AN ABSTRACT OF THE THESIS OF

Aaron H. VanderMeulen for the degree of Master of Science in Electrical Engineering presented on June 14, 2007.

Title: Novel Control of a Permanent Magnet Linear Generator for Ocean Wave Energy Applications.

Abstract approved: ____________________________________________

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                             Annette von Jouanne

Wave energy conversion devices are a rapidly growing interest worldwide for the potential to harness a sustainable and renewable energy source. Due to the oscillatory nature of ocean waves, the power generated from a permanent magnet linear generator (PMLG) for ocean wave energy conversion is pulsed. Focusing on direct drive technology, the PMLG directly translates the motion of the waves into electrical energy. The power generated, left unconditioned, is not easily used or stored.

With conventional diode rectification topologies, line currents cannot be controlled easily, resulting in an uncontrolled generator output and force. With an active rectifier topology, the real and reactive power from the PMLG is fully controllable. This thesis will investigate the generator modeling and design of a novel three-phase active rectifier topology and force controller with a dc-dc converter for bus voltage regulation. An in-depth analysis for the controller design and simulations are presented. Hardware for the three-phase active rectifier is specified and built with initial lab test results. The controller design is implemented with National Instruments’ LabView and compiled on a CompactRIO real-time controller.
Novel Control of a Permanent Magnet Linear Generator for Ocean Wave Energy Applications

by
Aaron H. VanderMeulen

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APPROVED:

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Dean of the Graduate School

I understand that my thesis will become part of the permanent collection of Oregon State University libraries. My signature below authorizes release of my thesis to any reader upon request.

Aaron H. VanderMeulen, Author
I would like to thank my major professor, Dr. Ted Brekken for his guidance, enthusiasm and sincerity during my experience with the Energy Systems Group. I would also like to thank my co-major professor Dr. Annette von Jouanne for her support, guidance and passion with my research and time with the Energy Systems Group. Special thanks go to the late Dr. Alan Wallace. His passion and knowledge are greatly missed and will always be remembered.

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1 INTRODUCTION

1.1 Background

Wave energy conversion devices are a rapidly growing interest worldwide for the potential to harness a sustainable, predictable and almost unlimited energy resource. Water has a much higher density than that of air thus the dimension of the energy converting device takes up less space compared to that of wind turbines. Wave energy is a form of concentrated solar power originating from the uneven heating of the earth creating wind and wind in turn creating waves. The waves gather energy across vast stretches of ocean resulting in high power energy sources near coastal shores. The wave energy system presented utilizes the heave (vertical) motion of the wave. Therefore the output power will be modulated at the wave frequency, approximately 5 to 10 seconds. This pulsed power needs to be conditioned and regulated for connection to a utility grid. Advancements in power electronics technology has made wave energy power production possible with maximum efficiency and maximum power extraction from the wave.

1.2 Wave Energy

Ocean energy conversion encompasses ocean waves, ocean tides and ocean currents as a source to extract electrical energy. There are various mechanical devices currently deployed that convert ocean waves into electrical energy. Such devices include Ocean Power Delivery’s Pelamis Wave Energy Converter and Ocean Power Technology’s PowerBuoy. These devices translate ocean wave motion into electrical energy mechanically via a hydraulic system to a rotary generator. This added intermediary step of mechanical components adds to system losses and maintenance with increased moving parts. At Oregon State University, the Energy Systems group is
focusing on wave energy converters that eliminates the mechanical linear to rotary conversion altogether. This thesis primarily focuses on direct drive technology employing linear electric machines.

Ocean wave devices translate kinetic motion into linear motion from the wave excitation force. This force moves the buoy float vertically along the spar, creating the relative motion between generator components in the heaving float vs. the stationary spar. The spar is moored to the sea floor, making it relatively stationary.

The excitation force will move the float linearly with a velocity. The relative motion between the permanent magnets and the coils will generate the electrical energy. Faraday’s law explains how a change in a magnetic field relative to a coil will induce a voltage within the coil. The relative motion of the permanent magnets relative to the coils in a direct drive linear generator is the basis on which electrical energy is created. Len’s law describes the magnetic field produced by the coils acting in the opposite direction of the changing magnetic field which produced it. This creates a constant magnetic flux within the active region and produces an opposing generator force. For the direct drive linear generator, such devices are built to generate high voltages to reduce the amount of current drawn through the coils.

Ocean waves have varying wave periods and height determined by winds and the distance traversed. The height of a wave is defined by the distance from the crest (peak) to the trough (low point). The period of the wave is determined by the distance from crest to crest. Data collected off of the Oregon Coast by NOAA (National Oceanographic and Atmospheric Association) buoys show a trend seen in figures 1.1 and 1.2.
Fig. 1.1 – Average wave period

Fig. 1.2 - Significant wave height
For computer simulations of the generator (generator and buoy system) interface with the power electronics, vertical velocities will be varied in order to generate different voltage levels. The maximum vertical velocity will determine the maximum output voltage and thus the power electronics will need to be designed to handle this output voltage.

Figure 1.3 shows the progressive surface wave parameters for a monochromatic wave traveling at a phase celerity (phase velocity), C. Other defining parameters are the wave height, H, in meters, wave length, L, in meters, and wave depth, d, in meters. The wave velocity is defined by the wave length, L, and wave period, T. [1]

\[
C = \frac{L}{T} \quad (1.1)
\]

As the wave front travels from left to right, the motion of the particles are shown by the arrows in figure 1.3. The orbiting dimensions decrease to zero as depth increases. At the surface a water particle will experience an upward vertical velocity from the incoming wave front. The velocity is represented by equation 1.1. The generator will be considered a particle on the surface of the wave, where \( z = 0 \) and any damping or
phase shifting is neglected. Therefore, the generator will be considered to be a wave follower. The vertical velocity of a particle is shown in equation 1.2.

\[ w_s = \frac{\pi H}{T} e^{kz} \sin(kx - \sigma t) \]  
(1.2)

\[ k = \frac{2\pi}{L} \]  
(1.3)

\[ \sigma = \frac{2\pi}{T} \]  
(1.4)

Equation 1.3 is the wave number and equation 1.4 is the wave angular frequency. For investigation, the maximum velocity of the generator will be at position \( z = 0 \). The vertical velocity is then reduced to:

\[ w_c = \frac{\pi H}{T} \sin(kx - \sigma t) \]  
(1.5)

At an arbitrary position, \( x = 0 \), the velocity profile of a particle on the surface of a wave is shown in figure 1.4. The wave height is \( H=1.5\text{m} \), wave period of \( T=6\text{sec} \) and water depth of \( 45\text{m} (150\text{ft}) \).

Fig. 1.4 - Surface particle velocity
For our investigations, H ranges from 1m to 3m. Keep in mind that the maximum buoy travel for the 1kW generator is 1m, but the velocity of the buoy will change with wave height. The wave length will be fixed at 91 meters, the average wave length. For ideal monochromatic wave generation, the wave period will be varied to generate a range of output voltages from the generator. The reason for varying the wave period is explained in the ideal wave generator model section.

The peak linear velocity can be found using wave height, $H_o$, and wave period $T_o$, using equation 1.6.

$$w_c = \frac{\pi H_o}{T_o} \quad (1.6)$$

<table>
<thead>
<tr>
<th>Wave Period (s)</th>
<th>Wave Height (m)</th>
<th>Velocity (m/s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>1.5</td>
<td>0.785</td>
</tr>
<tr>
<td>8</td>
<td>1.5</td>
<td>0.589</td>
</tr>
<tr>
<td>10</td>
<td>1.5</td>
<td>0.471</td>
</tr>
</tbody>
</table>

Table 1.1 – Peak velocity at 1.5m wave height

<table>
<thead>
<tr>
<th>Wave Period (s)</th>
<th>Wave Height (m)</th>
<th>Velocity (m/s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>3.0</td>
<td>1.571</td>
</tr>
<tr>
<td>8</td>
<td>3.0</td>
<td>1.178</td>
</tr>
<tr>
<td>10</td>
<td>3.0</td>
<td>0.942</td>
</tr>
</tbody>
</table>

Table 1.2 – Peak velocity at 3.0m wave height

### 1.3 Power Electronics

The field of power electronics has rapidly expanded allowing for the construction of new devices that were not possible even a decade ago. New materials and production methods have allowed for higher switching frequencies, increased current, high voltage, and higher power capabilities. High powered IGBTs (Insulated Gate
Bipolar Transistors) now have faster switching frequencies which makes them competitive with fast-switching FET (Field Effect Transistor) devices. However, the FET devices do not allow the higher power handling capabilities of the IGBT; the IGBT still is ideal in higher power switching topologies.

With different active rectifier front-end topologies, it is possible to control the real and reactive power flow in and out of a generator. The generator variable voltage variable frequency output is not readily usable since the power output is pulsed due to the low frequency excitation force. For example, if an incandescent light bulb is placed on the terminals, it would flash on and off with twice the electrical frequency output of the generator. The power electronics described in this thesis will interface between the generator terminals and a dc-dc converter. The dc-link will provide a stiff bus voltage, temporary energy storage and an interface to a loading system.
2 GENERATOR MODELING

The generator model will interface with the power electronics and controls components for ideal and dynamic system simulations. The ideal wave model will interface with the SimPowerSystems blocks in MATLAB/Simulink as well as the average model of the power electronics. The dynamic generator model will interface with the average switching model of the power electronics and the stochastic wave environment.

2.1 Ideal Model

The permanent magnet linear generator (PMLG) is designed for a maximum of 1 meter vertical displacement, limited by the active magnetic region, and a speed range from 0 to 3 m/s. These parameters are used in the construction of a MATLAB/Simulink ideal wave source model used to test all power electronic topologies. The ideal source produces a monochromatic wave used as a baseline. The monochromatic wave output only represents a single wave envelope frequency versus a stochastic ocean wave environment where many harmonic frequencies exist. The monochromatic wave output is considered for understanding of the generator. The ideal wave model input variables required are changes in wave period and changes in output voltages.

