DEVELOPMENT OF A POWER ELECTRONICS FOR A FLYWHEEL ENERGY STORAGE SYSTEM

by

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Thesis submitted to the Faculty of the
Virginia Polytechnic Institute and State University
in partial fulfillment of the requirements for the degree of
Masters of Science
in
Electrical Engineering

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September 18, 1995
Blacksburg, Virginia
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(ABSTRACT)

The development of a power electronic circuitry for a flywheel energy storage system is discussed in the following aspects. First, due to the nature of permanent magnet brushless DC motor/generator, the operation of three-phase voltage source inverter/rectifier can be simplified to that of a bi-directional DC-DC converter, allowing the use of mostly analog control. Second, there is a problem associated with the existing six-step brushless DC motor/generator control in the generator mode. A twelve-step control scheme is proposed to solve this problem. Third, high-switching frequency is necessary for the flywheel charger/discharger in order to reduce the size/weight of the system and to synthesize the high-frequency motor/generator current waveforms. A working prototype demonstrates that a high efficiency can be achieved at 100-kHz switching frequency by the innovative ZVT soft-switched three-phase inverter/rectifier.
To my parents
Acknowledgments

I would like to express the most sincere gratitude to my advisor, Dr. Fred C. Lee, for his support and encouragement during my studies and research. I would also like to thank my other committee members: Dr. Dusan Borojevic and Dr. Dan Y. Chen for their valuable suggestions and comments. I thank all faculty, staff and graduate students in VPEC who have enriched my experience.

I also acknowledge the supports provided by Virginia Power Technologies, Inc. and Virginia Center for Innovative Technology under a NASA SBIR contract.

This work was based on the project which has been collectively accomplished by a team. I am especially indebted to professor Xinfu Zhuang, who is a visiting scholar from P.R. China for the control circuit design, and Dr. Borojevic for his suggestion of the twelve-step control. I would like to thank Dr. Dan Sable and Dr. Gui-chao Hua for their advice and help.

With much love and gratitude, I thank my parents Ruichang Zhang and Suzhe Tian for their love, support and prayers.
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1. INTRODUCTION

1.1 Overview

Battery has been used for energy storage for many years, and it would still be the primary energy storage device in electronic equipment in the future, especially in portable electronics. However, the battery system does have many problems. Its lifetime is limited, weight is heavy, especially that of the lead acid battery, the disposal of chemical battery causes environmental concerns.

With the development of power electronics, many new energy storage systems such as the superconducting magnetic energy storage, the flywheel energy storage, and the capacitive energy storage, etc. are being intensively studied recently in order to replace battery in some special applications. Among these innovative energy storage systems, the flywheel system exhibits some unique features such as high-power density, easy maintenance, and longer lifetime.

Figure 1.1 shows a flywheel rotating at an angular speed of $\omega$. According to [1], the moment of inertia of the flywheel is:

$$I = \frac{1}{2} mR^2,$$  \hspace{1cm} (1.1)

where $m$ is the mass of the flywheel, and $R$ is radius of the flywheel.

The kinetic energy stored in this flywheel is:

$$K = \frac{1}{2} I \omega^2,$$  \hspace{1cm} (1.2)
Figure 1.1. A Rotating Flywheel and its Kinetic Energy Storage.
where I is moment of inertia, and \( \omega \) is the angular speed.

The angular momentum of the flywheel is:

\[
L = I\omega .
\]  

(1.3)

It is seen that kinetic energy stored in the flywheel is quadratically proportional to the rotational speed, and linearly proportional to the mass. Therefore, for the same energy storage, less mass is needed if the rotational speed is increased, and angular momentum does not change with the mass. In order to reduce weight and size, it is necessary to run the flywheel at high speeds.

As a matter of fact, the flywheel has been used for centuries as a cheap short-term energy storage, especially in machinery applications. Using today's technology, the flywheel is directly mounted to the rotor of the motor/generator, and an inverter/rectifier is used to transform kinetic energy to electrical energy, or vice versa. Figure 1.2 illustrates a flywheel energy storage system. The DC bus is connected to both power supply and load, and needs to be regulated all the time. When the power provided from the power supply is more than what the load actually needs, the surplus energy will be transferred into kinetic energy by spinning up the flywheel. When the load requires power larger than what the power supply can provide, the flywheel will release its energy by spinning down.

The flywheel storage systems have many potential applications. One important application is to replace battery in the uninterruptible power system (UPS) [2], making it ideal for today's environmentally-conscious market.
Figure 1.2. Block Diagram of a Flywheel Energy Storage System.
Future regulation of vehicular emission levels has prompted a recent increase of interest in electric and hybrid vehicles. Electric vehicles have been limited in range and performance by the storage capability and size/weight of currently available batteries. Flywheel is deemed an attractive alternative to batteries [3,4].

Another application is electric utility load leveling. The flywheel is used to store energy during times when energy is inexpensive to produce, and then return it to the customer during times of peak power demand when generated energy is most expensive [5].

This paper discusses the flywheel energy storage system in Low Earth Orbit Satellite (LEO) applications, where spacecraft power is traditionally provided by photovoltaic cells and batteries. The batteries store excessive photovoltaic power from solar array when the satellite is out of eclipse and sees a 60 minute of sunlight, and provide all the power when satellite is in eclipse and sees a 30 minute of darkness. If a lifetime of 17 years is required for the satellite, the energy storage system must be capable of sustaining approximately $10^5$ charging/discharging cycles. Currently, the Nickel-Cadmium battery (at 1.3 volts per cell) or Nickel-Hydrogen battery (at 1.3 volts per cell) is used for energy storage. To achieve the necessary bus voltage (about 120V), cells are placed in series, and the resultant system has a cyclical energy density of 6.6 Wh/kg for the Nickel-Cadmium battery, and 12.1 Wh/kg for the Nickel-Hydrogen battery. Electrochemical batteries are well known to suffer from limited cycle lifetimes, reliability problems of numerous cells in series, difficulties in measuring the state of
charging, and the inability to test the actual batteries in the spacecraft [6]. Because of these difficulties with batteries, NASA has been engaged in an active ongoing effort to look into a magnetically suspended composite flywheel energy storage system. At this time, a flywheel and a 500-W motor/generator are under construction in University of Maryland [7]. The flywheel is made of graphite/epoxy composite, and is suspended by magnet biased active magnetic bearings. A motor/generator is a 3-phase, 4-pole, permanent magnet brushless DC design, and provides the means of transferring power to and from the system [8]. The final flywheel system will have a usable energy density greater than 20 Wh/Kg. The research project conducted in Virginia Power Electronics Center is to develop a power electronics for this flywheel energy storage system.

1.2 Design Target

The design target of the flywheel charger/discharger is derived to be compatible with a spacecraft bus such as on the NASA EOS-AM spacecraft:

- **Spacecraft DC bus voltage:** 120 V
- **Maximum motor/generator back EMF:** 100V_{\text{line-line}}
- **Maximum average power:** 500 W
- **Peak transient power:** 1 kW
- **Maximum motor/generator speed:** 90,000 rpm
- **Maximum motor/generator electrical frequency:** 3 kHz
- **Switching frequency:** >100 kHz
- **Bus regulation loop bandwidth:** >1 kHz
Efficiency of full power operation: >97% at nominal operation (80-V back EMF voltage, 500-W output)

The charge and discharge functions of flywheels are significantly more complicated than battery charger/dischargers. Most importantly, the flywheel charger/discharger is a three-phase AC/DC converter, rather than a DC/DC converter as in a battery power system. The charger/discharger has to provide the following features:

-------- Bi-directional power flow.

-------- Precisely shaping motor/generator phase currents. In order to minimize the pulsating torque and mechanical vibration of the flywheel, which is rotating at extremely high speeds, the charger/discharger has to provide a constant power flow.

-------- Variable frequency waveforms. Due to the variable flywheel speed, the converter must provide currents and voltages in the frequency range from 0 Hz to 3 kHz.

-------- DC voltage regulation. The charger/discharger has to provide tight regulation of the DC bus voltage.

-------- High efficiency and high power density. These requirements are imperative for reduction of overall system weight and volume.

1.3 Thesis Outline

Chapter 2 presents a control strategy of the flywheel energy storage system. Trapezoidal brushless DC motor/generator structure greatly simplifies the control of the three-phase inverter/rectifier to that of a bi-directional DC/DC converter. Since the six-
step control exhibits some problems in the rectifier mode, twelve-step control is proposed to solve these problems.

In Chapter 3, small-signal analysis of the flywheel energy storage system is performed, and compensator design example is given. Physical explanation of the peculiar small-signal behavior of the bi-directional converter is provided.

Chapter 4 discusses the problems associated with the hard-switched voltage source inverter/rectifier using MOSFET anti-parallel diodes as rectifier diodes. A newly developed three-phase Zero-Voltage-Transition (ZVT) soft-switched inverter/rectifier is considered particularly suitable for this application. The comparative study and design consideration have been provided.

Conclusions are given in chapter 5. A brief summary of this project is introduced, and remaining issues which need to be addressed in the future are also discussed.
2. CONTROL STRATEGY OF THE FLYWHEEL ENERGY STORAGE SYSTEM

The flywheel energy storage system is intended to replace the battery system in satellite applications. The flywheel is mounted to the rotor of a three-phase motor/generator, and a three-phase voltage source inverter/rectifier is employed to drive the motor/generator. As discussed in Chapter 1, the kinetic energy storage in a flywheel is quadratically proportional to the angular speed of the flywheel. Derivation of Equation 1.2 is

\[ K' = I\omega' \]  \hspace{1cm} (2.1)

Equation 2.1 shows that when a flywheel is rotating at a extremely high speed, a small amount of speed variation can significantly change the kinetic energy storage of the flywheel. In other words, the rotational speed of a high speed flywheel changes in a very slow fashion with a finite amount of charge/discharge power. For example, in an LEO satellite, it takes about 30 minutes to release three quarters of the stored kinetic energy by reducing the flywheel speed by half. The back EMF of the motor/generator, which is linearly proportional to angular speed of the flywheel, changes very slowly, and can be practically treated as a three-phase voltage source, regardless of the flywheel dynamics. Therefore, the control of the flywheel system can be simplified to the control of a three-phase voltage source inverter/rectifier. The shape of back EMF waveforms is determined
by the structure of the motor/generator, and can significantly affect the way
inverter/rectifier is controlled.

2.1 Introduction

2.1.1 Classification of Brushless Permanent Magnet Motors

Brushless permanent magnet motors are becoming increasingly attractive in servo
and variable-speed applications, since they can produce the torque-speed characteristics
similar to that of a permanent magnet dc motor, while avoiding the problems of brushes
and mechanical commutation. For integral horsepower (hp) sizes, brushless permanent
magnet motors have been shown to be more efficient than induction machines of
comparable size, due to reduced copper losses resulting from the lack of rotor current
[9,24,31]. This is one of the reasons why permanent magnet brushless dc structure is
selected for the design of the three-phase motor/generator for the flywheel energy storage
system [8].

