commercial arena. Despite this, the concept behind the switched reluctance machine has been in existence since 1838 when it was introduced as an “electric engine.” The idea of that time was that by energizing and de-energizing a series of electromagnets positioned around the stator to generate torque on a steel or iron rotor, rotation could be induced. Unfortunately there were several challenges presented by the design which could not be overcome by the technologies of that era.

As with the “electric engine,” Switched Reluctance Motors operate by firing electromagnetic coils positioned around their stator poles to generate torque and “pull” the passive rotor poles into alignment with the stator poles. Once aligned, the currently active stator pole or poles should be de-magnetized and the next set in position to generate torque in the direction of rotation energized. There are several challenges in this to generate optimal performance.

First and most significantly, it is desirable to avoid having stator poles magnetized when the rotor poles are positioned such that the magnetic pull will generate torque in the direction opposite of that which is desired. This “negative torque” as it is commonly referred to obviously causes significant losses and inefficiency. Furthermore, it is a generator of acoustic noise as well as a major source of torque ripple and mechanical stresses placed on the motor. As the rise and fall of current in the electromagnetic coils is restricted by inductance it is necessary to begin excitation and commutation before a specific rotor pole has reached the limits of the arc which the operator wishes to have a given stator pole acting on it. This is further complicated by the non-linearity of inductance in switched reluctance motors. As inductance in SRMs varies with both position and with electrical current one set of advance and commutation angles might not
be sufficient to prevent negative torque for all operating speeds. Coupled with the need to open and close current paths through the various stator coils is the desire to be able to recover and use the energy in the coils being de-energized. Other problems inherent in switched reluctance motors include: the need to know rotor position and electrical current present in each stator coil, and difficulties in starting from at rest conditions when all rotor poles are fully aligned or fully unaligned with the stator poles.

While SRMs have begun to gain acceptance and see use in a wide range of applications, the two-phase system has lagged behind single-phase designs in low cost home appliance applications, and behind three or four phase drive systems for higher performance industrial applications. Yet with smaller zero torque regions than found in single phase systems, and lower converter part counts than needed by three and four phase systems, two-phase designs represent an attractive option for mid-range applications. Their one main drawback is that they possess much the same startup difficulties as single phase SRMs because while smaller than those found in single phase systems, they do still have zero torque regions.

1.2 Objectives

The objectives of this project are the development of a controller for a 4:6 (four stator pole, six rotor pole) two phase switched reluctance motor and analysis and comparison of two separate SRM power converters.

The SRM controller should enable the motor to function in all four quadrants of operation: forward motoring, forward regeneration, reverse motoring, and reverse regeneration. It will make use of a three signal encoder for position and speed feedback information. Electrical current will be monitored through LEM current transducers embedded in each power converter. User commands for the controller will be to increase the motor’s set speed, decrease the set speed, and to turn power to the motor on or off. These commands will be generated and transmitted to the controller by way of a wireless
device. Finally the controller is to function with both of the two different SRM power converters. The power converter designs will be: an N+1 switch (N = number of phases of the motor) converter invented by R. Krishnan and P. Materu to be used as a baseline, and an N switch converter invented by W. A. Harris.

Verification of the controller functions will be carried out by comparison of current in the motor stator coil with rotor position and rotor speed. Performance of the two converter designs will be evaluated by measuring and analyzing the electrical current wave forms in the motor stator windings. Performance of the controller and the two separate SRM power converter designs will be discussed and suggestions for future investigation made.

1.3 Thesis Outline

The rest of the thesis paper will be organized as follows. Chapter 2 talks about the SRM motor used in the course of this project. Chapter 3 discusses the two differing power converters investigated in the project. Chapter 4 discusses design and implementation of the controller. Chapter 5 focuses on experimental results and discusses the outcome and performance of the various components being developed and investigated during the project. Chapter 6 wraps up the thesis by stating the conclusions reached as a result of the experimentation recorded in chapter 5 and by suggesting avenues of future research and development.
Chapter 2 – Switched Reluctance Motor

In order to design and implement a working controller for the system, it is necessary to understand the design of the motor being controlled. Firing the stator coils over the proper angles to “pull” the rotor in the desired direction requires knowledge of the rotor and stator pole angles, and the stator coil winding configuration. Optimizing the firing angles further to reduce torque ripple and acoustic noise caused by negative torque requires the calculation of electrical current rise and fall times in the stator coils.

This chapter discusses of this necessary information. In section 2.1 it discusses the mechanical design of the motor along with operational limits used in later calculations. In section 2.2 the mechanical design is related to regions of positive and negative torque and rising and falling inductance, and actual inductance curves from experimental data are set up. This in turn is used to set up initial firing arcs, and to calculate electrical current rise and fall times for the stator coils which will be used to generate optimized firing arcs in later sections. Section 2.3 discusses the encoder used to provide position feedback information to the controller.
2.1 Mechanical Design

The motor used in this project was a 4/6 two phase SRM drive. The machine was designed to operate on a 240 Vdc supply at up to 8 Amps in its stator coil windings. Its maximum rated speed is 12,000 rpm. It also makes use of stator pole shaping to generate self starting capability through air gap tapering.

The mechanical dimensions of the rotor and stator laminations are shown in figure 2.2. From these calculations can be done to determine the basic angular values needed for setting up the controller and the arcs over which the stator coils will be energized (to be referred to henceforth as the firing arc).

These angular values are:

- Rotor Pole Arc \( (\theta_r) = 30 \) degrees
- Rotor Pole-to-Pole Angle = 60 degrees
- Stator Pole Arc \( (\theta_s) = 17.5 \) degrees
- Stator Pole-to-Pole Angle = 90 degrees
Additional important values include the maximum rated current of 8 Amps and rated speed of 12,000 rpm.

2.2 Electro-magnetic Properties

Average torque for a given torque value in SRM drives is simply:

\[
T_{av} = \frac{1}{2} i^2 \frac{\partial L}{\partial \theta}
\]

(2.1)

However, inductance itself is a non-linear function of both current and position. Still, it can be approximated for a given current as being flat in positions at which the stator and rotor poles completely overlap one another and when there is no overlap between rotor and stator poles whatsoever [1]. As the rotor and stator pole arcs for this motor are known, it is possible to develop rough approximations of the inductance and torque curves used for generating the preliminary firing arcs of each phase. The approximated inductance and torque curves are shown in figure 2.3.

Figure 2.3: Approximate inductance and torque with unaligned rotor position and aligned rotor position shown.
Setting the fully unaligned position ($\theta_u$) as 0 degrees, we know the fully aligned position ($\theta_a$) will be one half the angle between rotor poles. For a 4/6 machine this equates to 30 degrees. The start of the firing arc in general equates to the rotor position at which the rotor pole begins to overlap the stator pole, and the end of the firing arc generally equates to the point at which the rotor pole completely overlaps (or is overlapped by) the stator pole.

In the case of this machine, initial overlapping position and final overlapping position are:

\begin{align*}
\theta_i &= \theta_a - \frac{1}{2} \theta_r - \frac{1}{2} \theta_s \\
&= 30 - \frac{1}{2} (30) - \frac{1}{2} (17.5) = 6.275 \text{ degrees} \\
\theta_f &= \theta_i + \text{smaller of($\theta_s$ or $\theta_r$)} \\
&= 6.275 + 17.5 = 23.775 \text{ degrees}
\end{align*}

However, it should be noted that the stator poles as shown in figure 2.1 are shaped such that the air gap from one side to the other is tapered. This will have the effect of shifting the maximum inductance point and altering the inductance curve near the maximum. The goal of this pole shaping is to enable the motor to self start regardless of position by both shifting and limiting the overlap of zero torque regions. Because of this start up capability, start speed firing arcs needed to be developed which covered all 360 degrees of mechanical rotation. To develop these firing arcs exact knowledge of the zero torque (maximum and minimum inductance points) were needed for both phases. Additionally, knowledge of maximum and minimum inductance values enable SRM stator coil current rise and fall times to be estimated. This in turn can be used to improve high speed operation by setting advance and commutation angles to the stator coil firing arcs.

In order to determine this information, the motor was set in a test fixture where the rotor position could be fixed at known angular positions and inductance measured as outlined
in [2]. These measurements were done by applying a known current level (1 ampere) to one stator coil at a time, then cutting off the current abruptly and measuring the time needed for the current through the coil to fall to zero. This test was performed at positions approximately every 2.8 degrees apart from zero to 60 degrees. Each position was tested five times to attain a reasonably certain average value. The results of these tests are given on the table in table 2.1 and graphed in figure 2.4.

<table>
<thead>
<tr>
<th>position</th>
<th>Current</th>
<th>InductA</th>
<th>InductB</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1</td>
<td>6.158</td>
<td>6.084</td>
</tr>
<tr>
<td>2.8125</td>
<td>1</td>
<td>6.582</td>
<td>5.408</td>
</tr>
<tr>
<td>5.625</td>
<td>1</td>
<td>6.98</td>
<td>4.956</td>
</tr>
<tr>
<td>8.4375</td>
<td>1</td>
<td>7.436</td>
<td>4.514</td>
</tr>
<tr>
<td>11.25</td>
<td>1</td>
<td>7.938</td>
<td>4.574</td>
</tr>
<tr>
<td>14.0625</td>
<td>1</td>
<td>8.04</td>
<td>4.1</td>
</tr>
<tr>
<td>16.875</td>
<td>1</td>
<td>8.048</td>
<td>4.228</td>
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<td>19.6875</td>
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<td>22.5</td>
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<td>5.236</td>
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<td>25.3125</td>
<td>1</td>
<td>6.824</td>
<td>5.644</td>
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<td>28.125</td>
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<td>6.278</td>
<td>5.746</td>
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<td>30.9375</td>
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<td>8.094</td>
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<td>47.8125</td>
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<td>6.496</td>
</tr>
<tr>
<td>60</td>
<td>1</td>
<td>6.158</td>
<td>6.084</td>
</tr>
</tbody>
</table>

Table 2.1: Table of average inductance values at 1 ampere from zero to 60 degrees.
As can be seen, the stator pole shaping has indeed had the desired affect, shifting the minimum and maximum inductance values equating to the zero torque positions of each stator pole off from one another. From this data, start speed firing arcs were developed for all four quadrants as shown in table 2.2.

<table>
<thead>
<tr>
<th>Angles</th>
<th>Forward Motoring</th>
<th>Forward Regeneration</th>
<th>Reverse Motoring</th>
<th>Reverse Regeneration</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 – 15 degrees</td>
<td>A</td>
<td>B</td>
<td>B</td>
<td>A</td>
</tr>
<tr>
<td>15 – 16 degrees</td>
<td>A</td>
<td>A</td>
<td>A</td>
<td>A</td>
</tr>
<tr>
<td>16 - 46 degrees</td>
<td>B</td>
<td>A</td>
<td>A</td>
<td>B</td>
</tr>
<tr>
<td>46 – 47 degrees</td>
<td>A</td>
<td>B</td>
<td>B</td>
<td>A</td>
</tr>
<tr>
<td>47 – 60 degrees</td>
<td>A</td>
<td>B</td>
<td>B</td>
<td>A</td>
</tr>
</tbody>
</table>

Table 2.2: Phase firing arcs at machine startup.
It is also worth noting that the minimum inductance value for either stator coil is approximately 4 mH and the maximum 8 mH. These values correspond to the unaligned inductance ($L_u$) and aligned inductance ($L_a$) respectively.

