The studies on one generator and one motor marine ship' s propulsion system control

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NANYANG
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**THE STUDIES ON ONE GENERATOR AND ONE MOTOR
MARINE SHIP'S PROPULSION SYSTEM CONTROL**

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The Studies on One Generator and One Motor Marine Ship's Propulsion System Control

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SUMMARY

The electric marine ships have been deployed for many years, but they are still facing a variety of problems and need to be perfected. Electricity makes it more efficient than consuming raw energy resources like gasoline. There are many rectifiers and inverters which are composed of conventional thyristors and diodes causing inherent drawbacks, such as harmonics and reactive power, which contribute to the power system pollution on the marine ships. In this thesis, a one generator and one motor marine ship's propulsion system is studied for simplicity and efficiency.

In my research, a one generator and one motor marine ship power system model is set up in which a new control method of PWM rectifier based on the power ripple minimization is introduced and applied. It will reduce the power loss and improve the power quality of the marine ship's power system. The marine ship's propulsion motor control is also demonstrated and the Fuzzy-PID controller is applied. What's more, an energy storage system is designed to stabilize the one generator and one motor marine ship's power system and store energy when it is necessary.

My main contributions in this thesis can be summarized as follows:

1. Modelling of the one generator and one motor marine propulsion system.
2. The generator model and its simulation in which a rectifier based on power ripple minimization DPC control strategy is applied.
3. The design of a double closed loop control system of DC motor and its improvement with Fuzzy-PID control.
4. An energy storage system is designed and applied in the one generator and one motor marine propulsion system.

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Chapter 1

Introduction and Literature Review

1.1 Backgrounds and Motivations

The electric marine ships emerged quite a few years ago, and the electrification of the marine ships has drawn much interest from quite a number of companies and customers. While they are still facing a number of problems such as power loss, power quality and power system stability. With respect to ship design and operation, minimizing power loss associated with power electronics harmonics and keeping power system operate in stable situation are critical goals. In the future, the electric ships will have more reliable power system and less energy consumption. Electricity has brought a new era for the marine ship industry and it needs a lot of work to be done in this new field.

The marine ship's electric power system is an independent power system, and is playing a crucial role on the ship.[1] Nowadays, the tonnages of ships are getting bigger and more ships are chosen to be installed with electric power system and propelled by electric motor. The capacity of electric generators on the marine ship is increasing with time. The power system on the marine ship has to be improved with higher level of stabilization because electric propulsion is the keystone for the ship's cruising. The electric power which is generated by electric generators also powers other electric appliances on the ship apart from the electric propeller. It is of great importance to stabilize the power system voltage at a rated value, otherwise it will damage marine ship's power grid and electric appliances.

The marine ship propulsion system applied with electric power is getting more and more popular in the shipping industry. In the electric propulsion system, the power source doesn't connect directly with the propellers, while the mechanical propulsion system does. In fact, the power source, such as a diesel turbine engine, is connected to a large electric generator and the generator will supply the electric motors driving the propellers with electricity.[2]

Figure 1.1 shows a typical configuration of a marine ship's electric propulsion system, which consists of a couple of components.

Figure 1.1 A Typical Configuration of a Marine Ship's Electric Propulsion System

All-electric or hybrid electric marine ship propulsion has become a leading and promising field of research, which fosters the investigation in hybrid propulsion system design, as well as demonstration of concept vessels. What's more, new and existing ships are being subject to regulatory requirements, specifically regarding emissions and energy efficiency[3].

It is a crucial goal that providing continuous and reliable electric power to the electric power system on the marine ship.

The electric propulsion system on the marine ship generally utilizes power electronic converters to drive variable-speed electric motors that are mechanically linked with the propellers. There are different kinds of power electronic converters applied on the marine ship, and a lot of engineers are spending plenty of time selecting the suitable converters that are the optimal ones which have a compromise between, like steady state and dynamic performance, volume and efficiency, quality and cost [4]. While in my research, a new control method of PWM rectifier is studied and it is based on the power ripple minimization. This will improve the power quality and reduce the power loss of the power system on the electric marine ships, and the dynamic stability is also strengthened.