The ideal source is derived mathematically based on magnetic and electrical properties, as well as wave mechanics. Vertical displacement of the generator depends on the maximum range associated with a specific generator, $d$ in equation 2.1. The maximum distance traveled for the PMLG in consideration is 1m. The generator will be displaced vertically due to the wave excitation force. This force in ideal conditions is sinusoidal. The vertical displacement, $y(t)$, is shown in equation 2.2, where $\omega_m$ is
the ocean wave frequency in rad/sec and the maximum generator travel, $d$, in meters.

$$y(t) = \frac{d}{2} \sin(\omega_m t)$$  
$$\omega_m = 2\pi f_m = \frac{2\pi}{T_m}$$  

(2.1)

The flux seen by the coils within the spar, respect to time (zero initial conditions) is shown in equation 2.2. The variable $\Phi$ is the peak flux produced by the permanent magnets in Tesla and $\lambda$ is the magnetic wavelength in meters. The pole pitch for the linear generator is half of the magnetic wavelength.

$$\phi(t) = \Phi \sin\left(\frac{2\pi}{\lambda} y(t)\right)$$  

(2.2)

![Fig. 2.1 - PMLG Cross-sectional area.](image)

The voltage induced in the coils can be described by Faraday’s Law by equation 2.3, where $N$ is the number of turns per coil, and the change in flux. Differentiating the flux associated with time, equation 2.4, results in the per-phase voltage. $\hat{V}$ is the peak phase-to-neutral voltage. Since the linear generator is a three-phase machine, each phase is electrically phase shifted 120 degrees, shown in equation 2.4 and 2.5.
\[ v(t) = N \frac{d\phi}{dt} \]  \hspace{1cm} (2.3)

\[ v(t) = \dot{V} \cos(\omega_m t) \cos\left(\frac{\pi d}{\lambda} \sin(\omega_m t) + \vartheta \right) \]  \hspace{1cm} (2.4)

\[ \vartheta = 0, +/ - \frac{2\pi}{3} \]

\[ v_a(t) = \dot{V} \cos(\omega_m t) \cos\left(\frac{\pi d}{\lambda} \sin(\omega_m t) \right) \]

\[ v_p(t) = \dot{V} \cos(\omega_m t) \cos\left(\frac{\pi d}{\lambda} \sin(\omega_m t) - \frac{2\pi}{3} \right) \]  \hspace{1cm} (2.5)

\[ v_e(t) = \dot{V} \cos(\omega_m t) \cos\left(\frac{\pi d}{\lambda} \sin(\omega_m t) + \frac{2\pi}{3} \right) \]

The peak electrical frequency is calculated by dividing the peak speed of the translator by the magnetic wavelength. Equation 2.6 shows the peak electrical frequency calculation where \( d \) is in meters and magnetic wavelength \( \lambda \) is in meters. The peak electrical frequency associated with equation 2.6 is expected because the magnetic wavelength represents a complete cycle from north to south. By increasing the velocity of this transition, the cycle time decreases.

\[ \hat{\omega}_e = \frac{2\pi}{\lambda} \left( \frac{dx}{dt} \right)_{\text{max}} \]  \hspace{1cm} (2.6)

\[ \hat{f}_e = \frac{\text{velocity}_{pk}}{\lambda} \]

The ideal wave source was assembled in Simulink with the corresponding parameters, where the ocean wave frequency, \( \omega_m \), is the variable:
\( d = 1m \)
\( \lambda = 144mm = 0.144m \)
\( \omega_m = 2\pi(f_m) \)

Fig. 2.3 – Ideal wave mathematical model

Fig. 2.4 – Per-phase voltage to SimPowerSystem Block interface

Figure 2.4 shows the monochromatic wave model interfaced with the SimPowerSystems dependent voltage source blocks. SimPowerSystems is an add-on to Simulink that allows circuits to be simulated. The SimPowerSystems blocks are used to output a voltage dependent on the input reference. The SimPowerSystems blocks are similar to circuit simulation layout, where node voltages and currents can be easily measured.
2.2 Dynamic Model

A dynamic generator model will allow faster simulation performance times since the switching model can be verified using an average model. With the ideal generator model, it is easy to select some desired current reference based on the output voltage from the generator; however there is no feedback to the generator system. By developing a dynamic linear generator model, verification that the switching control works in conjunction with it will transition into a full hardware based test.

The dynamic linear generator equations are similar to those of a rotary permanent magnet synchronous generator that were used to develop the control system. The equations however differ slightly because of the torque and force representation. The rotational mechanical angle in a rotary machine is dependent upon the angular velocity, whereas the mechanical ‘angle’ of the linear generator is dependent upon the linear velocity.

The dq-axis equations for a linear generator are presented below, where $R_s$ is the coil resistance, $\omega_m$ is the electrical angular frequency, $i_q$ is the q-axis current, $i_d$ is the d-axis current and $\lambda_{fd}$ is the excitation linkage flux of the stator due to flux produced by the magnets. Also, $V_d$ is the d-axis voltage and $V_q$ is the q-axis voltage. [3]

\[
\begin{align*}
    v_{sd} &= R_s i_{sd} + \frac{d}{dt} \lambda_{sd} - \omega_m \lambda_{sq} \\
    v_{sq} &= R_s i_{sq} + \frac{d}{dt} \lambda_{sq} + \omega_m \lambda_{sd} \\
    \dot{\lambda}_{sd} &= L_s i_{sd} + \lambda_{fd} \\
    \dot{\lambda}_{sq} &= L_s i_{sq} \\
    L_s &= L_{ss} + L_m
\end{align*}
\]
Combining both parts of equation 2.7 and 2.8, results in the cross coupled dq-voltage equations.

\[
\begin{align*}
\nu_{sd} &= R_s i_{sd} + \frac{d}{dt} \left( L_s i_{sd} + \lambda_{fd} \right) - \omega_m L_s i_{sq} \\
\nu_{sq} &= R_s i_{sq} + \frac{d}{dt} L_s i_{sq} + \omega L_s i_{sd}
\end{align*}
\] 

(2.9)

\[
\omega_m = \frac{P}{2} \omega_{mech}
\]

(2.10)

In equation 2.11, the rotational mechanical frequency relates to the electrical frequency, both in rad/sec, by the number of poles of the machine.

\[
T_{em} = \frac{P}{2} \left( \lambda_{sd} i_{sq} - \lambda_{sq} i_{sd} \right)
\]

(2.11)

Substituting in the dq-flux linkage from equation 2.8, results in equation 2.12 giving the output torque related to the q-axis current and magnet excitation flux linkage.

\[
T_{em} = \frac{P}{2} \left( (L_s i_{sd} + \lambda_{fd}) i_{sq} - L_s i_{sq} i_{sd} \right) = \frac{P}{2} \lambda_{fd} i_{sq}
\]

(2.12)
Fig. 2.5 – Dynamic generator/buoy Model

The dynamic system layout is shown in figure 2.5. The hydrodynamic model and dynamics model will generate forces created by an ocean wave. Optimal Force Controller block will intelligently compute the optimal generator loading. The wave excitation force will exert a force upon the buoy and the generator will prescribe a force to exert upon the wave. This generator force is determined by the current output of the generator.

The q-axis current substituted into equation 2.11, resulting in torque. The torque is force times the radius of a machine. Equation 2.13 expresses the length of the stator for a linear generator is the pole pitch, $\tau$, times the number of poles, $p$. The circumference of a rotary synchronous generator is expressed in equation 2.13, where $r$ is the mean radius of the rotor. [6]

$$l = \tau \cdot p \ phases = 3\tau \cdot p$$  \hspace{1cm} (2.13)

$$c = 2\pi \ r$$  \hspace{1cm} (2.14)
Substituting in equation 2.13 into equation 2.14 where the length and circumference are equal:

\[ r = \frac{3\pi p}{2\pi} \quad (2.15) \]

Assuming a 2 pole machine, 1 pole pair, the radius of a machine is equal to:

\[ r = \frac{3\pi}{\pi} \quad (2.16) \]

For a rotary machine with 2 poles, the torque output is equal to:

\[ T_{em} = \lambda_{jd} i_{sq} \quad (2.17) \]

The torque is equal to force times radius thus relating torque and force, results in:

\[ F = \frac{T_{em}}{r} = \frac{\pi}{3\pi} \lambda_{jd} i_{sq} \quad (2.18) \]

The force output of a linear synchronous machine of multiple pole pairs will increase linearly with the number of pole pairs, like a rotary machine the poles pairs will linearly increase the torque. A general equation for force output is equation 2.19:

\[ F = \frac{p\pi}{6\pi} \lambda_{jd} i_{sq} \quad (2.19) \]
The Simulink model for the permanent magnet linear machine is shown in figure 2.6.

![Simulink model diagram](image)

Fig. 2.6 – PMLG dynamic Simulink model

The force output is computed from a measured $I_q$ current, this force is then fed back into the Optimal Force Controller. The force block, labeled ‘$f(u)$’ is shown in figure 2.6 after the ‘Lambda_dq→idq’ block.
3 PASSIVE RECTIFIER INVESTIGATIONS

3.1 Passive Rectifier Overview

Line commutated passive rectifiers in this investigation will be used as a reference with which three-phase active rectifier results will be compared. The passive rectifier operation is based on the line-to-line voltage and the dc-bus voltage. The diode rectification investigation is vital knowledge, since in the event of switching failure of an active rectifier, the buoy will still generate power in this manner and thus the electronics will need to be designed to handle such events.

3.2 Passive Rectifier Simulations

For the passive rectifier simulations, a model was built using MATLAB/Simulink with models from the SimPowerSystems Library. The passive rectifier was arranged in a three-phase six-pulse full-bridge topology. The diodes each have snubber circuits utilizing a series capacitor and resistor to reduce high voltage spikes during switching modes which can cause the diodes to fail. The di/dt time can be calculated using equation 3.1. [4]

![Diode reverse recovery charge](image)

Fig. 3.1 - Diode reverse recovery charge
\[
\frac{di}{dt} = -\frac{V_d}{L_o}
\]  

(3.1)

Equation 3.1 relates the change in current with the voltage and inductance connected to the device. A curve similar to figure 3.1 shows what visually happens when current is quickly switched. \(V_d\) is the voltage across the device, the worst case scenario for a diode bridge is with zero dc bus voltage and full input voltage. The voltage \(V_d\) is selected as the maximum output voltage per phase. The peak phase voltage at velocity 2m/s is 655\(V_{LN}\). The inductance is dominated by the source inductance of the permanent magnet linear generator, thus stray inductances are neglected. [5]

\[
\frac{di}{dt} = -\frac{1134V}{2*24mH} = -23625 \text{ A/s} \quad (3.2)
\]

\[
I_{rr} = \left(\frac{di}{dt}\right)_{t_{rr}} \quad (3.3)
\]

The reverse recovery current, \(I_{rr}\), can be defined by equation 3.3 above where the reverse recovery time is \(t_{rr}\). The reverse recovery time specification is available on most IGBT/diode packages. For the IGBT/module CM75DU-24F, the reverse recovery time measured under inductive load testing at full rated current and dc bus voltage is 150ns. Using this time required in equation 3.3, the maximum reverse recovery current is \(I_{rr} = -3.54 \text{mA}\). The snubber capacitance is defined by equation 3.4 below, where \(L_s\) is source inductance and \(V_{LL}\) is the line-to-line RMS voltage.

\[
C_s = L_s \left(\frac{I_{rr}}{V_{LL}}\right)^2 \quad (3.4)
\]
Since the source inductance, the line-to-line voltage, and the reverse recovery current are all known, the computed snubber capacitance can be calculated as $C_s = 2.34e-13F$. The required snubber resistance is then found with equation 3.5.

$$R_s = 1.3 \frac{V_{LLpeak}}{I_{rr}} \quad (3.5)$$

Using 1134V, the peak line-to-line voltage produced by the generator at 2m/s, and the previously computed reverse recovery current, the snubber resistance is found to be 416 kΩ.