Brushless permanent magnet motors can be classified as of a sinusoidal or
trapezoidal type, according to the rotational voltage induced (back EMF) [10,11]. Motors
with a distributed winding are referred to as permanent magnet synchronous motors.
Motors with a concentrated winding are referred to as brushless dc motors. The winding
distribution does not necessarily correspond to the waveform of the induced EMF, since
the permanent magnet arrangement also affects the waveform of back EMF. The surface
magnet design produces a trapezoidal distribution of airgap flux, and a concentrated
stator winding in this type of machine will have a trapezoidal induced EMF waveform.
Interior magnet machines with distributed stator windings will more likely have a sinusoidal EMF. The waveform of induced EMF has a significant impact on the type of excitation method selected.

The flat top of the trapezoidal induced EMF waveform is well matched to the square wave voltage and current waveforms that can be produced by a voltage source inverter (VSI), and pulse width modulation (PWM) is used with an essentially constant pulse width. This is called trapezoidal excitation or brushless dc control scheme.

For motors with sinusoidal back EMF, sinusoidal excitation is favored, and a current regulated inverter is used to supply sinusoidal currents to the stator windings. Good torque performance with minimum torque ripple can be achieved by this method.

The trapezoidal scheme is easier to implement, and allows the use of mostly analog control. Therefore, it is also a common practice, even for sinusoidal PM motors [11]. One disadvantage of the trapezoidal excitation is the need for accurate electrical commutation, which is controlled by rotor positions. There are many methods which can be used for rotor position detection, such as magnetic position encoder, absolute optical encoder, incremental optical encoder, and resolver [12].

The motor/generator in the flywheel energy storage system is under construction in the University of Maryland, and is intended to be a permanent magnet brushless dc design. The permanent magnet is mounted to the inner surface of ring structure rotor, and produces trapezoidal flux distribution along the airgap [8,24]. As will be discussed in Chapter 3, the motor/generator armature inductance needs to be minimized for wide
control loop bandwidth, therefore, the stator winding has to be wound with few turns and with concentrated winding distribution. Under these circumstances, the motor/generator will have a trapezoidal induced EMF waveforms, which are drawn in Figure 2.1.

The advantages of the PM brushless dc motor over a PM synchronous motor are up to 15% higher torque and power density, simpler power converter control and simpler construction [26].

2.1.2 Control of Brushless DC Motor/Generator

Figure 2.1 shows the voltage and current waveforms of the 120° inverter system [27,28,29]. Each phase winding conducts current only during the 120° top flat portions of its induced EMF. Only two phases conducts current at any time; the third phase current is zero. In the shaded 60° duration shown in Figure 2.1, phase A and phase B conduct current, while phase C has zero current. Using 120° control, the power flow into the motor is constant if the quasi-square armature current is provided; therefore, it theoretically yields a torque-ripple free system, which is required in the satellite applications.

The voltage source inverter/rectifier used to drive the three-phase motor/generator is shown in Figure 2.2. Three boost inductors are motor/generator armature inductors, and the back EMF of the motor/generator is represented as a three-phase voltage source, as discussed before. The DC bus voltage $V_{bus}$ needs to be regulated all the time, and solar
Figure 2.1. Waveforms of a Brushless DC Motor with 120-Degree Drive.
Figure 2.2. Voltage Source Inverter/Rectifier Used to Drive the Motor/Generator.
array is approximated as a current source. Figure 2.3 shows the six-step brushless dc control; the hatched areas indicate that switches are pulse width modulating. Two switches on one switch leg turn on and off complementarily with a small dead time in order to simplify the implementation of the control.

The 360° electrical cycle is divided into six 60° segments. Commutation occurs every 60°. During the 60° section shaded in Figure 2.1 and in Figure 2.3, the equivalent circuit of the three-phase inverter/rectifier is drawn in Figure 2.4. Switch S4 is always on during this 60° section, and switches S1 and S2, which are on the same leg, turn on and off complementarily. This is a well-known dc/dc bi-directional converter, is the same circuit as the battery charger/discharger currently used in the satellite battery energy storage system [13,14,15].

2.1.3 Current-mode Control of the Bi-directional Converter

Figure 2.5 shows a current-injection control, i.e. peak current control, for the bi-directional dc/dc converter, which is the equivalent of the three-phase inverter/rectifier during the 60° section shaded in Figure 2.3. The DC bus of the flywheel system is connected to both solar array and load. The solar array is approximated by a current source Isa, and the load of satellite is denoted as RL. When the solar array provides current larger than what the load needs, the extra current from the solar array will be used to charge the flywheel. When the solar array provides current less than what the load needs, the flywheel will be discharged to provide power to the load. The outer loop is the
Figure 2.3. PWM Scheme of the six-step control.
Figure 2.4. The Equivalent Bi-directional DC/DC Converter
Figure 2.5. Current Injection Control of the Bi-directional DC/DC Converter
voltage loop, which has a relatively small bandwidth and a large DC gain. The inner loop is the current loop using Current Injection Control (CIC), which has a relatively large bandwidth. The current-injection control is a commonly used approach of controlling a boost converter, and has a much better dynamic performance than a single-loop control. For the current loop alone, traditionally, the current of switch S2 needs to be sensed in the boost operation, i.e. in the discharge mode; while the current of switch S1 needs to be sensed in the buck operation, i.e. in the charge mode. During this design, the current of phase A inductor is sensed and used as CIC control signal in both the charge mode and the discharge mode.

Figure 2.6 illustrates how the current-injection control works in a bi-directional converter. Current peak is needed in the discharge mode, while current valley is needed in the charge mode [19]. For example, when the bus voltage, \( V_{\text{bus}} \), is larger than reference, voltage compensator output \( V_e \) will decrease. In the flywheel discharge mode, inductor current \( I_a \) is positive, and duty cycle of switch S2 will decrease due to reduced \( V_e \). As a result, \( I_a \) will reduce. Therefore, less power flows out of the motor/generator, and bus voltage drops.

In the flywheel charge mode, inductor current \( I_a \) is negative, and \( V_{\text{bus}} \) larger than reference causes voltage compensator output \( V_e \) to decrease into the negative region. Duty cycle of S2 decreases, and the absolute value of \( I_a \) increases. Therefore, as more power is drawn from the \( V_{\text{bus}} \) into the motor/generator, the bus voltage drops.
Figure 2.6. Illustration of the Current Injection Control.
In the discharge mode, the current peak is limited by \( V_e \); in the charge mode, the current valley is limited by \( V_e \). Since the inductor current ripple can be designed relatively small, common current reference can practically limit inductor current in both the charge mode and the discharge mode with an offset of current ripple.

Switches S1 and S2, which are on the same leg, turn on and off complementarily with small dead time. External ramp \( S_e \) shown in Figure 2.5 is necessary for eliminating subharmonic oscillation when duty cycle of the boost switch S2 exceeds 50% [16].

The dc/dc bi-directional operation discussed in Figure 2.5 and Figure 2.6 is valid for the 60° section indicated in Figure 2.3. In the next 60° section, phase A and phase C will be used for constructing a bi-directional dc/dc circuit, and phase B current will be forced to zero by turning off switches S3 and S4.

**2.2 Control Electronics of the Flywheel System**

The motor/generator control electronics is designed to regulate a +120-V spacecraft bus with equal and potentially better dynamic performance than is ordinarily obtained with a conventional battery charger/discharger system. Figure 2.7 shows a block diagram of the control electronics. The control compensator design and ZVT soft-switched inverter/rectifier will be discussed in Chapter 3 and Chapter 4 respectively, and the detailed schematic diagram will be provided in Appendix.

The pulse-width modulation and commutation scheme is based on the six-step brushless dc control, illustrated in Fig. 2.3. The position sensor output is three phase-
Figure 2.7. Control Block Diagram of the Flywheel System
shifted square-wave signals; the unique combination of these signals determines rotor position with 60° resolution. Figure 2.8 shows how position sensing signals distinguish these 60° sections. \( V_{ab}, V_{bc} \) and \( V_{ca} \) are three TTL signals from the rotor position sensor. These TTL signals are uniquely combined as six states, i.e. 001, 101, 100, 110, 010, 011. The commutation decoder and the commutation logic is made of a GAL device and used to determine the proper state for each of the six switches in the inverter/rectifier according to the rotor position of the motor/generator. The bi-directional motor/generator currents are detected with Hall-effect current sensors. The current sensor outputs are used for overcurrent protection and for current injection control. The switching frequency of the converter is 100 kHz, and three-phase current signals are summed with a 100-kHz ramp to prevent subharmonic oscillations. The spacecraft bus voltage is sensed and compared to a reference voltage. The difference is amplified and provides the error voltage, \( V_e \). The error voltage is summed with each of the current sense signals to the pulse width modulator (PWM). There are three current loops, one for each phase. Therefore, three independent PWM signals are generated, of course, during each 60° segment, only one PWM signal will be selected, and steered to the proper switches of the inverter/rectifier by the commutation logic.

The control electronics is designed for a trapezoidal PM brushless dc motor/generator. The current waveforms provided to the motor/generator are quasi-square. This greatly simplifies the design of the control electronics, allowing the use of
Figure 2.8. Position Sensor Signals Used to Distinguish 60-Degree Sections.
mostly analog control. At present time, this is the preferred type of motor/generator in this application. A permanent magnet synchronous motor/generator with air gap modulation would require sinusoidal current waveforms and a more advanced digital controller.

2.3 Test of the Charger/Discharger with a Simulated Motor/Generator

Since a motor/generator with the proper characteristics could not be located, the power and control electronics were tested by a motor/generator simulator as shown in Figure 2.9. The motor/generator simulator consists of a high-power three-phase amplifier together with a control circuit designed to output a three-phase signal at varying frequency and amplitude. The motor/generator armature inductance is simulated with three 60μH inductors. Since the amplifier can only source power, a power resistor was placed across each of the three phases to allow the simulated flywheel charging. Voltage source Vs is used to power the inverter/rectifier in the charge mode, and resistance Rs is used to increase the voltage source impedance so that V_bus can be regulated during the charge mode. Diode D is used to prevent the inverter/rectifier from flooding the power source Vs. This test setup allows the simulation of three-phase brushless dc motor/generator back EMF waveforms up to 3-kHz electrical frequency, and up to 100-V line-line voltages. The DC bus voltage, V_bus, needs to be regulated, no matter whether the system is in the charge mode or in the discharge mode.
Figure 2.9. Test Setup for the Flywheel Charger/Discharger with the Simulated Motor/Generator.
TECHRON model LVG-7200 line voltage generator is used as a three-phase amplifier, which has sufficient bandwidth for 3-kHz square-wave output. Figure 2.10 shows the output of three-phase amplifiers. The edges are rounded up by RC filters in the control circuit in order to simulate the soft-edged back EMF voltages. With the 120° operation discussed above, these soft edges should have no influence on the system. Based on the information of three-phase back EMF waveforms, the rotor position TTL signals are generated by another electronic circuitry.

2.4 Problems Associated with the Six-Step Control

The six-step control shown in Figure 2.3 is a standard method of controlling the brushless dc motor, and works very well for conventional brushless dc motor applications, where the three-phase inverter/rectifier works primarily in the inverter mode, i.e. the motor mode. Figure 2.11 shows the experimental waveforms taken in the inverter mode. The bottom curve is the rotor position signal of phase A. Electrical frequency is 300 Hz, DC bus voltage is 120 V, and back EMF line-to-line voltage is 70 V. The charge power is 450 W.