1.2 Objectives and Major Contributions

The objectives of this thesis are shown as follows:

(1) To study the one generator and one motor marine ship's propulsion system and it is the preliminary research project supported by the Rolls-Royce Corporate @NTU Lab.

(2) It is aimed to develop a new control strategy for the PWM rectifier in the one motor and one generator marine ship's propulsion system. This new control strategy will help diminish the harmonics caused by the thyristors and diodes, as well as improve the power factor of the marine ship's power system.

(3) To design a better control method for the electric motor of the marine ship when the load is fluctuating and to design a DC voltage stabilizer in the one generator and one motor marine ship's power system.

The major contributions of this thesis are:

- (1) A one generator and one motor marine ship's propulsion system is modelled and studied.
- (2) A new control strategy based on the Power Ripple Minimization for the PWM rectifier direct power control is introduced. It is a good method which is applied in the AC/DC rectifier to restrain harmonics and increase the power factor in the marine ship's power grid. Based on the structure of three-phase PWM rectifier, the mathematical model in $d-q$ coordinate is set up. The calculation methods of instantaneous reactive power and active power are demonstrated in order to establish the PWM rectifier power model. A zero vector and a non-zero vector whose acting time is optimized are applied in the rectifier to realize the power ripple minimization for the marine ship's power system.
- (3) A Fuzzy-PID controller is designed and applied in the DC motor double closed loop control system. This controller will help make the speed at constant when the load is changing with time.
- (4) An energy storage system is introduced in order to stabilize the DC voltage between the AC/DC rectifier and DC/AC inverter, and it also can store and release energy when necessary. This energy storage system based on a reverse parallel boost-buck voltage stabilizer is an additional device which helps keep the one generator and one motor marine ship's power system stable.

1.3 Organization of the Thesis

This thesis is following the organization as:

Literature review is expressed in Sections 1.4, 1.5 and 1.6, in which the research background of PWM rectifiers is explained and the generally utilized control strategies are demonstrated as well.

Chapter 2 presents the modeling of one generator and one motor system. Chapters 3 and 4 describe the PWM rectifier DPC (Direct Power Control) based on the power ripple minimization strategy in details and its simulation compared with the conventional DPC's on marine ship's power system.

Chapter 5 demonstrates the DC motor double closed loop control system, and the Fuzzy-PID controller is studied and designed to improve the motor's performance in transient state in this chapter. What's more, the comparisons between the PID controller and Fuzzy-PID controller are demonstrated in simulations.

Chapter 6 illustrates the design of an energy storage system in the one generator and one motor marine propulsion system. It helps stabilize the DC voltage and store energy if necessary in the one generator and one motor marine propulsion system.

Chapter 7 presents a summary of this thesis and draw major conclusions with respect to the thesis objectives. And it also refers to the future research where there are possibilities and potentials.

1.4 Literature Review

Electric marine ship has been widely developed in these years and it has shown a good future in the new era because of the reduction of oil deposit and the flexibility of electrical control. About seven to nine out of ten power units from the generators in the marine ship's power system is consumed in the propulsion system. [5-7]

AC/DC converter, also known as rectifier, is the power circuit device capable of converting AC power to DC power, and it is also the earliest invented, with largest number and capacity power device in the area of practical application of the power electronics converters. It has been playing an important role in the electric marine propulsion ship.

For the sake of simplicity and efficiency, this thesis mainly focuses on the study of one generator and one motor marine ship's propulsion system. It is derived from Figure 1.1. So the one generator and one motor marine ship's propulsion system can be shown as a schematic diagram in Figure 1.2.

Figure 1.2 The One Generator and One Motor Marine Ship's Propulsion System

Figure 1.3 Categories of the Rectifiers

The topology of traditional rectifiers is made up of thyristors and diodes, and this kind of rectifiers is reliable and has easy control methods, which make it popularly applied in the industrial field. However, these thyristors' phase controlled circuits and diodes' uncontrolled rectifier circuits can pour lots of harmonics and reactive power into the power grid, which will destabilize the marine ship's power grid heavily. The main bad effects on the power grid are as follows:

(1) Adding tremendous reactive power, and making the power factor very low. Reactive power also leads to the increases of currents and apparent power, which results in increasing the loss of equipment and lines, and reducing the utilization efficiency of power generation equipment and transmission equipment.