### 3.2 Ideal Source with Passive Rectifier

The three-phase diode bridge is shown in figure 3.2. Each diode has a turn-on voltage of 3V and a turn-on resistance of 1 mΩ. Each diode has a parallel RC snubber.

![Simulink Passive Rectifier Model](image)
circuit with values calculated previously. The dc load resistance is $156\Omega$ to produce peak 1kW. The wave period is $T=6s$ and voltage levels for 0.7m/s velocity.

Voltage Input: $V_{LL_{rms}} = 280V$, $V_{LN_{pk}} = 228V$ (0.7m/s)

![Voltage Input](image)

**Fig. 3.3(a) – Rectifier input voltage (no dc bus capacitance)**

![Generator Back EMF](image)

**Fig. 3.3(b) – Generator back EMF (no dc bus capacitance)**
Figure 3.3(a) shows the input voltage; the peak line-to-neutral voltage is 228V with the eight pulses on the upstroke and eight pulses on the down-stroke. The generator back EMF has a slightly higher voltage due to the drop across the line resistance and source inductance. The current draw at peak voltages results in the notches seen in figure 3.3(c).

Figure 3.3(c) shows the time when the peak electrical frequency occurs. The electrical frequency calculated is:

$$\frac{1}{t_2 - t_1} = \frac{1}{3.14s - 2.883s} = 3.8Hz$$

This electrical frequency is less than the anticipated 5.5Hz. This is due to limitations of the source model used. The model has a maximum travel of 1m. This results in having only a 1 meter wave height, when this is not the case. The limitations
of the model for the stroke length result in off peak electrical frequencies. The peak electrical voltages, however, are correct.

Fig. 3.3(d) – Line current (no dc bus capacitance)

Peak current levels are shown in figure 3.3(d) and (e) at approximately 2.5A.
The line current with no dc bus capacitance is seen in figures 3.3(d) and figure 3.3(e). The double peaked currents are expected due to line-to-line commutation twice per electrical period.
The dc bus voltage in figure 3.3(f), zoomed in figure 3.3(g) shows the peak dc bus voltage. This voltage level at approximately 400V is from the peak line-to-line voltage at the rectifier input, verification is shown below.

\[ V_{LL\text{peak}} = V_{LL\text{rms}} \sqrt{2} = 395V \]
Fig. 3.3(g) – Peak dc bus voltage

Fig. 3.3(h) – dc bus current
The dc bus current figures 3.3(h) and 3.3(i) show the peak current at 2.5A. This current draw results in a peak power dissipation of approximately 1kW.

\[ P_{\text{peak}} = \hat{V} \hat{I} = 391.6V \times 2.5A = 977.5W \]

To stiffen the dc bus voltage, a capacitor is placed across the resistive load. However, a specific capacitance will be required depending upon desired maximum voltage ripple. The ripple percentage is specified by the difference between the highest and lowest voltages divided by the RMS voltage level. The capacitance needed for a specific voltage ripple and energy requirement is shown in equation 3.7. Voltage ripple is defined by equation 3.8.
\[ E = \frac{1}{2} C (V_{\text{ripple}})^2 \]  
\[ C = \frac{2E}{(V_{\text{ripple}})^2} \]  
\[ \%V_{\text{ripple}} = \frac{V_{\text{peak}} - V_{\text{low}}}{V_{\text{RMS}}} \]  

Energy dissipation of 1000W for 3 seconds, half the average wave period, requires 3000 Joules of energy storage. With a combined voltage ripple of 10% (V), requires a total capacitance of 2.2 Farads. This large capacitance would result in a long transient period for the simulation to reach steady state. A large capacitance would also increase the inrush current of the rectifier which may exceed the rating of the armature wire. A solution around this is to pre-charge the dc bus capacitor to the steady-state voltage before connecting the generator to the rectifier. Considerations in connecting the generator to the rectifier need to be done to ensure that the large unloaded generator voltages do not exceed the ratings of the power electronics. The capacitance chosen for the next simulation is 0.5F

Voltage: \( V_{\text{LLrms}} = 280\text{V}, V_{\text{LNpk}} = 228\text{V} \) (0.5F dc capacitance)
Fig. 3.4(a) – Generator back EMF

Fig. 3.4(b) – Rectifier Input Voltage
Figures 3.4(a) and (b) show the generator back EMF and the input voltage waveform into the rectifier. The zoomed in figure 3.4(c) shows an even more pronounced distorted waveform due to the double peak input current.

![Fig. 3.4(c) – Rectifier input voltage (zoom)](image-url)
Fig. 3.4(d) – Line current

Fig. 3.4(e) – Line current (zoom)
Figures 3.4(d) and (e) show the line input current. The double peak is expected with the high dc bus voltage with peaks of 20A at maximum voltage input.
Figures 3.4(f) and (g) show the dc bus voltage and current. The ripple is low at approximately 2% (7V). The average dc bus voltage is 350V with an average dc bus current of 2.86A. The average power dissipated is 1001W with a load resistance of 122Ω.

The next simulation is with a dc bus capacitance of 1100uF. This capacitance will be used in hardware testing and is a good estimate for those results. Input voltages from figures 3.5(a) and (b) are similar to the no dc bus capacitance results.

Voltage Input: \( V_{\text{LLrms}} = 280V, V_{\text{LNpk}} = 228V \), (dc bus capacitance 1100uF)

![Generator Back EMF](image)

Fig. 3.5(a) – Generator back EMF
Fig. 3.5(b) – Rectifier input voltage

Fig. 3.5(c) – Input line current
The double peak currents in figure 3.5(c) and (d) are much more pronounced than the previous simulations. This is due to a varying dc bus voltage that will effect the commutation of the diodes.
Fig. 3.5(e) – Rectifier dc bus voltage

Fig. 3.5(f) – Rectifier dc bus current
The dc bus voltage and current, figures 3.5(e) and (f) show that the dc bus voltage does not fall to zero like the no dc bus capacitor simulation. This is from the small 1100uF capacitance that results in small amount of energy storage available to the load when there no input power.

### 3.3 Summary of Passive Rectifier Results

The double peak is due to the line-to-line diode commutation. As the voltage on the anode of the diode is at peak it conducts, the other two top diodes are reversed biased and do not conduct. The bottom diode with the largest negative voltage on the cathode is forward biased and conducts, the other two diodes do not conduct due to reverse bias. [5, 7]

![Fig. 3.6 – Three-phase diode bridge rectifier](image)

The double peak current results in a non-linear loading and can result in a flat topping of the line voltage, in the simulations above, only a notch is present due to very little current draw, but can drastically increase. The double peak current has a large harmonic content which is injected back to the generator. These harmonics could potentially damage the linear generator under high loading and continuous operation.
in a large scale generator. Also, since each diode conducts based on the dc bus voltage, the buoy will not be controlled and therefore may not be extracting optimum energy from each wave. Passive rectification would represent a worst-case scenario in which gating signals from the active rectifier were disabled. Energy extraction from the generator is still possible and represents a fail-safe mode of operation.

Ideal buoy/generator model was initially simulated with a generator displacement of 1m, a magnetic wavelength of 144mm and wave period in order to obtain the desired output phase voltage. The limitation of the ideal model results in an electrical frequency for a 1m/s wave regardless of actual wave height. If the distance traveled was changed in the ideal model, more electrical pulses would result. Only eight electrical pulses are obtainable on the up and eight on the downward-stroke due to machine design.

<table>
<thead>
<tr>
<th>Wave Period (s)</th>
<th>Wave Height (m)</th>
<th>Velocity (m/s)</th>
<th>Peak Electrical Frequency (Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>1.0</td>
<td>0.524</td>
<td>3.6</td>
</tr>
<tr>
<td>8</td>
<td>1.0</td>
<td>0.393</td>
<td>2.7</td>
</tr>
<tr>
<td>10</td>
<td>1.0</td>
<td>0.314</td>
<td>2.2</td>
</tr>
</tbody>
</table>

Table 3.1 – Wave height and linear velocity

The electrical parameters are selected to provide baseline results for peak 1kW power dissipation. The resistive load is selected to dissipate 1kW at the peak of the dc bus voltage. For example, the 280VLLrms will have a dc load resistance of 156 Ohms since the dc bus voltage is the peak line-to-line voltage of each phase. Table 2.2 shows the voltages and the resulting load resistance.

<table>
<thead>
<tr>
<th>VLNpk</th>
<th>VLLrms</th>
<th>VLLpk</th>
<th>dc bus</th>
<th>Load (Ohms)</th>
</tr>
</thead>
<tbody>
<tr>
<td>228</td>
<td>280</td>
<td>396.6</td>
<td>408</td>
<td>156</td>
</tr>
<tr>
<td>228</td>
<td>280</td>
<td>396.6</td>
<td>350</td>
<td>122</td>
</tr>
</tbody>
</table>

Table 3.2 – dc bus resistive loads
The peak line currents for the no dc bus capacitance and the 1100uF dc bus capacitance have low peak current values of 2.5A and 6A, respectively. This is within the current capabilities of the generator windings (12 AWG), however the sustained 1kW simulation with dc bus capacitance of 0.5F results in high currents of 20A. Sustained currents exceeding the gauge recommendations are detrimental to the survivability. Results from simulations are summarized in table 1.3.

<table>
<thead>
<tr>
<th>VLLrms</th>
<th>dc bus</th>
<th>dc current</th>
<th>Capacitance</th>
<th>Load (Ohms)</th>
<th>Power (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>280</td>
<td>391V</td>
<td>2.5A</td>
<td>X</td>
<td>156.0</td>
<td>977.500</td>
</tr>
<tr>
<td>280</td>
<td>350V</td>
<td>2.86A</td>
<td>0.5F</td>
<td>122.0</td>
<td>1001.000</td>
</tr>
<tr>
<td>280</td>
<td>408V</td>
<td>2.61A</td>
<td>1100uF</td>
<td>156.0</td>
<td>1064.880</td>
</tr>
</tbody>
</table>

Table 3.3 – Passive rectifier simulation results
4 DQ CONTROL

4.1 dq Overview

The concept of dq control is derived from a mathematical transformation to obtain independent variables of interest: real and reactive power. Real power results in a change in output torque or for a linear generator, force. The reactive power will change the flux.

Controlling a full-bridge active rectifier using dq allows independent changes in real and reactive power. The transformation, from three-phase to two-phase dq, results in control values that are dc quantities vs. ac control where all control signals are sinusoidal. The use of dc values allow for elimination of steady state errors.