Figure 2.12 shows the phase currents in the rectifier mode. The bottom curve is the rotor position signal of phase A. Electrical frequency is 300 Hz, DC bus voltage is 120 V, and back EMF line-to-line voltage is 70 V. The discharge power is 500 W. The phase currents are supposed to be quasi-square, however, it is seen that the negative portion of phase current does not have a sharp edge as the current in the inverter mode.
Figure 2.10. Outputs of the Three-Phase Amplifiers.
Figure 2.11. Phase Current Waveforms in the Inverter Mode Operation.

Time scale: 1 ms/div, current scale: 10 A/div.
Figure 2.12. The Phase Current Waveforms in the Rectifier Mode Operation.

Time scale: 1 ms/div, current scale: 10 A/div.
2.4.1 Detrimental Impact on the Flywheel System

In order to facilitate the analysis, the slow current transition in the rectifier mode is simplified as a linear curve as shown in the shaded area of Figure 2.13. The shaded area indicates the current transition between phases, i.e. from phase A to phase B, and the width of the shaded area represents how much time the transition takes, and is defined as delay angle.

The operation of PM brushless dc motor/generator heavily relies on the accurate commutation, delay of the commutation causes torque ripple. If the dynamic response of the armature inductors can be ignored and conversion efficiency is 100%, the power released from the flywheel is equal to the output power of the motor/generator, then the instantaneous torque in the motor/generator can be calculated as

$$\Gamma = \frac{1}{\omega} (I_a V_a + I_b V_b + I_c V_c) ,$$

(2.2)

where $\Gamma$ is the torque, $I_a$, $I_b$ and $I_c$ are motor/generator phase currents, $V_a$, $V_b$ and $V_c$ are the magnitude of motor/generator back EMF, and $\omega$ is the angular speed of the flywheel.

The relationship between peak torque ripple and transition delay angle is plotted in Figure 2.14, and the torque ripple in this case is a dip as shown in Figure 2.13.

Because of the torque ripple, the average output power of the motor/generator is reduced, and the relationship between the average output power reduction and delay angle is plotted in Figure 2.14.
Figure 2.13. Simplified Current Transition Waveforms in the Rectifier Mode.
Figure 2.14. Impact of the Slow Current Transition on Torque Ripple and Output Power.
Torque ripple in this flywheel system influences spacecraft attitude control and magnetically suspended bearing, and should be avoided as much as possible. Output power reduction causes motor/generator less utilized, and overdesign of the motor/generator increases the weight/size of the flywheel system.

It is seen that both torque ripple and output power reduction due to the slow current transition are strong functions of delay angle $\alpha$. In Figure 2.12, where electrical frequency is 300 Hz, the delay angle is about $40^\circ$. Based on Figure 2.14, the torque ripple is about 17%, and output power reduction is about 3.8%.

Figure 2.15 shows the 3-kHz phase currents in the rectifier mode. The bottom curve is the rotor position signal of phase A. Electrical frequency is 3 kHz, DC bus voltage is 120 V, back EMF line-to-line voltage is 70 V. The discharge power is 500 W. It's seen that the delay angle is about $60^\circ$. Based on Figure 2.14, the torque ripple is about 25% and output power reduction is about 8%. It’s seen that actual influence of the slow current transition covers more than $60^\circ$.

The cause of current distortion arises from the assumption that when both switches on one leg of three-phase inverter/rectifier are turned off, the current of the phase which is connected to this leg should decay to zero automatically. In reality, the current in the phase to be switched off may take on various waveforms. An analysis of this phase current in the inverter mode was performed in [17], and it shows that this current may be negative, positive or zero, depending upon the rotor speed, the DC source

2. CONTROL STRATEGY OF THE FLYWHEEL ENERGY STORAGE SYSTEM
Figure 2.15. The Phase Current Waveforms in the Rectifier Mode Operation.

Time scale: 100 us/div, current scale: 10 A/div.
voltage, and the advance in the firing angle. In the flywheel system, the operation of the inverter/rectifier is simpler; there is essentially no DC source voltage change, and no advance in the firing angle. Therefore, more specific analysis regarding the uncontrolled current is needed to see how the uncontrolled phase current behaves in both the flywheel charge mode and the flywheel discharge mode.

2.4.2 Transition among 60° Segments in the Inverter Mode

Figure 2.16 shows the first case in the inverter mode, when the flywheel is charged. During the 60° portion before the transition point, switch S1 is pulse-width modulated, switch S4 is normally ON. When switch S1 is turned on, armature inductors of phase A and phase B are charged by the DC voltage source; when switch S1 is turned off, the energy stored in armature inductors of phase A and phase B is discharged by back EMF voltages. At the transition point, switch S4 should be turned off, and switch S6 should be turned on. The current of phase B flows through the antiparallel diode of S3; then, the DC voltage is applied across the node B and node C. Since the DC voltage is larger than line-to-line voltage $V_{BC}$, which is zero at the transition point, the current in phase B will decay, and the current in phase C will increase. The current transition from phase B to phase C is rapid. Simulation result in Figure 2.17 proves the rapid current transition.

During the analysis, a sinusoidal waveform is used instead of a trapezoidal waveform just for convenience. The original scheme is that both switches in one leg
Figure 2.16. Transition between 60-degree Segments in the Inverter Mode.
Figure 2.17 Simulation Results of the Inverter Mode Operation---Case I

2. CONTROL STRATEGY OF THE FLYWHEEL ENERGY STORAGE SYSTEM
should turn on and off complementarily for the bi-directional operation. However, if the current is known to flow through a diode, turning on or off the antiparallel switch is actually an extra switching action. All extra switching actions have been omitted in the analysis for the sake of simplicity, and the result should not be affected.

Figure 2.18 shows another case in the inverter mode. During the $60^0$ portion before the transition point, switch $S1$ is pulse-width modulated, while switch $S6$ is normally ON. When switch $S1$ is turned on, armature inductors of phase A and phase C are charged by the DC voltage source; when switch $S1$ is turned off, the energy stored in armature inductors of phase A and phase C is discharged by back EMF voltages. At the transition point, switches $S1$ and $S2$ should be turned off, and switch $S3$ should start to pulse-width modulate. The current of phase A flows through the antiparallel diode of $S2$. When $S3$ is turned on, the DC voltage is applied across the nodes B, A and nodes B, C, the current of phase A decreases in value, and the currents in phase B and phase C increase. When switch $S3$ is off, back EMF will cause the currents of the three phases to decrease. Therefore, the current of phase A can be transferred to phase B rapidly. Simulation result in Figure 2.19 proves the rapid current transition.

2.4.3 Transition among $60^0$ Segments in the Rectifier Mode

Figure 2.20 shows the first case in the rectifier mode, when the flywheel is discharged. During the $60^0$ portion before the transition point, switch $S2$ is pulse-width modulated, and all other switches are OFF. When switch $S2$ is turned on, armature
Inverter Mode -- Case II

Figure 2.18. Transition between 60-degree Segments in the Inverter Mode.
Figure 2.19 Simulation Results of the Inverter Mode Operation—Case II
Figure 2.20. Transition between 60-degree Segments in the Rectifier Mode.
inductors of phase A and phase B are charged by the back EMF; when switch S2 is turned off, the energy stored in armature inductors of phase A and phase B is discharged by the DC voltage source. At the transition point, the phase B current flowing through D4 should be transferred to phase C by flowing through D6. However, the only switching action comes from S2, which has no influence on the current transition between phase B and phase C. The driving force comes from the voltage difference between phase B and phase C, which is zero at the transition point. Of course, the voltage difference between phases B and C will gradually increase after the transition point, but in a very slow fashion. Simulation result in Figure 2.21 proves this slow transition. For the trapezoidal back EMF, the voltage difference between phases B and C after the transition point increases with a slightly different speed, but the current transition would be similarly slow anyway. This is the reason why the phase currents in Figure 2.12 do not have sharp transition edges in the negative portion.

Figure 2.22 shows another case in the rectifier mode. During the 60° portion before the transition point, switch S2 is pulse-width modulated, and all other switches are OFF. When switch S2 is turned on, armature inductors of phase A and phase C are charged by the back EMF voltage; when switch S2 is turned off, the energy stored in armature inductors of phase A and phase C is discharged by the DC voltage source. At the transition point, switch S2 should be turned off, and switch S4 should be pulse-width modulated. The current of phase A flows through the diode of S1; then the DC voltage, which is larger than line-to-line voltage $V_{AC}$, is applied across node A and node C until
Figure 2.21 Simulation Results of the Rectifier Mode Operation---Case I
Figure 2.22. Transition between 60-degree Segments in the Rectifier Mode.
the current in phase A decays to zero. The current transition from phase B to phase C should be rapid. Simulation result in Figure 2.23 proves it.

The six-step operation of the inverter/rectifier can be summarized into four cases discussed above. Of the four cases, one case in the rectifier mode has difficulty with the commutation among $60^\circ$ segments when six-step control is used.

Although this phenomena exists in the traditional brushless dc system using six-step control, the application requires the motor operation most of the time, and the distorted currents in the generator mode rarely do any harm to the system, since this mode is used only temporarily. However, in this flywheel system, the generator mode is equally important as the motor mode, since the tight DC bus regulation and precise current shaping should be maintained in both modes. The commutation delay problem has to be solved.

2.4.4 The Twelve-Step Control of Brushless DC Motor/Generator

One method of solving the slow current transition problem is to modify the six-step control into the twelve-step control. The twelve-step control involves dividing the $360^\circ$ into twelve $30^\circ$ segments, as shown in Figure 2.24. The hatched area represents the pulse-width modulation. The control is completely symmetrical, and does not consist of two cases, either in the inverter mode or in the rectifier mode, as it did for the six-step control. The transition among these $30^\circ$ segments can be summarized into only one case in both the inverter mode and the rectifier mode. In the inverter mode, the transition of
Figure 2.23 Simulation Results of the Rectifier Mode Operation---Case II
Figure 2.24. The Twelve-step Control of Brushless DC Motor/Generator.
the twelve-step scheme is the same as in case II of the six-step scheme shown in Figure 2.18, and there is no transition problem, as the above discussion indicates. In the rectifier mode, the transition of the twelve-step scheme is also the same as in case II of the six-step scheme shown in Figure 2.22, and, likewise, there is no transition problem, as the above discussion explains. Therefore, by using the twelve-step control, the current transition problems associated with the six-step control are solved.

One difficulty associated with the twelve-step control is the need for sensing rotor electrical position with $30^\circ$ resolution, for which a more accurate rotor sensor is required [12], and the logic table of the encoder in the control circuitry shown in Appendix needs to be rewritten. Except for this, the rest of the circuit in Figure 2.7, such as the voltage compensator network and current control loop, will remain unchanged.

Therefore, the twelve-step control can be implemented to solve these problems associated with the six-step control scheme with little addition of complexity.

2.5 Summary

A brushless dc motor/generator with trapezoidal back EMF waveforms will be used in the flywheel energy storage system. The six-step control scheme based on this type of motor/generator can simplify the control of three-phase inverter/rectifier to that of a bi-directional dc/dc converter.