Given the minimum and maximum inductance values, a measured resistance across the stator coils of approximately 1.5 $\Omega$, a peak operating current of 8 amperes, and DC-link voltage that can be set at up to 200 Volts an approximated current rise and fall time can be calculated as follows [1]:

\[
T_r = \tau_u \ln \left( \frac{1}{1 - \frac{R_s I_p}{V_{dc}}} \right), \text{s}
\]

\[
T_f = \tau_a \ln \left( 1 + \frac{R_s I_p}{V_{dc}} \right), \text{s}
\]

where $\tau_u = \frac{L_u}{R_s}, \text{s}$

\[
\tau_a = \frac{L_a}{R_s}, \text{s}
\]

For this system, $\tau_u = 0.00267$ seconds and $\tau_a = 0.00533$ seconds. Then, operating at a full 200 $V_{dc}$ link voltage the rise time ($T_r$) will be 1.65 x $10^{-4}$ seconds, and the fall time ($T_f$) will be 3.11 x $10^{-4}$ seconds. (At 60 V, the maximum voltage that is used for testing in the CRTS lab, the rise and fall times will be 5.96 x $10^{-4}$ seconds and 9.72 x $10^{-4}$ seconds respectively.)

In order to achieve optimal performance of the SRM, stator coil current levels should be at the commanded value as the rotor pole being acted on moves into the firing arc, and should be at zero as the rotor pole being acted on leaves the firing arc. In order to achieve this at speeds greater than zero it is necessary to begin the process of exciting and commutating the stator coil before $\theta_i$ and $\theta_f$ are actually reached. The angle between the point at which excitation is started and $\theta_i$ is referred to as the advance angle ($\theta_{adv}$).
Similarly the angle between the start of commutation and $\theta_f$ is referred to as the commutation angle ($\theta_{comm}$). The advance and commutation angles necessary to achieve full excitation at $\theta_i$ and full commutation at $\theta_f$ can be found using the following equations:

\begin{align*}
\theta_{adv} &= T_r \omega_r \\
\theta_{comm} &= T_f \omega_r
\end{align*}

where $\omega_r$ is the rotor’s angular velocity.

With the rise and fall times for various DC link voltages known, $\theta_{adv}$ and $\theta_{comm}$ can be calculated for the planned speed ranges. Typically the speed $\omega_r$ is just the upper limit for the speed range.

For this project, three speed ranges are planned. The first is for startup and will be from zero to 100 rpm. No advance or commutation will be used in this range and the firing arcs will be extended all the way from the minimum inductance to the maximum inductance of the phase in order to facilitate startup regardless of position. Mid-speed will be from 100 rpm to 1500 rpm. High speed operations will be for anything over 1500 rpm, through the firing arcs will be calculated for 2500 rpm. The calculated advance and commutation angles for these speeds, given the maximum lab testing voltage of 60 V, are given in table 2.3.

<table>
<thead>
<tr>
<th>Motor Speed</th>
<th>Advance Angle ($\theta_{adv}$)</th>
<th>Commutation Angle ($\theta_{comm}$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 – 100 rpm</td>
<td>0 degrees</td>
<td>0 degrees</td>
</tr>
<tr>
<td>100 rpm – 1500 rpm</td>
<td>5.36 degrees</td>
<td>8.75 degrees</td>
</tr>
<tr>
<td>1500 rpm +</td>
<td>8.94 degrees</td>
<td>14.58 degrees</td>
</tr>
</tbody>
</table>

Table 2.3: Advance and commutation angles calculated for low, mid, and high speed operation.
2.3 Interface

There are three primary interfaces with the SRM drive being used in this project. The current supply and return points with the stator coil designated as phase A, the current supply and return points with the stator coil designated as phase B, and the encoder mounted to one of the end bells and the rotor which provides digital signals used to feedback position and speed information to the controller. Current and current feedback information in this project is handled on the SRM power converter boards and in the current control loops and thus will be addressed in chapter 3 and chapter 4. However as the encoder was mounted as part of the motor, it will be discussed here.

The encoder used for providing position feedback was a DH48 model produced by Daido Corporation. It provided three channel digital feedback. Two of the channels each provide 1024 positive signals per revolution. These two channels will be referred to as channel A and channel B. The third channel, channel Z, has only one positive tick per revolution. The encoder operated with a $5 \text{ V}_{dc}$ supply and $0 \text{ V}_{dc}$ return. The high voltage level in the digital signal was approximately $3.5 \text{ V}_{dc}$. The pattern of signals provided by the encoder, assuming a constant motor speed is shown in figure 2.5.

![Figure 2.5: DH48 encoder signals.](image-url)
Chapter 3 – Power Converters

The basic purpose of the power converter in conjunction with SRM machines is electronic control of current flow through the independent stator coil windings (also referred to as the phase windings) in the motor. Unlike in induction machines, torque production in SRM drives is independent of current polarity, and hence it is possible to sufficiently control phase winding current with only a single switch per phase. However, limiting the number of switches may restrict the converter in its modes of operation and in this manner limit motor performance. This has led to an abundance of converter designs for use with SRM machines, each with a different topology intended to achieve different modes of operation and differing levels of performance.

Two power converter designs were examined in the course of this project. One, a design invented by R. Krishnan and P. Materu [3] makes use of one switch per phase with an additional shared switch common to all phases. The other, a design by W.A. Harris [4] uses a single switch per phase. For ease of discussion the Krishnan and Materu design will be referred to as the N+1 switch topology converter, and the Harris design will be referred to as the N switch topology converter throughout the rest of this paper, where N denotes the number of phases in the SRM machine being driven.

This chapter examines these two converters. Section 3.1 is devoted to the N+1 switch topology converter. Section 3.2 focuses on the N switch topology converter. Each of these sections is subdivided further into subsections. The first subsection in both 3.1 and 3.2 looks at the converter designs and modes of operation. The second subsection examines how the designs were actually implemented, from component selection to controller interface. The third subsection discusses converter operation and performance of the given SRM converter design observed during the course of the project.
3.1 N+1 Switch Topology Converter

3.1.1 Design

The N+1 Switch Topology Converter design is intended to minimize the total number of components while achieving a fairly wide range of modes of operation allowing normal conduction, free wheeling, and commutation. Intended for low cost, low power applications this drive has the following advantages and disadvantages [3]:

<table>
<thead>
<tr>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>Low part count</td>
<td>One extra switch (not truly single</td>
</tr>
<tr>
<td>Easy to implement</td>
<td>switch per phase)</td>
</tr>
<tr>
<td>Fast regeneration</td>
<td>Limited firing arc</td>
</tr>
<tr>
<td>Energy efficient</td>
<td>Common switch emitter voltage floats</td>
</tr>
</tbody>
</table>

Table 3.1: Advantages and disadvantages of the N+1 switch topology converter.

For a motors possessing a relatively low number of phases where only one phase should be conducting at a time it offers good performance at a relatively low cost.

![Figure 3.1: N+1 Switch Topology Converter.](image)
The N+1 switch converter has four possible modes of operation. One for normal conduction, two free-wheeling, and one for commutation.

The first mode, shown in figure 3.2 represents normal conduction. In this mode of operation, the shared common switch is conducting from the DC-source and the energy recovery capacitor into the active phase winding (stator coil) and then through the phase specific switch associated with the active phase winding back out to the DC-return.

![Figure 3.2: Mode 1 - Normal conduction.](image)

The second mode of operation, shown in figure 3.3 is one of the two freewheeling modes. In this mode, electrical current circulates through the active phase winding, through the phase specific diode, and then back to the phase winding through the shared common switch. The phase specific switch is not conducting in this mode.
The third mode of operation shown in figure 3.4 is also a freewheeling mode. In this mode, electrical current circulates through the active phase winding, through the phase specific switch which is conducting, then back to the phase winding through the common diode. The common switch is not conducting in this mode.

![Figure 3.3: Mode 2 - Freewheeling.](image1)

The fourth mode of operation provides commutation and is shown in figure 3.5. In this mode, electrical current passing through the phase winding (zero voltage differential), is conducted through the phase specific diode, then the energy recovery capacitor (where the voltage differential is greater than or equal to the DC-link voltage), then is returned to

![Figure 3.4: Mode 3 - Freewheeling.](image2)
the phase winding through the common diode. None of the switches in the converter conduct during this mode.

![Figure 3.5: Mode 4 - Commutation and recovery of electrical energy.](image)

### 3.1.2 Implementation

There were three types of issues facing implementation of the N+1 switch converter: mechanical implementation of the design itself, control signal generation, and controller interfacing. Mechanical implementation is concerned with sizing and selection of the
components shown in the schematic in figure 3.1, and addition of necessary safety features. Control signal generation looks at selection of the modes of operation to use, selection of how to activate those modes. Controller interfacing is primarily concerned with interfaces between the converter and the controller – in this case, selection and placement of the current feedback sensors and conditioning of the signals sent from the controller to the converter so they have the desired effect.

3.1.2.1 Mechanical Implementation

The DC voltage supply was created by feeding a variable AC voltage supply (8 amps 120 V<sub>ac</sub> max) into an H-bridge rectifier. Accordingly it was assumed the dc-link supply would provide up to 8 amps at 120V<sub>dc</sub>. The energy recovery capacitor selected was a 1000 µF 250V capacitor. This large capacitance was selected to avoid any significant voltage fluctuation during commutation. For switching operations, IRG4PC40U IGBT’s from International Rectifier were selected for both common and phase specific switches. These switches rated for 40 amps at 600V and are optimized for switching frequencies from 8 to 40 kHz. The diodes selected for both common and phase specific switching diodes were IR40EPF04 diodes from International Rectifier. These fast, soft switching diodes rated for up to 40 amps at 400V, with a 180ns reverse recovery time.

The safety features added were: surge absorbers, a capacitor drain resistor, and an on/off indicator LED. The surge absorbers were rated for 240 V and 600 amps. They were placed at the AC inputs to the H-bridge rectifier, and across the capacitor. The capacitor drain is a 25 kΩ 10 W power resistor connected in parallel with the energy recovery diode to allow the capacitor to drain when the system is shut off. The on/off indicator is a simple red LED connected in series with the capacitor drain resistor. It remains on both when the system is on, and as long as a significant charge remains in the capacitor.
3.1.2.2 Control Signal Generation

For the N+1 switch converter design, the phases are not entirely independent. When the current in one phase is being driven (mode 1), it is not possible to commutate (mode 4) as the driving mode requires the common switch (T_c) to be conducting, while commutation cannot occur with the common switch conducting.