The three-phase stator in a permanent magnet synchronous machine can be realized using an equivalent two-phase machine with the phases orthogonal to each other. This equivalent model decouples the d-axis (direct) and q-axis (quadrature) allowing full independent control of real torque or reactive power, respectively. The PMLG d-axis will be aligned with the magnetic north axis and the q-axis will be orthogonal to the flux. This arrangement controls current with the q-axis and reactive power with the d-axis. Figure 4.1 shows the projection of these two axes. [3]

\[ \tilde{i}_z = i_a(t) + i_b(t)e^{j2\pi/3} + i_c(t)e^{-j2\pi/3} \] (4.1)

Fig. 4.1 – Three-phase to two-phase projection
At time $t$, the three-phase currents $i_a(t)$, $i_b(t)$ and $i_c(t)$ can be represented by an equivalent current space vector $\vec{i}_r$. The MMF (Magneto-Motive Force), is linearly dependent on $\vec{i}_r$ by $N_s / p$, where $N_s$ is the turns per phase and $p$ is the number of poles. A pole is defined as north or south, thus two poles would include a north and south direction. The two orthogonal equivalent windings are each sinusoidally distributed with $\sqrt{3/2} N_s$ turns.

$$\frac{\sqrt{3/2} N_s}{p} (i_{sd} + j i_{sq}) = \frac{N_s}{p} \vec{i}_r^d$$

(4.2)

The $d$-axis and $q$-axis currents are scaled when projected on their respective axis. This can be seen from equation 4.2 shown above. The square root terms relating to current and turns are used in order to ensure the same MMF distribution as the three-phase equivalent.

The $d$ and $q$ axis windings are now decoupled magnetically since the flux linkage of two orthogonal windings is zero. Also, since inductance is proportional to the number of turns squared the magnetizing inductance is the same as the three-phase equivalent.

$$L_{m-dq} = (\sqrt{3/2})^2 * L_{m-1,\text{phase}}$$

$$L_{m-dq} = (\sqrt{2}) * L_{m-1,\text{phase}}$$

(4.3)

$$L_{m-dq} = L_m$$
With this, the inductances of the d and q axis can be calculated using equation 4.4.

\[ L_{sd} = L_{m-dq} + L_{ls} \]
\[ L_{sq} = L_{m-dq} + L_{ls} \]  

(4.4)

With the same magnetizing inductance and self-inductance, each dq-winding has the same inductance as each phase of a three-phase machine. No scaling is necessary.

The relation between the three-phase windings and dq windings needs to be determined in order for the MMF to be equivalent in both reference frames. The space vector \( \vec{i}_s \) can be represented from the stator space-vector \( \vec{i}^a_s \) seen in equation 4.5, where \( \theta_{da} \) is the angle between the stator current space-vector and the dq space vector.

\[ \vec{i}^d_s = \vec{i}^a_s e^{-j\theta_{da}} \]  

(4.5)

In a three-phase system:

\[ \vec{i}^a_s = i_a(t) + i_b(t) + i_c(t) \]  

(4.6)

Substituting in equation (4.5) into (4.7) results in:

\[ \vec{i}^d_s = i_a(t)e^{-j\theta_{da}} + i_b(t)e^{-j(\theta_{da}-2\pi/3)} + i_c(t)e^{-j(\theta_{da}+2\pi/3)} \]  

(4.7)

Separating the real and imaginary terms in equation 4.7, results in the transformation in equation (4.8).
This transformation, known as Park’s Transformation, will be used extensively to transform three-phase measurements into the dq-axis form. Note however that the input $\theta_{da}$ is the angle difference between the two reference frames and is fixed, but the magnitudes of the inputs due to varying amplitude input. The sinusoidal shape of the q-axis reference is a result of the generator force changing polarity as the generator velocity changes direction. Thus the dq controller will need to track ac values unless another transformation could theoretically decouple the amplitude modulation altogether. This ac component term will become apparent in the simulations.

### 4.2 Transfer Function of Generator

The three-phase active rectifier will apply voltage to the terminals of the permanent magnet linear generator. The difference in voltage between the back EMF and the applied terminal voltages determines the current out of the generator. This can be seen visually in figure 4.2, the per-phase equivalent circuit of the generator/rectifier front end.
In order to control the generator, the transfer function of the system is needed. The transfer function relates the output of a system to an applied input. The transfer function of the PMLG can be obtained from the equations for a Permanent Magnet Synchronous Machine (PMSM). From the dq stator windings: [1]

\[
V_{sa} = R_s i_{sa} + \frac{d}{dt} \lambda_{sa}
\]
\[
V_{sb} = R_s i_{sb} + \frac{d}{dt} \lambda_{sb}
\]

(Multiplying \((j)\) by both sides of the second equation in equation 4.9 combining the real and imaginary components produces:

\[
V_{sa} + j V_{sb} = R_s i_{sa} + j R_s i_{sb} + \frac{d}{dt} \lambda_{sa} + j \frac{d}{dt} \lambda_{sb}
\]

(Transforming from the alpha/beta stationary coordinates to the dq rotating coordinates, substitute the following equations (4.11) and (4.12) in to equation (4.10).

\[
\tilde{V}^\alpha_s = V_{sa} + j V_{sb}
\]
\[
\tilde{i}^\alpha_s = i_{sa} + j i_{sb}
\]
\[
\tilde{\lambda}^\alpha_s = \lambda_{sa} + j \lambda_{sb}
\]
\[
\tilde{V}^d_s = \tilde{V}^\alpha_s e^{j \theta_{\omega}}
\]
\[
\tilde{i}^d_s = \tilde{i}^\alpha_s e^{j \theta_{\omega}}
\]
\[
\tilde{\lambda}^d_s = \tilde{\lambda}^\alpha_s e^{j \theta_{\omega}}
\]

(4.11) (4.12)
The result is equation 4.15, after differentiation (equation 4.14) with applied chain rule, can be separated into their respective $d$ and $q$ components. This will allow the controller to independently control the real power and reactive power.

\[
\tilde{V}_s^d e^{j\theta_{\text{ua}}} = R_s i_s^d e^{j\theta_{\text{ua}}} + \frac{d}{dt} (\tilde{\lambda}_s^d e^{j\theta_{\text{ua}}}) \tag{4.13}
\]

\[
\tilde{V}_s^d e^{j\theta_{\text{ua}}} = R_s i_s^d e^{j\theta_{\text{ua}}} + \frac{d}{dt} (\tilde{\lambda}_s^d e^{j\theta_{\text{ua}}}) + j\omega \tilde{\lambda}_s^d e^{j\theta_{\text{ua}}}
\]

\[
\tilde{V}_{sd} = V_{sd} + jV_{sq}
\]

\[
V_{sd} + jV_{sq} = R_s (i_{sd} + ji_{sq}) + \frac{d}{dt} (\lambda_{sd} + j\lambda_{sq}) + \omega (j\lambda_{sd} - \lambda_{sq}) \tag{4.14}
\]

\[
V_{sd} = R_s i_{sd} + \frac{d}{dt} \lambda_{sd} - \omega \lambda_{sq} \tag{4.15}
\]

\[
V_{sq} = R_s i_{sq} + \frac{d}{dt} \lambda_{sq} + \omega \lambda_{sd}
\]

Taking equation (4.15) and transforming it into the frequency domain via the Laplace transform results in:

\[
V_{sd} = R_s i_{sd} + s\lambda_{sd} - \omega \lambda_{sq} \tag{4.16}
\]

\[
V_{sq} = R_s i_{sq} + s\lambda_{sq} + \omega \lambda_{sd}
\]

Substituting the flux linkage with the respective dq inductance values and currents:

\[
\lambda_{sq} = L_s i_{sq}
\]

\[
\lambda_{sd} = L_s i_{sd} + \lambda_{fd}
\]

\[
V_{sd} = R_s i_{sd} + s(L_s i_{sd} + \lambda_{fd}) - \omega (L_s i_{sq}) \tag{4.17}
\]

\[
V_{sq} = R_s i_{sq} + sL_s i_{sq} + \omega (L_s i_{sd} + \lambda_{fd})
\]
The control scheme will have cross-coupling effects due to the d-axis stator voltage dependent upon the q-axis current as well as the q-axis voltage dependent upon the d-axis current. This will introduce disturbances that the controller will need to compensate for.

4.3 Controller Design

The controller design will be based on a single-input single-output (SISO) control topology. This controller is in series with the plant and is shown in figure 4.18. The plant is the device being controlled; in this case it is the PMLG. [8]

![Control topology diagram](image)

**Fig. 4.3 –Control topology**

\[
Y(s) = G_p(s) \ast U(s) \\
U(s) = G_c(s) \ast E(s) \quad (4.18) \\
E(s) = R(s) - Y(s)
\]

Substituting and solving for \( Y(s)/R(s) \), where \( Y(s) \) is the output and \( R(s) \) is the input in the s-domain:
The result is the total closed-loop gain of the system.

The plant and controller transfer functions are shown in equation 4.22 and 4.23. The controller is PI (proportional-integral) controller. The proportional gain value accelerates the error increasing convergence time, reducing the rise time but increasing overshoot. The integral gain will decrease rise time and increase overshoot, however it will eliminate steady-state errors. The elimination of steady-state errors is highly desired in precise controls. There exists a derivative part for a PID controller, which reduces overshoot, however if the controllers performance does not have substantial overshoot, the derivative term may not be necessary.

The plan transfer function is derived from equation 4.17 in the previous section. Since generator force is related to generator current, current control is desired for this topology. Current referenced will be the input to the controller; the output will be applied terminal voltages to the generator.

\[
Y(s) = G_p(s)G_c(s)\left( R(s) - Y(s) \right)
\]

\[
Y(s) = G_p(s)G_c(s)R(s) - G_p(s)G_c(s)Y(s)
\]

\[
Y(s) \left( 1 + G_p(s)G_c(s) \right) = G_p(s)G_c(s)R(s)
\]

\[
Y(s) = \frac{G_p(s)G_c(s)}{R(s)} = \frac{G_p(s)G_c(s)}{G_p(s)G_c(s) + 1}
\]

Equation 4.20 is rewritten terms of the generator output current to the terminal voltages, equation 4.21. Since both the stator resistance and inductance are equivalent in the dq reference frame, the same plant can be used for the controller design.
The plant transfer function is described in equation 4.22, where $L_s$ is the stator inductance and $R_s$ is the stator coil resistance.

\[
\frac{i_{sq}}{V_{sq} + \omega (L_s i_{sd} + \lambda_{fd})} = \frac{1}{(R_s + sL_s)} \quad (4.21)
\]

\[
\frac{i_{sd}}{V_{sd} + \omega L_s i_{sd}} = \frac{1}{(R_s + sL_s)}
\]

The transfer function of a PI controller is shown in equation 4.23. It is typical to expand the integral term to aid in the calculations of the gain values. The total loop gain can be expressed as the multiplication of both the controller and plant transfer functions. It is common to cancel out the pole of the first term with the zero of the second term.

\[
G_p(s) = \frac{1}{sL_s + R_s} \quad (4.22)
\]

\[
G_c(s) = \frac{K_i}{s} + K_p
\]

\[
G_c(s) = K_i \left( \frac{K_i}{s} + 1 \right) \quad (4.23)
\]

The transfer function of a PI controller is shown in equation 4.23. It is typical to expand the integral term to aid in the calculations of the gain values. The total loop gain can be expressed as the multiplication of both the controller and plant transfer functions. It is common to cancel out the pole of the first term with the zero of the second term.