Current-mode control of a bi-directional dc/dc converter is introduced to the design of the control electronics of the flywheel energy storage system.
The test of the flywheel energy storage system is conducted by using a simulated brushless dc motor/generator.

When the 120° control scheme is used, there is always one phase conducting zero current during each 60° segment. However, in rectifier mode, it is sometimes difficult to keep one phase conducting zero current by using the six-step control. The slow current transition will cause torque ripple and reduce average output power.

The twelve-step control is introduced to solve these problems associated with the six-step control with little addition of complexity.
3. SMALL-SIGNAL ANALYSIS OF THE FLYWHEEL ENERGY STORAGE SYSTEM

The bus regulation loop bandwidth requirement is intended to improve the DC bus impedance. In order to avoid the system interaction problem, the output impedance of the energy storage system should be kept as small as possible, and the closed-loop bus impedance is equal to its open-loop impedance divided by the voltage loop gain plus one [32]. There are two kinds of methods to reduce the bus impedance, one is to increase the DC bus capacitance and reduce the capacitor ESR, the other is to increase the voltage loop gain and bandwidth. Obviously, the second approach is better in terms of reducing the system size/weight. A typical satellite battery charger/discharger has a bus regulation loop bandwidth of about 1 kHz [15,20], therefore, the flywheel system should have at least 1-kHz voltage loop bandwidth in order to achieve the same or better dynamic performance.

In order to design the flywheel energy storage system with a bus regulation loop bandwidth greater than 1 kHz, it is necessary to establish the system model, on which the compensator design will be based. Due to using the brushless DC motor/generator control scheme, the spacecraft bus control dynamics will be independent of the motor/generator rotational speed, and the operation will become similar to that of a bi-directional DC/DC converter, except during the transitions among 30° segments. As shown in Figure 2.19 and Figure 2.23, the transient disturbance due to the current commutation among phases
lasts no more than a couple of switching cycles, and well beyond the system bandwidth. Therefore, due to using twelve-step control, transitions among $30^\circ$ segments do not have any significant influence on the system dynamics, and can be ignored. The small-signal model of the bi-directional DC/DC converter has been intensively studied in [19,20], and will be directly applied to the flywheel energy storage system.

3.1 Small-Signal Analysis of the Flywheel System in the Discharge Mode

During each $30^\circ$ segment, the equivalent circuit of the three-phase inverter/rectifier is represented as a bi-directional DC/DC converter, such as the circuit in Figure 2.5. The small-signal model of an equivalent bi-directional DC/DC converter has been discussed in [19], and is drawn in Figure 3.1, where inductor $L$ is the combined armature inductance of two phases, $V_{\text{EMF}}$ is the back EMF of the motor/generator, $D$ denotes the steady-state duty cycle of the buck switch such as S1 shown in Figure 2.5, $\hat{d}$ denotes duty cycle perturbation of the buck switch S1, $F_m$ is derived from the inductor current slope when buck switch S1 is turned on, and the polarity of $I_L$ is positive when the converter operates in the buck mode. Due to the complementarily switching of buck switch S1 and boost switch S2, the control circuit design can be based on either buck switch S1 or boost switch S2, and the duty cycle perturbation of the boost switch S2 is $-\hat{d}$. $r_{eq}$ denotes the combined impedance of the solar array and a constant power load in parallel,
Fig. 3.1. Small-signal Model of the Flywheel Charger/Discharger
\[
\frac{1}{r_{eq}} = \frac{1}{r_0} - \frac{1}{r_s}, \tag{3.1}
\]

\(r_s\) is solar array impedance, \(r_0\) is constant power load impedance, and both of them exhibit negative dynamic characteristics at low frequencies. Negative sign in Equation 3.1 is due to the definition of the solar array current direction [19]. \(K_f\) and \(K_r\) are feedforward terms,

\[
K_f = -\frac{D R_i}{L f_s} \left(1 - \frac{D}{2}\right), \tag{3.2}
\]

\[
K_r = \frac{R_i}{2L f_s}. \tag{3.3}
\]

\(R_i\) is the current sensing gain, and \(F_m\) is the modulation gain,

\[
F_m = \frac{f_s}{S_n + S_e}. \tag{3.4}
\]

\(S_n\) is the external ramp. \(S_n\) is the inductor current ramp. \(H_e(s)\) contains a double complex pole pair at one-half of the switching frequency due to the sampling nature of the current loop [22], and it can be properly damped by adding an external ramp,

\[
H_e(s) = 1 + \frac{s}{\omega_n Q_p} + \frac{s^2}{\omega_n^2}, \tag{3.5}
\]

\[
\omega_n^2 = \frac{\pi}{T_s}, \tag{3.6}
\]

\[
Q_p = \frac{1}{\pi(m_e D - 0.5)}, \tag{3.7}
\]

3. SMALL-SIGNAL ANALYSIS OF THE FLYWHEEL ENERGY STORAGE SYSTEM
\[ m_c = 1 + \frac{S_Z}{S_n}. \] (3.8)

In designing the bus voltage control amplifier, it is important to determine the control-to-output transfer function. Due to feeding back the inductor current and summing the current sense signal with an external ramp voltage, the control-to-output transfer function does not contain a moving double pole pair, which can significantly reduce control bandwidth without proper damping. This is the benefit of current injection control (CIC).

The control-to-output transfer function with current loop closed can be approximated as [19]:

\[ \frac{v_{bus}}{v_c} = G_x \frac{(1 - \frac{s}{\omega_x})(1 + sR_cC)}{(1 + \frac{s}{\omega_x})He(s)}, \] (3.9)

where

\[ G_x = \frac{F_m \frac{V_{bus}}{D}}{1 + \frac{1}{D^2} + \frac{I_k}{D} F_m R_i + K_f F_m \frac{V_{bus}}{D}}, \] (3.10)

\[ \omega_x \approx \frac{V_{EMF}}{I_k L}, \] (3.11)

\[ \omega_v \approx \frac{1}{V_{bus} C}. \] (3.12)

The term \( I_k \) is given by:
\[ I_k = \frac{V_{\text{bus}}}{r_{\text{eq}}} + (I_o - I_{sa}) \equiv I_o - I_{sa}, \]  

where \( I_o \) is the spacecraft load current, and \( I_{sa} \) is the spacecraft solar array current, as shown in Figure 2.2.

When the load current is larger than the solar array current, \( I_k \) is positive, and the flywheel will be discharged to supply the load. Under these circumstances, the zero, \( \omega_y \), is in the right half-plane (RHP). The system bandwidth is limited to about one decade below the RHP zero. The frequency of the RHP zero is indirectly proportional to the motor/generator armature inductance. A typical boost converter in a spacecraft battery charger/discharger has a bandwidth of about 1 kHz. Hence, the RHP zero must be greater than 10 kHz under the worst case condition which is when the battery voltage is at minimum and the load is at a maximum. In the flywheel system, this case corresponds to the minimum motor/generator speed when the system must still supply the maximum spacecraft load. Assume that the lowest flywheel back EMF voltage is 50 V, the spacecraft bus voltage is 120 V, and the peak load power delivered by the flywheel is 500 W. To achieve an RHP zero with frequency greater than 10 kHz under these conditions, the motor/generator armature inductance should be below about 80\( \mu \)H. Thus, in order to have the flywheel charge/discharge system that has characteristics comparable to those of a conventional spacecraft battery discharger, the motor/generator armature inductance per phase should be less than 40\( \mu \)H.

### 3.2 Small-Signal Analysis of the Flywheel System in the Charge Mode
When the load current is smaller than the solar array current, $I_k$ is negative, and the flywheel will be charged while regulating the spacecraft bus. In the charge mode, the bi-directional converter appears as a buck converter. The zero, $\omega_y$, is shifted into the left half-plane (LHP). The pole, $\omega_x$, comes from splitting of the complex double pole pair due to the current mode control. From Equation 3.12, it is seen that this pole is shifted into the right half-plane in the charge mode and occurs at a very low frequency for practical designs. Then the system becomes unstable under open-loop condition, and conditionally stable when the feedback loop is closed. The compensator design for a system with a RHP pole is usually based on Nyquist stability criterion, and Bode diagram can not be directly applied.

For a bi-directional charger/discharger, the RHP zero in the discharge mode is normally the most important factor to limit the overall system bandwidth, a compensator design based on the worst case condition in the discharge mode can normally guarantee system stability in the charge mode [19,20]. In other words, the compensator design can start from the worst case condition in the discharge mode, then, Nyquist criterion is employed to check the system stability in the charge mode.

3.3 **Compensator Design**

Based on the power stage design which will be discussed in Chapter 4, the motor/generator armature inductance is 30$\mu$H, the bus capacitor is 470$\mu$F with 50m$\Omega$ ESR, the switching frequency is 100 kHz, the bus voltage is regulated at 120 V, and the
back EMF of motor/generator is in the range from 50 V to 100 V. The worst case condition is 50 V back EMF with 500-W charge/discharge power, and the following discussion will be based on this worst case condition. The DC bus is connected to load converters, which behave like a negative resistance up to the crossover frequency of its voltage loop gain. To simplify the analysis, the input impedance of a load converter is represented as

$$ R_{AC} = -\frac{V^2}{P} , $$

(3.14)

ignoring the dynamics of its input filter. For 120-V DC bus, the 500-W load impedance is $-28.8\Omega$, and inductor current is 10 A.

The inductor current is sensed by using Hall devices, and the sensing gain is

$$ R_i = 0.2V/A . $$

(3.20)

The external ramp is selected as $117\times10^3$ V/s, and this results in a worst case $Q_p$ of 0.85. The modulation gain $F_m$ is 0.286.

Figure 3.2 shows the control-to-output transfer function of the bi-directional converter with above circuit parameters in the discharge mode. From Equation 3.11, the RHP zero moves beyond 10 kHz with 60μH inductance. A compensator is needed to boost the DC gain and crossover frequency.

The compensation network is shown in Figure 3.3. A pole at DC is used to obtain tight regulation, and the gain of the integrator is selected in order to cross over 0 dB with acceptable gain and phase margins. A second pole is placed at the location of the RHP...
Fig. 3.2. Control-to-Output Transfer Function of a Bi-directional Converter in the Discharge Mode
Fig. 3.3. Compensation Network
zero or ESR zero, whatever comes first, in order to attenuate the noise. The compensator zero is placed at low frequency in order to boost the phase. The transfer function of the compensator is

$$F_c(s) = \frac{30(1 + \frac{s}{13.8})}{s(1 + \frac{s}{11700})}, \quad (3.21)$$

and the voltage loop gain is shown in Figure 3.4. The crossover is about 1.3 kHz, and has about 40° phase and 22-dB gain margins.

Figure 3.5 shows the control-to-output transfer function of the boost converter in the charge mode. There is a RHP pole located around 20 Hz. The system becomes open-loop unstable. Using the same compensator shown in Equation 3.21, the polar plot of the voltage loop gain in the charge mode is shown in Figure 3.6, and counterclockwise encircles (-1,0) once, therefore, the Nyquist stability criterion is met, and the system is stable.

