VIRTUAL MOVING AIR GAP
FOR THE SPEED RANGE IMPROVEMENT
OF A DUAL STATOR AXIAL FLUX MOTOR

A Thesis
Presented to
The Graduate Faculty of The University of Akron
In Partial Fulfillment
of the Requirements for the Degree
Master of Science

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ABSTRACT

The Virtual Moving Air Gap (VMAG) is a concept to improve the speed range of a dual stator axial flux permanent magnet synchronous motor (PMSM). The VMAG concept utilizes a dual stator axial flux motor where the stators are wired in series. As the speed increases the controller activates switches that remove one of the stators from the circuit thereby reducing the back EMF of the motor and allowing the motor to continue accelerating to a much higher speed than was obtainable with the original two stator machine. This concept can be employed in cases where both air gaps are equal as well as cases where the air gaps are unequal. In the unequal air gap case, the stator with the shorter air gap is removed from the circuit during acceleration. The VMAG concept can increase the no load speed by 100% in the case of equal air gaps and by a much greater amount in the case of unequal air gaps.
DEDICATION

Dedicated to my wonderful wife Paula and our two fantastic sons Brad Jr. and Frank.
I wish to thank Dr. Iqbal Husain and Dr. Yilmaz Sozer for their support as my academic advisors for this project. Being successful is easy when one surrounds themselves with great people. Their assistance and wisdom were a big part of my success on this project and with my coursework as well. I would also like to thank Dr. Malik Elbuluk who has always been a great source of inspiration to me as an engineer from the days when I was in his class as an undergraduate in 1991. He was also a very big help in my transition back to graduate school after so many years away from college. Dr. Tom Hartley and Dr. Robert Veillette were also a tremendous help to me in my pursuit of this degree. Thank you to Mrs. Gay Boden who managed to keep me on track throughout the entire program.

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<tr>
<td>$T$</td>
<td>Torque</td>
</tr>
<tr>
<td>$\pi$</td>
<td>Pi (3.14)</td>
</tr>
<tr>
<td>$r$</td>
<td>radius</td>
</tr>
<tr>
<td>$L$</td>
<td>Length of rotor</td>
</tr>
<tr>
<td>$\sigma$</td>
<td>Air gap shear stress</td>
</tr>
<tr>
<td>$B_g$</td>
<td>Air gap flux density</td>
</tr>
<tr>
<td>$r_a$</td>
<td>Average air gap radius</td>
</tr>
<tr>
<td>$r_o$</td>
<td>Outside air gap radius</td>
</tr>
<tr>
<td>$r_i$</td>
<td>Inside air gap radius</td>
</tr>
<tr>
<td>$K_e$</td>
<td>Back EMF constant</td>
</tr>
<tr>
<td>$K_t$</td>
<td>Torque constant</td>
</tr>
<tr>
<td>$\theta$</td>
<td>Commutation angle</td>
</tr>
<tr>
<td>$P$</td>
<td>Number of motor poles</td>
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\( \lambda_f, \lambda_{FD} \)  
Rotor (field) flux

\( L_d \)  
d axis inductance

\( L_q \)  
q axis inductance

\( i_d \)  
d axis current

\( i_q \)  
q axis current

\( \phi_g \)  
Air gap flux

\( \phi \)  
Magnet flux

\( \phi_r \)  
Remnant flux

\( K_l \)  
Leakage factor

\( K_r \)  
Reluctance factor

\( \mu_r \)  
Relative permeability

\( g \)  
Air gap length

\( A_g \)  
Air gap area

\( A_m \)  
Magnet area

\( l_m \)  
Magnet length

\( B_g \)  
Air gap flux density

\( N \)  
Conductors per slot
$L_{st}$  
Axial length of the motor

$R_{ro}$  
Air gap radius at the surface of the magnets

$\omega_{nl}$  
No load speed

$K_{e,original}$  
Original Back EMF constant of motor without VMAG

$K_{e, stator1}$  
Back EMF constant of Stator1

$K_{e, stator2}$  
Back EMF constant of Stator2

$\alpha$  
Percentage of original BEMF

$f_q$  
$q$ axis function variable

$f_d$  
$d$ axis function variable

$f_0$  
$0$ variable

$f_a$  
a phase variable

$f_b$  
b phase variable

$f_c$  
c phase variable

$V_d$  
d axis voltage

$V_q$  
$q$ axis voltage

$R_s$  
Stator resistance

xx
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<tr>
<td>$\omega_e$</td>
<td>Electrical frequency</td>
</tr>
<tr>
<td>$\lambda_d$</td>
<td>d axis flux</td>
</tr>
<tr>
<td>$\lambda_q$</td>
<td>q axis flux</td>
</tr>
<tr>
<td>$L_s$</td>
<td>Stator inductance</td>
</tr>
<tr>
<td>$V_{L-LRMS}$</td>
<td>Motor line-to-line voltage in Volts RMS</td>
</tr>
<tr>
<td>$\omega_r$</td>
<td>Rotor speed</td>
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<tr>
<td>$\psi$</td>
<td>Phase angle between voltage and current</td>
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CHAPTER I
INTRODUCTION

1.1 Background

Permanent magnet synchronous machines (PMSM) or brushless DC (BLDC) machines tend to have very high power and torque densities relative to other electric motor types. This makes them well suited for many industrial and commercial applications such as Electric Vehicles (EV) and Hybrid Electric Vehicles (HEV). The EV/HEV application requires not only a high power and torque density, but a wide speed range of operation. This is the area where the PMSM and BLDC motor tend to fall short. If they are designed with an optimization for high torque, then the back EMF (BEMF) constant is high and the speed that the motor can achieve on a fixed DC bus is limited. Likewise, if the motor is designed to operate at high speed, then the torque constant will be low and the motor will not produce much torque. What is required for the EV/HEV (and many other applications) is a motor with a wide speed range. The motor needs to be able to produce high torque at low speeds for vehicle launch and acceleration while being able to also achieve high speeds (at a lower torque) for highway driving.

The axial flux permanent magnet synchronous machine (AFPMSM) is a topology that can yield very good torque densities due to the large surface area of the air gap. This
torque density advantage is maximized at large diameters with short axial lengths [1] and high pole counts [1][15]. Although its physical geometry is quite different from the typical radial flux machine, the electrical characteristics of an AFPMSM are essentially the same as the electrical characteristics of the radial flux PMSM. AFPMSM suffer from the same speed range issues as any other type of PMSM or BLDC motor. The axial flux motor can be designed and constructed with various topologies, such as 1 stator/1 rotor, 2 stators/1 rotor, 1 stator/2 rotors, and multiple stators and rotors.

We have studied the 2 stators/1 rotor concept. With this topology, one of the stators can be selectively “turned off” or removed from the circuit when the BEMF begins to get high due to increasing speeds. Switching one of the stators off above a certain speed would reduce the BEMF constant of the motor and allow for the motor to achieve a higher speed without altering the available bus voltage.

1.2 History of Axial Flux Motors vs. Radial Flux Motors

Many of the earliest motors were axial flux motors, beginning with the Faraday Disc Motor invented by Michael Faraday in 1831 (Figure 1.1), followed by several others in 1832-1834. The radial flux machine which is today’s most common motor configuration was not invented until 1837 by T. Davenport [3].
The radial flux machine was easier to commercialize due to a variety of manufacturing challenges with axial flux machines [3]. For this reason, axial flux machines were largely forgotten. With the advent of rare earth magnets in recent years, many novel motor concepts are receiving much attention, including some that have axial flux structures [3].

1.3 The Brushless Motor

The brushless motor comes in several different forms such as PMSM with a sinusoidal BEMF or BLDC with a square wave or trapezoidal wave BEMF. The PMSM is sometimes referred to as a Brushless AC (BLAC) motor. The brushless motor in any of the above forms can be made with the radial flux or axial flux geometries. The basic electrical machine characteristics are the same whether radial flux or axial flux
geometries are used. In any form, the brushless motor has a rotor containing permanent magnet material and a stator containing electrical coils. An electronic motor controller is required to switch the stator currents at the appropriate times in relation to the rotor position so that the flux induced by the stator currents interacts with the rotor flux to create torque. A rotor position measurement is necessary to properly commutate the phase excitations.

1.4 The Axial Flux Motor

While the radial flux motor has flux lines that flow across the air gap in a radial direction (outward from the center of the motor shaft) as shown in Figure 1.2, the axial flux motor is so named because the flux flows across the air gap in the axial direction or parallel to the motor shaft. Figure 1.3 shows the axial flux motor geometry. The torque developed by an axial flux machine is proportional to the cube of the radius, while radial flux motor torque is a function of the square of the radius. This gives axial flux motors great torque advantage in large diameters.

Figure 1.2 Radial Flux Motor Geometry
1.5 Speed Range Limitations

Whether radial flux or axial flux, all brushless motors suffer from the same speed range limitations. When the motor design is optimized to produce high torque, the speed that the motor can achieve is limited. This is due to the higher BEMF that is generated with a high torque motor. In the EV or HEV applications, if the vehicle has a lot of torque for launch and acceleration, then it may not be able to reach highway speeds.

1.6 Motivation for This Research

The EV and HEV markets along with many other industries require a motor or motor control concept that allows for the extension of the usable speed range of the highly power dense brushless motor. The lack of a wide speed range for the brushless motor is
one of the primary technology barriers for the growth of the EV/HEV industry. The
permanent magnet brushless motor provides the power density needed, but not the speed
range. In order to provide the required torque density for vehicle launch, motors need to
be wound such that they can reach their base speed at a speed that is too low for highway
use. Many solutions exist for extending the speed range of the PMSM [2][6]. Most of
these solutions to address this problem are costly and tend to negate the advantages of
brushless motors by reducing their efficiency or their power density. The flux weakening
techniques that are used above base speed reduce the drive efficiency significantly.
Placing a gearing mechanism between the electric motor and the wheels is another option
but that requires additional cost and maintenance. One of the goals of the EV is to
eliminate this gearing that is used on internal combustion engine vehicles today. This is
the root of the need for a wide speed range. This thesis proposes using a unique Virtual
Moving Air Gap (VMAG) technique to increase the speed range of the AFPMSM motor
used in EV/HEV.

1.7 The Virtual Moving Air Gap Concept Overview

Axial flux motors give an added ability to modify the motor’s BEMF characteristics
during operation. In this thesis, the VMAG concept will be developed. This concept
involves the use of a two stator, one rotor axial flux motor configuration wherein one of
the stators is selectively bypassed at higher speeds to reduce the BEMF constant of the
motor and extend the achievable speed range. The VMAG concept allows a speed range
improvement of at least 100% and provides a much more effective solution than flux weakening since VMAG does not reduce efficiency the way that flux weakening does.

1.8 Thesis Organization

This thesis begins with an introduction to both radial flux and axial flux brushless motor technologies along with the strengths and shortcomings of brushless motors.

Chapter II discusses the engineering principles and details behind the radial flux and axial flux motor technologies. Design equations are presented that show the weaknesses in flux weakening that will later be addressed by this thesis.

Chapter III shows the theoretical analysis that defines the VMAG concept.

Chapter IV presents the modeling and simulation of the VMAG concept applied to an actual motor.

Chapter V provides the details of the experimental setup and the test results.

Chapter VI discusses the results of the testing as compared to the simulations and the theoretical analysis. From this, conclusions are presented and discussed.
2.1 The Axial Flux Advantage

The axial flux motor can have advantages over the more popular radial flux machine. The primary advantage of axial flux can be seen by looking at how torque is produced in an electric motor. Torque in any electric machine is produced by the following process. The interaction of the stator magnetic field and the rotor magnetic field causes shear stress in the air gap. This shear stress in the air gap is multiplied by the air gap area to produce a force in the air gap acting on the rotor. This air gap force when multiplied by the radius of the air gap produces a torque on the rotor. For a radial flux machine, the torque developed [7] is given as

\[ T = 2\pi r^2 L \sigma, \]  

(2.1)

where \( \sigma \) is the air gap shear stress, \( L \) is the active length of the rotor, \( 2\pi r \) is the air gap area, and \( r \) is the air gap radius (moment arm).

The air gap shear stress is given as the product of the electrical and magnetic loadings [1]. Air gap shear stress is shown by

\[ \sigma = B_g A, \]  

(2.2)
where $B_g$ is the air gap flux density and $A$ is the electric loading (A/m).

The equation (2.1) shows that torque is a product of the magnetic force generated in the air gap and the radius of the air gap (length of moment arm). The force in the air gap is equal to the air gap shear stress multiplied by the air gap area. As a result of these two equations we can say that the torque is proportional to the square of the motor’s radius.

The following analysis adapts equation (2.1) to axial flux geometry. In axial flux motors, the air gap is a planar surface forming an annular ring with an inside radius $r_i$ and an outside radius $r_o$ as shown in Figure 2.1. The area of the air gap is $\pi(r_o - r_i)^2$. The force generated in the air gap again is the air gap area times the air gap shear stress and is given as $\pi(r_o - r_i)^2 \sigma$. The air gap force acts at the average radius of the air gap, which is given as $r_a = \frac{r_o + r_i}{2}$. The resulting torque becomes a function of radius cubed shown as

$$T = \frac{\pi(r_o - r_i)^2(r_o + r_i)\sigma}{2}. \quad (2.3)$$

This is where the axial flux geometry becomes an advantage. When the axial flux radius is increased, the torque production increases dramatically.

The AFPMSM is a topology that can yield very good torque densities due to the large surface area of the air gap. In [15] the sizing equations of (2.1) and (2.2) are adapted to axial flux geometry and expanded to show the affecting factors such as electric loading, pole count, pole arc factor, speed, stator core length in the machine design. These sizing equations are then adapted for radial flux and axial flux machines and used
to compare the torque densities of each machine type. The effects on torque density of the factors listed above are then summarized. The findings of [15] are as follows:

- Axial flux motors in all cases examined have better torque density than radial flux machines.

- The torque density advantage of axial flux motors increases as pole count increases.

- Much of the advantage of axial flux over radial flux is the result of the radial flux rotor volume not being fully utilized. Much of the radial flux rotor volume does not contain a high flux density.

Figure 2.1 Axial Flux Motor Geometry
A thermal analysis is conducted in [1] to compare the torque densities of the axial flux motor and the radial flux motor. The thermal analysis considers the ratio of the fixed losses to the thermal wasting surface. The effects of pole count and motor dimensions are examined in this analysis. The main findings of [1] are as follows:

- The advantage of axial flux motors over radial flux motors is greater when pole count is higher.
- Axial flux has a greater advantage over radial flux when the ratio of motor length/diameter is less than 0.3.

Both [1] and [15] arrive at the same conclusions. The torque density advantage is maximized at large diameters with short axial lengths [1] and high pole counts [1][15]. Further analysis of the torque density advantages of axial flux can be reviewed in [14], [13]. All research reviewed shows the same basic results: Axial flux motors can have a significant advantage in torque density over radial flux motors.

Although the physical geometry of an axial flux machine is quite different from that of the typical radial flux machine, the electrical characteristics are essentially the same. Axial flux brushless PM Motors suffer from the same speed range issues as any other type of PMSM or BLDC motor.
The Speed Range Issue with Permanent Magnet Brushless Motors

The permanent magnet brushless motor (PMSM or BLDC motor) creates torque through the interaction between the permanent magnet field of the rotor and the electromagnetic field of the stator. Other types of motors may use electromagnets to create the rotor field. The use of permanent magnets for the rotor field gives the brushless motor much greater torque density and power density than a motor that utilizes electromagnets to create the rotor field. This is especially true when the permanent magnets used are rare earth magnets like Samarium Cobalt (SmCo) or Neodymium Iron Boron (NdFeB). These high energy magnets allow motors to have extremely high flux densities in the air gap without using electrical current to create that flux. The result is a motor with a very high torque constant and a lower mass. The motor’s torque constant is the amount of torque the motor will produce for a given amount of current. It is most often expressed in SI units of Newton Meters / Ampere (Nm/A). Permanent magnet brushless motors tend to have high torque constants when compared to comparably sized motors with electrically generated fields.