The plant transfer function shown in figure 4.4 is stable with the phase margin at 34.5 rad/sec. With no controller, this system will oscillate towards steady state values. The 3dB point at $Ra/La = 23.33$ rad/sec is noted. The pole at this frequency will pull the phase margin towards -90 degrees. The pole of the integrator is at zero rad/sec (1dB) and adding any more poles will bring the phase shift down more. By canceling out the plant pole with controller zero, the phase margin will stay at -90 degrees.
To calculate the PI values, the plant transfer function and the controller transfer function are used in the MATLAB SISO tool. The SISO (single input single output) tool quickly develops a new transfer function from a given plant and controller transfer function. Gain margins, phase margins, poles and zeros can be altered within the graphical SISO tool. The controller and plant transfer function bode plot are shown below in figure 4.4 with initial $K_i$ and $K_p$ values equal to 1 as an initial guess. It is
noted that by canceling out the pole of the plant with the zero of the controller, $K_p/K_i$ will be known. $K_p/K_i$ is calculated to be 0.0428.

![Bode Diagram](image)

**Fig. 4.5 – Bode plot of plant and controller open-loop transfer function**

\[
G(s) = \frac{s + 1}{s\left(s\left(\frac{L_s}{R_s}\right) + 1\right)R_s}
\]

\[R_s = 0.56\Omega\]

\[L_s = 24mH\]  \hspace{1cm} (4.25)

\[
G(s) = \frac{s + 1}{s(s(0.04286)+1)0.56}
\]

From equation 4.25, there should be a zero at $\omega = 1$ rad/sec and a pole at $\omega = 1/0.04286 = 23.3$ rad/sec.
Verification from the mathematical derivation, there is a zero located at \( w = 1 \) rad/sec and a pole at \( w = 23.3 \) rad/sec, this can be seen in figure 4.5. Poles add instability by decreasing the phase margin and gain margin, whereas zeros add stability by increasing the phase margin and gain margin. At 1 rad/sec, the gain margin falls at -20dB/dec then flattens out. The phase margin at the zero is close to -45 degrees. This is expected since a zero has phase margin of -45 degrees at the zero frequency, and 0 at the w/10 and 90 at 10w. The true is also for a pole, except the phase margin decreases.

The crossover frequency is chosen to be 10% of the switching frequency. It is common for the crossover frequency to be this low to allow the system to react faster than changes are made. If the controller was much faster than or as fast as the switching frequency, ripples in the controller output would not be recognized by the PWM generator. [9]

![Bode Diagram](image)

**Fig. 4.6 – Controller crossover frequency, phase and gain margin**
By canceling out the zero and pole or by moving the zero over the pole location in the SISO tool GUI, the results are shown in figure 4.5.

The new controller transfer function now has a crossover frequency of 3150 rad/sec which is approximately 10% of the 5kHz (31krad/sec) switching frequency. The phase margin is -90 degrees, ensuring system stability. Any phase margin larger than 180 degrees will be unstable and any phase margin close to 180 can have large oscillations before steady-state is reached. The SISO tool gives the controller a new transfer function shown in equation 4.26.

\[
G_c(s) = 1760 \frac{1 + 0.043s}{s} \tag{4.26}
\]

From the previous controller equation, the values for \(K_i\) and \(K_p\) can be found. Looking at equation 4.26, \(K_i\) is the overall gain equal to 1760 and since \(K_p/K_i = 0.043\), \(K_p\) is 75.6. These values will be used in the PI controller for the simulations.

An average model can be constructed to evaluate how effective the controller is. The average model will eliminate the switching dynamics from the model that are present with simulations done with active elements. The switching dynamics add another level of complexity to the model that does not necessarily help evaluate the performance of the controller. The switching elements also involve slower computation time and different simulation solvers in order to compensate for active switching elements.

To construct the switching model, the transfer function of the controller and the plant with feedback are arranged as seen in figure 4.7.
The switching model takes in terminal dq-voltages, normalizes the value and sent to the PWM (switching) block. The PWM will regulate the average voltage applied to the generator terminals. This is a physical model representation.

The average model eliminates the switching block and the dynamics created from the IGBT switching. This assumes ideal PWM generation and no losses.

The per-phase equivalent circuit of the three-phase converter is shown above in figure 4.9. The left hand side represents the dc bus and the right hand side represents
the permanent magnet linear generator. The $[v]_{abc}$ is the rectifier applied PWM voltage signals, the $[Vs]_{abc}$ is the PMLG back EMF. The controller outputs the necessary terminal voltage, so for the average model, no PWM block is necessary.

Fig. 4.10 – Controller layout

Fig. 4.11 – Simulink controller layout

Implementing the dependent voltage source is done with the SimPowerSystems dependent dc voltage source block. The direct output of the controller is fed into the dependent voltage source. The decoupling terms are connected to the output of the PI controller to eliminate the coupling terms within the plant. This will provide increased performance for the system.
4.4 Controller Verification

Testing the verification of the controller is done using the step feature in MATLAB. The cascade controller transfer function is expressed in equation 4.27. The step response of the close loop system $T(s)$ is shown below. The step response test will show how quickly the system will converge to unity, the referenced input.

$$G(s) = \frac{1}{s \left( \frac{L_i}{R_s} + 1 \right) R_s} \left( s \left( \frac{K_p}{K_i} + 1 \right) \right)$$  \hspace{1cm} (4.27)
\[ T(s) = \frac{G(s)}{G(s) + 1} \]

\[ T(s) = \left( \frac{1}{s \left( \frac{L_s}{R_s} + 1 \right)} s \left( \frac{K_p}{K_i} + 1 \right) \right) K_i + 1 \]

\[ T(s) = \frac{1}{s \left( \frac{L_s}{R_s} + 1 \right)} s \left( \frac{K_p}{K_i} + 1 \right) + 1 \]

\[ T(s) = \frac{1.806s^3 + 84.14s^2 + 980s}{0.000576s^4 + 1.833s^3 + 84.45s^2 + 980s} \]

Fig. 4.13 – Step response of closed-loop controller and plant
From figure 4.13, the system reaches unity output at 2.5ms. This convergence time is very acceptable since the switching period is 200us (5 kHz). Also, no oscillation is present.
5 THREE-PHASE SYNCHRONOUS ACTIVE RECTIFIER

5.1 Active Rectifier Overview

Three-phase active rectifiers allow for a greater range of control of the real and reactive input power, resulting in better control of the PMLG. Three-phase active rectifiers are built using six IGBT switches which are modulated on and off. This modulation controls the magnitude and frequency of the rectifier phases. Contrast this to the passive rectifier where no control is obtained since current draw is determined by the dc bus voltage. Since line currents can be controlled, the generator force is controlled. [5]

The active rectifier topology is similar to an inverter for motor control. An inverter voltage output is limited to the dc bus voltage. The dc bus voltage used in the testing of the PMLG is 1000V, allowing full current control up to $612V_{\text{LLrms}}$. If the input voltage exceeds this, the current cannot be controlled well.

Each IGBT module has built in anti-parallel diodes for current to conduct when gating signals are disabled. In the event of gating signal failure, the generator/rectifier will still be able to supply power to the dc load under passive rectification mode. This is important in a scenario when sustained power is critical.

5.2 Ideal Wave Model Simulations

The ideal wave model interfaced with the SimPowerSystems block will establish the base testing for the three-phase active rectifier. The controller accuracy will be tested using a step change in current and sinusoidal current reference. Generator back EMF voltages of $V_{\text{LLrms}}=280V$ will be simulated. This voltage represents a 0.7 m/s linear velocity.
5.2.1 Switching Model

The three-phase IGBT active rectifier is connected to a variable load which is represented by the dc-dc converter. The dc-dc converter is controlled by a hysteretic controller which maintains the dc bus voltage at an average 1000V. The load resistor is 100 Ohms. The IGBT Simulink model contains snubber circuits to reduce high voltages spikes during switching events due to the IGBTs. The snubber circuits are parallel RC components designed to handle switching power dissipation. Switching frequency is set at 5 kHz.

Fig. 5.1 – Three-phase active rectifier with dc bus regulator

Fig. 5.2 – dq-control Simulink model
The $I_{q,\text{ref}}$ block is the input reference q-axis current. The first simulation will set $I_{q,\text{ref}}$ to a sinusoidal current draw at $T = 6s$, 0.167Hz, with a peak amplitude of 10A.

![Active Rectifier Applied Voltage](image)

**Fig. 5.3(a) – Active Rectifier Applied Voltage (280V_{LL\text{rms}})**

In figure 5.3(a), the applied voltage from the active rectifier is a PWM voltage. The voltage seen is measured from phase-to-neutral.
Fig. 5.3(b) – Generator back EMF (280V_{LL\text{rms}})

Fig. 5.3(c) – Line input current (280V_{LL\text{rms}})
In figure 5.3(c), the current waveform input is not as smooth as expected. The current is in phase with the voltage input waveform giving unity power factor, however the slow switching frequency at 5kHz has caused large ripples in the line current. Increasing the switching frequency to 10 kHz or more will reduce the current ripple yielding a cleaner input waveform and no reverse dc bus current.

![DC Bus Voltage](image)

**Fig. 5.3(d) – dc bus voltage**

The dc bus voltage is shown in figure 5.3(d). The voltage ripple is 20V from 1010 to 990. This is maintained by the dc-dc converter.
Figure 5.3(e) shows the current into the dc capacitor. There is no reverse current until the simulation reaches 1.25 seconds and is sustained for half a second. This reverse current is from the noisy line input current during this time, allowing current to flow back to the generator.
Controller $I_{sq}$ measured vs. reference

Fig. 5.3(f) – $I_{sq}$ measured vs. reference

Controller $I_{sd}$ measured vs. reference

Fig. 5.3(g) – $I_{sq}$ measured vs. reference
Figures 5.3(f) and (g) show excellent tracking of the reference currents $I_{sq}$ and $I_{sd}$. The measured values show a rippled that is due to the switching IGBTs and would be reduced by increasing the switching frequency.

### 5.2.2 Average Model

Running the average switching model will eliminate the current ripples seen in the switching model simulations. A step function will change the input reference current level at time $t = 1$s into the simulation from 2A to 5A. The input reference and the measured current level can be seen in figure 5.4.

![Isdq Output](image)

**Fig. 5.4(a) – $I_{sdq}$ current output**

Above in figure 5.4, shows the $I_{sdq}$ current output of the PMLG. The step change applied at 1 second can be seen from 2A to 5A. The low frequency component from the $I_{sq}$ measurement is due to controller tracking an ac value from the decoupling compensation terms. The output of the dq transform is normally dc values for a constant amplitude signal, however the variable amplitude nature of the generator
output results in a sinusoidal reference. The ac fluctuation is very small compared to the reference and is negligible for our application. Future tests could disregard the decoupling compensation terms.