In order to achieve self starting it might be necessary to fire the phase windings even at non-optimal low torque positions. Consequently the controller in this project was designed with a start-up speed range without any gaps between the firing arcs of the two phases. As a result, there was an advantage to allowing commutation of an outgoing phase to occur during the freewheeling periods of the newly active phase. This requires freewheeling to utilize mode 3 instead of mode 2 of operation. Consequently, the modes of operation for the N+1 switch topology converter design were mode 1, mode 3, and mode 4. Mode 2 was not utilized in this project at all.

In order to achieve these modes of converter operation, the phase specific switches, T_1 and T_2, were used for controlling the firing arc. The shared common switch was used to alternate between driving and freewheeling as called for by the PWM signal generated for the currently active phase.

3.1.2.3 Controller Interfacing

The two controller interfacing issues facing the N+1 switch converter were selection and placement of the current sensors for each phase, and “conditioning” of the gate control signals to the switches so that switching operation would actually take place.

In the N+1 switch topology converter, current sensing and feedback was performed using LA25-NP current transducers manufactured by LEM Inc. The LA25-NP sensor makes use of the Hall effect to generate a measurable amount of output current for each ampere flowing through its primary windings. This small output current is then sent through a
known resistor to ground and the voltage at the resistor input is measured, filtered and amplified as necessary to produce the feedback signal. This is discussed in greater detail in the controller chapter of this thesis.

The current transducers were placed in position 1 above the phase windings as shown in figure 3.7. While position 2 could also have been chosen, A was selected as the inductance of the phase windings helped to smooth out and reduce the impact of transients from switching operations in the phase specific switches.

In order to protect the control circuitry, all three switches were driven using opto-isolating gate drivers to producing a +15 V\textsubscript{dc} gate signal with complete galvanic isolation. Additionally, it was necessary to ensure that this gate signal floated as needed to be 15 V greater than the switches’ emitter pin voltages (the maximum threshold voltage for V\textsubscript{GE} of the IRG4PC40U is +6 V\textsubscript{dc}, +15 V\textsubscript{dc} is recommended). This was achieved by creating an independent power supply for each switches’ gate driver using low power NMD05515 dc-dc converters, tying the output reference to the switch emitter pin voltage. The final design is shown in figure 3.8.
This gate driving circuit achieved the desired galvanic isolation, and voltage float needed to ensure that $V_{GE}$ floated with $V_E$.

### 3.1.3 Operation

The N+1 switch converter performed extremely well over all speed and torque ranges tested during the project. Response time during phase excitation and during phase current commutation was good. Some offset from the command value was observed but this was more a function of the controller gains than converter performance.
Figure 3.9: N+1 switch topology converter response to a current step command.

3.2 N Switch Topology Converter

3.2.1 Design

The N Switch Topology Converter designed by W.A. Harris [4] is intended to be a low cost, high efficiency design. While it is not intended to have free wheeling capability, the premise of the design is that it will make up for this through recovery of the energy in the phase winding electrical currents being commutated into its energy recovery capacitors and use of this energy to drive the subsequent phases. Advantages and disadvantages to this drive are:
<table>
<thead>
<tr>
<th>Advantages</th>
<th>Disadvantages</th>
</tr>
</thead>
<tbody>
<tr>
<td>• Low part count</td>
<td>• Two diodes per phase</td>
</tr>
<tr>
<td>• Single phase per switch</td>
<td>• Dedicated capacitor for every phase</td>
</tr>
<tr>
<td>• No floating emitter voltages</td>
<td>• <strong>Current recirculation</strong> (discussed below)</td>
</tr>
</tbody>
</table>

Table 3.2: Advantages and disadvantages of the N switch topology converter.

While the true single switch per phase nature of this converter design may make it attractive from a low cost standpoint, the current recirculation which was observed in its use limit it to only the least performance concerned applications.

![Figure 3.10: N Switch Topology Converter design.](image)

The two anticipated modes of operation for the N+1 switch topology converter are normal conduction and commutation. However, during the course of experimentation in this project a third mode of operation not discussed in the patent filing was discovered and the second, commutation, was observed to not actually take place. For purposes of discussion, this third mode will be referred to as recirculation.
The first mode of operation, shown in figure 3.11 is normal conduction. In this mode, current is drawn from the active phase’s capacitor (if its voltage exceeds the DC-link voltage) or the DC-link voltage itself, flows through the phase winding, and then is conducted through the active phase switch back to the DC return.

![Figure 3.11: Mode 1 – Normal conduction.](image)

The second mode of operation, shown in figure 3.12 is commutation. During this mode of operation, the switch for the phase winding in question does not conduct. Instead, current travels from the phase winding, through the drain diode to source rail for the next phase where it is dumped into the energy recovery capacitor for that phase. It is blocked from passing through the next phase’s stator coil by the coil inductance. As energy is captured by the capacitor, the voltage across the capacitor rises. This in turn causes the voltage at the outlet of the commutating stator coil to rise, increasing the reverse voltage bias across the coil and in turn increasing the rate of commutation.
Unfortunately, the inductance in the next phase’s stator coil will not block current for more than an instant. As energy recovery takes place, the voltage across the second phase’s energy recovery capacitor (also the voltage at the inlet to the stator coil) will rise, and eventually will exceed the voltage in the source rail of the next phase, forward biasing the second phase’s drain diode. At this point, current will begin flowing through second phase’s stator coil into the source rail of the third phase in series where it begins charging that phase’s energy recovery capacitor. This process is shown in figure 3.13 and is the third mode of operation mentioned for this converter design.
In the case of this project where only two phases existed, the “next phase” of course was the first phase which was still trying to commutate. When this happened the reappearance of this energy would drive the source rail voltage back up, slowing commutation and eventually stopping it entirely at which point it the first phase’s current would begin to rise once more. As the only way in which current was able to drop was through inefficiencies and stator coil resistances the source rail voltages (energy recovery capacitor voltages) were observed to rise and fall in a slowly decaying cyclical manner. Current would likewise decay in what appeared almost sinusoidal manner as can be seen in figure 3.14, never disappearing entirely before being driven back upwards when the arrival of a rotor pole in the phase’s firing arc would result in the switch being activated and mode 1 being re-initiated.

![Figure 3.14: N switch topology converter displaying current recirculation.](image)

Scale: 1 Amp / div

Even in motors with higher numbers of phases where the re-circulating energy might not reach the outgoing phase before full commutation had taken place. It is likely this re-
circulating energy will cause negative torque, acoustic noise, heat generation, and other inefficiencies regardless of the motor being driven.

3.2.2 Implementation

![Image](image.png)

Figure 3.15: N switch topology converter implementation.

3.2.2.1 Mechanical Implementation

Mechanical implementation of the N Switch Topology Converter followed similar paths to that of the N+1 Switch Topology Converter. The DC source used was exactly the same (a 120 Vac 8 amp variable supply driving a H-bridge rectifier). Identical IRG4PC40U IGBT’s from International Rectifier were used for switching and likewise IR40EPF04 diodes were also used. The only real variation in selection of major components was in the sizing of the energy recovery capacitors.

In order to size the capacitor, the equations for energy in a capacitor and energy in an inductor were used.

\[
(3.1) \quad U = \frac{1}{2}CV^2 \quad \text{energy in a capacitor}
\]

\[
(3.2) \quad U = \frac{1}{2}LI^2 \quad \text{energy in an inductor}
\]
Taking these equations the change in energy in the inductor as the current was
commutated from its peak value to zero was assumed to be equal to the gain in energy of
the capacitor.

\[ \Delta U_{\text{capacitor}} = \Delta U_{\text{inductor}} \]

\[ \frac{1}{2} C \Delta V^2 = \frac{1}{2} L \Delta I^2 \]

Then, by plugging in assumed values for the current (assumed to be the maximum current
rated for the machine i.e. 8 A), inductance (assumed to be the peak measured inductance
(8 mH), and readily available capacitor sizes resulting values for the change in voltage
could be found. It was desirable to have a fairly large voltage increase to promote
commutation. At the same time it was necessary to keep the change in voltage low
enough that there would not be a risk of exceeding the voltage ratings of our components.

By using this method it was determined that a 33 \( \mu \text{F} \) capacitor would experience an
voltage rise of 124 Vdc. Assuming a peak operating voltage of 120 V this would exceed
240 V on the source rail of the incoming phase and was deemed to be too high.
However, connecting two 33 \( \mu \text{F} \) capacitors in parallel would create a 66 \( \mu \text{F} \) capacitance
which would only rise 88 V, resulting in an acceptable 208 Vdc on the incoming phase’s
source rail while still providing a suitably large commutating voltage.

Like the N+1 switch design, this converter was also implemented with surge absorbers,
capacitor drain resistors, and LED’s to indicate the presence of charge across the
capacitors. Essentially the same components were selected as outlined in section 3.1.2.1.
3.2.2.2 Control Signal Generation

Since there was no shared switch, and only two controllable modes of operation for each phase, control signal generation was done slightly differently for the N switch converter from the N+1 converter. Instead of having the PWM signal for each phase sent to one switch and an active phase selection signal being sent to another switch, there was just the PWM signal for each phase (the supposedly inactive phase PWM signal being fixed at zero through a method discussed in the controller chapter) being sent to that phase’s IGBT gate.

3.2.2.3 Controller Interfacing

Current sensing and feedback was also virtually identical in the manner it was performed to the method used with the N+1 switch topology capacitor. Again LA25-NP current transducers from LEM were selected. They also were placed in a near identical manner, as shown in figure 3.16. Locations 1 and 2 for each phase were the possibilities, and as in the N+1 design, location 1 was selected for the same reason as discussed in subsection 3.1.2.1.

![Figure 3.16: transducer placement (location 1) and alternate placement (location 2).](image-url)
As with the N+1 switch converter design, opto-isolating gate drivers were used to provide the IGBT gate pins with the switching signal. However as the emitter for all switches in the N switch converter design was at the converter ground which in turn was tied to the control ground the controller +15 Vdc power supply was able to power the gate drivers and there was no need for dc-dc converters to provide isolated floating power supplies.

3.2.3 Operation

The N switch converter, as discussed had serious recirculation problems. While the current could be controlled during the active phase’s firing arc, once the command input switched to zero this recirculation prevented the current from commutating. Additionally it would cause second phase firing to drive the current back up in the first phase. This could result in current during the inactive periods occasionally exceeding the controlled value seen during active phase firing.

This inability to control the current through inactive positions resulted in negative torque, acoustic noise, and greater heat generation than seen with the N+1 switch topology converter. It also limited the motor to a maximum speed of just slightly above 1000 rpm when running with the N switch topology converter.
Figure 3.18: N switch topology converter response to a current step command.
Chapter 4 – Controller Design

Due to the manner in which SRM machines operate, it is necessary to have a relatively sophisticated electronic control mechanism for them to operate. They cannot simply be connected to a DC voltage source like a DC motor can and be expected to run. It is necessary for the stator phase windings to be conducting current within specific rotor position ranges in order to achieve either positive or negative torque to provide the desired motivation to the rotor. Additionally inductance of the stator phase coils varies with both rotor position and current. Thus in order to achieve optimal performance, it is necessary to adjust the firing arc (when excitation and commutation of the current in the stator coils takes place) according to speed, current, and operating quadrant.