The brushless motor also has another constant called the Back EMF constant or BEMF constant. This is the amount of Back EMF (voltage) that is generated by the motor for a given motor speed. A higher BEMF constant equates to a higher voltage generated for a given speed. The motor’s BEMF will act as a speed limiter. As the motor spins faster, the BEMF increases and approaches the voltage of the DC bus that is supplying power to the inverter. When the difference between the DC bus voltage and
the BEMF is very low, then the current cannot flow into the motor and the motor cannot
spin any faster.

While a high torque constant is desirable, a high BEMF constant is not. Unfortunately, the torque constant and the BEMF constant are the same constant with the proper SI units [4]; that is,

\[ K_e = K_t, \]  

(2.4)

where \( K_e \) is the BEMF constant and \( K_t \) is the torque constant.

If \( K_e \) is expressed in Volts/RPM it must be converted to Volts/(rad/sec). \( K_t \) must be expressed in units of Nm/A. If the value of \( K_e \) is measured line-to-line on a three phase motor, a \( \sqrt{3} \) correction must be done to get \( K_t \) since current is measured per line (per phase). If the proper units are not used, then the values for \( K_e \) and \( K_t \) are not the same.

It is desirable to design the motor so that the \( K_t \) is high. This gives a large amount of torque for a small amount of current. Unfortunately, if the \( K_t \) is high, then the \( K_e \) is also high. This would result in a motor that can produce a lot of torque, but cannot spin very fast. If \( K_e \) is reduced to allow higher speed operation, then the motor will not produce as much torque.

It would be desirable to have a motor that can produce high torque at lower speeds, and still achieve a higher speed range. Presently, the primary way to accomplish this objective is to use a motor that has a greater power rating than is needed in order to get both high torque and high speed. The over powered motor is often also larger and
heavier than what is desired. These tradeoffs are especially an issue for some applications such as EV and HEV. Several other techniques for increasing speed range presently exist as well and are discussed in the next section.

2.3 Prior Art for Increasing the Speed Range of Radial Flux Brushless Motors

There are several methods commonly used for increasing the speed range of a PMSM as well as many novel concepts. Many concepts for increasing the speed range of a PM brushless motor involve reducing the flux linkage between the rotor and the stator through either mechanical or electromagnetic manipulation. The most common method is known by several different names as “field weakening,” “flux weakening” and “angle advance.”

In field oriented control, the stator currents are timed in relation to the rotor position such that there is always a 90 degree phase angle between the stator flux and the rotor flux. This yields a maximum torque production. Figure 2.2 shows the relationship between the stationary abc reference frame and the rotating dq reference frame. The abc reference frame is the reference frame traditionally used to describe the stator current as the sum of the a phase, b phase and c phase component vectors. Each of the phase axes is displaced by 120° in phasor domain. The current vector \( i_s \) is the resultant of the abc component vectors. This vector representing instantaneous stator current is a vector in two dimensional space but expressed as the sum of three component vectors separated by 120°. If the abc reference frame is translated into an alternate two dimensional reference
frame having a d and q axis (direct and quadrature) then the stator current vector can be
defined as the sum of two orthogonal component vectors $i_d$ and $i_q$. With additional phase
transformation the dq reference frame is allowed to rotate synchronously with the rotor
and the d axis is aligned with the rotor flux. In this case, the q axis current $i_q$ produces
motor torque. The d axis current produces stator flux that either adds to or subtracts from
the rotor flux. For maximum torque production the d axis current should be zero.

If the phase angle is advanced, as shown in Figure 2.3 the angle between $i_s$ and the
d axis is no longer 90 degrees. The stator current vector $i_s$ has a component on the q axis
and a component on the d axis. This d axis component of the stator current creates flux
that opposes the flux of the rotor and therefore weakens it. This has the effect of
reducing the BEMF constant and allowing the motor to spin faster. The penalty of this
method is having lower efficiency in the motor. The d axis current that is used to weaken
the flux causes both copper loss and iron loss in the motor, but does not produce torque.
Using this method often results in severe heating in the motor. Additionally, the large d
axis currents can in some cases permanently demagnetize the rotor. A more in depth
discussion of field oriented control is given in Section 4.1.

The shortcoming of the flux weakening method is primarily due to its limited
effectiveness on a rotor with surface permanent magnets (PM). For rotors with surface
PMs, the d and q axis inductances are nearly the same. This limits the effects of flux
weakening. This can be seen by looking at the equation for torque as a function of $L_d$, $L_q$,
$i_d$, $i_q$ given in [5] as
\[ T = \frac{3}{2}P \left[ \lambda_f i_q + (L_d - L_q)i_d i_q \right], \tag{2.5} \]

where \( P \) is the number of poles, \( \lambda_f \) is the rotor flux, \( L_d \) and \( L_q \) are the d and q axis inductances while \( i_d \) and \( i_q \) are the d and q axis currents.

As shown in equation (2.5), when the q axis inductance and the d axis inductance are equal, the d axis current does not influence torque production, and therefore does not affect the BEMF. If the values of q axis and d axis inductance are nearly the same, then the effect of adding d axis current is minimal. In order to improve the flux weakening operation, the rotor can be manufactured with the PM material buried in the rotor beneath magnetic steel. This introduces saliency in the rotor resulting in different values for q.
and d axis inductances. When a buried magnet rotor is used, the effect of flux weakening is much greater but the manufacturing cost is increased [5].

Another concept for increasing the speed range of a PM brushless motor is a proprietary solution by a company called Variable Torque Motors. Their concept is a simple one by which the normal radial flux motor design is modified to allow the rotor to slide on the rotor shaft in the axial direction. At low speeds, the rotor is slid into position on the rotor shaft so that the PM material is fully inside of the active portion of the stator. This gives the maximum flux linkage and therefore the greatest torque constant. As the speed of the motor increases, the rotor is slid axially such that some of the PM material is outside of the active region of the stator. This reduces the flux linkage, and therefore reduces the BEMF constant allowing for higher speed operation. The rotor can slide until all of the PM material lies outside of the active portion of the stator. This allows the motor to be spun loss free. Their system is powered by ultracapacitors and is used primarily for vehicle applications with a lot of start and stop events. This allows for the motor to provide launch assist and energy recovery through regenerative braking without any efficiency degradation at highway speeds. The graphic of Figure 2.4 shows this concept.

2.4 Prior Art for Increasing the Speed Range of Axial Flux Brushless Motors

Similar to the VTM solution presented above is a concept for axial flux motors developed and patented by Dr. Dantam K. Rao of Precision Magnetics Inc. (Premag)
[12] and shown in Figure 2.5. What is unique about this particular axial flux motor is that the PMs on the rotor are able to move in the radial direction. At low speed, they are in position such that the magnet material is aligned with the active portion of the stator core. As the speed increases the magnets are allowed to move outward in the radial direction. In this condition, some of the magnet material falls outside of the active region of the stator thereby reducing the BEMF constant of the motor. This concept is shown in Figure 2.4 and described in detail in [12].

Figure 2.4 Variable Torque Motors Concept
Both of the previous two examples are very similar in advantages. They allow the BEMF constant of the motor to change as speed increases without a loss of efficiency. They do however have a drawback. They both require a significantly increased volume to allow for the space to accommodate the motion of the magnet material. In the case of the Variable Torque Motor concept, the motor housing must be nearly twice the length of the active portion that actually produces torque. This is only a shortcoming for certain applications. The application where VTM promotes this technology is for the replacement of a bus, delivery truck or van driveshaft with their motor. In this case, a
relatively long slender motor is well suited for replacing the driveshaft of the vehicle. In
the case of the Dr. Rao concept, the motor housing must have an increased outer diameter
beyond the active portion of the motor. This increases the overall motor diameter while
not increasing the active torque producing portion of the motor. Since, in an axial flux
motor, torque output increases proportional to the cube of the change in diameter, adding
diameter that does not produce torque is a significant drawback.

Due to the planar nature of the air gap, some unique opportunities exist in axial flux
motors for mechanical flux weakening. One such method has been employed very
successfully in motors for solar cars. In this method, a company called NuGen Mobility
(formerly New Generation Motors) uses a single rotor single stator configuration of the
axial flux motor. They utilize a mechanical actuator to physically move the rotor axially
away from the stator as speed increases. As the air gap length is increased, the flux
linkage is reduced, and in turn, the BEMF constant is reduced. This allows the motor to
have high torque for launch and acceleration, while being able to achieve highway
speeds. In [4] the equation is given for a radial flux motor as

\[
\phi_g = K_r \phi = \frac{K_r}{1 + K_r \frac{\mu_0 g A_m}{l_m A_g}} \phi_r .
\]

This shows the relationship between air gap length \(g\) and the air gap flux as derived from
a magnetic circuit model analysis. As \(g\) increases, the air gap flux \(\phi_g\) decreases
proportionally. The same type of relationship between air gap length and air gap flux
exists in axial flux motors. The variable air gap or “moving air gap” concept of NewGen
Mobility inspired the concept that became known as “Virtual Moving Air Gap.”
3.1 Basis of the Invention

The Virtual Moving Air Gap (VMAG) is named such because the effects of it are similar to those of the moving air gap described at the end of Chapter II. In this case however, the air gap does not actually change. The VMAG concept utilizes the two stators and one rotor configuration of the axial flux motor. In this configuration, there is one rotor located between two stators resulting in two air gaps in the motor. In the dual stator axial flux motor design, the air gaps are typically identical, as this balances the forces of magnetic attraction between the rotor’s magnet material and the stator core material.

The VMAG concept can be applied to the typical motor with equal air gaps, but greater benefit is realized if unequal air gaps are employed. In the two stator configuration, the stators can be wired in series or in parallel. If the stators are in parallel, the air gaps must not be equal. Also, if the stators are in parallel, the benefit is not as great as if they are in series. This thesis will focus on the wye connected series implementation. The VMAG concept is covered under US Patent 7,888,904 [9]. Figure
3.1 shows the equivalent circuit for the wye connected three phase dual stator axial flux motor.

In the series implementation, current flows in series through one stator coil into the other one for each phase. The neutral is buried in the motor and not connected outside of the motor. Each phase coil from the first stator is in series with the corresponding coil of the second stator. The equivalent circuit in Figure 3.1 shows that the BEMF of the first stator adds with the BEMF of the second stator to comprise the total BEMF of each phase of the motor. If the air gaps are equal, the flux linkage between the PMs of the rotor and the stator is also equal. This means that at any given speed, the BEMF of each stator will be the same as the BEMF of the other stator. If the air gaps for the two stators are equal,
different then the flux linkages as well as the BEMFs would be different. The stator with
the longer air gap will have a lower BEMF than the stator with the shorter air gap. In
equation (2.6) it was shown that the air gap flux $\phi_g$ was approximately inversely
proportional to the air gap length $g$. The relationship between air gap flux and air gap
flux density is

$$\phi_g = B_g A_g,$$  \hspace{1cm} (3.1)

where $\phi_g$ and $B_g$ are directly proportional.

The dependency in a brushless motor between BEMF constant $K_c$ and air gap flux
density $B_g$ is given in [4] as

$$K_c = K_t = 2NB_g L_{st} R_{ro},$$  \hspace{1cm} (3.2)

where $N$ is the conductors per slot, $B_g$ is the air gap flux density, $L_{st}$ is the axial length of
the motor and $R_{ro}$ is the air gap radius at the surface of the magnets.

Equations (2.6), (3.1) and (3.2) together show that the BEMF constant and torque
constant are approximately inversely proportional to the air gap length. The BEMF
constant and torque constant are directly dependent on air gap flux density $B_g$ which is
inversely proportional to air gap length [4].

Figure 3.2 shows the VMAG circuit. The switches S1 and S2 can be selectively
switched such that Stator1 is bypassed leaving Stator2 as all that remains in the circuit.
This will have the effect of reducing the BEMF constant. If the air gaps are equal, then
the final BEMF constant after Stator1 is bypassed will be half of the full motor BEMF
constant. If the air gaps are not equal, the stator designated as Stator1 should be the stator with the shorter air gap. This will leave the stator with the lower BEMF constant in the circuit after the VMAG bypass is activated. In this case, the BEMF constant after Stator1 is bypassed will be less than half of the full motor BEMF constant. Figure 3.3 shows the VMAG circuit after VMAG bypass is activated. Notice that the BEMF source of Stator1 is removed from the circuit, leaving only Stator2 to generate BEMF. Stator2 is also the only stator that can produce torque once Stator1 is removed from the circuit as in Figure 3.3. A drawing of the stators with unequal air gaps is given in Figure 3.4 where g1 and g2 are the two air gaps. In Figure 3.4, the stator on the left with air gap g1 (Stator1) is the one to be removed from the circuit when the VMAG switching occurs.
Figure 3.3 Axial Flux Motor Equivalent Circuit with VMAG Engaged

Figure 3.4 Dual Stators with Unequal Air Gaps
3.2 The Effect of VMAG on \( K_e \) and on No Load Speed

Figure 3.5 shows a plot of BEMF vs. speed which depicts a switching strategy to employ VMAG. The slopes of the lines in this plot represent the BEMF constant \( K_e \). We can call the \( K_e \) of the original unaltered motor without VMAG \( K_{e\text{-original}} \). As shown in the equivalent circuit of Figure 3.2, each stator has a BEMF source with its own \( K_e \). For the original unaltered motor with equal air gaps, the \( K_e \) of each stator is half of the total \( K_e \) of the motor and shown as

\[
K_{e\text{-stator1}} = \frac{K_{e\text{-original}}}{2}, \quad (3.3)
\]

\[
K_{e\text{-stator2}} = K_{e\text{-stator1}}. \quad (3.4)
\]

Before VMAG switching occurs, the total \( K_e \) is given by

\[
K_{e\text{-original}} = K_{e\text{-stator1}} + K_{e\text{-stator2}}. \quad (3.5)
\]

After VMAG switching occurs, \( K_e \) can be shown as

\[
K_e = K_{e\text{-stator2}}. \quad (3.6)
\]

Increasing one of the air gaps (Stator2) causes the BEMF constant for that stator to be reduced to some percentage \( \alpha \) of the original as shown by

\[
K_{e\text{-stator2}} = \alpha K_{e\text{-stator1}}, \quad (3.7)
\]

where \( 0 < \alpha \leq 1 \).
With unequal air gaps, when both stators are active the new equation for $K_e$ is

$$K_e = (1 + \alpha)K_{e\_stator 1} = \frac{1 + \alpha}{2} K_{e\_original}. \quad (3.8)$$

After VMAG switching when only one stator is active, $K_e$ becomes

$$K_e = \alpha K_{e\_stator 1} = \frac{\alpha}{2} K_{e\_original}. \quad (3.9)$$

In order to have the value of $K_e$ become less than half of $K_{e\_original}$ the stator with the longer air gap must be Stator2, the stator used for single stator operation.

From equation (3.8) it can be seen that there will be some reduction of $K_e$ for low speed (two stator) operation. This unfortunately means that there will also be an equal loss of torque constant $K_t$ for low speed (two stator) operation. It is assumed that based on the value of $\alpha$ in equations (3.8) and (3.9), the benefits of speed range improvement by the reduction of $K_e$ will outweigh the negative effects of the loss of $K_t$ at low speeds.