(2) Reactive power's augment will make the line voltage increase and the impact of reactive loads also causes severe voltage fluctuations.

(3) Harmonics can cause protection devices malfunction, making conventional electrical measuring instruments not accurate.

(4) Harmonics affect the normal operation of electric equipment, leading to accelerated aging of the capacitors, cables and other equipment from overheating, which will shorten the life span of electric equipment.

(5) Harmonics can cause series and parallel resonant vibration in a few parts of the power grid, so that harmonics will amplify, which makes the harms caused by harmonics increase, and even leads to serious faults.

The marine ships' various power electronic devices and nonlinear loads and shock loads increase a lot in recent years, so this worsens power quality, and directly has a bad effect on the power system economical operation. In order to reduce harmonic injection, the fundamental measures are to require the AC devices not to generate harmonics, to achieve the line currents sinusoidal and unity power factor operation. The appearance of pulse width modulation (PWM) technology provides an important method for solving the harmonic "pollution" problem. Bring the PWM technology into the rectifier control, will make currents in the marine ships' power grid side of the rectifier sinusoidal and work in unity power factor, as well as achieving two-way flowing of energy, which is truly "green power conversion". So this technology is widely used in high power factor rectifier (HPFR), active filter (APF), new energy generation and AC drive systems, and has broad application prospects.

This thesis mainly focuses on the new control strategy of the PWM Rectifier and its application in the one generator and one motor marine ship's propulsion system. With respect to the AC/DC converter in Figure 1.2, based on the traditional voltage-oriented direct power control (VO-DPC), the thesis optimizes the voltage switching state table and the acting time of voltage vectors in traditional direct power control system, which will reduce the marine ship's power grid's power ripple and improve the steady-state performance. The ultimate goal is to achieve unity power factor and to meet the "green energy transformation" requirement. The marine ship will have a power system with higher power quality and much more stable operation due to the application of the PWM rectifier based on power ripple minimization control method.

1.5 Overview of PWM Rectifiers

1.5.1 PWM Rectifiers Develop and Status Quo

The rectifier that converts AC to DC is applied earlier than other inverters or converters, and its development process has gone through 3 stages including non-controlled rectifiers (diode rectifier), phase controlled rectifiers (SCR) and the PWM rectifiers.

The research of PWM rectifier began in the 1980s [8], during which time the development and widely use of the full-controlled switching devices [9] promotes the application and research of PWM technology. PWM Rectifier, in the fields of main circuit topology optimization and control technology update, has entered a new era and become more and more mature.

In 1982, it was the idea that a three-phase full-bridge PWM rectifier topology based on turn-off devices' and the control strategy of the AC side current amplitude and phase was firstly come up with by scholars[10], and this control strategy basically realized the goal that the current source PWM rectifier AC side are sinusoidal currents and unity power factor. A control strategy based on the PWM rectifier topology's reactive power compensator was put forward in 1984 [11], which provided a keystone for voltage source PWM rectifier research. By the late 1980s, a few scholars proposed a PWM rectifier continuous-discrete dynamic mathematical model based on the coordinate transformation and its control strategy, thus research and development of the PWM rectifier reached a new height [12].

Since the 1990s, PWM rectifier has always been the hot spot of research and focus by researchers and scientists. With further research on PWM rectifier, based on the PWM rectifier topology and its control strategy expanding, as well as many related applications developing fast, such as the active filter [13], AC drive [14], superconducting magnetic energy storage [15], Unified Power Flow Control (UPFC) and High Voltage Direct Current Transmission (HVDC). Research and development of these applications have promoted the advancement and improvement of PWM rectifier and its control technology. After decades of development, the rectifier circuit which is considered as the rectifier circuit in the traditional sense evolving into the PWM rectifier in the modern sense, and its topology is evolving continuously, its performance improving as well.