The primary goal of this project was to develop a closed loop speed and current controller to provide four quadrant operation for a two-phase SRM machine. This controller was to provide some basic firing arc adjustment according to rotor speed. Additionally it was to have a wireless device that would enable the user to adjust the operating speed as well as turn the motor on and off while not positioned within arms reach of the motor, and not holding a device tethered to it by a wire or wires.

This chapter discusses the design and implementation of this controller. Section 4.1 discusses the basic controller design. Section 4.2 looks at the wireless input device design and how it interfaced with the rest of the control system. Section 4.3 examines the speed and current control elements both in design and implementation. Section 4.4 looks at four quadrant operation and discusses how it was achieved.

4.1 Basic Controller Design

All switched reluctance motor systems share a relatively common architecture from a control system standpoint. Acceleration and deceleration are provided by the torque generated by the motor. Torque is dependent upon position and current. Thus current
control is generally the inner control loop, while speed control takes place in the outer loop. Phase switching to control the firing arc for each stator coil takes place in the current loop just in front of the power converter block. This basic system concept is shown in figure 4.1.

![Diagram of basic system block diagram for a SRM machine.](image)

**Figure 4.1:** Basic system block diagram for a SRM machine.

### 4.2 Four Quadrant Control

While the basic concept shown in figure 4.1 is simple enough, moving from this to an actual working four quadrant controller for a two phase SRM requires the addition of a great amount of detail.

First and most obviously is the need to control both of the two separate motor phases. In cases where no phases are conducting current simultaneously, a single current sensor could be used. While in most cases only one phase will be receiving a command to conduct time as the two phases will be in opposite regions of their respective torque profiles, the tendency of current in the N switch topology converter to recirculate from one phase to the other even before the first had ceased conducting prevented this. In order to control at least the peak current level in each phase with both power converter designs it was decided that the controller would have a separate, independent current control loop for each motor phase each of which would possess its own sensor to provide feedback.
With the control of both motor phases addressed, the next step is to determine what quadrant the SRM is operating in at any given time. Accomplishing this requires that the controller be capable of discerning whether the motor needs to increase or decrease speed to reach the command reference, and be able to tell which direction it is spinning in. To do this, it was decided that the controller would monitor the speed input command to see whether it was positive (indicating the motor should spin in the clockwise direction) or negative (spin in the counter-clockwise direction), the speed error which would let us know whether it was spinning too fast or too slow, and the actual direction of rotation. With the quadrant of operation known, the next most important information for handling phase switching is rotor position, which was easily determined by use of the encoder attached to the two phase SRM used in this project. Finally, it is desirable to have a means of watching the actual motor speed and adjusting the firing arcs (advance and commutation angles) to improve performance. With all of these in place, it is now possible to come up with a design for a four quadrant controller with firing arc adjustment.

### 4.2.1 Implementation

![Image of circuit board](image.png)

Figure 4.2: Implementation of speed and firing arc controls.

The controller developed in this project was primarily analog based, fabricated using CMOS devices and discrete components. The core of the design centered around the
phase switching control device. Phase switching would be based upon feedback received about rotor position, rotor speed, and quadrant of operation. The phase switching control device would take the output signal from the speed controller and pass it to the active current control loop, while giving the inactive current control loop a zero input. It would also give phase firing arcs directly to the converter as needed (for cases in which there are more than one switch per phase in the converter design being used). The baseline block diagram used for design and implementation of the controller is given in figure 4.3.

Phase Switching Control was implemented using an EEPROM chip. It was decided that use of an EEPROM to store a lookup table that would point to the proper active phase would be the simplest way of handling switching in an analog controller. As there were only two phases that could be active, only two data bits were necessary per memory address. However the address itself would need to be 15 bits wide so that it could accommodate the 10-bit position resolution from the encoder, the 1 bit indicating rotational direction, the 2 bits needed to monitor speed command and error signals, and having 2 bits remaining enabling the controller to have multiple speed ranges for variation of advance and commutation angles to improve performance. The EEPROM selected was the AT25C256 produced by Atmel. It is a parallel 256K EEPROM which provided a 15 bit wide, parallel loading address register along with 8 parallel loading data bits per address location with a 150ns access time. While there were EEPROM chips with similar numbers of address and data pins that were faster, 150 ns was deemed more than adequate for the purposes of this project. 150 ns equates to 0.01 degree at 12000 rpm which is much greater than the 0.35 degree resolution of the encoder and counter used.
The lookup table would be organized based upon the frequency with which the address bits were expected to change. The quadrant indicating bits expected to change the least (speed command polarity, speed error polarity, and direction) often would be used as the three most significant bits of the address register. The motor speed indicator would be next, and then finally the position which would be changing at an extremely high rate of speed forming the least significant bits.
A14 = speed error polarity (+ = 1, - = 0)
A13 = speed command polarity (+ = 1, - = 0)
A12 = direction indicator (CW = 1, CCW = 0)
A11 = most significant speed range bit
A10 = least significant speed range bit
A9  = position bit 10
A8  = position bit 9
A7  = position bit 8
A6  = position bit 7
A5  = position bit 6
A4  = position bit 5
A3  = position bit 4
A2  = position bit 3
A1  = position bit 2
A0  = position bit 1

Table 4.1: EEPROM addressing system

The quadrant indication bits were generated directly from the analog control circuitry using Schmitt Triggers designed to transition from low to high at 0.03 V and from high to low at -0.03 V. This provided an adequate window to avoid noise from causing inadvertent jittering between operational quadrants in the lookup table. The output of the Schmitt trigger circuits was limited to 5.1 V through use of a zener diode to enable them to be directly connected to address pins A13 and A14 on the EEPROM.
Figure 4.4: Schmitt trigger with output limiter.

The trigger threshold equations for a non-inverting Schmitt-trigger circuit are:

\begin{align}
V_{TH} &= -\frac{R_1}{R_2}V_{OL} \\
V_{TL} &= -\frac{R_1}{R_2}V_{OH}
\end{align}

Where $V_{TH}$ and $V_{TL}$ are the high and low trigger voltages respectively, $V_{OL}$ is the low saturation voltage (-13 V in this case since an LM324 amplifier with +/- 15V supplies was used), and $V_{OH}$ is the high side saturation voltage (13 V). The values used for $R_1$ and $R_2$ were 205\,\Omega and 90.9k\,\Omega respectively. Actual quadrant corresponded to the values given in table 4.2. The addition of the diode and zener for the Schmitt-trigger circuits used in this project limited the circuit output to 0 V (when low) and 5.1 V (when high) allowing the output to be sent directly to the appropriate EEPROM address pin.
Direction was determined from the encoder providing position feedback. The encoder had three channels for providing feedback. Channels A and B each provided 1024 positive return signals (“high return signals”) per revolution, separated by 1024 zero return signals (“low return signals”). The high and low signals were equal in duration, and channels A and B were offset from one another by one half of a high signal duration. Channel Z as it was referred to provided a single positive return signal once per revolution that was equal in duration to one of the positive signals from channels A or B. Channels A and B were used to determine direction through use of a simple digital circuit which essentially looked at whether the B signal was transitioning from low to high when channel A was high. The circuit used for this operation is shown in figure 4.5a and relative signal values are shown in figure 4.5b.
Figure 4.5b: Direction monitoring circuit inputs and output.

Speed information was monitored and provided to the EEPROM using Schmitt triggers with limited outputs similar to that shown in figure 4.4. The only difference being the negative input pin was connected to a positive voltage reference which corresponded to the speed feedback voltage at 100 rpm (0.1 V) or at 1500 rpm (1.5 V). This shifted the threshold voltages up to 0.1 V +/- 0.03 and 1.5 +/- 0.03 respectively. In this manner we were able to have three speed ranges in which different firing arcs would be used to improve performance and reduce acoustic noise.
Position information was taken from the same encoder as the directional information. However, as previously stated, the encoder only has three channels by which feedback was provided. Thus in order to get specific rotor position information it was necessary to add a digital counter that could reference the Z-channel positive signal as a reset bit used to set the counter to zero through its parallel loading feature, and using either channel A or B to count up or down depending on the direction indicated by the directional circuit. In the case of this project, two 74F269 bi-directional counters were used. They were connected together as shown in figure 4.6 to gain full 10-bit counting capability.

![Figure 4.6: 10-bit bidirectional counter implementation.](image)

### 4.2.2 Firing Arc Table Generation

The actual firing arcs used to ensure current reached desired levels prior to the entry of the rotor into positive torque regions and commutation of said current prior to entry into
negative torque regions were generated offline and loaded into the EEPROM. Address locations were selected as outlined in table 4.1 keeping in mind that the 10 address bits corresponded to merely 360 degrees and so some conversion from decimal to binary was needed when creating the tables.

The tables were created one for positions ranging from 0 to 60 degrees and then copied as needed taking the 60 degree symmetry for the 4:6 motor into account. Advance and commutation angles were taken from table 2.3. Low speed firing arcs were lifted directly from table 2.2. Mid-range speed firing arcs and high speed firing arcs were generated by adding and subtracting the advance and commutation angles from \( \theta_i \) and \( \theta_f \) positioned about \( \theta_a \) and \( \theta_u \) values realigned from 30 and 0 degrees to the estimated high and low inductance value positions found looking at table 2.1 and figure 2.4 (reflecting the fact that the position sensor is not actually aligned with the poles rotor and stator poles). Since estimated peak and minimum inductance points varied from one another, the point from which \((\theta_i - \theta_u)\) and \((\theta_a - \theta_f)\) were added and subtracted was set midway between the two (at 15.5 degrees and 46.5 degrees respectively).

<table>
<thead>
<tr>
<th>Angles</th>
<th>Forward Motoring</th>
<th>Forward Regeneration</th>
<th>Reverse Motoring</th>
<th>Reverse Regeneration</th>
</tr>
</thead>
<tbody>
<tr>
<td>0 – 9.23 degrees</td>
<td>A</td>
<td>B</td>
<td>B</td>
<td>A</td>
</tr>
<tr>
<td>9.23 – 21.78 degrees</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>21.78 – 40.23 degrees</td>
<td>B</td>
<td>A</td>
<td>A</td>
<td>B</td>
</tr>
<tr>
<td>40.23 – 52.78 degrees</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>52.78 – 60 degrees</td>
<td>A</td>
<td>B</td>
<td>B</td>
<td>A</td>
</tr>
</tbody>
</table>

Table 4.4: Baseline \((\theta_{adv} = 0 \text{ and } \theta_{comm} = 0)\) firing arcs.

For forward motion (motoring and regeneration) the advance and commutation angles are both subtracted from these baseline values to generate the actual firing arcs. For reverse motion, the advance and commutation angles are added to reflect the fact the rotor is approaching these positions from the opposite direction. As an example this A-phase firing arc in forward motoring at mid-speed (5.36 degree advance, 8.75 degree
commutation) would shift from 52.78 thru 9.23 degrees (note: 60 degrees = 0 degrees due to symmetry) to 47.42 thru 0.48 degrees.