Since no load speed is determined by the speed when the BEMF reaches the level of the DC bus that is supplying the motor, then reducing the BEMF by some amount should increase the no load speed by the inverse of that amount.

If the no load speed of a dual stator axial flux motor with equal air gaps before VMAG is called $\omega_{NL\_original}$ then the no load speed for a motor with VMAG (either equal or unequal air gaps) could be shown as

$$\omega_{NL} = \frac{2}{\alpha} \omega_{NL\_original}. \quad (3.10)$$
For a motor with equal air gaps, $\alpha = 1$ and the motor would be able to spin twice as fast as without VMAG. If unequal air gaps were utilized, then $\alpha < 1$ and the motor would be able to spin more than twice as fast as without VMAG.

### 3.3 Speed Dependent Switching

The VMAG concept is based on the switching event being speed dependent. When the motor speed is low, the maximum torque is desired from the motor. As the motor speed increases, the BEMF rises. As shown in Figure 3.5, as the BEMF approaches the operable limit of the DC bus voltage, the VMAG bypass is activated. After the switching action the BEMF constant of the motor drops significantly allowing the motor to continue accelerating. The slopes of the lines in Figure 3.5 represent the changing value of $K_e$ in the VMAG system. It is important to note that the torque constant also decreases when the bypass occurs, so in order to maintain the present torque output, current will have to increase at the same time. The field oriented controller that is controlling the motor commutation is based on modeled motor parameters. When the VMAG operation is activated, the motor parameter values in the motor controller must be changed to accurately reflect the new parameters of the motor with one stator bypassed. More details related to the effect on motor parameters will be given as part of the modeling and simulations in Chapter IV. While operating on one stator, if the motor decelerates to a sufficiently low speed, it will be desirable to switch back to two stator configuration. This will allow the motor to again have a high torque constant to facilitate the possibility for acceleration should it be required. In order to prevent oscillation between VMAG
active and inactive states (one and two stator operation), a hysteresis function is employed such that there is a difference between the speed when the VMAG is activated and the speed when it is deactivated. This hysteresis function is shown in Figure 3.5.

![Figure 3.5 VMAG Switching Speed with Hysteresis](image)

3.4 Driving the Switches with Pulse Width Modulation (PWM)

The original concept for VMAG was to drive the switches using a pulse width modulation (PWM) approach as shown in Figures 3.2. and 3.6. In this approach, the switches are assumed to be active semiconductor switches and are pulsed at a high frequency with PWM. The S1 and S2 switches work in complementary fashion. Initially at low speed, S1 would be 100% while S2 is 0%. This essentially leaves the motor in its original circuit configuration with both stators fully active. As the speed nears the point
when the BEMF is too high compared to the DC bus, the duty cycle of S1 is slowly increased as a function of motor speed while S2 is decreased to maintain a combined total between S1 and S2 of 100%. Eventually S1 will be at 0% and S2 will be at 100%. This will gradually decrease the BEMF constant of the motor from that of the dual stator motor to that of the single stator machine. This methodology would tend to yield a smooth transition from two stator operation to one stator operation. It is believed however that transient voltage spikes due to continuous switching of the phase currents would cause this concept to fail. Future research could be done to find a way to overcome this risk.

The focus of this thesis is to use a single discrete switching event to change between two stator and one stator operation. This will yield the instantaneous change in $K_e$ as shown in Figure 3.5. During the single switching event, switching could be done at the zero crossings of the stator currents to minimize voltage transients.

![Figure 3.6 PWM Operation of VMAG Switches](image-url)
3.5 Switch Type Selection

The VMAG circuit of Figure 3.2 is drawn with generic switch symbols. For this circuit, there are a variety of choices for switching devices. The most desirable switch would be a semiconductor switch such as an Insulated Gate Bipolar Transistor (IGBT) or a Metal Oxide Semiconductor Field Effect Transistor (MOSFET). This would allow for virtually infinite numbers of cycles without any degradation of the switch quality. Conversely, a metal dry contact switch such as an industrial contactor or motor starter would be very easy to implement, but would not have the robustness of the IGBT or MOSFET. The dry metal contact would be prone to degradation from arcing and would have a limited number of cycles before replacement is required. Additionally, the industrial contactor is much more bulky than the semiconductor switches. There is also a very loud audible click that occurs when the contactor changes its state.

The IGBT or MOSFET would seem to be the better choice. Unfortunately, large high current IGBTs are manufactured with an integral body diode for transient suppression. Attempts to find a commercially available high current IGBT without a body diode were unsuccessful. If an IGBT with a body diode were used, the circuit would be adversely affected by the presence of the diodes. Figures 3.7 and 3.8 show what the VMAG circuit would look like if IGBTs with integral body diodes were used. Figure 3.7 shows the condition when S1 is on and S2 is off. In this case, S2 electrically is replaced by the diode as shown. The S2 diode would short out Stator1 for half of each cycle. The AC BEMF generated by Stator1 with the S2 diode shorted across it would cause very high currents that would likely result in a catastrophic failure of the diode in
S2. In Figure 3.8, S1 is off and S2 is on. In this case, the diode in S1 would conduct for half of each cycle. During the half cycle when the diode is conducting, the diode would again complete a short circuit around Stator 1 resulting in excessive currents and causing failures of both S1 and S2. Even if S1 and S2 did not fail as a result of the excessive currents, the short circuit around Stator 1 would act as a brake during the BEMF half cycle when the diode was conducting.

Figure 3.7 IGBT / MOSFET with S1 On and S2 Off
Using a pair of N channel MOSFETs in series with one MOSFET reversed as shown in Figure 3.9 is also a possible switch solution. This would negate the issue of the free body diode acting as a rectifier as there would be 2 diodes in series, but one would be reversed. The diode pair blocks both halves of the AC cycle conduction unless both MOSFETs are on.

In the absence of an IGBT without body diodes, the dry metal contact is the switch that will be used for the VMAG circuit. In the future, there may be sufficient reason to
have custom made IGBTs without body diodes. Zero crossing switching may be sufficient to protect the IGBTs from failure due to voltage transients when the stators are turned off.

Figure 3.9 MOSFET Configuration for S1 and S2
CHAPTER IV

VMAG SIMULATION

4.1 Field Oriented Control

The VMAG concept requires a modified field oriented control (FOC) to function. FOC is often used for brushless PM motor applications. FOC is a control technique by which the three phase voltages and currents of the brushless PM motor are translated into a 2 phase system called d and q. The flux and torque could be controlled directly with d and q variables respectively.

The vector diagram of Figure 2.3 is the basis of the translation from the abc reference frame to the dq reference frame. From this vector diagram,

$$\begin{pmatrix}
    f_q \\
    f_d \\
    f_0
  \end{pmatrix} = \left( \begin{array}{c}
    \frac{2}{3} \\
    \sin(\theta) \\
    \frac{1}{2}
  \end{array} \right) \times \begin{pmatrix}
    \cos(\theta) & \cos(\theta - \frac{2\pi}{3}) & \cos(\theta + \frac{2\pi}{3}) \\
    \sin(\theta) & \sin(\theta - \frac{2\pi}{3}) & \sin(\theta + \frac{2\pi}{3}) \\
    \frac{1}{2} & \frac{1}{2} & \frac{1}{2}
  \end{pmatrix} \times \begin{pmatrix}
    f_a \\
    f_b \\
    f_c
  \end{pmatrix}$$

(4.1)

is derived to describe the translation from the abc reference frame to the dq reference frame [5][10][8].
Figure 4.1 presents the equivalent circuit representation of equation (4.1) [11]. A circuit analysis of the equivalent circuits of Figure 4.1 yields the differential equation relationships between the dq currents and the dq voltages [5] [11] as

\[ V_d = R_s i_d + \frac{d}{dt} \lambda_d - \omega_e \lambda_q, \]  
\( (4.2) \)

\[ V_q = R_s i_q + \frac{d}{dt} L_s i_q + \omega_e (\lambda_d + \lambda_{FD}), \]  
\( (4.3) \)

\[ \lambda_q = L_s i_q, \]  
\( (4.4) \)

\[ \lambda_d = L_s i_d, \]  
\( (4.5) \)
where $\lambda_q$ is the q axis stator flux, $\lambda_d$ is the d axis stator flux, $\lambda_{FD}$ is the rotor flux and $\omega_e$ is the electrical frequency of the stator current.

These differential equations are the basis for the motor model used in this simulation as well as for the controller model used in this simulation. These equations are also the basis for the FOC implemented in the experiment detailed in Chapter V.

4.2 The Modifications to Field Oriented Control

As described in Chapter III, the VMAG concept involves “turning off” or “bypassing” one of the two series stator coils in order to reduce the axial flux motor’s BEMF constant. When this occurs, the motor characteristics change. This change in motor parameters must be accounted by the motor controller. In FOC, motor parameters are used in the calculations to translate the variables from the abc reference frame into the dq reference frame. If these parameters used in the calculations are not representative of the real motor parameters, the FOC will not function properly.

The motor parameters used in the FOC are as follows:

- Stator resistance.
- Stator Inductance
- Rotor flux from the PM on the rotor.

When the VMAG system switches from two stators to one, the motor parameters listed above are all affected significantly. Stator resistance and inductance are reduced to
half of the value for two stators. The value of the rotor flux linkage is also reduced. If
the air gaps of the axial flux motor are not equal, and the stator with the smaller air gap is
removed, then the rotor flux will be reduced to less than 50% of the original value.
Changing the values of these parameters dynamically based on the VMAG state will be
required in order for the motor controller to function properly with VMAG.

4.3 The VMAG System Level Simulation

The Simulink model for the VMAG system is shown in Figure 4.2. The model
consists of the FOC or vector controller, the axial flux motor, the mechanical system
(load), the torque command profile, the VMAG switch and the scope to observe the
system output.

4.4 The Axial Flux Motor Simulation

The Simulink model for the VMAG system shown in Figure 4.2 includes several
subsystems. The axial flux motor is modeled as a basic PMSM in the dq rotor reference
frame. Figure 4.3 shows the axial flux motor model. The motor parameters modeled are
based on measurements taken from the actual motor to be used for this project. In the
motor model, the abc voltages that are applied to the motor are converted to voltages in
the dq reference frame according to the transformation in equation (4.1), as shown in
Figure 4.4. Based on the dq axis equivalent circuits shown in Figure 4.1 and described by
equations (4.2)-(4.5), the d and q axis currents are calculated as shown in Figure 4.5. In
Figure 4.6, the q axis current along with the rotor flux $\lambda_{FD}$ are used to calculate the torque developed by the motor as shown by

$$
T = \frac{3}{2} P \frac{1}{2} i_q \lambda_{FD},
$$

(4.6)

where $P$ is the number of motor poles and $i_q$ is the q axis current.

Finally, for the purpose of providing current feedback to the model of the motor controller, the d and q axis currents are converted to abc currents according to the inverse of the transformation in equation (4.1), as shown in Figure 4.7.
In the voltage to current model of Figure 4.5, the currents produced from the applied voltages are dependent on the motor’s stator inductance and stator resistance. When the VMAG is activated, one of the two stators will be bypassed. With one of the stators bypassed, the stator resistance and inductance values will be 50% of the values before the bypass. For this reason, the stator current model of Figure 4.5 has the values of resistance and inductance as input variables. In the system level model of Figure 4.2 the values of stator resistance and inductance come from a speed dependent selection in the function block labeled “VMAG Stator Switch”. This function will be discussed further in Section 4.7.

Figure 4.3 Axial Flux Motor Model
Figure 4.4 abc Voltages to dq Axis Voltages

Figure 4.5 dq Voltages to dq Currents
4.5 The Field Oriented Controller Simulation

The detail of the vector controller function block is shown in Figure 4.8. The vector controller model and the supporting equations are explained in detail in [5], [8], [10].
In the vector controller, a torque command reference \( T^* \) is the primary input to the current command function. This is shown in Figure 4.8 and in Figure 4.9. In Figure 4.9, the torque reference, \( T^* \) is used to derive the \( q \) axis command current \( I_{qs}^* \). The equation (4.6) is the inverse of this operation. In \( dq \) control of a PMSM, the \( q \) axis current is the component of the phase current responsible for producing torque. The \( d \) axis current is normally used to produce flux in an induction motor. In this case, the motor has a permanent magnet rotor so the \( d \) axis current \( I_{ds} \) should always be zero. The current command function sets the \( d \) axis current command \( I_{ds}^* \) to zero. Since the VMAG switching of the stators will cause the rotor flux \( \lambda_{FD} \) to change, \( I_{qs}^* \) needs to be adjusted accordingly to compensate this change.

Figure 4.8 Vector Controller
Figure 4.10 shows the block diagram of the current regulator which takes the current commands $I_{qs}^*$ and $I_{ds}^*$, actual phase currents $I_{qs}$ and $I_{ds}$, along with rotor speed $\omega_{\text{ref}}$, and produces necessary voltage commands $V_{qs}^*$ and $V_{ds}^*$ according to equations (4.2)-(4.5). Stator inductance $L_s$ and rotor flux $\lambda_{FD}$ are the parameters that can vary with the switching actions of VMAG.

Figure 4.11 shows the conversion block that takes voltage commands from the rotating dq reference frame to the stationary abc reference frame as determined from the inverse transformation of equation (4.1). The result is a voltage command for each of the three motor phases a,b,c.
Figure 4.10 dq Current Commands to dq Voltage Commands

Figure 4.11 V_{\text{qd}} to V_{\text{abc}} Command Function
The command voltage references are then fed into the function of Figure 4.12. This function derives three phase control signals from the three phase command reference signals. In the function block of Figure 4.13, the control signals that are output from Figure 4.12 will be compared to a triangular wave to determine the duty cycles of the three phase inverter switches. In Figure 4.12, the abc voltage command signals are scaled by a ratio of the triangular wave magnitude divided by the DC bus voltage and offset by the difference between the neutral voltage and the DC bus negative voltage.

In the duty cycle function generator of Figure 4.13, the control signals from Figure 4.12 are compared to a triangular wave function. When the control signal is greater than the triangular wave function, the duty cycle signal turns on. When the control signal is less than the triangular wave function, the duty cycle signal turns off.

Figure 4.12 $V_{abc}$ Command Reference to $V_{abc}$ Control Signals
The duty cycle signals are then fed into the inverter simulation function as shown in Figure 4.14. In this simulation block, the abc inverter output voltages are generated by multiplying the abc duty cycle signals by the DC bus voltage and offsetting the output by the difference between the neutral and the DC bus negative voltage. More details of the method of simulating the inverter as shown in Figures 4.12-4.14 are given in [8].

As shown in Figure 4.15, actual three phase currents are fed into the transformation block to convert the actual phase current into dq reference frame using the transformation of equation (4.1). These signals provide the current feedback to the vector controller.

Figure 4.13 $V_{abc}$ Control Signals to $V_{abc}$ Duty Cycles
Figure 4.14 Inverter Simulation

Figure 4.15 abc Currents to qd Currents Actual
4.6 The Mechanical System

The mechanical system model is shown in Figure 4.16. The mechanical system models the acceleration of the motor based on the net torque and the system inertia. The motor velocity is the output of this model. The net torque is the difference between the motor torque and the load torque. The electrical frequency $\omega_e$ of the motor resulting from the influence of the motor on the mechanical system is governed by the differential equation

$$\dot{\omega}_e = \frac{P}{2} \times \frac{T_{em} - T_{load}}{J},$$  \hspace{1cm} (4.7)

where $P$ is the number of motor poles, $T_{em}$ is the motor torque, $T_{load}$ is the load torque and $J$ is the system inertia.