In recent years, PWM rectifier study is focused on the following aspects:

1. Modeling and analysis of the voltage source PWM rectifier;

2. Current control of the voltage source PWM rectifier;
3. Study on the main circuit topology ;
4. Study on the control strategy of the system;
5. Study on the current source PWM rectifier.

1.5.2 The Category of the PWM Rectifiers

PWM rectifier technology is developing continuously, at present people have devised a variety of PWM rectifiers and the classification methods are also diverse. The classifications are shown in Figure 1.4 [16]:

Figure 1.4 Classifications of the PWM Rectifiers

Although there are a variety of PWM rectifier classification methods, but the most basic classification method is to divide PWM rectifier into voltage source and current source. The difference between voltage source and current source PWM rectifiers is that they have different forms of energy storage at the DC side. Voltage PWM rectifier's DC side has a large capacitor which stores DC energy, while the current PWM rectifier's DC side is an inductor that holds DC energy. Each of both types has its own characteristics with respects to topologies, control strategies and other aspects, and both of them have the duality in their circuits. Other classification methods, concerning circuit topology, can be attributed to the voltage source or current source PWM rectifiers.

At present, with respect to the current source PWM rectifiers, voltage source ones have received more research and are more widely applied. So in this thesis, the voltage source PWM rectifier is utilized in the study.

1.5.3 The Control Strategy of the PWM Rectifiers

PWM rectifier control technology not only plays an important role in its development process, but also is the hot research area of the PWM rectifier in recent years. In terms of the development of control technology at present, PWM rectifier control techniques consists of indirect current control, direct current control, space vector PWM control (SVPWM), nonlinear control and direct power control, etc.[17]

(1) Indirect Current Control:

Indirect Current Control can be divided into steady-state indirect current control and dynamic-state indirect current control, mainly represented by the phase and amplitude control (PAC) [18, 19]. The essence of indirect current control technology is that by utilizing PWM control, it generates the sinusoidal PWM voltage whose amplitude and phase are controllable on the AC side of the bridge. The PWM voltage and power grid electromotive force (emf) work together in the grid side and form a sinusoidal fundamental current in the grid side, harmonic currents can be filtered out through the filter reactors of the grid side. Because this kind of current control scheme indirectly controls the voltage in the grid side in order to achieve the purpose of controlling the current in the grid side, thus it is a control strategy of indirectly current control.

The advantage of indirect current control is that it is simple to control, and generally does not need a current feedback controller. The disadvantage is that the dynamic response of the current in the PWM rectifier is not fast enough, and what even worse is that the current may contained DC component in the grid side, what's more, it is sensitive to the parameters' fluctuations of the system. So the indirect current control is often suitable and applied in the system with low level dynamic response requirements and simple control structures.

(2) Direct Current Control:

Direct current control is put forth for overcoming the shortcomings of the indirect current control (dynamic response is slow, sensitive to fluctuations of the system parameters),

such as Hysteresis Current Control (HCC) [20, 21], Current Control with Fixed Switching Frequency (CCFF) [22] and Space Vector Current Control (SVCC) [23].

With respect to the indirect current control, direct current control that is featuring the fast current feedback control has a closed-loop control on the grid side concerning the structure. Since the application of the grid-side current closed-loop control, the current dynamic and static performances improve a lot on the grid side of the bridge, and also it makes grid-side current control insensitive to the system parameters' fluctuations, which enhances the robustness of the current control system.

The advantage of direct current control is that it can get higher quality current response, but its control structures and algorithms are much complicated than the indirect current control's.

(3) Space Vector PWM Control:

Space Vector PWM (SVPWM) control strategy is a newly developed control method and its principle is to control the rectifier by switching the rectifier's space voltage vectors or space current vectors. SVPWM control strategy is firstly put forward by a scholar when he studied the AC motor variable frequency drive in the 1980s [24]. He abandoned the popular Sine Pulse Width Modulation control strategy and achieved a circular rotating magnetic field through choosing the inverter's space voltage vectors. In a low switching frequency working situation, SVPWM can achieve better performances than the SPWM does, which represented by a less ripple of the AC motor's torque, the promotion of the motor's dynamic response performance and the increase of the bus voltage utilization rate of the voltage inverter. Also, the simple vector mode switching method contributes to the application of the single-chip microprocessor.