4.3 System Input

The input to the controller in this project became a significant block in and of itself due to the requirement for a wireless input device that would allow a user to operate the motor from a location removed from the motor and its controller. This device was required to perform three functions. It had to let the user increase the motor’s operating speed, decrease the motor speed, and bring the motor to a complete stop.

In order to accomplish this there were three elements that needed to be brought together. First there was the user interface through which the operator would enter these three commands. Second, there was the wireless link which would transmit these commands from the operator’s interface to the controller which was physically connected to the converter and SRM. Third, there was the controller side interface needed to interpret the signal from the wireless link and convert it into a signal useable by the actual controller.

4.3.1 Wireless Technology

While there are a myriad of wireless technologies available, it was decided to keep the input device as simple and inexpensive as possible. To this end, a radio transmitter and receiver pair operating on the 433 MHz band that was already available in the lab was selected along with a pair of 6 inch whip style antennas. If this proved to be ineffective due to electromagnetic interference or some other reason then other options such as infrared, or higher frequency RF systems with greater EMI rejection would be looked at, time allowing.
4.3.2 User Controls and Transmitter

The user interface to the system consists of two buttons and a switch. The two buttons generate a signal for the controller to increment the speed reference by one spot on the digital potentiometer, and a signal for the same speed reference to be decremented by one position. The switch cuts the off the speed reference resulting in the motor coming to a complete stop from whatever speed it is operating at and remaining at a halt until the switch is flipped back to its “on” position.

The signals from the user controls are fed through a Holtek HT-12E encoder which combines them with an address code which will be used on the receiving end to discriminate between a genuine command and some external unrelated signal. From the encoder, the signals are sent to the TWS-434 transmitter unit which converts them from a 5V level signal to a 434MHz amplitude modulated radio signal which is then transmitted over the 6 inch whip radio antenna.

Figure 4.7: User controls, encoder, and RF transmitter.
4.3.3 Receiver and Decoder

The RF receiver used was an RWS-434. It takes in the 434MHz AM signal from the whip antenna mounted on the speed controller / switching controller board and converts them to a 5V digital signal which is then fed to a Holtek HT-12D decoder. The decoder in turn examines the signal for the address bits set by the encoder. If they are present and match the address settings of the decoder, it then accepts the signal as valid and splits the 4 data bits, setting its data output pins D0-D3 to match those encoded and sent by the transmitting unit. These data out pins will remain at the received voltage level until another valid signal is received.

Data out pin D0 carries the increment speed command signal to the speed control. D1 carries the decrement speed command. D2 carries the run / stop (cutoff) signal.

![Diagram of RF receiver and decoder](image)

Figure 4.8: RF receiver and decoder.

4.3.3 Controller Interface

Interfacing the RF receiver to the speed control loop required that the controller be able to convert from a digital signal to an analog signal. One option considered to carry out
this digital to analog conversion was a binary counter to record the increment and decrement pulses and feed them to a digital-to-analog converter. The other option was to use a digitally controlled potentiometer for essentially the same purpose. As the digital potentiometer was simpler to implement, and required fewer parts and connections, it was selected.

The potentiometer used was a DS1804 from Maxim Integrated Circuits. It has 100 wiper positions which it can increment and decrement between and is capable of handling input voltages of up to 5 V_{dc}. Additionally, it has onboard memory which allows it to store wiper positions and automatically return to that wiper position every time it powers up. This was utilized to set the wiper position to its mid-point at every power up so that the default output voltage would be 2.5 V_{dc}. In order to provide the digital potentiometer with the needed signals, i.e. increment and up / down select, some intermediate digital circuitry was needed. The DS1804 is designed to step its wiper position in one direction or the other on the falling edge of the signal being sent to its INC pin. If the signal on the UP/DN pin is high at that point in time, then the wiper position moves one step towards the H pin. If the signal on the UP/DN pin is low, then the wiper moves one step towards the L pin.

Clearly, the INC pin needs to have a single high to low transition for both when the increase speed button is pressed on the input device, and when the decrease speed button is pressed (condition A). But the INC pin should not receive a falling signal if both buttons are somehow depressed simultaneously (condition B). The signal on the UP/DN pin on the other hand just needs to be high when the increase speed button is pressed (condition C). In order to accomplish this, it was decided that it would be acceptable to let the potentiometer carry out its operations on the release of the user control pushbuttons rather than requiring it act immediately when they are depressed. With this decided, the signal going to the DS1804’s INC pin was set as the output of the decoder’s output 1 and 2 pins fed through an XOR gate. This satisfied both conditions A and B. In order to meet condition C, output 1 (increase speed) was fed through a digital buffer and then passed by a passive analog delay circuit to prevent it from falling as quickly as the
signal to INC. This indeed achieved the desired result. The entire intermediate circuitry from D0 and D1 to the DS1804 is shown in figure 4.9.

![Diagram of decoder to digital potentiometer intermediate circuitry.]

To implement the run / stop (cutoff) command, a similar circuit was needed. Ideally, there would have been a reset feature on the digital potentiometer; however that feature did not exist. Accordingly it was decided that the output of the DS1804 would be fed through an additional circuit which would essentially turn the potentiometer wiper output voltage on and off before handing it off to the speed loop error generator. For this purpose, a precision SPST (single pole, single throw) analog switch in a MAX352 integrated circuit was selected and a pull-down resistor circuit designed. This circuit is shown in figure 4.10.
From the cutoff switch in the MAX352 IC, the $V_{\text{spd}_{\text{in}}}$ is fed through a difference amplifier where 2.5 V is subtracted to give it a range from -2.5V to +2.5V, the negative range equating to counterclockwise rotation, the positive range clockwise rotation. Then it proceeds through a unity gain first order integral circuit with anti-windup limiting which is used to generate a “soft-start” to the speed loop command. From there, it is fed through a gain amplifier where it is scaled to the desired voltage range and then a difference amplifier where a tune-able voltage level -1.5 V to 1.5 V can be added to trim the signal as needed to set the voltage to zero when the digital potentiometer wiper is at its middle position. Finally the signal is sent to the speed control loop as the speed command or reference voltage. Figure 4.11 displays this series of op-amp circuits.
4.4 Speed Loop

The speed control loop makes up the outer control loop of this system. It is designed to function with both positive and negative voltage signals, corresponding to operation of the motor in the clockwise and counter clockwise direction respectively. Elements in the speed loop that were included in the controller were: the speed error generator, the forward feed-through control blocks (proportional and integral), and the feedback signal conditioning elements. This section will examine these elements in that order.

4.4.1 Speed Error Generation

Speed error generation was carried out using a simple, non-inverting difference amplifier circuit (which can be viewed in appendix-B if desired). The difference amplifier had unity gain for both inputs.

4.4.2 Feed through control

From the error generator, the error signal was sent to the forward feed-through control elements. In this project, the speed controller made use of both proportional and integral control. A first order low-pass Butterworth circuit was used to provide integration with anti-windup. Proportional gain was combined with the summing circuit to minimize the
number of components needed for fabrication. The circuits used to achieve this are shown in figure 4.12.

![Forward feed through control circuits](image)

Figure 4.12: Forward feed through control circuits.

The idealized characteristics of the Butterworth filter being used as an integrator are:

\[
A_F = \left(1 + \frac{R_F}{R_1}\right) = \text{ideal passband gain}
\]

\[
f_c = \frac{1}{2\pi RC} = \text{high cutoff frequency of the filter}
\]

\[
t_c = \frac{1}{f_c} = 2\pi RC = \text{time constant}
\]

By making R_F very small (10 Ω), and R_1 large (1 M Ω) we can make the gain approach unity to give us an integral gain of one. As the maximum gain in this circuit is A_F, which in our case is approximately 1, we do not need to worry about saturation from integral windup.
The ideal output equation for the summing and proportional gain amplifier is:

\[ V_o = \frac{R_A}{R_B} \frac{R_A + R_B}{R_A} \frac{R_A R_B}{R_B + R_A} \left[ \frac{V_1}{R_B} + \frac{V_2}{R_A} \right] \]

From this we can see that the ideal gain values for the two inputs are:

\[ A_1 = \frac{R_A}{R_B} \frac{R_A + R_B}{R_A} \frac{R_A R_B}{R_B + R_A} \]

(4.6)

\[ A_1 = \frac{R_A}{R_B} \]

(4.7)

\[ A_2 = 1 \]

And the entire output equation can be restated as:

(4.8)

\[ V_o = A_1 V_1 + A_2 V_2 \]

In the project the final gain values used were a gain of 1 for the integral output, and a gain of 10 for the proportional output. To achieve this \( R_A \) was selected as 100k\( \Omega \) and \( R_B \) was selected as 10k\( \Omega \).

Leaving the speed control circuits, \( V_{\text{spd_ctl_out}} \) still needs to be conditioned for use in the current control loop. Because the current in either of the two converters always has a positive polarity due to the design of said converters, the input to the current control loop should be positive in all cases. Generating positive and negative torque will be handled through phase selection and switching instead of current polarity. Furthermore the signal needs to be limited to same the voltage level as that which corresponds to 8 A being
present in the stator coils. This is necessary to prevent the controller from allowing the currents in the stator coils to exceed the rated current of the motor.

In order to make the current controller input positive for both positive and negative polarity speed loop outputs, an absolute value circuit was used [5]. This circuit is shown in figure 4.13.

The ideal output of this circuit is simply:

\[
V_o = |V_i|
\]

The behavior of its various elements to achieve this is described in [5]. It is also paraphrased in Appendix-B.

Since the current feedback was scaled such that 8 A corresponds to a signal of 7.57 V dc, the output of the absolute value circuit was limited by use of a 7.5 V zener diode. This corresponds to a command for 7.9 A, ensuring that the controller should never call for increased current beyond the 8 A limit rating.
4.4.3 Feedback signal conditioning

Feedback in the speed control loop is provided through the 3 channel encoder previously described in section 4.2.1. In order to convert the digital pulses to a useable analog format, the output of channel A was fed to the input pin of a frequency-to-voltage converter circuit based around the LM2907 chip in a manner nearly identical to that described in [2]. This circuit can be seen in figure 4.14.