Figure 4.16 Mechanical System
The mechanical system model of Figure 4.16 also provides for applying torque loads at two different points in time, as well as a torque load that depends on the direction of rotation. For this simulation, the direction is positive and the load torque that is applied is a small 0.25Nm load to simulate friction in an unloaded motor. The values for torque and time are set in the initialization file shown in Appendix A.

4.7 The VMAG Stator Switch

The Simulink model for the VMAG system shown in Figure 4.2 includes a function block referred to as the VMAG stator switch. This function block simulates the speed dependent switching of the dual stator axial flux motor from running on both stators to running on only one stator. The VMAG Stator Switch Function can be seen in detail in Figure 4.17. If the stators have equal air gaps between each stator and the rotor, then the selection of which stator is removed from the circuit and which stator is left in the circuit does not matter. In either case, the value of all parameters, $R_s, L_s, K_e$ and $\lambda_{FD}$ will be half of their values when there were two stators in the circuit. However, if the stators do not have equal air gaps, then the stator with the longer air gap to the rotor is the stator that will be left in the circuit to obtain the greatest possible top speed. When the stator configuration is switched from the two stator configuration to the one stator configuration, the result will be values of $K_e$ and $\lambda_{FD}$ that are less than $\frac{1}{2}$ of the original values for two stators. The values of $K_e$ and $\lambda_{FD}$ in the one stator and two stator configuration will be a function of the air gap lengths. Also, $R_s$ and $L_s$ will be reduced to half of the value as when there were two stators in the circuit. $R_s$ and $L_s$ are not
dependent on air gap length. The relay function in Figure 4.17 sets its output to a one when the motor speed is above a switch threshold called VMAGspeed and to a zero when the motor speed is below VMAGspeed–100. This provides a hysteresis of 100 RPM in the selection of one stator versus two stators. The value of VMAGspeed is set in the initialization file given in Appendix A.

Figure 4.17 VMAG Stator Switch
4.8 The Simulation Parameters

A Matlab m file was used to store all of the simulation parameters used in the simulation. These parameters are shown in Appendix A. The simulation parameters used were taken from test data collected by the motor manufacturer for the same motor that was used in the experiment shown in Chapter V. One critical parameter used that was not taken from motor test data was the change in $K_e$ for a given change in air gap length. The degree by which this motor’s $K_e$ would change for a change in air gap length was determined by the motor’s manufacturer using their finite element analysis model. The details of that model cannot be shown here due to restrictions of a non-disclosure agreement. The results of this study are shown in Table 4.1. According to the predicted results of Table 4.1, if one air gap was changed from 1.5 mm to 5 mm, then the BEMF produced by that stator would be 59% of its previous value. The BEMF produced by the other stator would be unchanged.

Table 4.1 BEMF Constant ($K_e$) Reduction of Extended Air Gaps

<table>
<thead>
<tr>
<th>Gap (mm)</th>
<th>1.5</th>
<th>3</th>
<th>5</th>
<th>10</th>
</tr>
</thead>
<tbody>
<tr>
<td>% $K_e$ of original</td>
<td>100%</td>
<td>82%</td>
<td>59%</td>
<td>22%</td>
</tr>
</tbody>
</table>
If equations (3.8) and (3.9) were evaluated using the values from Table 4.1, then for the original motor with both air gaps at 1.5 mm the value of \( \alpha \) would be 1. The resulting equations would be the same as equations (3.5) and (3.6).

If that same motor were built with unequal air gaps, for example 1.5 mm and 5 mm air gaps, then \( \alpha \) from equations (3.8) and (3.9) would be 0.59 per Table 4.1. Referring to the BEMF constant of the unaltered motor with both air gaps at 1.5 mm as \( K_{e\_original} \), the expression for \( K_e \) would be found using the analysis in Section 3.2 as

\[
K_e = \frac{1 + \alpha}{2} K_{e\_original} = 0.795 K_{e\_original} \tag{4.8}
\]

before VMAG switching, and

\[
K_e = \frac{\alpha}{2} K_{e\_original} = 0.295 K_{e\_original} \tag{4.9}
\]

after VMAG switching.

Looking again at equation (3.10), for the case of the motor with unequal air gaps of 1.5 mm and 5 mm, the value of \( \alpha \) would be 0.59. As defined in Section 3.2, \( \omega_{NL\_original} \) is the no load speed of a dual stator axial flux motor with equal air gaps with no VMAG. The new equation for no load speed for the specific case of 1.5 mm and 5 mm air gaps with VMAG is

\[
\omega_{NL} = \frac{2}{\alpha} \omega_{NL\_original} = 3.39 \omega_{NL\_original} \tag{4.10}
\]
This means that with VMAG on a motor with unequal 1.5 mm and 5 mm air gaps, the no
load speed should increase by 239% over the motor with equal air gaps and no VMAG.

The purpose of this concept is to extend the speed range of the dual stator axial flux
motor. Since the motor that is simulated is designed with a no load speed of 5000 RPM
with a 115 VDC bus system, it was necessary to select a lower bus voltage to allow the
demonstration of VMAG without exceeding the motor’s maximum allowable speed of
6000 RPM. For this reason, a bus voltage of 28 VDC was selected. This would allow
the motor to reach no load speed at about 1200 RPM. Then when the VMAG is
employed, the speed should then reach around 2400 RPM.

4.9 Simulation Results

A baseline simulation was conducted with VMAG turned off and with equal air gaps
to establish a baseline of the motor’s performance. Then a speed was selected to engage
the VMAG stator switching. This speed is selected such that the switching occurs before
the motor reaches its BEMF speed limit. This first simulation of the VMAG switching is
done with equal air gaps. Then the simulation is repeated with unequal air gaps to show
the added benefit of VMAG when unequal air gaps are used. By altering two values in
the initialization file of Appendix A, the simulation that is performed is selected. The
gain variable G3 corresponds to $\alpha$ in equations (3.7), (3.8) and (3.9). By setting $\alpha =1$
equal air gaps are simulated. With $\alpha =0.59$, non equal air gaps of 1.5 mm and 5 mm are
simulated. By setting VMAGspeed=900000, VMAG is turned off since the simulation
will never reach that speed. Note that any large value of VMAGspeed will have this effect. Note that VMAGspeed is in units of RPM. To enable VMAG in the simulation, the VMAGspeed was set to VMAGspeed=900. This is a speed that is slightly lower than the speed of 1200 RPM where the motor is expected to max out.

In all simulations, a torque command of 35 Nm is set at t=0.3 sec. A very small load torque of 0.25 Nm is applied 0.2 sec later. This small load is intended to simulate friction losses.

The results of the baseline simulation of equal air gaps and no VMAG are shown in Figure 4.18. In this plot, the motor torque output precisely tracks the command torque until the motor starts to reach the no load speed. As the motor speed approaches no load speed, the q axis current that can be driven into the motor begins falling off as is seen in the current plot of Figure 4.19. Though the precise value of the no load speed is not clearly visible in Figure 4.18, the value was measured after simulation at 1473 RPM. This is the baseline no load speed for the motor with equal air gaps and no VMAG operation.

The next step was to simulate VMAG with equal air gaps. In this case, \( \alpha = 1 \) and VMAGspeed=900. This activates VMAG at 900 RPM and simulates equal air gaps of 1.5mm each. In the plot of Figure 4.20, the motor torque precisely tracks the command torque until the motor reaches 900 RPM where the VMAG switching occurs. Both \( K_e \) and \( K_t \) are reduced by 50% as can be seen in the BEMF plot. In order continue tracking the torque command, the controller needs to double the current from the level prior to VMAG switching. However, since the motor is only rated for 200 A, the controller
simulation has a max current limit of 200 A programmed. This 200 A current limit can be seen in the q axis current in Figure 4.21. The result is that the motor cannot track the torque command but instead produces a torque that is the result of a 200 A current and the new $K_t$. As discussed previously in this report, this reduction in torque capacity when VMAG is engaged was expected. A close inspection of the plot of Figure 4.20 reveals the no load speed achieved with equal air gaps and VMAG is 2957 RPM. This is 2.01 times the baseline no load speed of 1473.

Next the simulation was done for the case of unequal air gaps and no VMAG. In this case, the motor’s $K_t$ is approximately 20% lower than the equal air gap case and therefore, as can be seen in Figure 4.22, the motor torque does not track the torque command due to the 200 A current limit discussed previously and shown in Figure 4.23. If the torque command were reduced to a value requiring a current of less than 200 A, then the torque output would track the torque command. This case of unequal air gaps and VMAG not enabled is not really a scenario that would be desirable in the real world. This is because extending one of the air gaps reduces the motor’s low speed torque. By extending one of the air gaps but not enabling VMAG operation, there would be a reduction of torque with no improvement in speed range. There would be no reason to reduce the $K_e$ and $K_t$ of the motor by extending one air gap without intending to use VMAG to extend the speed range. This case is however shown here for reference only. In this case, the no load speed settles out at 1858 RPM.
Figure 4.18 Torque and Speed for Equal Air Gaps with VMAG Disabled

Figure 4.19 q and d Axis Currents for Equal Air Gaps with VMAG Disabled
Figure 4.20 Torque and Speed for Equal Air Gaps with VMAG Enabled

Figure 4.21 q and d Axis Currents for Equal Air Gaps with VMAG Enabled
Figure 4.22 Torque and Speed for Unequal Air Gaps with VMAG Disabled

Figure 4.23 q and d Axis Currents for Unequal Air Gaps with VMAG Disabled
The final case that was simulated is the case of unequal air gaps and VMAG enabled. In this case, \( \alpha = 0.59 \) and VMAG speed=900. This activates VMAG at 900 RPM and simulates a 1.5 mm air gap and a 5 mm air gap. Figure 4.24 shows the torque, BEMF and speed for this simulation. Like the previous simulations, there is a 200 A current limit. The torque output of the motor does not track the torque command due to this current limit and the reduced \( K_t \) of the motor with the extended air gap. The \( K_t \) is further reduced when the VMAG switching occurs. This can be seen in Figure 4.24 as a step change in torque output at the same time corresponding to the step change in BEMF of the motor. Also note that after the step change in BEMF, the slope of the BEMF plot is less than the slope before VMAG was activated. This indicates that the motor is running with a reduced \( K_e \). Figure 4.25 shows the q and d axis currents from the simulation. The 200 A limit can be seen in the plot of \( i_q \).

In all cases simulated, the same scale and limits were chosen for the output plots. In the case of unequal air gaps and VMAG enabled, the motor no load speed plotted is 4752 RPM but is still accelerating after the 4 second simulation is complete. When examining the speed plot of Figure 4.24 by zooming in, there is still significant slope to the speed plot line at the end of the plot. In all other cases, the slope was flat at 4 seconds. For this reason the simulation was repeated with a longer time duration to determine the steady state no load speed. With a 7 second simulation the no load speed settles out at 4950 RPM. This extended time simulation output is shown in Figure 4.26.
Figure 4.24 Torque and Speed for Unequal Air Gaps with VMAG Enabled

Figure 4.25 q and d Axis Currents for Unequal Air Gaps with VMAG Enabled
4.10 Transient Simulation

The simulation presented thus far has shown the predicted effect of VMAG on the motor BEMF and torque. What has not been examined is the impact of this switching on the motor circuit itself. Figure 4.27 shows the equivalent phase lead to neutral circuit for one phase of the motor with VMAG. In several cases, the average power for one phase delivered to the motor will be examined using

\[ P_{wr} = I_{RMS} \times V_{RMS} \times \cos(\psi), \]  

(4.11)

where \( \psi \) is the phase angle between the voltage and the current.
4.10.1 Transition from Two Stators to One Stator

A circuit simulation was conducted using LTspice simulation software to examine the transients that will occur when VMAG is engaged. There will be significant stored energy in Stator1 when it is suddenly isolated from the circuit. The circuit model that was simulated is shown in Figure 4.28.

![Figure 4.27 Circuit Simulation Schematic](image)

Sinusoidal voltage sources in phase with each other were used to simulate the BEMF’s induced in the two stators. The supply voltage has a slight phase shift of 0.733 degrees from the BEMF’s. This causes the stator currents to be in phase with the BEMF’s as would be the case in a motor driven by FOC. This model helps to understand the changes in the stator voltages and currents when the stator switching occurs. A frequency of 180 Hz was selected to simulate a motor speed of 900 RPM. The BEMF peak amplitude of 5.51 V with a 0 degree phase angle represents the BEMF at 900 RPM. The source voltage simulated was 11.0509 V peak with a phase angle of 0.733 degrees.
This voltage was selected so that the resulting current would be approximately 5 A peak with a 0 degree phase angle (in phase with the BEMF’s).

For the VMAG switches, spice models for a voltage controlled switch were used. The switch characteristics can be seen in the lower right corner of Figure 4.28. The “on” resistance is 0.0000001Ω, the “off” resistance is 2000000MΩ, the series inductance is 0 H and the series voltage is 0 V. A voltage of 0.5 V at the control input will cause the switch to turn on. With a hysteresis voltage of 0.1 V, the switch will turn off again when the control input drops below 0.4 V. A PULSE voltage source was used to control the VMAG switches. The timing of the pulses was selected to engage the VMAG switches.
at the desired time relative to the motor current. The effects of engaging the VMAG switches when the stator current was at a peak level was examined, as well as the effects of switching when the stator current was at a zero crossing. Altering the timing of the switching allowed both options to be examined. Also the effects of operating the two switches at the same time as well as overlapping (make before break) were examined.

The following scenarios were simulated for this circuit:

- Closing S2 while simultaneously opening S1 when current is not at a zero crossing.
- Closing S2 7 ms before opening S1 when current is not at a zero crossing.
- Closing S2 while simultaneously opening S1 when current is at a zero crossing.
- Closing S2 7 ms before opening S1 when current is at a zero crossing.

Not modeled was any scenario where S1 is opened before S2 is closed. Having a state where both switches were open simultaneously would result in a total disruption in motor current. This should be avoided as it would cause excessive voltage spikes at the motor leads and would likely damage the inverter driving the motor.

Current in both stators as well as the Stator1 voltage were examined for each scenario. The circuit simulation shows that opening the switch S1 and isolating Stator1 while the motor is running causes a significant voltage spike on the Stator1 voltage. The magnitude of this voltage spike is different for each of the scenarios simulated but is very high in all cases. The worst case is when S2 is closed 7 ms before opening S1 while the
current is at a peak and not a zero crossing. For this worst case scenario, the simulated plots for Stator1 voltage, Stator1 current, and Stator2 current are shown in Figures 4.29, 4.30 and 4.31 respectively. Figure 4.32 shows the current in switch S2 during the transition.

Figure 4.29 shows the voltage across Stator1. The switch S2 closes at 0.499 s and S1 opens at 0.506 s. What is of particular interest is the magnitude of the voltage spike that occurs when S1 opens. The exact value is not important, only that it is extremely high. The simulated magnitude is shown in Figure 4.29(a). Before and after the switching event, the voltage is sinusoidal based on the BEMF voltage of Stator1 and the current through Stator1. During the 7 ms when both S1 and S2 are on, the Stator1 voltage is zero due to the brief short circuit condition caused by both switches being closed. This can be seen in Figure 4.29(b).