Fig. 5.4(b) – $I_{sq}$ actual vs. reference

The step changes the $I_{sq}$ measurement and reference value. The reference value is a step change instantly and the controller accurately follows the step change. The ac component can be seen from the measurement value and is not an overshoot.
The $I_{sd}$ measurement and reference values can be seen in figure 5.4(c). The reference value for the d-axis current is set to zero. This will provide no reactive power draw from the PMLG. The actual $I_d$ value also exhibits the low frequency ac component. The error on the ac component compared to the reference value is negligible. The max error current value is 0.5mA, negligible compared to real q-axis component of 5A. The cross-coupling factor is exhibited at $t = 1s$, noted by the spike in current.

A sinusoidal current draw with the same frequency of the wave excitation force is shown in the simulations below. The peak current will correspond to the peak velocity and no current draw at zero velocity (the top of an ocean wave). $V_{LLrms}$ of 280V corresponding to 0.7m/s velocity will be simulated.

Voltage: $V_{LLrms} = 280V$
Fig. 5.9(a) – Generator output voltage

Fig. 5.9(b) – Active rectifier input voltage
The active rectifier input voltages and currents are in phase, producing unity power factor. This can be seen by comparing figure 5.9(b) with 5.9(c), noting the zero crossings for each waveform.
Fig. 5.9(d) – Sinusoidal $I_{sq}$ current output

Fig. 5.9(e) – $I_{sd}$ actual vs. reference
The sinusoidal reference current is tracked with only a slight phase shift seen in figure 5.6(f). The amplitude matches the reference value and a slight phase shift.

### 5.3 Dynamic Model

The dynamic model simulation combines the dynamic PMLG model and the dq controller. There are no SimPowerSystems block interfaces. The Simulink model is shown in figure 5.10 with a sinusoidal reference current. Note the de-coupling compensation terms.
The first simulation will step the input reference current $I_{sq}$ at $t = 1s$ from 2A to 5A to show the response time of the controller.

The step change seen in figure 5.11(a) shows the measured $I_{sq}$ tracking the reference value perfectly. No ac component seen due to de-coupling compensation terms.
Figure 5.11(b) showing the $I_{sd}$ current measured tracks well to the desired reference value. With exactly zero $I_d$ current, there is no reactive power draw from the generator producing unity power factor. No ac component with the de-coupling compensation terms.
The applied terminal voltages by the rectifier are shown in figure 5.11(c). The controller is tracking well to the reference values of $I_{sq}$ seen in figure 5.11(d).
Again, with the sinusoidal input reference current, there is little oscillation in the $I_{sd}$ measured values. This result is unity power factor and with no reactive power draw from the generator. The controller is matched perfectly with the generator model.

### 5.4 Stochastic Wave Simulations

The stochastic wave simulation will reflect more of a real world environment. The stochastic model will generate a random sea environment of various wave periods and frequency determined by a wave energy frequency spectrum. A designed optimum force controller will command a reference force for the generator to exert determined by the sea state. The current designed controller must track a prescribed force command accurately. This simulation will use a pre-existing wave generator and optimum force controller. [10]
The force input block will output the reference $i_{q\text{-ref}}$, from a commanded force input, seen in equation 5.1

\[ i_{q\text{-ref}} = \frac{2F \tau}{\lambda_{el}} \]  

The blue line in figure 5.13(a) shows the ocean wave height and the green line represents the buoy position. The optimum force controller will prescribe a force required by the PMLG. This force is seen in figure 5.13(b)
The controller performance is seen in figure 5.13(b). The controller tracks the prescribed generator force with very little phase shift or amplitude shift. The actual graphs are shown to be exactly on top of each other.
Fig. 5.13(c) – Commanded current reference and measured

The force command is computed to a current $I_{sq}$ reference. Fig. 5.13(c) shows that the controller tracks the prescribed current waveform well.

Fig. 5.13(d) – Active rectifier applied voltage
In figure 5.13(d) the active rectifier terminal voltages are shown. These terminal voltages control the current from the generator.
6 GATING SIGNAL GENERATION

6.1 Pulse Width Modulation

The gating signals applied to the IGBT switches are designed to generate a Pulse Width Modulated (PWM) signal output. This PWM output average is sinusoidal with added harmonic content depending on the switching frequencies and phase connections. This section will focus on two PWM generation techniques used widely in industrial drive applications.

6.2 Sine-Triangle PWM

SPWM (Sine-Triangle Pulse Width Modulation) is a technique for generating the necessary gating signals on a converter or inverter. Sine-Triangle PWM is easy to implement and requires little processing power for the gating signal generation.

By switching each required IGBT on and off, the voltage output with reference to the neutral of the wye-connected phases will generate a voltage pulse.

Fig. 6.1 – Three-Phase IGBT Bridge
In the above figure 6.1, the IGBT switches with labels are shown. The control signals are from the dq to abc output block. These control signals have a peak height, $V_{\text{control,peak}}$ for each phase. Taking a single phase A for example [5]:

$$v_{\text{control},a} \geq v_{\text{tri}} \rightarrow Q1(\text{on})$$

$$v_{\text{control},a} \leq v_{\text{tri}} \rightarrow Q2(\text{on})$$

When the control signal of phase-a is larger than the triangle comparison wave, the turn the top leg on, thus the voltage output $v_{an} = +V_{\text{dc}}$, if the control signal is less than the triangle wave, then $v_{an} = -V_{\text{dc}}$. The equivalent will happen with phase-b, legs Q3 and Q4, and phase-c, legs Q5 and Q6, except the control signals will be phase shifted by -120 and +120 degrees, respectively. When computing the line-to-line voltage, for example between phase-a and phase-b, $v_{ab} = v_{an} - v_{bn}$.

The PWM generator was implemented in Simulink shown in figure 6.1.

The $v_{\text{control}}$ signal input is from the dq to abc transformation. These values are the voltage control values for each phase. For an inverter the peak phase-to-neutral fundamental value is specified by equation 6.1.
(\hat{v}_{LNpeak})_1 = m_a \frac{V_{dc}}{2} \quad (6.1)

The modulation ratio index, \( m_a \) is the ratio of the peak control voltage level to the peak triangle wave value. For the generated triangle wave, \( V_{tri} \), the amplitude is 1V. The maximum peak output phase-to-neutral voltage is 500V or 612\( V_{LL_{rms}} \) for a dc bus of 1000V and 300V or 367\( V_{LL_{rms}} \) for a 600V dc bus.

The dead-time generator will space the gating on signals by shift the triangle waves by the amount within the dead-time block in figure 6.2. The dead-time will prevent both top and bottom IGBTs from turning on at the same time and shorting the dc bus. The dead-time is necessary due to the specified rise/fall time and turn-on/off delay of the IGBT device.
7 DC-DC Converter

The current generated from the PMLG will charge the dc bus voltage. For a single buoy test design, current will vary depending upon wave conditions, force prescribed and current controller requirements. The purpose of the dc-dc converter is to maintain the 1000V dc bus by regulating the current delivered to a resistive load bank.

For a single 1kW generator test setup dc-dc converter will be located on the same interface as the three-phase active rectifier. The resistive load bank will be located on the test vessel with data acquisition equipment. The resistive load bank will be used to dissipate the power generated. The active rectifier will be located inside the spar of the generator. The small test setup is shown in figure 7.1.

![Test system setup](image)

**Fig. 7.1 – Test system setup**

For a future 12kV high voltage dc (HVDC) transmission, one or more generators are arranged to a common 1000V dc bus. The dc-dc converter in this case is designed to regulate the 1000V dc bus and boost the voltage to 12kV for transmission to shore. If the future setup will include back driving the PMLGs, a bi-directional dc-dc converter would need to be implemented.

![Future test setup](image)

**Fig. 7.2 – Future test setup**
For the small setup, comparing it to the larger future setup, the dc-dc converter and load for the 1kW test setup will act like the dc-dc converter, HVDC transmission, inverter, and utility grid interface in the 12kV setup.

### 7.1 Model Descriptions

#### 7.1.1 Boost Circuit

![Fig. 7.3 – Boost circuit layout](image)

The boost circuit converts the input voltage \( V_{dc} \) to an output voltage \( V_a \) where \( V_a \) is greater than \( V_{dc} \). When the IGBT is switched on by a gating signal, the diode is reversed biased due to the voltage drop across the emitter-collector being of a smaller value than the output voltage, \( V_{dc} \). The current within the inductor will ramp up linearly with large inductance values. Note that energy is proportional to the square of turn on time [11].

\[
V_s = L \frac{di}{dt} = L \frac{\Delta I}{t_{on}} \tag{7.1}
\]

\[
\Delta I = \frac{V_{dc}t_{on}}{L}
\]

\[
E = \frac{1}{2} LI^2 = \frac{V_{dc}^2t_{on}^2}{2L} \tag{7.2}
\]

When the IGBT is switched off, the current in the inductor cannot change instantaneously, thus the voltage polarity is reversed to maintain a flowing current.
The inductor delivers its energy to the load, capacitor and diode until the IGBT is switched on again.

\[ V_{dc} - V_a = L \frac{di}{dt} = L \frac{I_s - I_i}{t_{off}} \]

\[ t_{on} = DT \tag{7.3} \]

\[ t_{off} = (D-1)T \]

\[ V_a = \frac{V_{dc}}{1-D} \tag{7.4} \]

\[ I_{dc} = \frac{I_a}{1-D} \]

When noted that the peak-to-peak current ripple through the inductor is the same then both equations can be combined to form the average output voltage \( V_a \) and \( I_{dc} \). For the \( V_{dc} \) bus to maintain 600V the buck circuit will need to regulate the current drawn from the capacitor. When \( V_{dc} \) drops below the 1000V reference, the PI controller reduces the duty ratio of the IGBT, drawing less energy from the capacitor and thus raising the capacitor voltage. When \( V_{dc} \) increases above the reference, the PI increases the error, increasing the duty ratio which draws more energy from the capacitor and decreases its voltage.

### 7.1.2 Buck Circuit

Fig. 7.4 – Buck circuit layout
The buck circuit converts the input voltage $V_{dc}$ to an output voltage $V_a$ where $V_{dc}$ is greater than $V_a$. The relationship of the output voltage to input voltage is described by equation 7.5.

$$V_a = V_{dc}D$$
$$0 < D < 1$$ (7.5)

When the IGBT is switched on, diode is reversed biased and current will begin to charge up the inductor since $V_{dc}$ is greater than the load voltage $V_a$.

$$v(t) = L \frac{di}{dt} = L \frac{\Delta I}{t_{on}}$$
$$v(t) = V_{dc} - V_a$$ (7.6)

The voltage across the inductor will change at a rate shown in equation 7.6. When the IGBT is switched off, current in the inductor cannot change instantaneously resulting in a polarity change in voltage across the inductor. This polarity change forward biases the diode allowing current to flow in the circuit. The energy in the inductor flows to the load until it is depleted.

$$-V_a = LI(t_1 - t_2)_{t_{on}} \Rightarrow V_a = L \frac{\Delta I}{t_{off}}$$ (7.7)

Peak to peak ripple current is equivalent in steady state as state in equations 7.8, below.

$$\Delta I = \frac{V_a t_{off}}{L} = \frac{(V_{dc} - V_a)t_{on}}{L}$$
$$t_{on} = DT$$
$$t_{off} = (1 - D)T$$
$$V_a = V_{dc}D$$ (7.8)
Since the input voltage needs to be maintained the amount of current flowing from $V_{dc}$ to the load is regulated. To regulate the current from the source to the load control of the IGBT switch duty cycle is changed with a feedback loop referencing the desired input voltage. A control loop references 1000V as the desired input voltage. If more current is delivered to the load than the generator can provide the input voltage will decrease past 1000V. The PI controller integrates the negative error, accumulating, and thus the duty ratio decreases; $v_{\text{control}}$ will decrease lowering the overall duty ratio until steady state has occurred.