SVPWM control is applied to the rectifier control, mainly for its advantages of high voltage utilization and fast dynamic response. SVPWM control has become a hot research topic in the field of PWM rectifier control strategies.

(4) Direct Power Control

Direct Power Control (DPC) strategy and current control strategy are different. DPC does not have to transform the power into corresponding current variables that can be controlled, while it introduces the instantaneous power theory. It makes the PWM rectifier

instantaneous output active power and reactive power as the controllable variables that are utilized for the power direct closed-loop control. The basic idea of DPC is: detecting and calculating the output of the PWM rectifier's instantaneous active power and reactive power, then sending the difference between the detected value and the given value into two corresponding hysteresis comparator; based on the hysteresis comparator's output and the judgment calculation of the grid voltage vector's position, afterwards determining the switching status that are used for driving each power electronics in the PWM rectifier bridge.

Compared with other control strategies, such as current control strategy and nonlinear control strategy, PWM rectifier direct power control strategy solves the control problems fundamentally(based on energy), and it achieves a higher power factor and a lower harmonics content, what's more, it has a simple structure, fast dynamic response, and a strong anti-interference ability, etc.

In the 1990s, the three-phase PWM rectifier Direct Power Control Strategy was firstly proposed [25], after which a few scholars made a deeper study on DPC strategy [26]. In recent years, scholars around the world have made significant achievements in the scope of PWM rectifier Direct Power Control research, and the Direct Power Control strategy is increasingly attracting widely attention of researchers.

1.6 The Disadvantages of the Traditional DPC Strategy

Direct Power Control (DPC) is a control method that controls the instantaneous power quickly in the three-phase voltage source PWM rectifier. The reference paper[27] gives the traditional three-phase PWM rectifier DPC system structure, and its schematic diagram is shown in Figure 1.5 (it has omitted AC side equivalent resistance R). The PWM rectifier DPC system consists of the main circuit and the control circuit.

The main circuit is made up of the three-phase AC power supply, filter reactors, PWM rectifier, DC Capacitor and DC load.

The control circuit is composed of the “outer loop” DC voltage and the “inner loop” power. The “outer loop” DC voltage is to stabilize or regulate the DC voltage and it consists of a DC voltage detecting circuit and the PI regulator. The “inner loop” is made up of the AC power voltage and current detection circuits, the sector divider, the power estimator, the power hysteresis comparator and a switch table.

Figure 1.5 The Schematic Diagram of the Traditional DPC Structure

The basic working principles of the traditional DPC system are:

- (1) Detecting power grid's three-phase voltages e_a, e_b, e_c and three-phase currents i_a, i_b, i_c ; and by utilizing the transformation matrix $T_{abc/\alpha\beta}$, then it can get the power grid's three-phase voltages and currents in the stationary α - β coordinate system: e_α, e_β and i_α, i_β .
- (2) According to e_α, e_β and i_α, i_β , the power calculator module estimates the instantaneous active power and reactive power p and q , and the given reference active power and reactive power are p_{ref} and q_{ref} , respectively; so the error signals between the estimation values and given ones are $p - p_{ref}$ and $q - q_{ref}$, and after inputting them into the power hysteresis comparator, it can get switch signals s_p and s_q ; p_{ref} is set by the product of the DC voltage outer loop PI regulator's output (represented as current) and the DC voltage, and q_{ref} is set to be zero in order to make the rectifier operate at the unity power factor state.
- (3) The sector divider is used to divide the sectors based on the e_α and e_β , then it achieves the sector signal θ_n .

- (4) The switch table can choose suitable voltage vectors on the basis of s_p , s_q and θ_n , afterwards drive the main circuit's power electronics and finally it can realize the goal of adjusting the instantaneous power.