Important equations relating to this circuit taken from the LM2907 datasheet are:

\[ V_{\text{out}} = V_{\text{CC}} f_{\text{in}} C_1 R_1 \]  
\[ R_1 \geq \frac{V_{\text{out}}}{I_{3\text{min}}} \]
\( f_{\text{max}} = \frac{I_2}{C_1 V_{CC}} \)

\( V_{\text{ripple}} = \frac{V_{CC}}{2} \left( \frac{C_1}{C_2} \right) \left( 1 - \frac{V_{CC} f_{in} C_1}{I_2} \right) \)

Equation (4.10) is used in sizing of \( R_1 \) and \( C_1 \), given a known supply voltage, frequency, and a desired output voltage at that frequency. For this project with the SRM rated for 12,000 rpm it was decided to scale the feedback to give a 10 V dc return at a speed of 10,000 rpm. 10,000 rpm at 1024 pulses per revolution gives us a pulse frequency of:

\[
f_{in} = 10000 \cdot \frac{1}{60} \cdot 1024 = 170,667 \text{ Hz}
\]

With this value known and \( V_{CC} \) set at 15 V, a capacitor size of 10 pF was selected for \( C_1 \). \( R_1 \) was then calculated using equation (4.10) giving a result of 390,550 \( \Omega \). (Select 390.6 k\( \Omega \) as a standard resistor size.) These values were then checked against equations (4.11) and (4.12) to ensure they satisfied operational requirements of the LM2907 IC.

\[
R_1 \geq \frac{V_{out}}{I_{3\text{min}}} = \frac{10 V}{150 \mu A} = 67.7 \text{ k}\Omega \\
390.6 \text{ k}\Omega \geq 67.7 \text{ k}\Omega \quad \checkmark
\]

\[
f_{\text{max}} = \frac{I_2}{C_1 V_{CC}} = \frac{190 \mu A}{10 \text{ pF} (15 V)} = 1.27 \text{ MHz} \\
1.27 \text{ MHz} \geq 170.7 \text{ kHz} \quad \checkmark
\]

Finally, \( C_2 \) was selected using the equations (4.13) using a maximum voltage ripple of 0.01 V which equates approximately to 10 rpm. Using all the previous values for \( V_{CC} \), \( f_{in} \), \( C_1 \), and \( I_2 \) a capacitance of 6800 pF was eventually settled on for \( C_2 \).

\[
V_{\text{ripple}} = \frac{V_{CC}}{2} \left( \frac{C_1}{C_2} \right) \left( 1 - \frac{V_{CC} f_{in} C_1}{I_2} \right)
\]
After converting the pulse frequency being returned from the encoder, it was still necessary to correct the signal polarity to match with the input voltage polarity scheme (positive voltage to CW rotation, negative voltage to CCW rotation). This is accomplished using the circuit shown in figure 4.15.

![Figure 4.15: Speed feedback polarity correction circuit.](image_url)

This circuit is composed of three primary parts. In the first, the feedback signal is sent through a non-inverting gain block where it is multiplied by 2. This doubled signal is then sent to a MAX352 analog switch where it is connected as the voltage source in a pull-down resistor so that when the direction signal is low the output of the circuit is zero.
and when the direction signal is high it will be passed through. The direction signal itself is generated by the circuit shown in figure 4.4a and then inverted and boosted to the 15V level needed to motivate the switches in a MAX352 IC. After the pull-down circuit, the final stage of the polarity correction circuit is a difference amplifier. The difference amplifier is set up to subtract the output of the pull-down circuit, from a non-amplified $V_{\text{spd_fb}}^*$ signal. The result of this is when the motor is rotating clockwise the difference amplifier will simply subtract zero from $V_{\text{spd_fb}}^*$, and when the motor is rotating counterclockwise the difference amplifier will subtract $2(V_{\text{spd_fb}}^*)$ from $V_{\text{spd_fb}}^*$ resulting in a signal of $-V_{\text{spd_fb}}^*$. Thus $V_{\text{spd_fb}}$ is given a full range of positive and negative values according to the speed and direction of rotating of the motor.

4.5 Current Loop

The two independent current control loops in the system each contain the following elements: a selection circuit to carry out part of the phase switching, an error generation circuit, the forward feed through control circuit, a PWM generation circuit, any necessary digital switching logic, the feedback generation circuit, and a filtering and scaling circuit for the current feedback. This section will examine these elements in this order.

4.5.1 Phase Loop Selection

After leaving the speed loop, the control signal is sent to both independent current control loops. Some of the phase switching done by the controller (all of the phase switching in the case of the N switch topology converter) is done at this stage. This is accomplished by switching the control signal on and off using precision SPST analog switches from a MAX352 IC in a pull-down arrangement as shown in figure 4.16.

From the phase selector circuit, the current loop input signals were sent to the error generator for each respective phase control loop.
4.5.2 Current Error Generation

Current error generation was carried out using a simple, non-inverting difference amplifier circuit. The difference amplifier had unity gain for both inputs.

4.5.3 Feed through control

The current controller was designed and implemented as a simple proportional controller. This was implemented through use of a simple non-inverting gain with a gain of 10. Such a gain amplifier can be seen in Appendix-B. The resistor sizes used to create this gain were 10kΩ and 90.9kΩ.
4.5.4 PWM Generation

PWM generation was carried out using LM3524’s produced by National Semiconductor. However before being sent to the actual PWM generator, the control signal needed to be limited and shifted to match the PWM input limits. In the PWM generator used in this project, a control input of approximately 1 V is the lower limit at which the generator begins creating a positive duty cycle square wave. A full 100 percent duty cycle is reached at an upper limit of approximately 3.5 V. However, both of these limits are very approximate and can vary from chip to chip. The actual circuit used to limit and shift the control signal is shown in figure 4.17.

![Figure 4.17: Current signal limit-shift circuit.](image)

This circuit first limits the output of the current controller to 7.5 V maximum through use of the zener diode. Following this, the limited signal is divided by three through a voltage divider and then sent to one of the inputs of a unity gain summing amplifier. The second input to the summing amplifier is simply the wiper output of a 20kΩ potentiometer with its high and low input voltages set to +15 V and – 15 V respectively. The potentiometer is used to create a trim voltage which will set the PWM input level to just below the zero duty cycle input limit when the input to the current control loop and the current feedback are both zero. As each current loop has its own PWM generator,
and the input limits can vary each loop’s limit shift tuning potentiometer needs to be adjusted separately.

From the limit-shift circuit, the control signals were sent to the input pin of the LM3524 PWM generation circuits shown in figure 4.18.

![Figure 4.18: PWM generator.](image)

This is a fairly standard circuit, outlined in the LM3524 datasheet and used in [2] as well. The variable resistor and capacitor between pins 6 and 7 and ground are used in adjusting the PWM oscillator speed according to a table provided in the datasheet. For this project a 20 kHz PWM switching frequency was selected. The resistor off of the PWM output line is actually a pull-down resistor used to ensure the output voltage does not float.
during “off” periods. The 10Ω resistor between pin 13 and 15 is simply to limit the output current.

### 4.5.5 Switching Logic

There was no need for additional switching logic to control the N switch topology converter, however as it had been decided that the shared IGBT of the N+1 converter would be used for PWM switching some logic was needed for the N+1 converter. The logic circuit used is shown in figure 4.19.

![Switching logic for N+1 converter common switch.](image)

This circuit actually outputs the same signal as a circuit made up of two AND gates receiving the phase select signal and PWM signal for each respective phase feeding their results to an OR gate. NAND gates were used as a substitute due to their lower cost.

### 4.5.6 Feedback

As discussed in Chapter 3, current feedback in both converter designs is provided by way of an LA-25NP current transducer manufactured by LEM Inc. Set up in a single winding configuration, the transducer produces 1 mA output per amp of winding current. This in turn is sent through a 300 Ω resistor to ground producing a 0.3 Volt / Amp voltage level which is then sent through a low-pass filter and gain block before being subtracted from the current command to produce the phase’s current error signal. The feedback circuit is shown in figure 4.20.
As is shown in the figure, the 300 Ω resistance used to convert the current output to a voltage level is composed of a 280 Ω resistor (actually two 560 Ω resistors in parallel) and a 100 Ω variable resistor used to trim as needed to get the desired accuracy.

The filter is a second order low-pass KRC filter designed for unity gain. Its cutoff frequency is simply:

\[
 f_H = \frac{1}{2\pi RC} 
\]

Scaling of the signal is handled by the non-inverting gain amplifier the gain of which is:

\[
 A = \left(1 + \frac{R_B}{R_A}\right) 
\]

In the case of this project, the input to the gain amplifier ranged from zero to 2.4 V as current increased from zero to 8 A, since our current command is restricted to 7.5 V, we wanted to scale the feedback for 8A as close to that as possible. Accordingly \( R_A \) and \( R_B \) were simply set as 5.1kΩ and 11kΩ respectively to give us a gain of 3.15 and a final feedback signal range from 0 to 7.57 V.
4.6 Observations

Overall, the speed and current control loops performed their job adequately. Variation present in the resistors and the op amps themselves led to the buildup of common mode error which was amplified by subsequent circuits, but this could be largely “trimmed” from the control loops through adjustment of the input trimming circuit just before the speed error generator and the trim potentiometer in the current control shift-limit circuits. A larger source of error which was impossible to trim out was the behavior of the MAX352 analog switches. When “off” or open, some current still bled through to the high side of the pull-down resistor resulting in a positive offset which would feed through subsequent op amp circuits. As it was still possible to have an actual zero volt signal when the MAX352 was “on” or conducting, this error was not always present and thus could not be tuned out without loss of some PWM duty cycle range. The digital elements present, generally in the phase selection and switching control portions of the system, worked flawlessly.

Another source of difficulty which could not be overcome with the time and resources that were available was a tendency for electromagnetic interference to get into the system at specific, short periods of time during testing. It is surmised that this EMI was the result of experimental work being done in nearby laboratories and was entering through the extensive mass of wire wrap connections traversing the prototyping boards from one connector pin to another. If more work is done building upon the groundwork laid by this project it will be highly advantageous to move to a digital implementation for the speed and current control loops, and to have a printed circuit board fabricated to eliminate or at least reduce the wire-wrap connections present as a potential entry for noise into the system.
Chapter 5 – Experimental Verification

Following the implementation of the converter and controller designs, the entire system was tested to verify the ability to carry out the stated objectives. Readings were taken of the speed command, speed feedback, current command, current feedback, and active phase selection (firing arcs). By examining graphs of these signals it was verified that the controller successfully drove the examined converters and the two phase switched reluctance motor through all four quadrants of operation. Additionally, the system’s ability to follow step commands to both current and speed were tested and observed.

This chapter discusses the various tests that were conducted and examines their implications.

5.1 Four Quadrant Operation: N+1 Switch Topology Converter

Relevant signals to show that the motor is operating in all four quadrants of operation are the speed command signal, the speed feedback signal, and the current waveform in at least one phase’s stator winding.

A positive speed command signal indicates the motor is being commanded to drive in the forward direction, while a negative speed command indicates driving in the reverse direction has been commanded. When the speed feedback signal is positive, but less than the speed command the motor should be forward motoring. When the speed feedback is positive and greater than the speed command the motor should be in the forward regeneration quadrant. When speed feedback is negative but greater (closer to zero… not greater in magnitude) than the speed command, then the motor is commanded to be in the reverse motoring quadrant. Finally, when the speed feedback is negative and less than (greater in negative magnitude) the speed command, the motor should be in reverse
regeneration. Presence of a current waveform verifies that the motor is being actively driven (torque is being applied) and is not just coasting at the recorded speed.