The Stator1 current is shown in Figure 4.30, along with the Stator1 BEMF for reference. Examining the Stator1 current in Figure 4.30 reveals that at 0.499 s, when S2 closes, the waveform of the Stator1 current flowing through R1 shifts. Before 0.499 s, the Stator1 current is in phase with the Stator1 BEMF voltage and 27 watts is being delivered to Stator1. During the 7 ms transition period, the Stator1 current is greatly increased and leading the voltage. Based on the impedance angle during the transition, the phase lead would be 102 degrees in the steady state. The power in Stator1 would be approximately −215 watts during the 7 ms transition period. This indicates that the power flow in Stator1 reverses direction during this brief 7 ms short circuit condition. There would be a torque transient from the motor as this power flow reversal would
create negative torque from Stator1 for the 7 ms that both switches were closed. The
effect of this torque transient on the net motor torque will also depend on the torque
produced by Stator2 during this 7 ms period. The Stator2 power will be examined below.
In addition to the Stator1 current phase shift that occurs during this 7 ms period with both
contacts closed, the Stator1 current has a very large current magnitude of approximately
750 A peak-to-peak. This is due to the fact that Stator1 has a BEMF proportional to the
rotating speed of 900 RPM as shown in Figure 4.30 when the 7 ms short circuit occurs.
When S1 opens, the Stator1 current instantly becomes zero. With the Stator1 current
changing instantaneously, the \(\frac{di}{dt}\) of the Stator1 inductance is very high resulting in the
high negative transient voltage of Figure 4.29(a). The transient is negative since the
current in the stator is positive at the moment when S1 opens. The smaller positive
transient in Figure 4.29(a) is a simulation artifact. The fact that the Stator1 current
becomes very high during the 7 ms overlap of the S1 and S2 switches is part of the reason
that this scenario has the highest level of transient on the Stator1 voltage.
Figure 4.29(a) Stator1 Voltage During Transition
Figure 4.29(b) Stator1 Voltage Detail During Transition
Figure 4.30 Stator1 Current and Stator1 BEMF During Transition

Figure 4.31 shows the Stator2 current during the same transition. The supply voltage is shown for reference. Here the current in Stator2 changes instantly from 5 A peak with a 0.73 degree phase shift as compared to the supply voltage to approximately 360 A peak with a 78 degree phase shift as compared to the supply voltage when S2 closes at 0.499 s. This is due to the sudden decrease in motor BEMF and impedance. The Stator2 current is also the total motor current as seen by the supply. Before S2 closes, the supply is delivering approximately 27 watts of power to the motor. After S2 closes at 0.499 s, the supply is delivering approximately 413 watts to the motor.
It is also important to note that when S2 closes, Stator2 begins consuming 413 watts of power, indicating positive torque from Stator2. During the 7 ms transition period, Stator1 is generating approximately 215 watts of power, indicating a negative torque from Stator1. The net power of both stators is approximately 198 watts, indicating that the motor has a positive torque during this 7 ms transition period. After the 7 ms transition period, the net motor power is 413 watts. This is a dramatic increase in power from the 27 watts of power being delivered to the motor before the transition. The sudden and dramatic increase in current after S2 closes would cause the motor to accelerate, as the torque being produced by Stator2 would increase proportional to this current increase. In a motor controlled by a field oriented controller, the control loops regulating the q axis and the d axis current to the motor would need to react quickly to prevent this large current increase. Any delay in regulating the current to the desired level would result in an acceleration of the motor.

The current through switch S2 during transition is shown in Figure 4.32. During the 7 ms period when both switches are closed, S2 has approximately 1500 A peak-to-peak flowing through it. This is due to the fact that the Stator2 current flows through S2 but during the 7 ms overlap, there is also a current of 750 A peak-to-peak flowing out of Stator1.

The other scenarios that were modeled but not shown here exhibited the same results, except for the Stator1 voltage and the S2 current. This case shown was the worst case in regards to the Stator1 voltage. The other scenarios resulted in a lower magnitude
of the voltage spike than what is shown in Figure 4.29(a). The cases where the S1 and S2 switches were not overlapping showed no large current transients on S2.

Figure 4.31 Stator2 Current During Transition
Reducing the Voltage Transient During Transition

In an effort to eliminate the large voltage transient during the transition, a capacitor was added in parallel with Stator1 as shown in the circuit simulation model of Figure 4.33. The added capacitor provides a means to dissipate the stored energy in Stator1 when Stator1 is suddenly isolated from the rest of the circuit. The capacitor provides a path for the transient current to flow, and exchanges energy with the stator inductance until it dissipates in the winding resistance. Several values of capacitance were simulated, 100 μF, 10 μF, 1 μF and 0.1 μF. In the analysis of the original
simulation in Figure 4.28, the scenario that resulted in the worst case voltage transient was presented. Since the goal of adding the capacitor is to reduce the switching transients during the transitions, the scenario that results in the lowest transients is of particular interest. The scenarios simulated with the added capacitor were the same scenarios as the original circuit without the capacitor. Each scenario was simulated with each of the capacitor values listed above. These scenarios were as follows:

- Closing S2 while simultaneously opening S1 when current is not at a zero crossing.
- Closing S2 7 ms before opening S1 when current is not at a zero crossing.
- Closing S2 while simultaneously opening S1 when current is at a zero crossing.
- Closing S2 7 ms before opening S1 when current is at a zero crossing.

The simulations showed that there were tradeoffs related to the choice of capacitor values. The highest capacitor value resulted in the lowest Stator1 voltage transient. The higher capacitor values also resulted in undesirable increased steady state capacitor current and an undesirable phase shift between the Stator1 current and the Stator2 current. This phase shift would result in reduced torque production. Using a 10 μF capacitor in the scenario of switching both switches simultaneously at the zero crossing of the stator current resulted in both a low stator voltage transient, a low steady state capacitor current, and low phase shift between the Stator1 and Stator2 currents. It is not certain that switching at the zero crossing of current is possible in the three phase system. For this reason, the scenario of switching both switches simultaneously while the
current is at a peak is presented. The simulations showed that the advantage of zero crossing switching was not very significant so the decision to switch at peak current instead of zero crossing will not have a significant impact. Simulations using other switch timing scenarios resulted in even higher voltage transients on the Stator1 voltage during the switching event. The choice of capacitor values affected the Stator1 voltage as well. Lowering the capacitor value resulted in a higher transient on the Stator1 voltage. With the capacitance value too low, the capacitor was not able to quickly absorb the energy from the Stator1 inductance, therefore the transient voltage was not as greatly reduced.

Figure 4.33 Circuit Simulation of Axial Flux Motor with VMAG with Added Capacitor
With a capacitor value of 10 \( \mu \text{F} \), the Stator1 voltage of Figure 4.34 shows a dramatic reduction in transient magnitude during the transition as compared to the nearly 20 MV spike in Figure 4.29(a). The voltage of Figure 4.34 has some high frequency content. This damped oscillation is the response of the resonance between the capacitor and the inductance of Stator1. Figure 4.35 shows the current in the Stator1 winding. There is a damped oscillation after the transition occurs before settling at zero amperes. Like the Stator1 voltage in Figure 4.34, this damped oscillation is the response of the resonance between the capacitor and the inductance of Stator1. The Stator2 current in Figure 4.36 is nearly identical to the Stator2 current for the original simulation shown in Figure 4.31. The supply voltage is again shown for reference. The final plot of this simulation, Figure 4.37, shows the current in the Stator1 capacitor. Once S1 opens at 0.4985 s, this current is the same current as the Stator1 current in Figure 4.35. In the post transition oscillations of Figures 4.34, 4.35, and 4.37, the damping of the oscillations is the result of the stator resistance.

Also of concern is the effect on steady state current from adding the capacitor. Figure 4.37 shows that there is no significant steady state current through the capacitor. A closer zoom of the current plot reveals a steady state current value of 60 mA peak at the 180 Hz corresponding to 900 RPM. For the purpose of determining the impact on steady state capacitor current, other speeds were simulated as well. At 10 Hz the steady state capacitor current was 3.5 mA peak. At 500 Hz the steady state capacitor current was 180 mA. Based on this simulation, it appears that adding the 10 \( \mu \text{F} \) capacitor in
parallel with Stator1 greatly reduces the voltage spike that occurs when Stator1 is switched off while introducing no significant steady state effects.

Figure 4.34 Stator1 Voltage During Transition with Added Capacitor
Figure 4.35 Stator1 Current During Transition with Added Capacitor
Figure 4.36 Stator2 Current During Transition with Added Capacitor
As stated previously, the worst case scenario in the original circuit of Figure 4.28 without the capacitor was when S2 was closed 7 ms before S1 was opened. How the circuit with the capacitor responded to this scenario was also of interest. The simulation of Figure 4.33 was repeated with S2 closing at 0.4985 s and S1 opening 7 ms later at 0.5055 s.

Like the simulation of Figure 4.28, when S2 closes 7 ms before S1 opens, there is essentially a short circuit across the capacitor and across Stator1 for that 7 ms period. This can be seen in the Stator1 voltage of Figure 4.38. At 0.4985 s when S2 closes, the
voltage across Stator1 is zero until S1 opens at 0.5055 s. The Stator1 current plot in Figure 4.39 shows that, like the simulation of Figure 4.28, there is a large current during the 7 ms short circuit period. During the 7 ms transition period, the current is greatly increased and leading the voltage by 102 degrees. During this time, power is flowing out of the motor rather than into the motor. Due to the 7 ms short circuit condition, the current in Stator1 during this transition is much greater than the current in Figure 4.35 when the S1 and S2 switches are not overlapping. The result is a higher $\frac{di}{dt}$ when S1 opens and therefore a higher transient voltage on Stator1 during the transition as seen in Figure 4.38.

The capacitor current can be seen in Figure 4.40. When S2 closes at 0.4985 s the short circuit across Stator1 and the capacitor causes the capacitor to discharge as seen by the large positive current spike at that time. This spike is approximately 17 kA. Using Ohm’s Law, this spike is consistent with the current caused by the negative capacitor voltage across the series resistance of S1 and S2. There is also a negative voltage spike on the capacitor voltage that is likely a simulation artifact given the fact that there is no explanation for it based on circuit analysis and the fact that the RC time constant of the capacitor and switch resistance is very short, $2 \times 10^{-12}$ s, which is likely shorter than the time step of the simulation. During the rest of the 7 ms period when both switches are closed, the capacitor voltage is zero and the Stator1 current is high. Then when S1 opens at 0.5055 s, the Stator1 current value is nearly 350 A. Once S1 is open, Stator1 is in series with the capacitor and the Stator1 current and the capacitor current are the same. This current resonates in a damped oscillation at 14.2 kHz, the resonant frequency of the
10 μF capacitor in series with the 12.5 μH inductance of Stator1. This oscillation is damped due to the Stator1 resistance. The initial value of the damped oscillation is very high, approximately 350 A peak. There is another voltage spike shown in Figure 4.40 at 0.5505 s when S1 opens. There is no reasonable explanation based on circuit analysis for this spike. Also, starting at 0.5505 s, the capacitor current of Figure 4.40 and the Stator1 current of Figure 4.39 should be the same. Figure 4.39 has no such current spike. Therefore, this spike in Figure 4.40 at 0.5505 s is believed to be a simulation artifact and not real. Allowing S1 and S2 to be simultaneously closed is shown by this simulation to cause higher transient voltages on Stator1 and higher transient capacitor currents than the simulation shown in Figure 4.33 where the switches S1 and S2 are not closed simultaneously.
Figure 4.38 Stator1 Voltage During Transition, Added Capacitor and S2 Switched First
Figure 4.39 Stator1 Current, Added Capacitor and S2 Switched First
4.10.3 Transition from One Stator to Two Stators

The circuit for the transition from one stator to two stators is also of interest. This LTspice simulation is shown in Figure 4.41. In this case, S1 and S2 are operated simultaneously and not at a zero crossing of the current. The supply voltage of 5.525 V peak with a phase shift of 0.733 degrees is selected to provide current of approximately 5 A when the motor is operating in the initial condition of only Stator2 in the circuit. The Stator1 voltage plot of Figure 4.42(a) shows that in this case there is also a large voltage transient when the switching event occurs. The large negative spike is the result of the
high $di/dt$ on the Stator1 inductance. The smaller negative voltage spike is likely a simulation artifact and not real. Figure 4.42(b) shows the detail of the Stator1 voltage before and after the transition. Before the transition, there is no current in Stator1 as shown in Figure 4.43. After the transition, Figure 4.43 shows that the motor current is very high. Figure 4.42(b) shows that before the transition the Stator1 voltage is simply the value of the Stator1 BEMF. After the transition, the Stator1 voltage is lower than before the transition. This voltage drop suggests that after the transition from one stator back to two stators, the high current is actually flowing out of Stator1. More evidence of this will be given later in this analysis.

Figure 4.41 Circuit Simulation of VMAG for One Stator to Two Stator Transition
Figure 4.43 shows that after the transition from one stator back to two stators, the motor current is very high, approximately 175 A peak. The current in Stator2 is approximately 5 A peak before the transition as seen in Figure 4.44. After the transition, Stator2 is in series with Stator1 and the current in Stator2 is the same as in Stator1.

Figure 4.42(a) Stator1 Voltage During Transition from One Stator to Two Stators
Figure 4.42(b) Stator1 Voltage During Transition from One Stator to Two Stators
Figure 4.43 Stator1 Current During Transition from One Stator to Two Stators
At first glance, it would appear that the high current in Stator1 and Stator2 after the transition from one stator back to two stators should result in a motor acceleration as in the transition from two stators to one stator previously shown. However, a quick look at the BEMF voltages in Figure 4.41 shows that after the transition the voltage source of 5.66 V peak is supplying a motor that now has a BEMF of 11.02 V peak. The motor voltage being higher than the supply voltage suggests that the motor may actually be generating after the transition from one stator to two stators. Figure 4.45 again shows the Stator2 current. It also shows the supply voltage. It can be seen that the phase
relationship between the motor current (Stator2 current) and the supply voltage changes after the transition. Before the transition the motor current and supply voltage are in phase and power is flowing into the motor. After the transition, the motor current is leading the supply voltage by 102 degrees and power is flowing out of the motor. This means that the motor switches from motoring to generating when the transition occurs. Without an adjustment of the supply voltage, the motor would decelerate to a speed where the BEMF of the motor was lower than the supply voltage.

In a motor controlled by a field oriented controller, the control loops regulating the q axis and the d axis current to the motor would need to react quickly to prevent this large current reversal. Any delay in regulating the current to the desired level would result in a deceleration of the motor.
4.11 Summary of Simulation Results

The simulations show that when using the VMAG on a dual stator axial flux motor with unequal air gaps, the speed range of the motor can be extended significantly. In the case of the 1.5 mm and 5 mm air gaps, the speed range increased by 236%. The analysis done in Section 4.8 and shown in equation (4.10) indicated an expected improvement of 239%.
Using VMAG on the dual stator axial flux motor with equal air gaps also showed a significant improvement in speed range. As shown previously, the speed range improved by a factor of 107% with equal air gaps, a close correlation to the expected 100% improvement from Section 4.8.

Section 4.10 used a circuit simulation to show the voltage and current transients during the transition from two stators to one stator. The circuit simulation shows that during transition the Stator1 voltage has a transient of nearly 20 MV. This is due to the fact that the switch opens instantaneously causing a very high $di/dt$. Simulations also show that adding a 10 $\mu$F capacitor in parallel with Stator1 does not significantly affect the steady state currents but does significantly reduce the voltage transients on Stator1.

Also shown in Section 4.10 is the fact that when the motor transitions from two stators to one stator there is a large increase in current. The supply voltage must be altered to compensate. This will likely result in an acceleration of the motor until the current controller properly regulates the current.