Traditional DPC control has the problem of power ripple. This problem results from the reason that during the interval between two samplings, the switch table only selects one voltage vector and the voltage vector works in the entire control period. So it is very likely to cause the regulation intensity of the instantaneous power too strong or too weak, which contributes to the failure of the goal that the instantaneous power tracking the given power, resulting in producing the PWM rectifier's input power ripple and reactive power.

1.7 Chapter Summary

This chapter has demonstrated the backgrounds and motivations for the studies on the electric marine ships. Then, the objectives and major contributions of this thesis are introduced. And for the sake of efficiency and simplicity, the thesis mainly studies the one generator and one motor marine ship's propulsion system control, which is a preliminary research project supported by Rolls-Royce Corporate @NTU Lab.

The literature review and the overview of PWM rectifiers are presented at length in order to get a clear understanding of the one generator and one motor marine ship's propulsion system and the PWM rectifier utilized within it. The disadvantages of the traditional DPC control strategy of PWM rectifiers are described as well. Thus, it becomes a stimulus to put forward the power ripple minimization control strategy for the AC/DC rectifiers.

In next chapter, the one generator and one motor marine ship's propulsion system is analyzed and modelled. And the mathematical model is established and it will help set up the simulation models of the marine ship's propulsion system.

Chapter 2

One Generator and One Motor System Modelling

In this chapter the marine ship's one generator and one motor mathematical model is set up to make the complicated ship power system in Figure 1.1 a much simplified and easy to be studied one. This will be a fundamental base for the following chapters' studies and simulations.

2.1 Dynamical Analysis of the Two-Machine System

2.1.1 Swing Equations of the Two-Machine System with No Control Device

Figure 2.1 is derived from Figure 1.2 and the system includes a synchronous generator and an electric motor connected by a finite line. The AC/DC converter and DC/AC inverter have been treated as a reactance.

Figure 2.1 One Generator and One Motor System

It is known to all that any difference between the shaft torque and electromagnetic torque of the generator will definitely lead to its acceleration or deceleration.

$$T_a = T_t - T_e \quad (2.1)$$

$$P_a = P_t - P_e \quad (2.2)$$

$$P_a = T_a \omega = I \alpha \omega = M \alpha \quad (2.3)$$

P_a is in mega watts if M is in mega joule-second per electrical degree. And the unit of angular acceleration α is in electrical degree per second².

H and M are defined as inertia constants. The relation between them is expressed as:

$$M = \frac{H}{\pi f} \quad (2.4)$$

H is in the unit of seconds and M is in second per electrical radian.

Defining θ as the angular position of the generator's rotor, then it has:

$$\alpha = \frac{d^2\theta}{dt^2} \quad (2.5)$$

And because θ is continuously changing with time, it would be much simpler to measure angular position with respect to a reference axis rotating at the synchronous speed.

If δ is the angular displacement in electrical degrees in the synchronous rotating reference axis, then ω_s is the synchronous speed in electrical degrees per second.

$$\theta = \omega_s t + \delta \quad (2.6)$$

$$\frac{d\theta}{dt} = \omega_s + \frac{d\delta}{dt} \quad (2.7)$$

$$\omega - \omega_s = \frac{d\delta}{dt} \quad (2.8)$$

$$\frac{d^2\theta}{dt^2} = \frac{d^2\delta}{dt^2} \quad (2.9)$$

So it can get:

$$M \frac{d^2\delta}{dt^2} = P_a = P_t - P_e \text{ (swing equation 1)} \quad (2.10)$$

where the angle δ of the generator is torque angle, also named as power angle.

By the same method, the motor's swing equation can be achieved as:

$$M_m \frac{d^2\delta_m}{dt^2} = P_{am} = P_{em} - P_l \text{ (swing equation 2)} \quad (2.11)$$

The angular momentum M can be treated constant since the generator's rotor speed is almost the same as the synchronous speed unless the stability limit is exceeded.