### 5.1.1 Forward Motoring

![Diagram of Forward Motoring Using the N+1 Switch Topology Converter]

In figure 5.1 we can see the system operating in the forward motoring quadrant using the N+1 switch topology converter. As the speed command signal and speed feedback were scaled such that a 1 V signal equates to 1000 rpm, the motor is running at approximately 1000 rpm in this figure. The current probe used to record the displayed current waveforms was set to a scale of 0.5 amps per 10 mV displayed on the oscilloscope. As can be observed there is a slight steady state offset from the command value, but this is fairly small, on the order of 100 rpm.
5.1.2 Forward Regeneration

In figure 5.2 we can see a case in which the speed command has rapidly dropped to almost zero, but the actual speed is only just falling past 2000 rpm. Current is being applied through the negative torque angles to slow the motor. The high error signal caused by the large difference between command and feedback signals is resulting in a high duty cycle through the entire firing arc for each phase. This is the reason for the continuously rising current waveforms which are being cut off by the phase selector signal rather than peaking and flattening out as was the case with the forward motoring and reverse motoring cases shown, both of which were recorded at steady state operation.
5.1.3 Reverse Motoring

Figure 5.3: Reverse motoring using the N+1 switch topology converter.

Figure 5.3 shows the motor running in the negative motoring quadrant at approximately 1000 rpm. As with forward motoring there is a steady state offset on the order of 100 rpm.
5.1.4 Reverse Regeneration

Figure 5.4 shows the motor slowing its reverse directional speed as the command signal is changed from 1.5 V (approximately 1500 rpm) to zero. As with figure 5.2 the current waveform is triangular in shape instead of flattening out at a commanded peak value because the firing arc being given by the EEPROM is too short for the current to reach its command value before entering commutation.

Figure 5.4: Reverse regeneration using the N+1 switch topology converter.
5.2 Four Quadrant Operation: N Switch Topology Converter

As with the N+1 switch topology converter design, the speed command, speed feedback, and one phase’s stator coil current feedback were recorded to demonstrate the controller’s ability to drive the motor in all four quadrants of operation using the N switch topology design for the power converter. The phase-2 selection signal (firing arc) was also included at the bottom of each graph to clearly show when the current shown was being actively driven to produce positive torque versus merely recirculating from phase-1.

As discussed in the converter chapter, significant recirculation of current occurred while using the N switch topology converter, reducing both efficiency and performance. However, this did not prevent the system from successfully demonstrating operation in all four quadrants at lower speeds.

5.2.1 Forward Motoring

![Graph showing speed response, phase A current response, and phase A EEPROM select signal.]

Scale: 1000 rpm / div
2 Amp / div
5 Volts / div

Figure 5.5: Forward motoring using the N switch topology converter.
In figure 5.5 the motor is running at a steady 500 rpm (with an 800 rpm command speed). From the current wave form and current selection signals shown its quite clear that the current signal is not commutating between active selection angles and in fact is rising after a brief period to almost its full active command level. Current is also scaled in this case to be 2 amps per 10mV, so rather than running at the 1.5 amps peak range shown in figures 5.1 and 5.3 with the N+1 converter, in this case the motor is drawing between 3 and 5 amps through its stator coils at all times while running at about half the actual speed.

### 5.2.2 Forward Regeneration

![Forward Regeneration Diagram](image)

From the fact that both speed command and feedback signals are positive, but the speed feedback is greater than the command figure 5.6 clearly is supposed to shown forward
regeneration. The fact that speed is in fact dropping and current being applied confirms this.

5.2.3 Reverse Motoring

Figure 5.7: Reverse motoring using the N switch topology converter.

Figure 5.7 shows the motor in the reverse motoring quadrant using the N switch topology converter. As in figure 5.5 the current drops very little across the inactive angles and in fact rises in some places to exceed its level during the active selection positions. Also as with the forward motoring data, this figure shows a steady state offset of about 300 rpm.
5.2.4 Reverse Regeneration

Figure 5.8: Reverse regeneration using the N switch topology converter.

Figure 5.8 shows the motor breaking to slow from slightly over 1000 rpm in the negative direction to zero. The waveform follows the characteristic behavior of current controlled using the N switch topology converter.

5.3 Speed Response

In addition to demonstrating four quadrant operation, the system was also tested to show its ability to control motor speed. This was accomplished by using the cutoff switch to turn a set command speed on and then off. It should be noted that the presence of the “soft start” circuit prevented a true step signal from being input, this is the source of the “rounding” shown in figures 5.9 and 5.10. However by swapping in a 0.1 μF capacitor in place of the 1 μF capacitor initially used the time constant of the soft start circuit was
reduced by a factor of 10 resulting in an input close enough to a true step to get a fairly characteristic response.

Figure 5.9: System response to a step speed command when running with the N+1 switch converter.

Figure 5.9 shows the speed response to a step input command of 1500 rpm when the N+1 switch converter is used. The motor here seems to have the characteristic response of an under damped system, but it settles quickly until the speed command is changed again. There is a constant negative offset error that is present, also the amount of overshoot from the steady state value is greater when slowing to zero than from when motoring up towards the positive command value. This is undoubtedly due to the dampening affect of air resistance and friction which will be greater at higher speeds.
In figure 5.10 we can see the motor operating using the N switch topology converter responding to the same 1500 rpm command step input. While there is no clear overshoot, the motor also never really settles to a constant speed, instead fluctuating almost constantly around an average value of about 1100 rpm (close to the system’s maximum speed using the N switch topology converter). In addition to the larger offset error and the settling issue, response time also appeared to be somewhat slower.

5.4 Current Response

Moving from the performance of the speed control loop to the current control loop, signals were also recorded of the current command value for one phase and the actual onboard, filtered, and amplified current feedback signal.
As can be seen in figure 5.11 the N+1 switch topology converter did a fairly good job of controlling current. There was a significant negative offset from peak command value present, but response time was good, overshoot was limited, and very little ripple was present. The offset could be lessened through control gain optimization.
Figure 5.12: Current step response of the system running with the N switch topology converter.

The N switch topology converter on the other hand as discussed in chapter 3 had serious issues controlling the current present in the phase. While the current level during the actual firing arc for the selected phase was comparable to that for the N+1 switch topology converter, the total inability to commutate and the recirculation of current through inactive phases meant that the current was never properly cut off, and was in fact essentially uncontrolled during what should have been inactive periods for the observed phase, occasionally rising to levels higher than it was commanded to reach during active periods.
Chapter 6 – Conclusion

As intended, this project resulted in development of an analog based control system for a two phase switched reluctance motor capable of driving the motor in all four quadrants of operation. Furthermore, the controller was able to control two different converter designs. While current control was difficult using the particular N switch topology converter selected, this was a function of the converter design itself rather than the controller. With slight modification to the logic used to select the active phase and ensure PWM signals are only sent during active positions this basic control design could easily be used with any SRM converter design from 1 switch per phase, to 1 switch per phase with a shared switch, to 1.5 switches per phase, and even 2 switches per phase.

The analog control circuitry performed its job adequately. There were some issues with current leaking through the MAX352 analog switches when they were supposed to be open, but the offset error introduced by this was trimmed out with only some minor loss of current response. Further work could be done to optimize performance, reducing steady state offset and improving settling times for both speed and current, however optimization is beyond the initial scope of this project.

Finally, the RF transmitter worked as expected demonstrating the ability to remove the wire tethering the motor operator to the motor’s physical location. In a noisier environment such as an actual industrial floor the 434 MHz system used might not be sufficiently resistant to interference, but for home use it probably would be.

The primary contributions made by this project are:

- Development of a four quadrant controller for the 4:6, self starting, two phase, switched reluctance motor
- Development of a control platform for testing multiple 2-phase SRM converters with varying switching topologies
- Experimental testing of the N switch topology converter invented by W.A. Harris
6.1 **Recommendations for Future Work**

Continuing from the baseline which this project has established there are several avenues of future work that can be pursued. The easiest and most beneficial would probably be the conversion of this controller design from analog to digital, implementing its features in a DSP. This would have several beneficial effects:

- Reduce the components needed.
- Reduce susceptibility to electromagnetic noise.
- Eliminate the need for troublesome polarity and analog on/off circuits.
- Eliminate error from variation in analog components.

After the implementation of a DSP based system, a way of implementing more speed ranges for generating phase firing arcs or else an algorithm to generate firing arcs completely on the fly should be looked into. The transitions from one speed range to another were some of the roughest points during operation of the system. If the jump in speeds at which the firing arcs were set up for were large enough these transitions could actually lead to a breakdown in control – as advance angles pushed the start of a phase’s firing arc into negative torque producing angles while the motor was running slow enough for the converter to reach peak current levels before positive torque angles were reached. Introducing more speed ranges or eliminating incremental speed ranges all together would reduce or eliminate this problem and allow higher speeds to be more quickly and smoothly attained by the motor.

Finally, different sensing methodologies should be looked at. While adequate for this project, the encoder used added to the acoustic noise in the system and was a source of potential reliability faults as well as cost. Furthermore, the fact that it only had one reset bit per 360 degrees of rotation meant that while the motor could start from any position (it was “self starting”) – when the controller was first turned on, the motor would have to make 1 full rotation before it could be controlled. As the rotor is symmetric there should be no real need to have absolute position so long as the position of each stator pole
relative to the closest rotor pole is known. While the current sensing transducers did a fairly good job once filtering was added, they too represented a source of cost, potential error (1 transducer had to be replaced due to the production of a constant negative offset error), and complexity.
Appendix A – System Level Schematics
Switching Control
N Switch Topology Converter

From
Phase-A
Stator Coil
Outlet

To DC Supply
Return

From
Phase-B
Stator Coil
Outlet

To DC Supply
Return

PWM Signal - A

PWM Signal - B

HCPL-3150

Vcc +15 Vcc

PWM Switch (Ts)
IRG4PC40U

5kΩ

Cathode

Anode

HCPL-3150

Vcc +15 Vcc

PWM Switch (Ts)
IRG4PC40U

5kΩ

Cathode

Anode
Appendix B – Op Amp Circuits and Idealized Characteristics
Non-inverting Gain Amplifier:

\[(B.1)\]

\[V_{out} = \left(1 + \frac{R_2}{R_1}\right)V_{in}\]

Difference Amplifier:

\[(B.2)\]

\[V_{out} = \frac{R_2}{R_1}(V_2 - V_1)\]
First order Low-pass Butterworth Filter:

\[
V_{out} = \frac{Af}{1 + \left(\frac{f}{f_H}\right)} V_{in}
\]

where

\[
Af = \left(1 + \frac{R_F}{R_1}\right) = \text{passband gain}
\]

\[
f_H = \frac{1}{2\pi RC} = \text{cutoff frequency (hertz)}
\]

\[f = \text{input signal frequency (hertz)}\]
Variable Gain non-inverting Summing Amplifier:

(B.7) $$V_{out} = \frac{R_A}{R_B} \cdot \frac{R_A + R_B}{R_A} \cdot \frac{R_B R_A}{R_B + R_A} \cdot \left[ \frac{V_1}{R_B} \right] \left[ \frac{V_2}{R_A} \right]$$

or

(B.8) $$V_{out} = A_1 V_1 + A_2 V_2$$

where

(B.9) $$A_1 = \frac{1}{R_B} \cdot \frac{R_A + R_B}{R_A} \cdot \frac{R_B R_A}{R_B + R_A}$$

which simply reduces to 1.