When the motor transitions from one stator back to two stators, the BEMF increases instantly making the BEMF voltage higher than the supply voltage. The motor current reverses and the motor generates, causing a motor deceleration until the current controller properly regulates the current.
CHAPTER V

THE VIRTUAL MOVING AIR GAP EXPERIMENT

5.1 Experiment Goals and Procedures

The goal of the VMAG experiment is to prove that the speed range of a dual stator axial flux motor can be extended by selectively switching the motor from a dual stator motor to a single stator motor. This is of greatest value when the system has a bus voltage constraint that is low compared to the $K_v$ of the motor being run. For this experiment the dual stator axial flux motor used had an advertised no load speed of approximately 5000 RPM on a 115 VDC bus. In order to provide a bus voltage constraint on the no load speed for this experiment, a bus voltage of 30 VDC was selected. It was assumed that this bus voltage would limit the motor to a speed of approximately 1300 RPM. Maximum benefit of the VMAG concept comes from applying the stator switching to a motor that has unequal air gaps. In the case of the motor used in this experiment, the simulations predicted that for unequal air gaps of 5 mm and 1.5 mm, the speed would increase by 239%. For equal air gaps of 1.5 mm, the simulations predicted a speed increase of 107%. Conducting the experiment with unequal air gaps was greatly desired to show the concept’s maximum benefits. However, there would have been significant labor involved in disassembling the motor to increase
one of the air gaps. There was also significant risk of injury due to the lack of the proper equipment. Disassembly of a permanent magnet axial flux motor of any significant size requires the use of a crane or similar device to separate the stators from the rotor. If hand assembly is attempted, the attractive forces of the rotor’s permanent magnets in to the stators can force the motor to slam together accidentally with enough force to sever fingers or cause other severe injuries. For simplicity and safety, the experiment was limited to the case of using equal air gaps.

During the experiment particular attention was given to the behavior of the motor in the period of transition from two stators to one stator and from one stator to two stators. The experiment to validate the VMAG involved the following steps.

1. Build a FOC and inverter to control the axial flux motor, including a control for the switches that will switch between two stators and one stator.

2. Test the motor BEMF with the motor hard wired in the two stator configuration.

3. Run the motor with the stators hard wired for two stator configuration.

4. Test the BEMF with the motor hard wired in the one stator configuration.

5. Run the motor with the stators hard wired for the one stator configuration.

6. Wire the stators with switches S1 and S2 per the schematic of Figure 3.2.

7. Run the motor with the stator switches in place and examine behavior during the transition between stator configurations.
5.2 Hardware Design

The VMAG experiment requires a field oriented controller and the inverter to drive the motor. The Texas Instruments TMS320F28335 digital signal processor was selected as the control processor to be used. A Semikron SKAI inverter was selected for the inverter power stage. A signal conditioning circuit was designed and built to interface the 28335 DSP to the inverter. An incremental encoder was used as the position feedback device. A signal conditioning circuit was designed and built to allow the DSP to read the incremental encoder device. For the VMAG switches, Allen Bradley Bulletin 100C43 three phase contactors with low current 24 VDC coils were selected. A MOSFET driver circuit was designed and built to allow the DSP to drive the two contactors. This hardware scheme is shown in the block diagram of Figure 5.4. A detailed description of the hardware including schematics and a bill of materials is given in Appendix B. A photo of the interface circuit is shown in Figure 5.1. Figure 5.2 shows the interface circuit and the inverter, while the motor and contactors are shown in Figure 5.3. An overall system block diagram is shown in Figure 5.4.

Note that the inverter and the motor are liquid cooled. In this experiment, the system was operated without a load. If the system were operated under load, the inverter and motor would have required a coolant system in order to not overheat. According to Semikron the current rating of the inverter is about 80 A for less than 10 seconds and 5 A continuous. This no load experiment remained within those limits.
Figure 5.1 Photo of Interface Circuit
Figure 5.2 Photo of Interface Circuit and Inverter
Figure 5.3 Photo of Axial Flux Motor and Contactors

Figure 5.4 System Block Diagram of VMAG Experiment
5.3 Software Design

The VMAG experiment utilizes a modified FOC to operate the axial flux motor. The FOC is modified to operate switches S1 and S2 from Figure 3.2. Additionally, the FOC is modified so that the motor parameters that are used in the current regulation loop are variables whose values change based on the state of S1 and S2. Rather than starting from a blank sheet on the FOC software, sample code provided by Texas Instruments (TI) was used. This sample code was written for the processor used in this experiment, the F28335. The code used was from the TI “Control Suite” and written to support the TI “High Voltage Kit” for the PMSM. The sample code was originally written as a sensorless application, but the code for running with an encoder was present and not utilized. Converting this code to operate the motor with encoder feedback was a simple code change. In order to adapt the sample code provided by TI into the code required to run the VMAG experiment, the following major changes were made:

- The original code utilized a sliding mode observer based rotor position estimator to provide rotor position angle $\theta$ for the dq transformations. The code was changed to utilize the $\theta$ from the encoder feedback function (qep1) instead.

- The current and voltage feedback scaling were changed to reflect the scaling of the Semikron SKAI inverter rather than the TI “High Voltage Kit”.

- The values of $R_s$ and $L_s$ were made variables rather than #define’s so that their values could be changed during motor operation as needed.
• Rotor flux \( \lambda_{FD} \) was added as a variable. See Appendix C for details on how \( \lambda_{FD} \) was determined.

• The TI sample code did not incorporate the rotor-stator flux coupling terms described in Chapter 9 of [8], so code was added to incorporate this coupling compensation.

• Encoder alignment calibration was required so that the \( \theta \) returned by the encoder function would be properly aligned relative to the rotor flux. This step is necessary for the FOC to be able to commutate the motor. See Appendix D.

• PID gains for the current regulation and velocity control loops were tuned for the axial flux motor rather than using the default values in the original code.

• A speed dependent variable called “VMAG_Active” was added to define two states when the motor is operating on one stator or two stators.

• Code was added to change the states of the switches S1 and S2 (Figure 3.2) and change the values of \( R_s, L_s \) and \( \lambda_{FD} \) based on the state of VMAG_Active.

Appendix E shows the motor parameters used for one stator and two stator operation. The block diagram shown in Figure 5.5 represents the FOC with VMAG that was used in the VMAG experiment. The diagram shows the interconnection of the major function blocks that make up the field oriented controller. The software is written with certain “build levels” in place. Each build level adds functionality to the system. This allows the user the ability to bring the system up in smaller pieces in order to isolate and debug the subsystems rather than trying to debug the entire system at once. The diagram
of Figure 5.5 represents the final build level, *Build Level 6*. The build level called *Build Level 5* was the build level modified to utilize the encoder feedback. This build level is a fully functional FOC system with encoder feedback. *Build Level 6* was added on as a copy of *Build Level 5* with the addition of the VMAG capability. Further documentation on the build levels is available from TI. One important element that is missing from Figure 5.5 is the ability to run the motor open loop. A software switch in the form of a variable lsw exists in the code so that when lsw=1 the motor will run with a rotor $\theta$ value that is arbitrarily generated by a *rampgen macro*. Lsw=1 also allows a value for iqRef to be entered through the Code Composer Studio’s watch window. This allows the motor to run without rotor position or speed feedback (open loop) at a fixed speed and a fixed current limit as set by iqRef. This open loop operation was omitted from Figure 5.5 for clarity.

Figure 5.5 Field Oriented Controller with VMAG
5.4 Motor Startup

The motor startup began with working through the different build levels in the software starting with Build Level 1 and working up to Build Level 5. In each step of the process, different subsystems of the diagram in Figure 5.5 are verified and if necessary, debugged. One of the final steps was to calibrate the alignment of the angle reported by the encoder function (qep1) so that a zero degree $\theta$ results in the rotating d axis being aligned to the a phase. Appendix D gives a detailed description of the encoder alignment process. Also it is important to mention that the encoder used in the experiment is an incremental encoder. The encoder does have a one pulse per rev channel (index). This index channel gives the encoder an absolute reference. When the controller is first started up, the control algorithm has no knowledge of the rotor position until the rotor moves far enough for the controller to see an index pulse. Once an index pulse is read by the FOC, the rotor position is accurately known. For this reason, after any restart of the controller software, the motor must first be briefly run open loop as discussed in Section 5.3 in order to have the rotor position known by the controller. Failing to run the motor open loop until an index pulse is read will result in erratic and unstable motor behavior.

5.5 BEMF Testing

The concept of VMAG is that the dual stator axial flux motor will have a lower $K_e$ when one of the stators is bypassed, allowing it to then achieve a higher speed for a given DC bus voltage. The first actual test completed was to test the motor’s BEMF for the
normal two stator operation. Then the motor windings were hardwired for the VMAG single stator configuration and the BEMF was tested again. In both cases, the motor was disconnected from the inverter and an oscilloscope was connected to the motor phase leads a and b. The line-to-line voltage, \( V_{a-b} \) was then observed on the scope while the motor was spun by hand with a wrench. The motor speed was then determined from the frequency of the BEMF sine wave on the scope. The BEMF was determined from the magnitude of the BEMF sine wave. The waveform in Figure 5.6 is the BEMF of the motor with the normal two stator configuration. It can be seen in Figure 5.6 that the BEMF for one cycle has a peak-to-peak magnitude of 11.4V and a period of 19.5 ms. The scope shows the electrical cycles of the BEMF voltage. Since the desired BEMF constant is related to mechanical cycles the number of pole pairs (12) is taken into account in the calculations. The equation for the BEMF constant in terms of RMS line-to-line voltage and rotor speed is given by

\[
K_e = \frac{V_{L-L}}{\omega_r}. \tag{5.1}
\]

The line-to-line RMS voltage is determined from the peak-to-peak voltage by

\[
V_{RMS} = \frac{V_{p-p}}{2\sqrt{2}} = \frac{11.4 \text{ V}}{2\sqrt{2}} = 4.03 \text{ V}. \tag{5.2}
\]

Rotor speed in RPM is determined from the measured period by

\[
\omega_r = \frac{T_{\text{BEMF}}}{\frac{60}{P} + \frac{P}{2}}, \tag{5.3}
\]
where $\omega_r$ is rotor speed, $T_{BEMF}$ is the period of the BEMF waveform and $P$ is the number of rotor poles.

Applying measured values to equation (5.3) gives us

$$\omega_r = \frac{0.0195 \text{s}}{12} \times 60 = 256 \ \text{RPM.} \quad (5.4)$$

Now the values of line-to-line voltage and rotor speed are applied to equation (5.1) to get

$$K_e_{full} = \frac{4.03 \text{ V}}{256 \ \text{RPM}} = 0.0157 \ \text{V/RPM} = 15.7 \ \text{V/kRPM.} \quad (5.5)$$

Figure 5.6 BEMF with Normal Two Stator Configuration

With the calculations in equations (5.1)-(5.5), $K_e$ for the normal two stator configuration is found as 15.7 V/kRPM.
The motor was later hardwired for the one stator configuration. The same BEMF test was repeated for this configuration. The plot of Figure 5.7 shows the BEMF for the single stator. Examination of this figure shows the BEMF for one cycle has a peak-to-peak magnitude of 4.56 V and a period of 24.4 ms.

Repeating the calculations of equations (5.1)-(5.5) for these values yields a $K_e = 7.87$ V/kRPM for the one stator configuration. As expected, this is half of the $K_e$ value obtained for the two stator configuration.

![Figure 5.7 BEMF with One Stator Configuration](image-url)
5.6 No Load Speed Test

The next test was to determine what the no load speeds would be for the two stator and one stator configurations. After the two stator BEMF test was conducted, the two stator no load test was conducted before the hardwired changes were made to the stator for the one stator tests. Running the two stator no load test involved running the motor with no load connected and determining the maximum achievable speed. The DC bus voltage was set at 30 VDC before the test began. After collecting data for the two stator no load test, it was discovered that while the motor was running, the DC bus voltage was actually 28 VDC. It was decided that the bus voltage would be set while the motor was at speed for all tests and the value would be set at 28 VDC for consistency. For this test, the FOC software was hard coded with the two stator values for $R_s$, $L_s$ and $\lambda_{FD}$. The maximum speed that could be reached with two stators on a 28 VDC bus was 0.248 PU as indicated in the Code Composer Studio’s watch window. With a base speed of 5000 RPM programmed, the 0.248 PU speed corresponds to 1240 RPM. Note that while a speed of 1300 RPM was expected for a 30 VDC bus, the actual bus voltage used was 28 VDC. Therefore for a 28 VDC bus, the expected speed was actually 1217 RPM. This means the expected speed was 1217 RPM and the actual speed was 1240 RPM. Figure 5.8 is the encoder $\theta$ waveform from the DSP. The $\theta$ value was programmed to be set at the diagnostic PWM_DAC output with an RC filter. This allows for the observation of rotor $\theta$ in relation to the other data being collected. The period of this signal also allows for the calculation of rotor speed.
The period of the encoder signal is measured at 4 ms. Using equation (5.3), this corresponds to a speed of 1250 RPM. This is a close correlation to the 1240 RPM determined from the Code Composer Studio’s watch window speed indication. It is possible that the speed was actually fluctuating slightly between the time that the separate measurements were taken.

After the motor was hardwired with the one stator configuration, the BEMF test for one stator described in Section 5.5 was conducted followed by the no load speed test for one stator. This time, the FOC software was hard coded with the single stator values for $R_s$, $L_s$ and $\lambda_{FD}$. With the same 28 VDC bus, the expected speed was twice that of the two stator test. The output of the DC power supply was adjusted when the motor was running near the new top speed to ensure that the bus was indeed 28 V. The bus voltage actually measured 27.5V while the motor was at speed for the one stator no load test. The power supply adjustment knob was too sensitive to achieve a voltage that was precisely 28 V.
For the one stator no load test, the highest speed achievable was 0.502PU. This corresponds to 2510 RPM. The one stator no load speed was 102% greater than the two stator no load speed. Figure 5.9 shows the encoder $\theta$ value for the motor while running at the one stator no load speed.

The period for the encoder signal in Figure 5.9 is 1.99 ms. Again from equation (5.3) the speed is calculated at 2512 RPM. This is a very close correlation to the 2510 RPM speed that was observed in the Code Composer Studio’s watch window. Using the encoder $\theta$ signals for the two stator and one stator open loop tests, the speed increase is 101%.

![Figure 5.9 Encoder $\theta$ Value with One Stator at Full Speed](image)

The currents for two stator operation and one stator operation are shown in Figures 5.10 and 5.11 respectively. Note that for the same speed, the current in the case of one stator operation is roughly twice the current as when two stators are used. This is due to the two stator $K_t$ being twice the value of the single stator $K_t$.  

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Figure 5.10 Phase Current with Two Stators at 900 RPM

Figure 5.11 Phase Current with One Stator at 900 RPM
5.7 VMAG Stator Switching Test

The last and perhaps most involved part of the VMAG experiment was to wire the stators with the S1 and S2 switch configuration of Figure 3.2. This will allow the modified FOC to switch the stators from a dual stator motor at low speeds to a single stator motor at high speeds. Based on the two stator no load speed of 1250 RPM that was observed in the no load speed test, a transition speed of 900 RPM was selected as the speed at which the motor would transition from two stators to one. A speed differential of 100 RPM was selected so that the motor would transition from one stator back to two at 800 RPM. This hysteresis function, which was shown in Figure 3.5, prevents the motor from oscillating between the two stator and one stator configurations if motor operation is attempted at the transition speed. The flowchart of Figure 5.12 shows the logical sequence used to control the switch states and motor parameters.
Figure 5.12 VMAG Switch Control Logic
5.7.1 Open Circuit Stator Transients

In order to observe if any large transients were present during the transition from two stators to one stator, the rotor phase leads were disconnected from the inverter and an oscilloscope was connected to the motor phase leads for the a and b phases. The controller was enabled but the DC bus was turned off. The FOC software was temporarily configured with a transition speed of 250 RPM. The motor was then turned by hand with a wrench. When the rotor speed reached 250 RPM, the switches S1 and S2 changed state and the motor transitioned from a two stator machine to a one stator machine. The BEMF waveform was captured on the oscilloscope. The captured waveform is shown in Figure 5.13 and Figure 5.14. The waveform of Figure 5.13 is magnified to get the image of Figure 5.14.