If damping is not neglected and the damping power is considered proportional to $\frac{d\delta}{dt}$, the equation of the motion is modified as:

$$M \frac{d^2\delta}{dt^2} + D \frac{d\delta}{dt} = P_a = P_t - P_e \quad (2.12)$$

where D is the damping constant.

Marine ship's generator system consists of speed governor, diesel prime mover, generator and compound excitation transformer.[28] The schematic of generator system structure is displayed in Figure 2.2. From Figure 2.2, it can be learned that the generator system includes two closed-loop feedback control systems: speed feedback control system, which is shown in Figure 2.3 and excitation feedback control system, which is shown in Figure 2.4. After combining the two systems, the marine ship's generator system can be shown in Figure 2.5.

Figure 2.2 The Schematic of Marine Ship's Generator System Structure

Figure 2.3 Speed Feedback Control System of Marine Ship's Generator

Figure 2.4 Excitation Feedback Control System of Marine Ship's Generator

Figure 2.5 Voltage and Frequency Feedback Control Systems of Marine Ship's Generator

The speed feedback system works as a detector which measures the generator's speed value. The compound excitation transformer that is also known as feedback of excitation system can detect the output voltages as well as currents of the generator. With respect to the power system running reliability, it is crucial to keep the frequency of power grid constant. With a constant frequency of the power grid, it can provide a solid base where the speed control of the synchronous generator and induction motor can be conducted. The frequency of power grid is determined by the speed of all the in-net generator sets. The speed control system of the generator set has direct effects on the active power output of the generator. With respect to the control of the marine ship's generator excitation system, its control

features directly affects the quality of electricity as well. The excitation system control will have a great influence on the reactive power output of the generator and voltage stability of the ship's power system.[28]

2.1.2 Governing Equations for the Transient Process

The electromechanical equation of the marine ship's generator is described as

$$Ms^2\delta + Ds\delta = P_t - P_e \quad (2.13)$$

where P_t is constant and P_e is independent on δ only, and all other marine ship's propulsion system variables keep constant.

Then it is to analyze the marine ship's power system by taking into consideration of the transient process in the field winding.

The field winding of the marine ship's generator is charged with a constant voltage V_f from an exciter. And the governing equation is expressed as:

$$V_f = r_f i_f + \frac{d\psi_{ffd}}{dt} \quad (2.14)$$

and (2.14) can be transformed as:

$$\frac{V_f}{r_f} = i_f + \frac{1}{r_f} \frac{d\psi_{ffd}}{dt} \quad (2.15)$$

so it can also expressed as:

$$i_{fe} = i_f + \frac{1}{r_f} \frac{d\psi_{ffd}}{dt} \quad (2.16)$$

where i_{fe} is the field current at the steady state.

By transforming (2.16) into:

$$x_{afd}(\dot{i}_{fe} - \dot{i}_f) = \frac{x_{afd}}{r_f} \frac{d\psi_{ffd}}{dt} \quad (2.17)$$

it can have:

$$E_{qe} - E = \frac{x_{afd}}{r_f} \frac{x_{ffd}}{x_{ffd}} \frac{d\psi_{ffd}}{dt} \quad (2.18)$$

$$E_{qe} - E = \frac{L_{ffd}}{r_f} \frac{\omega x_{afd}}{x_{ffd}} \frac{d\psi_{ffd}}{dt} \quad (2.19)$$

Defining,

$$E'_q = \omega \psi_{ffd} \frac{x_{afd}}{x_{ffd}} \quad (2.20)$$

then it can get:

$$E_{qe} - E = T_{d0} \frac{dE'_q}{dt} \quad (2.21)$$

$$T_{d0} = \frac{L_{ffd}}{r_f} \quad (2.22)$$

The basic differential equation of the transient process in the marine ship's generator's rotor winding is:

$$E_{qe} = E + T_{d0} \frac{dE'_q}{dt} = E + T_{d0} s E'_q \quad (2.23)$$

Finally, the governing equations for the transient process are shown as:

$$\begin{cases} Ms^2\delta + Ds\delta = P_t - P_e \\ E_{qe} = E + T_{d0} \frac{dE'_q}{dt} = E + T_{d0} s E'_q \\ P_e = \frac{EV}{X_{d\varepsilon}} \sin \delta \end{cases} \quad (2.24)$$

The above analysis will help set up the Simulink simulation models of the one generator and one motor marine ship's propulsion system.