In practice, it is convenient to use the loop gain in analyzing the stability of the closed-loop system, and the Nyquist criterion can be translated as “the crossover of the loop gain must occur with a phase larger than −180°” for Bode diagram analysis. The difference between the phase at the crossover and −180° is the phase margin [25]. Figure 3.7 is the Bode diagram of the loop gain, and shows a phase margin of 65°.
Fig. 3.4. Loop Gain of the Voltage Control Loop of a Bi-directional Converter in the Discharge Mode
Fig. 3.5. Control-to-Output Transfer Function of a Bi-directional Converter in the Charge Mode
Fig. 3.6. Polar Plot of the Voltage Loop Gain in the Charge Mode
Fig. 3.7. Loop Gain of the Voltage Control Loop of a Bi-directional Converter in the Charge Mode
Therefore, the compensator design is complete, good stability margins for both the discharge and the charge modes have been achieved, and the bus regulation loop bandwidth requirement has been met.

3.4 Physical Explanation of the Small-signal Behavior of the Bi-directional Converter

In the above discussion, a RHP pole has been found in the control-to-output transfer function of the bi-directional converter with current loop closed in the charge mode, then the system becomes open-loop unstable. The open-loop unstable system is unusual in the real world, the peculiar behavior of a bi-directional converter needs to be explained physically in order to provide a design insight of the flywheel energy storage system.

According to [30], a constant current source and a constant power load connected in parallel will result in an inherently unstable system. As shown in Figure 3.8, initially, the system is at an equilibrium point, no net current flows into or out of the capacitor, and the constant current source provides the exact amount of power which is needed by the constant power load. For some reason, the DC Voltage \( V_c \) is increased, the input current of the constant power load will decrease, based on the V-I characteristics of the constant power load, then there will be a net current \( I_c=I_s-I_L \) flowing into the capacitor. Therefore, the capacitor voltage will increase, a positive feedback is formed, and the system will move away from the equilibrium point, and become unstable.
Figure 3.8. An Unstable System with a Current Source and a Constant Power Load in Parallel
The solar array is usually represented as a constant current source, and the active load which is connected to the DC bus is normally a constant power load, which exhibits negative input impedance at low frequencies. By placing a solar array and a active load in parallel, it is not a surprise that the system would become open-loop unstable, based on the discussion above.

In order to facilitate the analysis, the bi-directional converter with current-injection control as shown in Figure 2.5 is redrawn in Figure 3.9.

3.4.1 Stability Analysis of a Bi-directional Converter in the Charge Mode

When the system is operating in the charge mode with the current loop closed, the flywheel is charged with a constant inductor current. As discussed in Chapter 2, the flywheel can be practically considered to have a constant rotational speed, and the back EMF of the motor/generator has a constant magnitude in the control response range of our interest, therefore, the inverter/rectifier together with the motor/generator can be represented as a constant power load, which is also connected to the DC bus.

In Figure 3.10, initially the system is at an equilibrium point, the current source provides the exact amount of power which is needed by both the flywheel and the load, and no net current flows into or out of the capacitor. For some reason, the DC Voltage Vc is increased, the input currents of the constant power load and the inverter/rectifier, i.e. IL and Im, will decrease, because of the V-I characteristics of a constant power load. Then there will be a net current \( I_c = I_s - IL - Im \) flowing into the capacitor. Therefore, the capacitor voltage will increase, a positive feedback is formed, and the system will move away from
Figure 3.9. Current-Mode Control of the Bi-directional DC/DC Converter
Unstable System

Figure 3.10. An Unstable System When the Flywheel System Operates in the Charge Mode
the equilibrium point, and become unstable. This explains why there is RHP pole in the open-loop transfer function, as shown in Figure 3.5.

3.4.2 Stability Analysis of a Bi-directional Converter in the Discharge Mode

When the system is operating in the discharge mode with the current loop closed, the flywheel is discharged with a constant inductor current, therefore, the inverter/rectifier together with the motor/generator can be represented as a constant power source.

In Figure 3.11, initially the system is at an equilibrium point, the current source and the motor/generator provide the exact amount of power which is needed by the load, and no net current flows into or out of the capacitor. For some reason, the DC Voltage $V_c$ is increased, the input current of the constant power load, i.e. $I_L$, will decrease, and the output current of the constant power source, i.e. $I_g$, will decrease as well.

The net current flowing into the capacitor is

$$I_c = I_s + I_g - I_L = \Delta I_g - \Delta I_L,$$

(3.22)

where $\Delta I_g$ and $\Delta I_L$ are current variations of $I_g$ and $I_L$. If $I_c$ is positive, the system will become unstable; if $I_c$ is zero or negative, the system will be stable. The polarity of $I_c$ is a function of the load input impedance and the inverter/rectifier output impedance, both of which are negative.

If the solar array provides no current, the inverter/rectifier will provide exactly the same amount of power as what the load needs. Then from Equation 3.14, the load will
Figure 3.11. An Conditionally Stable System When the Flywheel System Operates in the Discharge Mode
have an input impedance which is the same as the inverter/rectifier output impedance. Therefore, for the same voltage perturbation $\Delta V_c$, the $I_g$ will have the same variation as $I_L$, i.e. $\Delta I_g = \Delta I_L$, therefore, from Equation 3.22, $I_c$ will be zero. No positive feedback is formed, the system will be stable. This explain why the system is open-loop stable in the discharge mode, as shown in Figure 3.2.

If the solar array still provides power the DC bus, i.e. $I_s$ is positive, the inverter/rectifier will provide less power than what the load needs, then from Equation 3.14, the load will have an input impedance whose absolute value is less than that of the inverter/rectifier output impedance. Therefore, for the same voltage perturbation $\Delta V_c$, the $I_g$ will have less current variation than $I_L$, i.e. $\Delta I_g > \Delta I_L$. From Equation 3.22, $I_c$ will be positive. The capacitor voltage will increase, a positive feedback is formed, and the system will move away from the equilibrium point, and become unstable. Figure 3.12 shows a control-to-output transfer function of a bi-directional converter in the discharge mode, and the back EMF is 50 V, load power is 500 W, but the solar array still provides about 0.83-A current, and the flywheel is discharged with 400-W power. Figure 3.12 shows that the system has a RHP pole, and open-loop unstable, exactly as predicted.

To provide more complete analysis, let's consider another case, which is not applicable in the real system. If the solar array absorbs power from the DC bus, i.e. $I_s$ is negative, the inverter/rectifier will provide more power than what the load needs, then from Equation 3.14, the load will have an input impedance whose absolute value is greater than that of the inverter/rectifier output impedance. Therefore, for the same
Fig. 3.12. Control-to-Output Transfer Function of a Bidirectional Converter in the Discharge Mode with 0.83-A Solar Array Current
voltage perturbation $\Delta V_c$, the $I_g$ will have more current variation than $I_L$, i.e. $\Delta I_g < \Delta I_L$.

From Equation 3.22, $I_c$ will be negative. The capacitor voltage will decrease, a negative feedback is formed, and the system will be stable.

### 3.4.3 DC Gain Analysis of a Bi-directional Converter

From Figure 3.2, Figure 3.5 and Figure 3.12, the DC gain of the control-to-output transfer function of a bi-directional converter with current loop closed have different polarities, i.e. positive or negative, and this peculiar behavior can be explained in Figure 3.13.

In the discharge mode I, as shown in Figure 3.13, the solar array provides no zero, and the output power of the inverter/rectifier is exactly the same as the input power of the load. From Figure 3.9, if $V_e$ is increased, the inductor current will increase, then the output power of the inverter/rectifier increases. Since the load is a constant power load, the extra power from the inverter/rectifier will be used to charge the capacitor, then the DC bus voltage $V_c$ increases, this results in a positive DC gain of the system, as shown in the Figure 3.2.

In the discharge mode II, as shown in Figure 3.13, the solar array provides some current. From Figure 3.9, if $V_e$ is increased, the inductor current will increase, then the output power of the inverter/rectifier increases, and the output current of the inverter/rectifier increases, therefore, the input current of the load increases as well. Since the load is a constant power load, the DC bus voltage $V_c$ will decrease, this results in a negative DC gain of the system, as shown in the Figure 3.12. Opposite to case I, the extra
Fig. 3.13. DC Gain Analysis of the Bidirectional Converter in the Charge Mode and the Discharge Mode
power from the inverter/rectifier is offset by the reduced output power from the solar array.

In the charge mode, as shown in Figure 3.13, the solar array provides current to both the load and the inverter/rectifier. From Figure 3.9, if Ve is increased, the inductor current will decrease in value, i.e. increase in the negative direction, then the input power of the inverter/rectifier decreases, and the input current of the inverter/rectifier decreases, therefore, the input current of the load will increase. Since the load is a constant power load, the DC bus voltage Vc will decrease, this results in a negative DC gain of the system, as shown in the Figure 3.5.

3.5 Summary

When the brushless DC motor/generator control scheme is used, the small-signal model of the flywheel energy storage system is the same as that of the bi-directional dc/dc converter, if transitions between 30° segments is fast and negligible.

In the discharge mode, the small-signal model of the flywheel system is the same as that of a standard dc/dc boost converter, where the location of the RHP zero determines the control bandwidth. Small armature inductance is required for wide control bandwidth.

Current injection control (CIC) is used to split a moving double pole pair, which can significantly reduce control bandwidth without proper damping.
In the charge mode, the small-signal model of the flywheel system is the same as that of a boost converter with negative current flow. The RHP zero in the discharge mode moves into the LHP, and it helps in designing a more stable system. However, an RHP pole appears at low frequency, and Nyquist stability criterion is resorted for stability analysis.

A stable compensator design in the discharge mode will normally guarantee the stability in the charge mode. In other words, the compensator design should start from the worst case condition in the discharge mode, then, apply Nyquist criterion to check the system stability in the charge mode with the same compensation network.

A physical explanation of some peculiar behaviors of the bi-directional converter is provided.
4. DEVELOPMENT OF A THREE-PHASE SOFT-SWITCHED INVERTER/RECTIFIER

As has been discussed in the previous chapter, in a spacecraft flywheel energy storage application, minimizing the motor/generator armature inductance is necessary for good dynamic performance, therefore, a three-phase inverter/rectifier with high switching frequency is required to reduce current ripple and to synthesize the high-frequency motor/generator current waveforms. A conventional hard-switched inverter/rectifier with high switching frequency produces a large amount of switching loss and EMI radiation. An innovative soft-switching technique is introduced to solve these problems.

4.1 Topology Study of Three-Phase Inverter/Rectifier

4.1.1 Hard-Switched Inverter/Rectifier

As will be discussed in Power Stage Design section, 100-kHz switching frequency is considered a good design in terms of control loop bandwidth, efficiency and intention to demonstrate the capabilities of the soft-switching technique in high frequency applications.

The 500-W power level, 100-kHz switching frequency and 120-VDC bus voltage demand the use of MOSFET switching devices. A conventional three-phase voltage source inverter/rectifier, as shown in Figure 2.2, is not workable in this application due to the extremely slow reverse recovery of the MOSFET antiparallel diode.
One method of solving this problem is to replace the slow MOSFET body diode with an external fast diode, as shown in Figure 4.1. The Schottky diode is used to block the MOSFET body diode.