Determining the component values can be calculated based on mathematical analysis of the buck converter circuit. The peak to peak capacitor ripple voltage can be expressed in equation 7.9.

$$\Delta v_c = \frac{\Delta I}{8f_sC} = \frac{V_sD(1-D)}{8LCf_s^2} \quad (7.9)$$

7.2 Resistive Loading

For the boost circuit, the load resistance would need to be high enough such that when the regulating IGBT is turned off during low generator velocity, the time constant of the RC circuit is such that the voltage across the capacitor never decays below that of the Vdc bus. If the voltage across the load resistor is less than the Vdc bus, the diode will be forward biased and will collapse the dc bus voltage.

The load resistance connected to the dc bus needs to dissipate a peak power of 1kW. The average power will be defined as the power over a complete ocean wave cycle. A load resistance required to dissipate 1000W at a 1000V dc bus voltage is calculated below in equation 3.6, where $P$ is total power, $V_{dc}$ is dc bus voltage, $I_{dc}$ is dc bus current, and $R_{load}$ is the load resistance.
\[ P = V_{dc} I_{dc} \]
\[ V_{dc} = I_{dc} R_{Load} \]
\[ P = \frac{V^2}{R_{Load}} \quad (3.6) \]
\[ R_{x=Load} = \frac{1000^2}{1000} = 1k\Omega \]

In order to dissipate 1000W of peak power a load resistance of 1000 Ohms is required. For our application, no load capacitance will be used. The tested load resistance is 100 Ohms, resulting in higher peak currents to the load.

### 7.3 Hysteretic Controller

For the controller of the dc-dc buck converter, a hysteretic controller is used. The use of a hysteretic controller will eliminate the design for a PI controller for the dc-dc converter. The hysteretic controller is implemented in Simulink using the relay block. The relay block has inputs for upper and lower bound values. When the input exceeds the upper bound value, the switch turns on. When the input goes lower than the lower bound value, the switch will turn on.

The switch will regulate the dc bus voltage on the dc link capacitor and power flow to the resistive load bank.

![Simulink dc converter model](image)

Fig. 7.5 – Simulink dc converter model
Fig. 7.6 – Hysteretic controller for dc buck converter

The tolerance band for the hysteretic controller is designed for a 2% ripple on a 1000V average dc bus. The maximum dc bus allowable is 1010 V and a minimum of 990 V.
8 POWER TAKE OFF FROM BUOY

8.1 Configurations

For physical testing in the ocean environment, the desired testing arrangement will need to be considered. The power electronics components will need to be located near the real-time controller for data acquisition and gating signal generation. The power take off cable needs to be located near the load bank for dissipation of power from the generator.

With these items in mind, the location for the required equipment will be as follows:

- Power electronics in buoy spar
- Real-time controller in buoy spar
- Power take off cable from buoy spar to ocean vessel
- Load bank on ocean vessel

An initial proposal was for the power electronics to be located on the boat along with the real-time controller and load bank. This arrangement would require substantial three-phase power cables spanning the distance of the ocean vessel to the buoy. Each three-phase AC line would increase phase resistance resulting in additional losses. The controller would also have to be designed for additional phase inductance. With the designed arrangement, the power electronics and real-time controller would be located in the spar. The three-phase AC lines would be replaced with two conductors from the dc-dc converter. The dc load would still be located on the boat due to space restrictions and inadequate ability on the buoy to dissipate 1kW of thermal heat from the load bank. By eliminating a conductor, cost of deployment would reduce. With two conductors, fewer water tight connectors would be needed, increasing reliability and cost.
The load bank in consideration would be a 100 Ohm resistive bank, air cooled or assisted with fans for adequate thermal cooling. Fans would be powered via auxiliary power aboard the ocean testing vessel.

### 8.2 Marine Cables

Cables will need to span from the ocean vessel to the buoy/generator. The dc link cable will be attached to the bottom of the spar and is stationary, thus reducing cable flex, which would decrease cable life and reduce failure.

![Marine cables from the AmerCable Inc. brochure](image)

Fig. 8.1 – Marine cables from the AmerCable Inc. brochure

One solution for marine cables is the certified marine cable from AmerCable Inc. These cables are built to specifications with many options if desired. Conductor number from one to four (three-phase and return ground), with insulators and shielding to reduce EMI and withstand high voltage applications. Fig. 8.1 shows a sample cable provided by their brochure.
8.3 Control of Power Electronics

A National Instruments CompactRIO (cRIO) NI-9012 real-time controller will control the three-phase active rectifier and dc-dc converter. The cRIO is a controller/data acquisition unit that will implement the dq-control, PWM generator and data acquisition. The communication link between the ocean vessel and buoy might include wireless communication via the 802.11g port located on the real-time controller. The 54 Mbps (maximum), transfer rate will allow adequate data transfer.

The cRIO is a stand alone real-time controller requiring 24V external power and a maximum consumption of 20W with 8 add-on modules powered. The cRIO implements a 400Mhz Freescale processor and built in 128 MB of non-volatile ram and 64 MB of DRAM. The cRIO has a built in Field Programmable Gate Array (FPGA) board. The FPGA is a configurable controller that allows connectivity to the provided add-on modules. The FPGA is programmed to provide fast data acquisition and digital output signals.

![Fig. 8.2 – CompactRIO NI-9012 RT Controller](image)

The add-on modules needed for our controller needs are an analog input module and a digital output module.
The data acquisition will be implemented using the National Instruments NI-9205 analog input module. This analog input module has 32 channel inputs with a resolution of 16-bits and has a maximum acquisition rate of 250kSamples/sec. The maximum input voltage is rated at +/- 10V.

The gating signals will be sent to the driver board via the National Instruments NI-9474 Digital Output Module. The module has 8 channels and a logic level output, depending upon the voltage supplied to the module (range of 5-30V). The rise time of the digital output module is a rated 1us. This allows for fast PWM signal generation.
9 HARDWARE IMPLEMENTATION

9.1 Hardware Selection

Hardware verification will compare simulation results to real world testing. Simulations were completed in MATLAB/Simulink 7.1 SP3 with SimPowerSystems blocks. Hardware testing will need to be implemented with hardware that will be suitable for initial lab testing while being flexible enough for future PMLG interfacing.

The PMLG output specifications are important for high-power electronic component selection. The maximum sustained current through the coil windings of the PMLG is limited by the 14 gauge wire used. 14 gauge wire has a maximum current rating of 15A. At a maximum velocity of 2 m/s with and series a coil configuration, the maximum output no load voltage is $1133 V_{LN,pk}$. Higher output voltages results in less current for equivalent power generation, however, high voltage input requires specialized power electronics designed to that can handle these extremes.

The standards for most industrial voltages are $230V_{LLrms}$ and $480V_{LLrms}$ thus requiring a dc bus link voltages on inverters at least 800 volts. With the high output voltage of our machine, a dc bus voltage of at least 1000V is necessary, requiring specialized industrial IGBT module. Most power electronics devices designed for high voltages are rated for high current capacities of tens to hundreds of amps. The proposed linear generator is rated at 1kW to 3kW peak, and so high current capability is not necessary. Future linear generators at 10kW and 100kW will require this high current capability in their power electronics, which needs to be considered.

Snubbers will be used on the IGBT module across each individual IGBT switch to ensure that high voltage spikes are suppressed to prevent failure of the semiconductor devices. As the switch is turned off, the conducted current will fall in 6us or less, depending on the device. This change in current causes a high voltage potential across the switch. If the voltage potential exceeds the semiconductor device rating, it may
cause failure. RC snubbers will be utilized to dissipate the power from the IGBT turn off. The RC snubber consists of a series resistor and capacitor connected to the IGBT in parallel, between the emitter and collector. The snubbers will add loss to the conversion of ac to dc, however the protection they provide is necessary.

Next consideration is the dc bus capacitor voltage regulator. Simulink simulations show promise in the PWM switching as well as the dynamic PI controller. The physical testing will consider the hysteretic controller for simplicity and robustness. The layout of the dc-dc converter will comprise of an IGBT switch connected to the positive dc bus terminal and a diode in series with the IGBT. The anode of the diode will connect to the negative terminal of the dc bus. In parallel with the diode will be the power-take-off cables and a 100 ohm resistive load bank for power dissipation. The dc-dc converter will be subjected to pulsed current and full dc bus voltage. High voltage IGBT and diode components are selected to protect from damage.

The IGBT elements require a positive bias across the base and collector in order to be in saturation. Since there is high loss associated in the linear region of operation, the IGBT is either on or off, saturation or cut-off, respectively. The base requires a current and voltage to turn on the switching element. In the three-phase configuration, two IGBTs are in series, thus the emitter-collector junction between the two IGBTs is at a floating potential. A higher voltage is necessary to turn the IGBT on with forward biasing of the base and collector. A solution to this is a gate driver board. The gate driver board inputs 15V logic level voltage and will output the necessary voltage and current to turn the IGBT on. Designing, building and testing a driver board takes multiple design considerations, ranging from EMI (electromagnetic interference) and PCB layout. Due to time constraints and the numerous tests being considered, a pre-manufactured driver board will be selected.

The features required for each board will allow a selected product to best suit the application needs. Three-phase current sensors are needed to measure and easily interfaced with the CompactRIO analog input module. DC bus link voltage and output dc load voltage is needed for controller needs and data acquisition. TTL (Transistor-
Transistor-Logic) from 0-10 V will be considered since the real-time controller has this input capability. The TTL logic will control the switches on the IGBT and the dc-dc converter. The output current will not be sufficient to drive the IGBT alone and will need to be interfaced with the driver board. The board must be able to handle up to 10V logic for turn on and 0V logic for turn off. Some logic boards require a high side input for logic on and a high input for off, this would double the digital output requirement.