![BEMF During Transition from Two Stators to One Stator](image)

Figure 5.13 BEMF During Transition from Two Stators to One Stator
While the BEMF waveforms do not show any substantial spike in BEMF during the transition, it is important to note that during this test the motor is not running, the stator connections are an open circuit and there is no current flowing in the motor. This means that the stator coils are not energized during this test. The most useful information provided by this test is that there appears to be significant contact bounce when the S1 and S2 contactors change state. This is evident from the multiple transients shown in Figure 5.14.
5.7.2 Transition Under Power

For the remaining tests, the motor was wired to the inverter with a DC bus voltage of 28 VDC. The transition speed was set to the 900 RPM for the two stator to one stator transition and to 800 RPM for the one stator back to two stator transition.

5.7.2.1 Speed Transients

The motor starts and is brought up to a speed just below the 900 RPM transition speed. Then the speed reference increases to 900 RPM. When 900 RPM is reached, the contactors S1 and S2 change states and Stator1 is removed from the circuit. The motor instantly accelerates up to between 1200 RPM and 1500 RPM, and then settles back down to the desired 900 RPM within a second or two. The speed reference is then set to 750 RPM. This causes the motor to decelerate. When the speed reaches 800 RPM, the contactors S1 and S2 again change states and the motor returns to the two stator configuration. This time however, as the second stator is brought back into the circuit, the speed drops to less than 500 RPM, and within a couple of seconds returns to the desired 750 RPM. Much time and effort was spent in trying to determine the cause of these speed disruptions that occurred during the transitions. The following analysis and the simulations of Section 4.10 attempt to provide that explanation.

Figure 5.15 shows a plot of the speed signal from a diagnostic PWMDAC. There is a lot of high frequency noise on the signal, but the signal is still quite visible. As the speed signal increases, it suddenly jumps a substantial amount before settling back down
at the desired 900 RPM. If the two flat levels of this scope capture represent 750 RPM and 900 RPM, then the speed jump during transition represents a speed peak of approximately 1400 RPM. It appears to take approximately 1.6 seconds to settle back down to the desired 900 RPM setpoint.

Figure 5.15 Speed During Transition from Two Stators to One Stator

A similar plot of speed during the transition from one stator back to two stators is shown in Figure 5.16. In this plot, the speed dips during transition and then comes back up to the desired 750 RPM. Again, using the two flat parts of the plot as references for 900 RPM and 750 RPM, it appears that the speed dips to approximately 480 RPM. It appears to take approximately 2 seconds to come back up to the desired 750 RPM.
5.7.2.2 Phase Current Transients

The plots of Figure 5.15 and Figure 5.16 show us how much the speed is changing, but not why. Further examination of current and voltage was done. The plot of Figure 5.17 shows the phase current before, during and after the transition from two stators to one stator. As can be seen, the steady state current after transition is about twice the value as before. This is expected as the $K_t$ is reduced by half when the stator switching occurs. During transition, it can be seen that the current spikes dramatically. It appears from the plot that there is a 50 A peak-to-peak current that lasts for less than $\frac{1}{4}$ second.
The current plot of Figure 5.17 and the simulated current plot in Figure 4.31 show a dramatic rise in current after the transition from two stators to one stator. As shown in the analysis of Section 4.10, when the motor transitions from two stators to one stator, the motor’s BEMF and impedance are suddenly cut in half. The current regulator in the field oriented controller must reduce the supply voltage to the motor quickly in order to prevent a high current transient from occurring. The large current transient in Figure 5.17 would indicate that the controller is not fast enough as implemented to prevent this current transient. This current transient is most certainly the cause of the transient speed increase documented in Section 5.7.2.1.

Figure 5.18 shows the phase current when the motor transitions from one stator back to two stators. There is a significant current spike when the switching occurs, but the current dampens quickly to its steady state level. What is not clear from Figure 5.18 is
the phase relationship of this current compared to the supply voltage. Therefore, it is not clear from Figure 5.18 if this current transient is motoring current or regenerative current. The simulation of Figure 4.41 would suggest that this current transient is regenerative. The simulation results of Figure 4.45 shows that the motor is generating after the transition from one stator to two stators until the field oriented controller’s current regulation loops can increase the supply voltage to regulate the current to the desired setpoint. If this current transient were indeed regenerative, that would also explain the transient speed decrease documented in Section 5.7.2.1.

![Figure 5.18 Current During Transition from One Stator to Two Stators](image)

5.7.2.3 Voltage Transients

The next thing examined was motor phase voltage. It was believed that there might be BEMF transients associated with actually running a motor with some stator current that were not present when the BEMF transient test of Section 5.7.1 was performed. Figure 5.19 shows the line-to-line voltage before during and after the transition from two...
stators to one stator. The peak-to-peak magnitude of this PWM voltage represents the DC bus voltage magnitude. Notice that during the transition period, the bus voltage appears to decrease before coming back up to the previous value. This can also be seen in the DC bus voltage plot of Figure 5.20. Here too, the DC bus voltage decreases instantly and then slowly builds back up to the previous value. In Figure 5.20, the bus voltage drop can be measured as approximately a 17.5 volt drop. The differential voltage probe that was used had a scale of 50 mV/div on the scope corresponding to a real value of 25 V/div.

![Figure 5.19 Line-to-Line Motor Voltage During Transition from Two Stators to One](image)

The plots of Figures 5.21 and 5.22 show that when the motor transitions from one stator back to two stators, the DC bus voltage actually increases. From Figure 5.22, the measured bus voltage rise is approximately 11 V. Again, the differential voltage probe that was used had a scale of 50mV/div on the scope corresponding to a real value of 25
V/div. This rise occurs instantaneously, and takes approximately 280 ms to return to the previous value of 28 V.

Figure 5.20 DC Bus Voltage During Transition from Two Stators to One

Figure 5.21 Line-to-Line Motor Voltage During Transition from One Stator to Two
The drop in DC bus voltage on the transition from two stators to one indicates a sudden rise in current. This is consistent with the current measurement of Figure 5.17 and the simulated current of Figure 4.31.

The rise in DC bus voltage during the transition from one stator to two stators indicates either a sudden drop in motor current or regenerative current from the motor. The simulation results of Figure 4.45 indicate that there is regenerative current after the transition from one stator to two stators. This regenerative current is likely brief as the velocity loop of the field oriented controller adjusts the current loop command to regulate the motor current. Additionally, when the transition occurred, there was an audible noise from the motor that indicated the motor was under load. With the motor having no

Figure 5.22 DC Bus Voltage During Transition from One Stator to Two
mechanical load connected, this sound is an indicator of the motor being loaded as a generator.

5.7.2.4 The Open Stator

During the transitions, voltage and current measurements were made on the stator that is being removed from the circuit. This stator is Stator1 on the left hand side of Figure 3.3. The first measurement made was a line-to-line voltage measurement of the Stator1 leads. Figure 5.23 and Figure 5.24 show line-to-line voltage measurements for the two stator to one stator transition and the one stator to two stator transition, respectively. Examination of Figure 5.23 shows that there is considerable contact bounce that occurs. It also appears that the contactor S2 is closing before the contactor S1 releases. This can be seen from the fact that the PWM pulses end due to S2 closing, but the sinusoidal average is still present which is likely due to current that is still flowing in Stator1. Then at the end of the transition, there is a large voltage spike followed by a clean sine wave. This large spike is likely when the S1 finally opens. The large spike is due to the current disruption in Stator1, and the clean sine wave is the BEMF of the isolated stator after the transition is complete. A timing analysis of Figure 5.23 shows that on a transition from two stators to one stator, there is approximately 7 ms of overlap when both the S1 and S2 contacts are closed.
In Figure 5.24 Stator1 is brought back into the circuit. The voltage initially is a clean BEMF of the open circuit stator. Then as the contactor S1 closes (and bounces several times) the BEMF is shorted as it appears that S2 is still closed. Finally, S2 opens
and the voltage of the inverter is seen across the Stator1 coil. This final voltage is actually the inverter voltage minus the voltage drop of Stator2 as the voltage is being measured at the junction between Stator1 and Stator2. Also, a timing analysis shows that there is an overlap of the S1 and S2 contacts of approximately 3 ms when both S1 and S2 contacts are closed.

The current on one of the phase leads for Stator1 was then measured to see how it behaved during transition. The plots of Figure 5.25 and Figure 5.26 show the two stator to one stator transition and the one stator to two stator transitions respectively.

![Figure 5.25 Stator1 Phase Current for the Two Stator to One Stator Transition](image)

For the transition from two stators to one stator, the current starts out with the inverter fed current that is flowing through both stators. When contactor S2 closes and bounces, there are several large transients as S2 bounces. Then when S1 opens, the
current drops to zero and the stator is isolated from the circuit. The small transient at the end just before the current goes to zero is likely due to current still flowing through contact arcing. This would correspond with the very large voltage spike shown in Figure 5.23 at the end of the transition.

![Figure 5.26 Stator1 Phase Current for the One Stator to Two Stator Transition](image)

For the transition from one stator back to two stators, the Stator1 starts out with zero current because it is isolated from the rest of the circuit. When contactor S1 closes before S2 opens, there is a short circuit across the phase leads of Stator1. This causes current to flow out of Stator1 for approximately 3 ms during the time that S1 and S2 are both closed. This too would tend to cause a drop in motor speed as the short circuit across the Stator1 has a braking affect for this brief 3 ms period of time.
6.1 Problem Summary

Axial flux motors were one of the first motors invented. Due to manufacturing considerations, radial flux motors were easier to commercialize and therefore they have been much more popular. Now with the advent of rare earth magnets, there is renewed interest in axial flux technology.

It has been shown that permanent magnet brushless motors (both PMSM and BLDC) have a very high power density, but they suffer from a limited speed range. Motors optimized for high torque cannot spin very fast and motors optimized for high speed cannot produce high torque. A number of solutions and techniques exist today to overcome this shortcoming. Most of the alternative solutions have one thing in common; they reduce $K_r$ by reducing the flux linkage between the PM rotor and the stator. One such technique involves physically moving the rotor of an axial flux motor away from the stator to reduce the flux linkage. This method is primarily utilized in the one rotor, one stator axial flux configuration. This technique served as the inspiration of another technique called VMAG for the one rotor, two stator configuration of the axial flux motor. In the maximum implementation of the VMAG concept, the dual stator axial flux...
motor is manufactured with unequal air gaps. During motor operation, as the speed increases toward the value where the motor’s BEMF equals the voltage of the DC bus, one of the motor’s two stators is turned off and isolated from the circuit. The result is a motor with a reduced $K_e$ that can continue to accelerate beyond the no load speed of the original motor. The feasibility of the VMAG concept was simulated using Simulink, the experimental tests were carried to prove the concepts. Simulations and experimental results showed that VMAG technique successfully extend the operating range of the axial flux permanent magnet machines.

6.2 Data Summary and Conclusions

The data presented in Chapter V showing the improvements in speed range had a strong correlation with the simulation results of Chapter IV. It was shown that the motor had approximately half the BEMF constant $K_e$ when only one stator was in the circuit. This was confirmed both directly by measurement of the BEMF and by observation of the steady state current required running at a fixed speed in both stator configurations. Consequently, the motor could run approximately twice the speed with only one stator. This was confirmed through encoder frequency measurements with an oscilloscope as well as the speed variable observed in the Code Composer Studio’s watch window. Unfortunately there is a substantial speed disturbance when the transitions occur. The motor suddenly accelerates when transitioning to single stator mode and suddenly decelerates when transitioning to dual stator mode. This effect is related to the fact that during the transitions, current is not well regulated.
There appears to be a 3 ms and a 7 ms overlap of the S1 and S2 contactors used in this experiment. This means that during the two transition periods there is a period of either 3 ms or 7 ms when both contactors are closed. This induces a circumstance that is not modeled in the current loop control and the current control cannot regulate. This phenomenon contributes to the motor speed transients during the transitions.

Based on the experimental results, an additional circuit simulation was done and presented in Chapter IV. This simulation included several scenarios related to the timing of the S1 and S2 switches, including the 7 ms overlap observed in the experiment as well as the observed contact bounce. The simulation showed that a voltage transient of approximately 20MV was present when the S1 switch opened regardless of switch timing and contact bounce. In the case of the contactors used in the experiment, this high voltage transient would most likely result in arcing across the contacts and therefore some continued current flow for some duration. This current flow would allow Stator1 to discharge. The simulation also showed that if the supply voltage is held constant there are very large motor currents after a transition from two stators to one stator. These high motor currents would be most likely the cause of the transient speed increase that occurs when this transition occurs.

The simulation in Section 4.10 also showed that during the transition from one stator back to two stators, the BEMF suddenly doubles and becomes higher than the supply voltage. This causes the motor to regenerate until the motor controller adjusts the supply voltage to regulate the current to the desired level. This regenerative current is most
likely the cause of the transient speed reduction that occurs when the motor transitions from one stator to two stators.

The VMAG concept applies to motors with equal air gaps and with unequal air gaps. Due to lack of ability to safely modify the test motor, the experiment dealt only with a motor having equal air gaps. With the motor tested in the experiment, there was approximately a 100% increase in the no load speed. Motor torque was not measured during the experiment but based on the estimated $K_e$ and $K_t$ values, we can say that the low speed torque production capability should not be affected by the VMAG operation.

The simulations of Chapter IV show that when VMAG is applied to a motor with unequal air gaps of 1.5 mm and 5 mm that motor loses approximately 20% of its low speed torque but experiences an increase in no load speed of approximately 100% over a motor with equal air gaps and no VMAG.

It is clear from this research that some control issues with VMAG are yet to be resolved. However, the VMAG has been proven in this thesis to be a promising solution to the need for a motor that can operate over a wide speed range with high torque production capability at low speeds.

6.3 Future Research Recommendations

The simulations and experimental tests with the VMAG have produced some promising results. More research on this topic is needed. The following items could be pursued as future research projects to further develop this concept.
• The first issue to be resolved is the velocity control issue during stator transitions. Areas to investigate are:
  
  o Switch closure timing.
  
  o Switch type selection. (Semiconductor switches)
  
  o Switch closure techniques. (Operating switches briefly in the linear region might reduce transients.)
  
  o Adding a capacitor in parallel with Stator1.
  
  o Increasing the bandwidth of the current regulation loops.
  
  o Employing a “feed-forward” technique to create an anticipatory output of the current regulator to prevent current transients at the time of the transitions from two stators to one stator or from one stator to two stators.

• The second issue to address is actually testing the motor’s torque vs. speed capabilities with VMAG operational.
  
  o Using a dynamometer that can provide up to 70 Nm of torque and a DC bus that can supply up to 300 A DC to test torque vs. speed.

• The third issue to address is expanding the benefits of VMAG by controlling a motor with unequal air gaps.
  
  o Modifying the motor mechanically to extend one of the air gaps.
- Repeating the experimental tests of Section 5.5 and Section 5.6 to determine the motor’s BEMF and no load speed improvements.