4.1.2 ZVT Soft-Switched Inverter/Rectifier

There is another way of solving the problem. The slow body diode can still be used if the soft-switching scheme is adopted to help its turn-off. Figure 4.2 shows a zero-voltage-transition (ZVT) soft-switched three-phase inverter/rectifier [18], which is identical to the conventional circuit, except for the addition of a commutation circuit.

One soft-switching scenario is drawn in Figure 4.3. Initially, rectifier diodes D1, D4, and D6 are carrying current; the sudden turning on S2, S3 or S5 will cause a severe diode reverse recovery problem. In order to avoid this problem, two switches in the commutation circuit, Sx1 and Sx2, are turned on just prior to the turn-on of any of these main bridge switches. Then the current in the main bridge diodes will be transferred to the auxiliary inductor, Lx. Once the diode current has dropped to zero, the main bridge switch can be turned on without creating the reverse recovery problem in the diode. When the main bridge switch is turned on, the auxiliary switches are turned off. This allows the inductor current in Lx to be discharged into the source voltage, Vs.

4.1.3 Comparison between the Hard-Switching and the Soft-Switching Schemes

Both the hard-switching scheme and the soft-switching scheme are evaluated experimentally under DC/DC conditions. Figure 4.4. shows how the three-phase
Figure 4.1. A Method of Solving the Reverse Recovery Problem of the MOSFET Body Diode.
Figure 4.2. The Inverter/Rectifier with a Soft-Switching Auxiliary Circuit.
Figure 4.3. Soft Turn-off of the Rectifier Diodes.
Figure 4.4. Three-Phase Inverter/Rectifier Tested under DC/DC Conditions. Switches on the same leg turn on and off complementarily.
inverter/rectifier is tested under DC/DC conditions. The auxiliary circuit or Schottky diodes are not drawn in Figure 4.4 for simplicity.

In the test circuit, MOSFET devices, which are also been used as the ZVT auxiliary switches, are IXYS 200-V IXFH42N20; Schottky rectifiers are International Rectifier 15-V 19TQ015; Ultra fast recovery rectifier diodes are Philips 200-V 30-ns BYV79-200. The conduction loss characteristics of these devices are measured in room temperature and shown in Figure 4.5 for reference.

The efficiency comparison between the hard-switching and the soft-switching schemes in the rectifier mode is shown in Figure 4.6, and the soft-switching scheme is significantly better than the hard-switching scheme in terms of efficiency. Compared with ZVT circuit, hard-switching circuit has additional conduction loss of Schottky diodes and ultra fast recovery rectifier diodes which have larger voltage drop than MOSFET body diodes.

Since switches on the same leg turn on and off complementarily for bi-directional power flow, based on the conduction loss characteristics of Figure 4.5, the MOSFET has lower voltage drop than its body diode in the operation ranges of interest; therefore, in ZVT circuit, the MOSFET also serves as a synchronous rectifier, and provides low impedance path for current in the operation ranges of interest. This is another important advantage of the ZVT circuit over the hard-switching circuit, where Schottky diodes prevent the MOSFET working as a synchronous rectifier.
Figure 4.5. Conduction Loss of Various Devices.
Measurement is taken under room temperature.
Figure 4.6. Efficiency Comparison in the Rectifier Mode.
Measurement is taken under DC/DC conditions. Output power is 500 W, and switching frequency is 100 kHz.
The efficiency comparison between the hard-switching and the soft-switching schemes in the inverter mode is shown in Figure 4.7, and the hard-switching scheme is better than the soft-switching scheme in terms of efficiency. In the inverter mode, the extra turn-off switching action of S4 required by the ZVT scheme together with the core loss of saturable cores, which will be discussed later, probably contributes to the low efficiency of the soft-switching scheme.

Comparing Figure 4.6 and Figure 4.7, the ZVT circuit has similar efficiency in both the inverter mode and the rectifier mode, since bi-directional operation makes the MOSFET conducting current most of the time. The hard-switching circuit has a significant efficiency variation from the inverter mode to the rectifier mode. Circuit thermal design is usually based on the worst case condition, the worst case of the hard-switching circuit has much lower efficiency than the ZVT circuit. Therefore, the hard-switching circuit requires more size/weight for heatsink, and becomes less desirable for spacecraft applications.

Figure 4.8 shows the voltage waveform across switch S2 when diode D2 is hard turned off with 10-A current in the inverter mode. The time scale is 100 ns/div, the voltage scale is 50 V/div, the digital scope is set with 100-MHz bandwidth. The voltage ringing causes severe EMI radiation, which interferes the control circuit during the test. Furthermore, the voltage ringing peak exceeds 200-V rating of the MOSFET device, and makes the operation unsafe. A RC snubber can damp the voltage ringing. The 1XFH42N20 MOSFET output capacitance is about 1 nF, then a RC snubber needs about
Figure 4.7. Efficiency Comparison in the Inverter Mode.
Measurement is taken under DC/DC conditions. Output power is 500 W, and switching frequency is 100 kHz.
Figure 4.8. Voltage Ringing due to the Hard-Switching.

Diode D2, shown in Figure 4.4, is hard turned off with 10A current in the inverter mode. Scope is set with 100-MHz bandwidth. Scales: 50 V/div, 100 ns/div.
5-nF capacitor for each MOSFET device in the inverter/rectifier. The loss dissipated in RC snubbers is about 7.2 W under DC/DC conditions with 100-kHz switching frequency, and will further reduce the efficiency of the hard-switching scheme in Figure 4.6 and Figure 4.7 by about 1.4%; then, the soft-switching circuit will have better efficiency than the hard-switching scheme in both the inverter mode and the rectifier mode. Later discussion will show that the voltage waveform in the soft-switching circuit is much cleaner.

The soft-switching scheme is considered better than the hard-switching scheme in terms of efficiency, thermal dissipation and EMI radiation, and will be employed in this application.

4.2 Modified Soft-switching Scheme in the Flywheel Energy Storage System

The ZVT three-phase inverter/rectifier is designed for the type of control using space vector modulation, and there are many operation modes in which different switching actions are needed to realize the zero-voltage transition [18]. In this flywheel system, the operation of the inverter/rectifier is essentially that of a bi-directional converter; therefore, the original soft-switching scheme can be simplified. The soft-switching scenario drawn in Figure 4.3 is the operation mode which needs to be used in the flywheel system.
The operation in the inverter mode is slightly different from that in the rectifier mode, since extra switching action is needed to create the soft-switching scenario shown in Figure 4.3 in the inverter mode.

4.2.1 Soft-switching in the Rectifier Mode

Figure 4.9 shows the circuit operation in the rectifier mode. Assume that phase A has the highest back EMF voltage, and phase B has the lowest back EMF voltage. The flywheel is being discharged through pulse-width modulation of S2. When S2 is on, the current in the armature inductance of phase A and phase B ramps up by flowing through S2 and D4. When S2 turns off, the current in the armature inductance flows through D1, into the source voltage Vs and back through D4. Before S2 turns on again, switches Sx1 and Sx2 are turned on. Since voltage across nodes A and B is equal to the source voltage, the armature inductance current will start to flow through the auxiliary inductor Lx. The inductance of Lx is much smaller than the armature inductance. Once the current through Lx has reached the armature inductor peak current, there is no more current in D1, and it will naturally turn off. There will be a small resonant transition between Lx and the switch capacitance of S1, S2, S3 and S4. This will reduce the voltage across S2 to half of the DC bus voltage Vs. Then S2 can be turned on without creating the reverse recovery problem in the diode D1.

The ZVT soft turn-off of the rectifier diode, i.e. D1, can be achieved. But the turn-on of the PWM switch S2 occurs under the reduced voltage condition, instead of under the zero-voltage condition. Therefore, the ZVT soft-switching is only partially achieved
Figure 4.9. Soft-Switching in the Rectifier Mode.
In order to turn off the body diode of S1, auxiliary switches Sx1 and Sx2 turn on prior to the turn-on of switch S2.
for these bridge switches. Since the diode reverse recovery problem is the major problem in this MOSFET based circuit, partial ZVT for bridge switches should be acceptable due to the relatively low DC bus voltage.

The experimental photograph is shown in Figure 4.10, and was taken when the circuit was operating in the rectifier mode. The upper waveform is the voltage across S2, 50 V/div.; the bottom waveform is the current flowing in ZVT inductor Lx, 5 A/div. The DC bus voltage is 120 V, input inductor current is 5.38 A, time scale is 2μs/div. Switch S2 turns on when there is a voltage applied across D1, i.e. no current flowing through D1.

4.2.2 Soft-switching in the Inverter Mode

Figure 4.11 shows the circuit operation in the inverter mode. Again, the assumption is that phase A has the highest back EMF voltage, and phase B has the lowest back EMF voltage. The flywheel is being charged through pulse-width modulation of S1. When S1 is on, the current in the armature inductance of phase A and phase B ramps up by flowing from Vs through S1 and then S4 and back to the source. When S1 turns off, the current in the armature inductance freewheels through D2 and S4. Before S1 turns on again, switches Sx1 and Sx2 are turned on. However, since the voltage across nodes A and B is equal to zero, no current will flow through Lx unless S4 is turned off. When S4 turns off, phase B current flows through D3; then the soft-switching scenario shown in Figure 4.3 is created, and the armature inductance current starts to flow through Lx. Once the current through Lx has reached the armature inductor peak current, there is no more current in D2, and it will naturally turn off. There will be a small resonant transition
Figure 4.10. **Experimental Waveforms of the Rectifier Mode Operation.**

The upper waveform is the voltage across S2, 50 V/div; the bottom waveform is the current flowing in ZVT inductor Lx, 5 A/div. Time scale: 2 us/div.
Figure 4.11. Soft-Switching in the Inverter Mode.
In order to turn off the body diode D2, S4 turns off to build voltage across the commutation network, then auxiliary switches Sx1 and Sx2 turn on prior to the turn-on of switches S1 and S4.
between Lx and the switch capacitance of S1, S2, S3 and S4. This will reduce the voltages across both S1 and S4 to half of the DC bus voltage Vs. Switches S1 and S4 can be turned on simultaneously without creating the diode reverse recovery problems.

Again, the ZVT soft turn-off of the rectifier diode, i.e. D2, is achieved. But the ZVT soft-switching is only partially achieved for bridge switches S1 and S4. There is an extra switching action, i.e. the turning off of S4, which is not required in the PWM hard-switching scheme. This extra turn-off is terrible for a minority carrier device like IGBT, since the current tail of these devices can generate a significant amount of turn-off loss, which makes efficiency suffer. Even for a majority carrier device like MOSFET, which has much less turn-off loss, this extra hard turn-off action is not desirable, and is an important factor to the lower efficiency of the soft-switching scheme with respect to the hard-switching scheme in the inverter mode shown in Figure 4.7.