Protection of the digital output control signals is also important. In the case of a fault on the digital output side isolation will protect the driver board; a fault on the drive board would be isolated from the expensive digital output module by a high frequency transformer at its inputs. Isolation is important to protect both sides: digital output module providing control signals and the driver board supplying signals to the IGBTs. Under-voltage protection will turn off gating signals to the IGBT if the supply voltage to the driver board dips below a certain voltage. Over-voltage on the dc bus is also protected against. If the dc bus is unloaded for a prolonged amount of time, the voltage will increase rapidly. This voltage will exceed the rated voltage of the capacitors and damage them. Also, there is potential to destroy the diodes and IGBTs by exceeding their maximum voltage drop $V_{ce}$ across the collector-emitter.

Driver boards come in a single phase, half-bridge and three-phase arrangement. Ideally, the desired configuration would be a three-phase driver board for the active front end and a single-switch driver for the dc-dc converter. Depending on the availability of the driver boards and IGBT modules will be the deciding factor. A three-phase input would require three, single phase boards. This would add to cost and overall footprint. Considering the size limitations within the buoy spar which will hold all the necessary equipment, size is a key deciding factor. If a half-bridge arrangement was utilized, two boards would be necessary. The half-bridge drive board has inputs to drive four IGBT switches. This would allow use of one board to drive four IGBT switches and the remaining switches would be driven by a second board. The second board would also be used to drive the dc-dc converter.
IGBT modules are available in several different configurations: single leg (two switches), half-bridge (four switches), and three-phase (six switches). IXYS, IR (International Rectifier), Semikron, Toshiba/Mitsubishi/PowerEx and Infineon are several manufactures which produce IGBT modules in these configurations. Cost and size regulate the arrangement of the IGBT switches. A single three-phase module would be more costly and troublesome to replace if a failure with a single diode or IGBT occurs. A single leg containing two IGBT switches is a cheaper alternative and is also less costly to repair in the event of a switch failure.

The combining driver board and IGBT module that was selected is the PowerEx PP75T120 Pow-R-Pak. This is an H-Bridge configuration with 4 IGBTs per converter. This assembly provides the driver board and IGBT modules along with the capacitors in a pre-assembled configuration. The included capacitors are rated to 450V and will need to be oriented in a series configuration to handle the 1000V dc bus voltage. The two Pow-R-Pak assemblies required take up a very large footprint and will be modified to reduce overall size.

The Pow-R-Pak provides isolation from the digital control signal input to the IGBT driver circuits. Three-phase current measurements are available on the pin outputs.
The current measurements are from a LEM HAS model, which is supplied +15V from the driver board. The current output reading is a linear voltage representation of the current level, scaled at 20:1. The dc bus voltage measurement output is also linear with a 100:1 voltage scale. Temperature measurements are also made from a thermocouple mounted to the heat-sink and are provided in Celsius as a voltage output with a 20:1 conversion in Celsius.

To simulate the output of the three-phase PMLG, a programmable source will be utilized. The programmable source is a Behlmann 120kVA unit. The Behlmann is connected to a Tektronix/Sony AWG2005 Arbitrary Waveform Generator. This waveform generator can be programmed to output a variable frequency, variable amplitude signal similar to the output of the PMLG. The signal generated from the AWG is fed to the Behlmann source and amplified. The Behlmann output will input into the three-phase active rectifier. The specifications on the Behlmann output are: 0-264VLNrms (646VLLpk) output voltage, with frequency 10-2000 Hz. The output frequency will be limited to a 10 Hz peak electrical frequency as the lower limit however the output is amplitude modulated, thus saturation of the transformer within the programmable source will be an issue when emulating the voltage output of the PMLG at the crest or trough of a wave when the electrical frequency goes to zero. A way to work around this issue is to output variable-voltage amplitude with a fixed electrical frequency. This allows a modulated voltage level while not dropping below the lower limits of the programmable source.
The test setup pictures are shown in the figures below.

Fig. 9.2(a) – 4 IGBT modules mounted on heat-sink

Four IGBT modules are mounted on a heat-sink with thermal grease shown in figure 9.2(a). Bus bars are used on the input phases and the dc bus to reduce inductance in the system. Resistive snubbers are added for each IGBT/diode. Each module contains two IGBT switches. The left module is the dc/dc converter and the remaining three-phase inputs.

Fig. 9.2(b) – Assembled three-phase active rectifier with driver board
Fig. 9.2(b) shows the assembled three-phase active rectifier with mounted driver boards. The driver boards are mounted on top of each other under the metal top mounted plate. The ribbon cables interface with the driver board and Category 5 cable (Cat5). Cat5 cable provides excellent noise immunity over short distances where high speed data transfer is employed. The cabling utilizes twisted pairs, rejecting surround noise. This is important in both data acquisition and with the high-speed gating signals for the driver board.

![Fig. 9.2(c) – Reverse side of the three-phase active rectifier](image)

Figure 9.2(c) shows the reverse side of the three-phase active rectifier. The gating signals from the board to the desired inputs can be seen by the white arrows. The two stacked driver boards are shown to be above the IGBT modules with enough clearance concerning heat dissipation and proximity to the dc bus voltage.
The bus bars to and between the capacitors will reduce system inductance. Each capacitor is rated at 3300uF and 450V rated. In a series configuration, the voltage rating is increased to 1350V while the capacitance is reduced to 1100uF.
The Behlmann programmable source interfaced with the Sony/Tektronix AWG2005 is shown in figure 9.2(e). This 120kVA source will represent a PMLG with variable voltage output and will be interfaced with the three-phase active rectifier.

### 9.2 Passive Rectifier Testing

In order to stay within the limits of the programmable source while generating a representative voltage level, time scaling was necessary. The lower frequency limit for the programmable source is 10Hz. The baseline test is with a velocity of the generator at 0.7 m/s. The peak output electrical frequency is:

\[
\hat{f}_e = \frac{1}{0.144m} \left( \frac{dy}{dt} \right) = \frac{1}{0.144m} \left( 0.7 \text{ m/s} \right) = 4.86 \text{ Hz}
\]
This electrical frequency is well below the limits of the programmable source. A fixed frequency will be used. Scaling time by a factor of 4 results in a peak electrical frequency is 19.44 or approximately 20Hz. The wave period of the 0.7 m/s vertical velocity is from a wave with a period of 4.488 seconds. By reducing the period by a factor of 4, the wave period is changed to 1.122 seconds. Note: Line-to-neutral values are reduced from the voltages produced by the PMLG at 0.7m/s to 161V<sub>Lnpk</sub>. This is due to current limitations on the load bank. The required resistance is 72 Ohms; this would draw a peak 3.3A, which is the rating of the resistance coils used. The input voltage is changed scaled to 161V<sub>LLpk</sub>. Data acquired from a Tektronix TDS5104 Digital Phosphorous Oscilloscope, is exported to a dat file and plotted in Matlab.

Voltage Input: 161V<sub>LNpk</sub>, no dc bus capacitance

Fig. 9.3(a) – Variable-voltage rectifier input (161V<sub>Llpk</sub>)
Fig 9.3(b) – Diode Input Current ($161V_{LLpk}$)

Fig 9.3(c) – DC bus current ($161V_{LLpk}$)
Fig. 9.3(d) – DC bus voltage ($161V_{LLpk}$)

Fig. 9.3(e) – Phase-a voltage and current ($161V_{LLpk}$)
Voltage Input: $161V_{\text{LNpk}}$, dc bus capacitance 1100uF

Fig. 9.4(a) – Variable-voltages rectifier input ($161V_{\text{LLpk}}$ and dc capacitance 1100uF)

Fig. 9.4(b) – Line input current ($161V_{\text{LLpk}}$ and dc capacitance 1100uF)
Fig. 9.4(c) – Line Input Current (zoom) (161V_{LLpk} and dc capacitance 1100uF)

Fig. 9.4(d) – Phase-a voltage and current (161V_{LLpk} and dc capacitance 1100uF)
Fig. 9.4(e) – dc bus capacitor voltage and current ($161V_{\text{LLpk}}$ and dc capacitance 1100uF)

Fig. 9.4(f) – dc Bus capacitor voltage ($161V_{\text{LLpk}}$ and dc capacitance 1100uF)
Fig. 9.4(g) – dc load current (161V_{LLpk} and dc capacitance 1100uF)

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Table 9.1 – Summary of passive rectifier testing (hardware)

### 9.3 Active Rectifier Testing

The active rectifier testing setup was done implementing the CompactRio programmed in LabView. The controller and permanent magnet linear generator design and simulations was preformed using MATLAB/Simulink. The control design and dq-transforms are implemented in LabView. The three-phase current measurements transformed to the dq-reference frame layout are shown in figure 9.5(a) and (b).
Fig. 9.5(a) – Three-phase to dq-reference frame

Fig. 9.5(b) – dq-reference frame to three-phase
Fig. 9.5(c) – Pulse Width Modulated gating signal generator

Fig. 9.5(d) – PI controller
The “Controller Gains” block is a LabView graphical controller that accepts proportional gain, integrator time and derivative time values. Integrator time is defined as \(1/K_i\) and the derivative term is \(1/K_d\), however this is omitted from our controller.

Hardware testing with the CompactRIO was not completely successful due to a number of real-time controller issues. The timing of the real-time controller and the FPGA board would lose connection with the end-user GUI if a set loop interval was set on the FPGA. In order for the controller to properly run, a set loop interval needs to be set otherwise the CompactRIO will compute the math based on an arbitrary speed. For our controller, the loop needs to operate at 50 kHz (20us) or faster to generate the appropriate PWM triangle wave. While programming the FPGA and real-time controller, triangle wave generation was severely aliased resulting in unusable PWM signals.
10 CONCLUSION

Modeling of the permanent magnet linear generator was done using dq-flux equations and simulated in Simulink. An ideal model of the generator was also developed to provide a baseline for simulation comparisons. A current controller was derived from the PMLG machine equations and designed for our generator specifications. Various simulations were performed with the ideal, dynamic and average switching model to verify the controller design. A three-phase active rectifier was designed with protective snubbers and used as a platform to test the controller topology and configuration. A dc-dc converter is used to maintain the dc bus voltage controlled by a hysteretic controller. The dc-dc converter dissipates power into a resistive load bank.

Hardware selection of the IGBT modules and drive boards was specified to ensure reliability, minimize size, maximize power handling capabilities, and allow implementation with the real-time controller. A three-phase 120kVA fully programmable source was used to generate a variable-voltage fixed frequency output. The control signals were provided with the CompactRIO NI-9012 real-time controller with digital output module. All transforms and controller designs were converted into LabView code for compilation to the CompactRIO. During testing, the CompactRIO had problems acquiring data and generating correct triangle waves for the PWM generation. The timing issues, clock speeds and data acquisition on the real-time controller were the source of highly aliased and unusable signals that could not be used for controlling the IGBT gating signals. Hardware testing concluded with diode testing with and without dc bus capacitance.

Future work would include correcting the CompactRIO timing and clock issues to effectively run the compiled controller. An alternative would be a specific real-time controller, such as dspace or Opal-RT real-time controller designed for HIL (Hardware-In-Loop) systems. Lab testing with a 1kW linear generator on the designed
Linear Test Bed at Oregon State University would provide opportunity for controller verification.
Bibliography


Appendices