- Testing the motor’s torque vs. speed capability on a dynamometer.
REFERENCES


APPENDICES
APPENDIX A

SIMULATION PARAMETERS

%%%%%  Virtual Moving Air Gap Simulation
%%%%%  Init File
%%%%%  Brad Mularcik Masters Thesis
%%%%%  2011
poles=24

% J=0.5 %kg-m^2
J=0.05 %kg-m^2

Rs_base=0.012/2 %measured line-line converted to line-neutral for wye
Ls_base=50e-6/2 %measured inductance line-line converted to line-neutral for wye
ke_base_L_L=0.015 %volt/RPM measured line to line
ke_base_L_L_rad=(2/poles)*ke_base_L_L*30/pi %convert volts/rpm mechanical to
volts/(rad/s)electrical
ke_base=ke_base_L_L_rad/sqrt(3) %convert ke line to line to ke line to neutral since
%machine is connected in wye
lambda_fd_base=ke_base*(3/2)
Rs=Rs_base
Ls=Ls_base
ke=ke_base
lambda_fd=lambda_fd_base
damp_coef=0
dwell_time=.3

% torq_cmd=[135.3, 30.6, -74.1, -135.3, -30.6]
%torq_cmd=[45, 30.5, -65, -30.5, 65]
%torq_cmd=[65, 65, 65, 40, 25, 25]
torq_cmd=[35, 35, 35, 35, 35]
torq_time=[dwell_time, dwell_time+0.2, dwell_time+.6, dwell_time+.7, dwell_time+1, dwell_time+1.75]
torq_load=[-20, .25,.25]
load_time=[dwell_time+0.05, dwell_time+.75]

Kp=2
Ki=0
Kpq=Kp
Kpd=Kp
Kiq=Ki
Kid=Ki
Vd=28
fs=16000
Vtri=5

% With 1.5mm as the standard air gap, the following BEMF scaling percentages exist for extended air gaps. These are per air gap.
% 1.5mm 100%
% 3mm 82%
% 5mm 59%
% 10mm 22%
% G3=.59 %This is the percentage of BEMF for the extended air gap
% It is the flux density after VMAG is activated as compared
% to before vmag was activated. This would be 1 if the
% air gaps were =.

G3=.59
G2=G3*.5
G1=0.5+G2 %This gain represents the BEMF scaling before VMAG
        % is activated. With equal air gaps, G2 would =1 and
        % therefore G1 would = 1.
VMAGspeed=900  % Set this to an extremely high number to disable VMAG.
The following is a detailed description of the signal conditioning hardware that was designed and built for the VMAG experiment.

The TI F28335 DSP included more than enough general purpose IO (GPIO) for the control of the axial flux motor. The GPIO were all 3.3V logic level IO. There were also several analog to digital converter channels on the DSP that were used for analog measurements. These analog inputs had a 0V – 3V input range.

The Semikron SKAI inverter used required the PWM input signals to be a 5V logic level. As was stated previously, the DSP had GPIO that were 3.3V logic level IO. In order for the DSP to sufficiently drive the PWM signals to the inverter, a TI SN74LVC245 was used to level shift the 3.3V logic to 5V logic levels. The six PWM signals from the DSP to the inverter pass through the SN74LVC245 (U1). The inverter also has current sensing on two of the phases. These current sense channels have an analog output of -10 V to +10 V representing -1000 A to +1000 A. Again, the DSP had an input range of 0 V to 3 V. Therefore, an opamp circuit was designed to adjust the scale an offset of the current sense signals to scale the -10 V to +10 V down to a 0 V to 3
V signal. The voltage dividers R4,5 and R6,7 provide the required scaling and the 1.5 V reference voltage provides the required offset. The voltage divider is then followed by the unity gain buffer U2. The same circuit was repeated to provide the DC bus voltage feedback. The SKAI inverter had an analog output of 0 V to +10 V representing a DC bus voltage of 0V to 177 VDC. The voltage divider R8,9 provides the required scaling. The reference voltage of 1.5 V that was used was not really the ideal reference voltage, but this error was accounted for in the input offset scaling in the DSP software. The opamp U3 provides the unity gain buffer from the voltage divider to the analog input.

The incremental encoder used on the axial flux motor provided a differential output for A/A#, B/B# and Z/Z#. These three signals needed to be converted to 0-3.3 V single ended logic signals. To do this, the AM26LS323AC (U5) was used to convert the encoder signals to single ended 5V logic signals. A second SN74LVC245A (U4) was used to level shift these to 3.3 V logic signals that were connected to GPIO’s on the DSP that were designated for use with a quadrature encoder.

The DSP was also required to drive the two VMAG switches S1 and S2. In this experiment, two Allen Bradley Bulletin 100 contactors were used. These contactors used a 24 VDC coil to drive three contacts rated at 43 A. In order for the DSP to drive the 24 VDC coils, two GPIO signals were used. These were passed through U1 to level shift them to 5V logic level signals. The 5V logic level coil drive signals gates of Q1 and Q3 which drive Q2 and Q4. These turn on coils S1 and S2 respectively. Not shown on the schematic C.1 are the PWM DAC signals used for diagnostic purposes. Three GPIO signals GPIO8, GPIO10, GPIO11 are connected to an RC network of 470 Ω and 470 μF.
This allows the DSP to generate analog diagnostic signals by simply generating a PWM signal of varying duty cycle. At different times during the experiment, different variables in the DSP program were mapped to one of the three PWMDAC outputs so that their values could be observed on the oscilloscope.

The SKAI inverter also required a 24 VDC power supply. This power supply was connected to the interface breadboard and wired through the J10 connector to the inverter in a 20 pin ribbon cable along with the other interface signals. This same 24 VDC power supply was used for the drive circuits for the S1 and S2 contactors.
Figure B.1 DSP Interface Schematic
The Virtual Moving Air Gap Experiment utilized the following Bill of Materials.

Table B.1 Bill of Material

<table>
<thead>
<tr>
<th>Description</th>
<th>Value</th>
<th>Manufacturer</th>
<th>Part Number</th>
<th>Designator</th>
</tr>
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<tbody>
<tr>
<td>DSP</td>
<td></td>
<td>Texas Instruments</td>
<td>TMS320F28335</td>
<td></td>
</tr>
<tr>
<td>Experimenter Kit for TMS320F28335</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>including docking station with USB</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>emulation</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Octal Bus Transceiver, 3.3V to 5V</td>
<td></td>
<td>Texas Instruments</td>
<td>SN74LVC245A</td>
<td>U1, U4</td>
</tr>
<tr>
<td>Shift with 3-state outputs OpAmp</td>
<td></td>
<td>National Semiconductor</td>
<td>LMH6646MM</td>
<td>U2, U3</td>
</tr>
<tr>
<td>Quadruple Differential Line Receiver</td>
<td></td>
<td>Texas Instruments</td>
<td>AM26LS32AC</td>
<td>U5</td>
</tr>
<tr>
<td>N Channel MOSFET</td>
<td></td>
<td>Fairchild</td>
<td>BSS138</td>
<td>Q1, Q3</td>
</tr>
<tr>
<td>P Channel MOSFET</td>
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<td>Fairchild</td>
<td>NDT2955</td>
<td>Q2, Q4</td>
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<tr>
<td>Resistor 1K ohm</td>
<td></td>
<td></td>
<td></td>
<td>R1, R2, R3</td>
</tr>
<tr>
<td>Resistor 1.5K ohm</td>
<td></td>
<td></td>
<td></td>
<td>R4, R6, R8</td>
</tr>
<tr>
<td>Resistor 10K ohm</td>
<td></td>
<td></td>
<td></td>
<td>R5, R7, R9</td>
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<tr>
<td>Resistor 150 ohm</td>
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<td></td>
<td></td>
<td>R10, R11, R12</td>
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<tr>
<td>Resistor 100 ohm</td>
<td></td>
<td></td>
<td></td>
<td>R13, R16</td>
</tr>
<tr>
<td>Resistor 402 ohm</td>
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<td></td>
<td></td>
<td>R14, R15, R17, R18</td>
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<tr>
<td>Capacitor 108uF</td>
<td></td>
<td>Semikron</td>
<td>SKA14201MD20-1452W</td>
<td>C1</td>
</tr>
<tr>
<td>20pin Dsub (F)</td>
<td></td>
<td>Allen Bradley</td>
<td>100-C43EJ01</td>
<td>S1, S2</td>
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<tr>
<td>20 Pin Ribbon Cable</td>
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<td>Light Engineering</td>
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</tr>
<tr>
<td>Inverter</td>
<td></td>
<td>Encoder Products Co.</td>
<td>755A-02-S-4096-</td>
<td></td>
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<tr>
<td></td>
<td></td>
<td></td>
<td>R-HV-5-S-S-N</td>
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</tr>
<tr>
<td>Contactors</td>
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<td></td>
<td></td>
</tr>
<tr>
<td>Axial Flux PMSM Motor</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Encoder 4096 Quadrature</td>
<td></td>
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</tr>
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</table>
APPENDIX C

MOTOR SPECIFICATIONS

- $K_e$ 15 V RMS /kRPM  Line-to-Line
- Continuous Torque 48 Nm @1000 RPM, Peak Torque 65 Nm @1000 RPM
- Rated Current 200 A RMS, Peak Current 300 A RMS
- Rated Power 14.3 kW @3000 RPM, Peak Power 17 kW @1000 RPM
- Inductance Line-to-Line 50 µH
- Resistance Line-to-Line 12 mΩ
- $K_t$ 0.223 Nm/A
- 24 poles (12 pole pairs)
- Liquid Cooled

Field oriented control requires the value of $\lambda_{FD}$. The following calculations were used to determine $\lambda_{FD}$ from $K_e$. $K_e$ must be expressed in units of volts/(radians/second) or volt-seconds/radian.

The relationship between $K_e$ and $\lambda_{FD}$ is shown in [8] and given here in equation (C.1). However, this is the power invariant form. Review of [5] gives us equation (C.2).

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Plugging in a value of $K_e$ along with algebraic manipulation yields the resulting $\lambda_{FD}$ as given in equation (C.10).

\[ K_e = \frac{\sqrt{3}}{\sqrt{2}} \lambda_{FD} \]  
\[ (C.1) \]

\[ K_e = \frac{3}{2} \lambda_{FD} \]  
\[ (C.2) \]

\[ \lambda_{FD} = \frac{2}{3} K_e \]  
\[ (C.3) \]

\[ K_e = \frac{K_{el-L}}{\sqrt{3}} \]  
\[ (C.4) \]

From motor data $K_{el-L}$ is 15V/kRPM

\[ K_e = \frac{0.015 \text{ V/RPM}}{\sqrt{3}} = 0.00866 \text{ volt min/rev} \]  
\[ (C.5) \]

\[ K_e = 0.00866 \frac{60}{2\pi P} \text{ volts/(radian/second)} \]  
\[ (C.6) \]

\[ K_e = 0.0068916 \text{ volts/(radian/second)} \]  
\[ (C.7) \]

\[ \lambda_{FD} = \frac{2}{3} \times 0.0068916 \text{ volt-seconds} \]  
\[ (C.8) \]

\[ \lambda_{FD} = \frac{2}{3} \times 0.0068916 \text{ volt-seconds} \]  
\[ (C.9) \]

\[ \lambda_{FD} = 0.00459 \text{ volt-seconds} \]  
\[ (C.10) \]
APPENDIX D

ENCODER ALIGNMENT PROCESS

To properly commutate the motor, the θ from Figure 5.5 used for the Park and Inverse Park transforms must be aligned to the a phase of the motor. The encoder was randomly mated to the rotor shaft. The relationship between θ and phase a-b BEMF shown in Figure 5.6 is the uncorrected value that is the result of this random mounting. In order for to align θ to phase a, an offset must be added to the encoder raw θ. Since the motor is wired in such a way that the neutral is embedded and not accessible, the phase a to phase b must be used as a reference to adjust the commutation angle θ. To get the phase a BEMF from the phase a-b BEMF, a 30 degree phase shift is applied. As stated previously, the BEMF plot of Figure 5.6 is a plot of the phase a-b BEMF along with the electrical θ value. It can be seen that by coincidence, the electrical θ nearly aligns with the phase a-b BEMF. In order to have the correct alignment of the electrical θ, the raw theta must be adjusted by the amount corresponding to 30 electrical degrees. As stated previously, the software used for this experiment was a modified version of sample FOC code provided by TI. In that code, the value of electrical θ is the result of

\[ \text{Elec}_\theta = \text{polepairs} \times \text{mech} \_ \text{scaler} \times \text{raw} \_ \theta . \]  

(D.1)
Based on the encoder resolution used, the value of mech_scaler is given by

\[ mech\_ scaler = \frac{0.25}{4096}. \]  

(D.2)

The offset variable in the software is qep1.CalibratedAngle. This is the amount of offset to add to the raw \( \theta \) in order to achieve the desired offset in electrical \( \theta \). Solving equation (D.1) for raw \( \theta \) yields

\[ raw\_\theta = \frac{elec\_\theta \times \text{polepairs} \times mech\_\text{scaler}}{elec\_\theta}. \]  

(D.3)

Applying the specific values to equation (D.3) yields

\[ raw\_\theta = \frac{elec\_\theta}{12 \times \frac{0.25}{4096}} = elec\_\theta \times \frac{4096}{3}. \]  

(D.4)

An offset of -30 electrical degrees is desired to achieve proper commutation. A -30 electrical degrees is the same as 330 electrical degrees. From Appendix C, polepairs=12.

Note that elec\_\theta is a per unit variable. Therefore for 330 electrical degrees in PU,

\[ elec\_\theta = \frac{330}{360} \]  

(D.5)

and

\[ raw\_\theta = \frac{330}{360} \times \frac{4096}{3} = 1251.5. \]  

(D.6)
Based on this calculation, the value in the software for encoder offset (qep1.CalibratedAngle) was set to 1251. The result was proper commutation of the motor. For this experiment, the motor seemed to run well. Running the motor under load was not required and in fact was not practical. Therefore no tests were conducted to verify the proper encoder offset was being used. The best test of this is to test the value of $K_t$ in both the forward and reverse directions. If the motor produces the same torque in both directions for a given current magnitude, then the encoder offset is properly set.
### APPENDIX E

**MOTOR CONTROL PARAMETERS**

Table E.1 Motor Parameters for Two Stator Operation

<table>
<thead>
<tr>
<th>Pole Pairs</th>
<th>12</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator Resistance (line-to-neutral)</td>
<td>0.006 ohms</td>
</tr>
<tr>
<td>Stator Inductance (line-to-neutral)</td>
<td>$25 \times 10^{-6}$ henries</td>
</tr>
<tr>
<td>BEMF Constant $K_e$ (line-to-line)</td>
<td>0.015 volts/RPM</td>
</tr>
<tr>
<td>BEMF Constant $K_e$ (line-to-neutral) SI Units</td>
<td>0.0068916 volts/(radian/second)</td>
</tr>
<tr>
<td>$\lambda_{FD}$</td>
<td>0.00459 volt sec</td>
</tr>
</tbody>
</table>
Table E.2  Motor Parameters for One Stator Operation

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pole Pairs</td>
<td>12</td>
</tr>
<tr>
<td>Stator Resistance (line-to-neutral)</td>
<td>0.003 ohms</td>
</tr>
<tr>
<td>Stator Inductance (line-to-neutral)</td>
<td>$12.5 \times 10^{-6}$ henries</td>
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<tr>
<td>BEMF Constant $K_e$ (line-to-line)</td>
<td>0.0075 volts/RPM</td>
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<tr>
<td>BEMF Constant $K_e$ (line-to-neutral) SI Units</td>
<td>0.0034458 volts/(radian/second)</td>
</tr>
<tr>
<td>$\lambda_{FD}$</td>
<td>0.002295 volt sec</td>
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