From Figure 4.11, it can be seen that there is an abrupt voltage change across the ZVT commutation diode bridge when switches S1 and S4 turn on. Then there will be reverse recovery problems for these commutation diodes. Even though they are selected to be ultra fast diodes, the hard turn-off may still cause a lot of problems. In order to solve these problems, several saturable reactors are inserted into the ZVT commutation circuit, as shown in Figure 4.12. When reverse voltages are applied abruptly to the diodes, these saturable reactors will block the applied voltage until the storage charge in the diodes is completely recombined. During each 30° segment, these saturable cores can be designed with a small amount of flux sweep; therefore, the core loss should not be significant.
Figure 4.12. Saturable Reactors Used to Avoid Diode Reverse Recovery Problems in the ZVT Commutation Circuit
Since the input inductor current changes signs at the electrical frequency of the motor/generator, these saturable cores will have a full flux sweep from positive saturation to negative saturation, and vice versa, at the electrical frequency of the motor/generator. Full flux sweep can cause a much larger amount of core loss per cycle; however, the overall core loss should be still not significant, since the electrical frequency of the motor/generator is no more than 3 kHz.

The experimental photograph, shown in Figure 4.13, was taken when the circuit was operating in the inverter mode. The upper waveform is the current flowing in ZVT inductor Lx, 5 A/div; the middle waveform is the voltage across S4, 100 V/div; the bottom waveform is the voltage across S2, 100 V/div. Time scale is 2μs/div. Switches S1 and S4 turn on at the same time when there is voltage applied across diodes D2 and D3, i.e. no current flowing through these rectifier diodes.

The above discussion divides the ZVT operation into the inverter mode and the rectifier mode. In the bi-directional operation, it is sometimes difficult to tell which mode the circuit is running in, especially with mostly analog control. Therefore, it is necessary to reduce the multi-mode soft-switching operations to one-mode operation.

4.2.3 Practical Implementation of the Soft-switching Scheme

Before turn-on of any switch, all the six switches should be turned off first, thus forcing the inductor current to flow through rectifier diodes, and creating the general scenario shown in Figure 4.3. When the ZVT commutation network is activated, the
Figure 4.13. Experimental Waveforms of the Inverter Mode Operation.

The upper waveform is the current flowing in ZVT inductor Lx, 5 A/div; the middle waveform is the voltage across S4, 100 V/div; the bottom waveform is the voltage across S2, 100 V/div. Time scale: 2 us/div.
bridge switches should not be turned on until the inverter/rectifier reaches the equilibrium point, i.e. until nodes A, B and C have the same voltage potential.

The soft-switching operation of the converter will be still divided into the rectifier mode and the inverter mode, but in an automatic fashion.

As discussed before, both switches in one leg of the inverter/rectifier need to turn on and off complementarily for the bi-directional operation, as a result, there are some extra switching actions, such as turning on or off the switch whose antiparallel diode is conducting current. The extra switching action should not have any impact on the circuit efficiency, since it is taken under zero-voltage conditions. However, based on the simplified soft-switching scheme, the ZVT commutation circuit will be activated whenever a switch needs to be turned on. Therefore, extra switching actions will result in extra commutation actions, which require turn-off of all switches and current circulation in the ZVT auxiliary circuit. The extra commutation action can significantly affect circuit efficiency. In order to eliminate the extra commutation action, the current direction of the boost inductor needs to be detected. As shown in Figure 4.14, when inductor current flows into the switch leg, only the turn-on of switch S2 requires ZVT commutation; when inductor current flows out of the switch leg, only the turn-on of S1 requires ZVT commutation. If ZVT commutation is based on the current direction, all the extra commutation actions can be avoided.

4.2.4 Natural Zero-Voltage Soft-switching Mode
Figure 4.14. Commutation Conditions.
When the inductor current flows into the switch leg, turn-on of switch S2 requires the ZVT commutation; when the inductor current flows out of the switch leg, turn-on of S1 requires the ZVT commutation.
When the charging or discharging power is small, the average inductor current is also small. The inductor current can actually change sign during one switching cycle due to the current ripple. Then, natural zero-voltage soft-switching (ZVS) can be realized for both diodes and switches [15], and no commutation action will be taken from the ZVT auxiliary network, since all the extra commutation actions can be avoided by sensing inductor current direction. Figure 4.15 illustrates the described operation. Both switches S1 and S2 are pulse-width modulated complementarily with a small dead time. When switch S2 is on, the inductor current ramps up; when S2 is turned off, the inductor current flows through D1; therefore, S1 turns on under the ZVS condition. Then, the inductor current ramps down to negative region; when S2 turns off, the inductor current flows through D2; therefore, S2 turns on under the ZVS condition. Because the average inductor current is still positive, the system operates in the discharge mode.

The experimental photograph, shown in Figure 4.16, was taken when the inductor current was changing sign during one switching cycle. The upper waveform is the phase A inductor current, 2 A/div.; the bottom one is the voltage across switch S2, 50 V/div.. Time scale is 2μs/div.. The natural zero-voltage soft-switching is realized for both switches and diodes. As a result of the soft-switching, the waveforms are very clean.

One design consideration is to expand the natural ZVS mode to the whole operation range, and greatly simplify the circuit operation. Figure 4.17 shows the efficiency estimation of the inverter/rectifier operating in the natural ZVS mode. The charge/discharge power is 500 W, the inductor current ripple is selected as 20 A for the
Figure 4.15. Natural Zero-Voltage Soft-Switching.

Natural zero-voltage soft-switching is achieved when the inductor current changes sign during one switching cycle. No ZVT commutation action is needed.
Figure 4.16. Experimental waveforms of Natural ZVS Soft-Switching.
The upper waveform is the phase A inductor current; 2 A/div.; the bottom one is the voltage across switch S2, 50 V/div. Time scale: 2 us/div.
Figure 4.17. Efficiency Estimation of the Natural ZVS Mode Operation.

Charge/discharge power is 500 W, and MOSFET on-resistance is 0.04 ohms.
worst case condition, i.e. 50-V back EMF, and MOSFET on-resistance is $40\,\text{m}\Omega$. Figure 4.17 shows that the full-range natural ZVS operation can achieve very good efficiency, and has a excellent potential for the 500-W power level application.

There are several concerns about the full-range natural ZVS operation. For the 100-kHz switching frequency, the motor/generator armature inductance has to be lower than $73\,\mu\text{H}$ per phase; for the 50-kHz switching frequency, the motor/generator armature inductance has to be lower than $14.5\,\mu\text{H}$ per phase. Later discussion in section 4.3.3 shows that the realistic value for the motor/generator armature inductance is about $30\,\mu\text{H}$ per phase, then the 500-W power level will require the switching frequency below 24 kHz. One concern is that 24-kHz switching frequency may not be sufficiently high for a wide control loop bandwidth and the minimization of the DC bus filter size/weight, another concern is that large high frequency current ripple may induce a significant amount of eddy current loss inside the motor/generator, whose efficiency has already suffered from the cycling magnetic field of the 3-kHz electrical frequency with 6-mil lamination structure [7]. Further investigation is needed regarding the feasibility of this design option in this application. The following discussion about power stage design will only analyze the operation range of the natural ZVS mode.

4.3 Power Stage Design

4.3.1 Selection of Switching Frequency
As discussed earlier, the system control loop must have at least 1-kHz bandwidth. With current-injection control, the current loop should be well separated from voltage loop, and has a crossover at least above 6-kHz. However, in this application, where the frequency of the motor/generator current can be as high as 3 kHz, much higher current-loop bandwidth is required to synthesize the high-frequency quasi-square current waveforms. The current loop crossover should be at least ten times larger than 3 kHz. Therefore, the 100-kHz switching frequency is probably the minimal requirement for the 30-kHz current loop. For the 30-kHz current loop, the voltage loop seems to be able to have a much higher bandwidth than 1 kHz. However, the determining factor for the voltage loop is the RHP zero in the worst case of the discharge mode, and the armature inductance determines the locations of the RHP zero. Therefore, the voltage loop can not be arbitrarily increased by increasing the switching frequency alone.

There is another strong motivation to increase the switching frequency in order to reduce the DC bus filter size and weight [20]. The innovative ZVT soft switched inverter/rectifier is capable of maintaining high efficiency at high switching frequency; 100 kHz is selected in this design, and is intended to demonstrate the capabilities of soft-switching techniques.

4.3.2 Selection of MOSFET Devices

The ZVT scheme can softly turn off a diode, but does not eliminate the reverse recovery current. The longer the transition time, the smaller the reverse recovery current, but the circulation current loss in the commutation circuit will increase, and effective duty
cycle, especially with high switching frequency, will be affected by the commutation process. For a faster diode with less reverse recovery time, less reverse current is produced to flow into the commutation circuit; therefore, shorter transition time is needed, and less circulation loss will be generated in the auxiliary circuit. Even though the MOSFET body diode can be used as a rectifier with the help of the ZVT commutation circuit, it is still desirable to select a MOSFET with a faster body diode for high efficiency and less duty cycle loss.

The MOSFET IXFH42N20 has lower on-resistance, 0.06Ω, and faster diode, less than 200-ns reverse recovery time, than IRFP250, which has 0.085Ω on-resistance and a body diode with 360-ns reverse recovery time. IXFH42N20 is selected in this design, and used as both bridge switches and auxiliary switches.

4.3.3 Inductor Design

The armature inductance of flywheel motor/generator is simulated with three external inductors. As discussed in Chapter 3, the motor/generator armature inductance per phase should be less than 40μH in order to design the system with 1-kHz control loop bandwidth. The prototype motor/generator in University of Maryland has 52μH armature inductance, and 170-V peak back EMF voltage, which is higher than 120-V bus voltage [24]. Modifications of flywheel motor/generator with fewer winding turns and laminated structure is under way, 30μH armature inductance seems more realistic.

In the 500-W flywheel system, maximum inductor average current is 10 A at 50-V back EMF low line, and can surge to 20 A at 1-kW peak transient power. TDK ferrite
core PC40EER42-Z is used to construct 30\(\mu\)H inductors, and is wound with 19 turns of Litz65/31 Litz wire. Air gap per leg is about 2.8 mm, and results in about 30\(\mu\)H inductance and 2500 gauss at 30A peak current.

The current ripple of inverter/rectifier is calculated as

\[
\Delta i = \frac{V_{EMF}T_s(120 - V_{EMF})}{120L}
\]  

(4.1)

Where \(V_{EMF}\) is back EMF voltage, \(T_s\) is switching period, \(L\) is 60\(\mu\)H. The largest current ripple is 5A at 60V back EMF. For average current less than 2.5A and charge/discharge power less than 150W, inductor current will change sign during one switching cycle, then, the inverter/rectifier will operate in the natural ZVS mode. When the average current is larger than 2.5A, the inverter/rectifier will operate in the ZVT soft-switching mode, ZVT commutation is needed to help the turn-off of MOSFET body diodes.

The selection of the DC bus capacitor is mainly based on the hold-on time and voltage ripple requirement on the DC bus. A 470\(\mu\)F aluminum capacitor, which has 50m\(\Omega\) ESR resistance, is selected. At the worst case condition, when back EMF is 50 V and inductor current is 10 A, voltage ripple due to 470\(\mu\)F capacitance is 40 mV, and 50m\(\Omega\) ESR resistance causes another 500-mV voltage ripple, then total voltage ripple on the DC bus is about 540 mV. It’s seen that excessively large capacitor does not reduce the voltage ripple due to the large ESR of this type capacitor. Low ESR capacitors and
secondary stage EMI filter are needed to attenuate voltage ripple. 470\mu F capacitance provides about 7-ms hold-on time at 500-W power level.