(B.10) $$A_1 = 1$$

(B.11) $$A_1 = \frac{R_A}{R_B} \cdot \frac{R_A + R_B}{R_A} \cdot \frac{R_B R_A}{R_B + R_A}$$

which in turn reduces to

(B.12) $$A_2 = \frac{R_A}{R_B}$$
Absolute Value Circuit:
(provided courtesy of Texas Instruments)

According to the Burr-Brown application note written by D. Jones and M. Stitt [5]:

![Figure 1: Precision Absolute Value Amplifier has High Input Impedance and Requires Only Two Matched Resistors.](image1)

The circuit shown in Figure 1 is a split supply circuit preferred when high input impedance is desired. To understand how the circuit works, notice that for positive input signals D1 becomes reverse biased resulting in the active circuit fragment shown in Figure 2. A1 drives the non-inverting input of A2 through forward biased diode D2. The feedback to the inverting inputs of A1 and A2 is from the output of A2 through resistors R1 and R2. Since no current flows through resistors R1 or R2, in this condition, VOUT is precisely equal to VIN.

![Figure 2: Positive Input Voltages to the Figure 1 Circuit Result in This Circuit Fragment. The circuit operates as a precision unity gain voltage follower. No errors are produced by the forward-biased diode, D2, or the resistors.](image2)
When the input voltage to the absolute value amplifier shown in Figure 1 becomes negative, $D_2$ becomes reverse biased resulting in the active circuit fragment shown in Figure 3. $A_1$ drives $R_1$ through forward biased diode $D_1$ to a voltage equal to $V_{IN}$. $A_2$, $R_1$, and $R_2$ form a simple unity gain inverting amplifier. $R_1$ and $R_2$ must be carefully matched to provide accurate gain $= -1V/V$ to match the $+1V/V$ gain for a positive input signal. Compensation capacitor $C_1$ ensures the circuit is stable with $A$ in the feedback loop. For good stability and best speed, set the $C_1 \cdot R_1$ pole equal to about $1/4$ the unity gain bandwidth of $A_2$.

![Circuit Diagram](image)

Figure 3: Negative Input Voltages to the Figure 1 Circuit Result in This Circuit Fragment. The circuit operates as a simple inverting amplifier. Resistors $R_1$ and $R_2$ must be matched to achieve a precise gain of $-1V/V$. 


Appendix C – Partlist
## Section C.1  Speed Controller & Phase Selection Controller

<table>
<thead>
<tr>
<th>Part No.</th>
<th>Description</th>
<th>No. Required</th>
<th>Unit Cost</th>
<th>Total Cost</th>
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### Section C.2  N+1 Switch Topology Converter / Current Control / Switching

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<td>MFR-25FBF-11K0</td>
<td>11kΩ resistor (1%)</td>
<td>2</td>
<td>0.11</td>
<td>0.22</td>
</tr>
<tr>
<td>MFR-25FBF-10K0</td>
<td>10kΩ resistor (1%)</td>
<td>6</td>
<td>0.11</td>
<td>0.66</td>
</tr>
<tr>
<td>MFR-25FBF-4K99</td>
<td>4.99kΩ resistor (1%)</td>
<td>25</td>
<td>0.11</td>
<td>2.75</td>
</tr>
<tr>
<td>MFR-25FBF-560R</td>
<td>560Ω resistor (1%)</td>
<td>4</td>
<td>0.11</td>
<td>0.44</td>
</tr>
<tr>
<td>MFR-25FBF-160R</td>
<td>160Ω resistor (1%)</td>
<td>4</td>
<td>0.11</td>
<td>0.44</td>
</tr>
<tr>
<td>MFR-25FBF-100R</td>
<td>100Ω resistor (1%)</td>
<td>3</td>
<td>0.11</td>
<td>0.33</td>
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<tr>
<td>MFR-25FBF-10R0</td>
<td>10Ω resistor (1%)</td>
<td>2</td>
<td>0.11</td>
<td>0.22</td>
</tr>
<tr>
<td>1N5236B</td>
<td>+7.5V zener diode</td>
<td>2</td>
<td>0.21</td>
<td>0.42</td>
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</table>

**SUBTOTAL** $222.73
## Section C.3 N Switch Topology Converter / Current Control / Switching

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<th>Part No.</th>
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<th>Unit Cost</th>
<th>Total Cost</th>
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<td>45P80-1</td>
<td>prototyping board</td>
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<td>22.75</td>
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<td>MKDS 3/2</td>
<td>2 connection terminal block</td>
<td>3</td>
<td>1.35</td>
<td>4.05</td>
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<td>22-23-2041</td>
<td>4 pin locking header connection</td>
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<td>3 pin locking header connection</td>
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<td>22-23-2021</td>
<td>2 pin locking header connection</td>
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<tr>
<td>123-93-314-41-001</td>
<td>14 pin IC socket</td>
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<td>1.60</td>
<td>4.80</td>
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<tr>
<td>123-93-316-41-001</td>
<td>16 pin IC socket</td>
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<td>1.83</td>
<td>5.49</td>
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<td>KBL04-ND</td>
<td>H-bridge rectifier</td>
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<tr>
<td>UVZ2W330MHD</td>
<td>33μF 450V electrolytic cap</td>
<td>4</td>
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<td>20J10K</td>
<td>10kΩ 10W power resistor</td>
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<td>2.88</td>
<td>5.76</td>
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<td>SSL-LX100133XGC</td>
<td>red LED</td>
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<td>0.72</td>
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<tr>
<td>ERZ-V05D241</td>
<td>240V 600A surge absorber</td>
<td>3</td>
<td>0.25</td>
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<td>FR606-T</td>
<td>rectifying diode</td>
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<td>0.92</td>
<td>1.84</td>
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<td>IRG4PC40U</td>
<td>IGBT</td>
<td>2</td>
<td>4.98</td>
<td>9.96</td>
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<tr>
<td>40EPF04</td>
<td>fast recovery switching diode</td>
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<td>3.71</td>
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<tr>
<td>LA 25-NP</td>
<td>LEM current sensing transducer</td>
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<td>HCPL-3150</td>
<td>opto-isolating gate driver</td>
<td>2</td>
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<td>LM324AN</td>
<td>quad op amp IC</td>
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<tr>
<td>LM3524DN</td>
<td>PWM generator</td>
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<tr>
<td>MAX352CPE</td>
<td>quad analog switch IC</td>
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<tr>
<td>3362P-1-203</td>
<td>20kΩ potentiometer</td>
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<td>3362P-1-103</td>
<td>10kΩ potentiometer</td>
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<tr>
<td>3362P-1-101</td>
<td>100Ω potentiometer</td>
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<td>0.83</td>
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<tr>
<td>ECU-S1H104MEA</td>
<td>.1µF 35V monolithic capacitor</td>
<td>6</td>
<td>0.43</td>
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<tr>
<td>ECU-S1H103MEA</td>
<td>.01µF 35V monolithic capacitor</td>
<td>2</td>
<td>0.32</td>
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<td>0.22</td>
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<tr>
<td>MFR-25FBF-11K0</td>
<td>11kΩ resistor (1%)</td>
<td>2</td>
<td>0.11</td>
<td>0.22</td>
</tr>
<tr>
<td>MFR-25FBF-10K0</td>
<td>10kΩ resistor (1%)</td>
<td>6</td>
<td>0.11</td>
<td>0.66</td>
</tr>
<tr>
<td>MFR-25FBF-4K99</td>
<td>4.99kΩ resistor (1%)</td>
<td>25</td>
<td>0.11</td>
<td>2.75</td>
</tr>
<tr>
<td>MFR-25FBF-560R</td>
<td>560Ω resistor (1%)</td>
<td>4</td>
<td>0.11</td>
<td>0.44</td>
</tr>
<tr>
<td>MFR-25FBF-160R</td>
<td>160Ω resistor (1%)</td>
<td>4</td>
<td>0.11</td>
<td>0.44</td>
</tr>
<tr>
<td>MFR-25FBF-100R</td>
<td>100Ω resistor (1%)</td>
<td>3</td>
<td>0.11</td>
<td>0.33</td>
</tr>
<tr>
<td>MFR-25FBF-10R0</td>
<td>10Ω resistor (1%)</td>
<td>2</td>
<td>0.11</td>
<td>0.22</td>
</tr>
<tr>
<td>1N5236B</td>
<td>+7.5V zener diode</td>
<td>2</td>
<td>0.21</td>
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**SUBTOTAL** $173.15
### Section C.4 Wireless / User Controls

<table>
<thead>
<tr>
<th>Part No.</th>
<th>Description</th>
<th>No. Required</th>
<th>Unit Cost</th>
<th>Total Cost</th>
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<tbody>
<tr>
<td>TWS-434A</td>
<td>Transmitter</td>
<td>1</td>
<td>8.50</td>
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<tr>
<td>RWS-434A</td>
<td>Receiver</td>
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<tr>
<td>TWS-ANT</td>
<td>50Ω Whip Antenna</td>
<td>2</td>
<td>9.00</td>
<td>18.00</td>
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<tr>
<td>HT-12E</td>
<td>Holtek Encoder</td>
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<td>1.70</td>
<td>1.70</td>
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<tr>
<td>HT-12D</td>
<td>Holtek Decoder</td>
<td>1</td>
<td>1.70</td>
<td>1.70</td>
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<tr>
<td>MFR-25FBF-33K0</td>
<td>33kΩ resistor</td>
<td>1</td>
<td>0.11</td>
<td>0.11</td>
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<tr>
<td>MFR-25FBF-750R</td>
<td>750Ω resistor</td>
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<tr>
<td>100SP1T1B1M1QE</td>
<td>Switch</td>
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<td>TP11SHZQE</td>
<td>Pushbutton</td>
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<td>3.96</td>
<td>7.92</td>
</tr>
<tr>
<td>123-93-318-41-001</td>
<td>18 pin IC socket</td>
<td>2</td>
<td>2.06</td>
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</table>
References:


Justin Morse was born on June 18, 1974 in Champaign, Illinois to Wayne and Linda Morse. He received his high school diploma from Potsdam High School in Potsdam, New York in June of 1993. Upon completion of his high school education, he continued his studies at Clarkson University as an undergraduate student in the Department of Mechanical and Aeronautical Engineering. He received his Bachelor’s of Science degree in Mechanical Engineering from Clarkson in May of 1996. After graduating he accepted a position as a mechanical design engineer with Lockheed-Martin’s “Skunk Works” facility in Palmdale, California. In 1999, he left Lockheed-Martin to pursue his master’s degree in Electrical Engineering at Virginia Polytechnic Institute and State University. After completing his master’s degree, he intends to pursue a position in industry performing research and development of electromechanical systems and controls. His professional interests include: electromechanical systems, controls, energy efficiency, and alternative energy sources.