4.3.4 ZVT Inductor Design

The ZVT inductor design is basically the selection of current di/dt at the turn-off of rectifier diodes. For MOSFET IXFH42N20, which has less than 200-ns reverse recovery time, 400-ns transition time for turn-off of 20-A diode current is selected. 120-V bus voltage will be applied across ZVT inductor until the diode current is completely diverted into ZVT circuit. Therefore, based on following equation,

\[ L \frac{\Delta i}{\Delta t} = 120 \]  \hspace{1cm} (4.2)

the ZVT inductor is 24\mu H.

TDK ferrite core PC40EER28L-Z is used to construct the 24\mu H ZVT inductor, and is wound with 5 turns of Litz65/31 Litz wire. Air gap per leg is about 1 mm, and results in about 25\mu H inductance and 2500 gauss at 40A peak current.

ZVT auxiliary switches turn-on time is selected as 500 ns.

There are four saturable reactors in the ZVT inverter/rectifier. TOSHIBA Spike Killer cores are used. SA 14x8x4.5 with 4 turns is used for these three saturable reactors in front of ZVT diode bridge, SA 10x6x4.5 with 7 turns is placed in series with ZVT inductor.

4.4 Efficiency of Three-Phase Soft-Switched Inverter/Rectifier
The efficiency of the three-phase inverter/rectifier with quasi-square current waveforms and trapezoidal back EMF voltages is very difficult to measure directly. However, due to using twelve-step brushless DC motor/generator control, the inverter/rectifier is very similar to the bi-directional DC/DC converter, except for commutations every 30°. Therefore, the efficiency of the inverter/rectifier with brushless DC control should be similar to the efficiency measured under static conditions (DC-DC operation), if the transitions among those 30° segments can be ignored. Figure 4.18 shows the efficiency measured under the DC-DC condition. An efficiency of more than 97.4% at nominal conditions, i.e. 80V back EMF, is recorded, and meets the requirement set forth in Chapter 1.

Because of bi-directional operation and low voltage drop of the MOSFET device, MOSFET devices carry current most of time, therefore, inverter/rectifier should have similar efficiency in both rectifier mode and inverter mode. However, due to the extra switching action and saturable core loss, the ZVT soft-switched circuit has slightly lower efficiency in the inverter mode.

4.5 Summary

MOSFET devices are considered most suitable for the flywheel energy storage system application. An innovative soft-switched three-phase voltage source inverter/rectifier is introduced to solve the reverse recovery problems of the MOSFET antiparallel diodes.
Figure 4.18. Efficiency of the Inverter/Rectifier.
Measurement is taken under DC/DC conditions. Output power is 500 W, and switching frequency is 100 kHz.
Original soft-switching is simplified for the twelve-step brushless DC motor/generator control. By detecting directions of inductor current, the control circuit does not need to distinguish the inverter mode or the rectifier mode for taking commutation actions, and all extra commutation actions are eliminated.

There is a benefit from a large inductor current ripple. If inductor current changes sign during one switching cycle, natural ZVS can be realized without taking any switching action from ZVT commutation circuit.

The power stage design is discussed.

The efficiency of the three-phase inverter/rectifier is similar to the efficiency measured under static conditions (DC/DC operations). Using the ZVT soft-switching scheme leads to a good performance of the three-phase inverter/rectifier.
5. CONCLUSIONS AND FUTURE WORK

A design of power electronics for the flywheel energy storage system in satellite applications has been presented.

The flywheel energy storage system has weight/size and lifetime advantages, and will probably replace the battery energy storage system in satellite applications.

High speed flywheel is used to store kinetic energy and mounted to the rotor of a three-phase motor/generator. The motor/generator has a permanent magnet trapezoidal brushless DC structure, and 120° brushless DC control strategy simplifies the control of the three-phase inverter/rectifier to that of a bi-directional DC/DC converter.

A small-signal analysis has been conducted, and compensator design example is given. The peculiar behavior of a bi-directional DC/DC converter has been explained physically. Minimizing armature inductance of motor/generator field winding is necessary for wide control loop bandwidth.

High-switching frequency is required to reduce current ripple due to reduced inductance and to synthesize the high-frequency motor/generator current waveforms. An innovative ZVT soft-switched inverter/rectifier is introduced to solve the diode reverse recovery problem of MOSFET antiparallel diode. The prototype circuit demonstrates that high performance can be achieved by using ZVT soft-switching scheme. An example of the power stage design of the inverter/rectifier has been given.

The following research areas are of interest for future work:
**Test of Flywheel Energy Storage System.** At the time of this work being conducted, the construction of the magnetically suspended flywheel and the three-phase motor/generator has not been completed, and the power electronic circuitry design is based on the simulated motor/generator. Integrated flywheel system will have to be tested, and some practical issues such as start-up and fault protection need to be addressed.

**Optimization of Motor/Generator Design.** Since the structure of motor/generator can significantly affect the power electronic circuitry design, more collaborative work is needed to optimize motor/generator design. High switching frequency current ripple induces significant amount of eddy current loss inside motor/generator. Laminated structure is necessary to tackle this problem.

**Sinusoidal Motor/Generator Drive.** At this stage of flywheel system design, trapezoidal brushless DC motor/generator is selected to drive a flywheel, and all power electronic circuitry design is based on the trapezoidal type of motor/generator. Once sinusoidal brushless DC motor/generator is selected, DSP control will be used, and small-signal analysis of d-q model [23] of the three-phase inverter/rectifier needs to be performed.

**Design of a Magnetic Bearing Switching Amplifier.** The University of Maryland flywheel system employed a linear amplifier for the magnetic bearings. This would lead
to unacceptable losses at high speeds. This can be overcome by employing a switch-mode amplifier.
REFERENCES


APPENDIX

Schematics of the Control Electronic Circuitry

The control electronic circuitry of the flywheel energy storage system was designed by professor Xinfu Zhuang, who is a visiting scholar from P.R. China.

The detailed circuit diagrams of the control electronics of the flywheel system shown in Figure 2.7 will be described as followings.

The block diagram of the control electronics is shown in Figure A-1. There are three main function blocks, whose detailed circuit diagrams are drawn in Figure A-2, Figure A-3 and Figure A-4 respectively. Voltage compensator block includes the voltage loop compensator, which was designed in chapter 3 in order to achieve the desired bandwidth and gain-phase margins. Current loop block implements the current-injection control, and three PWM signals are generated in this block by sensing phase currents. The commutation logic block distinguishes the six 60° segments. During each 60° segment, one PWM signal will be selected and steered to the proper switches.

The six-step control was implemented in the control circuitry to control the motor/generator, and the twelve-step was verified only by simulation due to the limited time-frame of this work.

Figure A-2 shows the voltage compensator block. Vc is the voltage compensator output, and Vdc is the DC bus voltage, which needs to be tightly regulated. The voltage
of Zener diode is 6.8 V, and provides reference for the compensator. The operational
amplifier of the compensator is MC33074. The detection of the phase current directions is
implemented with three comparators, i.e. LM339NA. IM_A, IM_B and IM_C are sensed
phase currents from the Hall devices, and are based in order to have single voltage
polarity. 1.98-V indicates zero current. Ia, Ib, Ic are TTL signals, which indicate the
inductor current directions.

Figure A-3 shows the current loop block. IL_A, IL_B and IL_C are inductor
currents sensed directly by Hall devices, and IM_A, IM_B and IM_C are the sensed
inductor currents with a 1.98-V DC bias in order to use a single power supply. The PWM
generator is actually a 100-kHz square-wave oscillator implemented by using
Im555CNB. The output of Im555 is used to generate a ramp, which is injected as the
external ramp in order to eliminate the subharmonic oscillation. The biased phase A
current signal is compared against the output of the voltage compensator, Vc, and a PWM
signal is generated for the phase A, i.e. PWM_A. There is an additional comparator,
which is used to limit the phase current. The PWM signals, i.e. PWM_B and PWM_C for
phase B and phase C respectively, are generated by another two similar circuits, which
are not drawn here for simplicity.

Figure A-4 shows the commutation logic block. PWM_A, PWM_B and PWM_C
come from the current loop block, and will be selected according to the rotor positions.
Ia, Ib and Ic are phase current direction signals, which come from the voltage
compensator block. P_A, P_B and P_C are three rotor position signals, which are used to distinguish the six 60° segments. All these signals go into the decoder, whose logic determines which PWM signal is selected, how the PWM signal is steered to the six switches and when the ZVT network needs to take action. Au, Bu, and Cu are gate signals for the three upper switches, Ad, Bd and Cd are gate signals for the three bottom switches, and Aux is the gate signal for the auxiliary switches. Two comparators are used to provide the time delay for the ZVT switches. IM_A, IM_B and IM_C are biased phase current signals, and are used for the positive and negative overcurrent protections.

Figure A-5 is the logic table for the decoder, which is programmed into a GAL device EP610. The Boolean equations are based on the Altera Hardware Description Language (AHDL). The change from the six-step control to the twelve-step control is mainly exhibited here. More accurate rotor position signals are needed, and the EP610 needs to be re-programmed, but the rest of the circuit remains the same.
Figure A-1. Circuit Block Diagram of the Control Electronics.
Figure A-2 Schematics of Voltage Compensator Network
Figure A-3 Schematics of Peak Current Control Network
Figure A-4 Schematics of the Commutation Logic Network
SUBDESIGN decoder1
(
    Wa, Wb, Wc, Pa, Pb, Pc, Ia, Ib, Ic, Pro : INPUT;
    A, B, C, Au, Ad, Bu, Bd, Cu, Cd, Aux : OUTPUT;
)
BEGIN
    A  =  Pa & !Pc ;
    B  =  !Pa & Pb ;
    C  =  !Pb & Pc ;
    Au =  Pro & Wa & A ;
    Ad =  Pro & ( Wb & A # !Pa & Pc & !Aux ) ;

    Bu =  Pro & Wa & B ;
    Bd =  Pro & ( Wb & B # !Pa & !Pb & !Aux ) ;
    Cu =  Pro & Wa & C ;
    Cd =  Pro & ( Wb & C # Pb & !Pc & !Aux ) ;
    Aux =  Pro & ( Wb & Wc & ( A & Ia # B & Ib # C & Ic) # Pro & !Wa & !Wc & ( A & !Ia # B & !Ib # C & !Ic ) ;
END;

Figure A-5 Logic Table of the Commutation Decoder
Vita

The author received the Bachelor of Engineering degree in Automation Department from Tsinghua University, Beijing, China, in July, 1990. From July 1990 to December 1991, he was employed in Chengdu Chemical Engineering Corporation of China as a process control engineer. From December 1991 to May 1993, he worked in Legend Computer Group Corporation. In 1993, he joined the Virginia Power Electronics Center of Virginia Tech as a research assistant and to work on an M.S. degree. His research interest includes the high-frequency dc/dc power conversion, magnetics, and soft-switched three-phase inverter/rectifier.

Zhuy Jr.