2.2 Voltage and Speed Regulators of Marine Ship's Generator

2.2.1 The Voltage Regulator

From Figure 2.4, the voltage regulator can be expressed as:

Figure 2.6 Block Diagram of the Voltage Regulator

The equation that controls the performance of the marine ship's generator system is:

$$\frac{E_{qe}}{V_r - V_t} = \frac{K_A}{(1 + T_A s)(1 + T_e s)} \quad (2.25)$$

Usually amplifiers work very fast and therefore the time delay in amplifying process can be omitted. So (2.25) can be simplified as:

$$\frac{E_{qe}}{V_r - V_t} = \frac{K_A}{(1 + T_e s)} \quad (2.26)$$

Linearizing (2.26), it can achieve:

$$\Delta E_{qe} = \frac{K_A (-\Delta V_t)}{(1 + T_e s)} \quad (2.27)$$

The voltage regulator's complexity can be improved by means of adding more stabilization loops, such as exciter stabilization loop, and more amplifier stages into the regulator. In the case of marine ship's generator, it only considers one stage.

Figure 2.7 One Stage Amplification and Excitation Stabilization

So the governing equations for the one stage amplification and exciter stabilization are expressed as:

$$E_{qe} = \frac{K_A(V_r - V_t - V_s)}{(1 + T_e s)} \quad (2.28)$$

$$V_s = \frac{K_s s E_{qe}}{(1 + T_e s)} \quad (2.29)$$

2.2.2 The Speed Regulator

A speed regulator usually consists of Speed Governor(SG), Speed Control Mechanism(SCM), Governing Controlled Values(GCV) and Speed Changer(SC). And they have the following features:

1. The Speed Governor consists of only the devices that have direct responds to the speed and position, as well as influences on action of other devices of the speed governing system.
2. The Speed Control Mechanism comprises devices, like relays, pressure or power amplifying devices, servo motors, levers and linkages between speed governor and governor controlled valves.[29]
3. Governor Controlled Valves is composed of the valves that control the input energy of the turbine and that are usually actuated by speed governor via the medium of speed control mechanism.
4. Speed Changer works as an assistive device by which the speed governing system can be adjusted to change the turbine's speed or its output power when it is working.

The speed signal and speed changer position are imported into the speed governing system. The signal U_G which the governor speed changer produces is decided by the automatic generating control system. The signals results in a change of the governor controlled valve's position which can alter the input into the turbine. The input changes of the turbine can lead to changes of the turbine torque which is linked with the generator's load torque and marine ship's power system.[30]

Figure 2.8 displays the speed governing system (SGS) of the steam turbine, which is derived from the speed feedback loop in Figure 2.2.

Figure 2.8 Schematic Diagram of a Turbine Speed Governing System

Time lag T_{g2} is the dynamic response of a non-reheat turbine. So it can get:

$$\Delta P_t = \frac{1}{1+T_{g2}s} \Delta P_v \quad (2.30)$$

where ΔP_v = the change in valve position.

ΔP_t = the change in turbine power.

The speed governor response can be expressed as:

$$\Delta P_v = \frac{-K_g}{1+T_{g1}s} \Delta \omega \quad (2.31)$$

So, the overall transfer function of the turbine's governing system is

$$\Delta P_t = \frac{-K_g}{(1+T_{g1}s)(1+T_{g2}s)} \Delta \omega \quad (2.32)$$

2.3 Chapter Summary

This chapter has demonstrated the modelling process of the one generator and one motor marine ship's system. It mainly focuses on the marine ship's generator and the motor model will be presented in Chapter 5.

The mathematical modelling of the one generator and one motor system will help set up the simulation models of the marine ship's propulsion system.

The next chapter will start to present the mathematical modelling of the PWM rectifier and it is the base of the proposed power ripple minimization control strategy.

Chapter 3

The Theory, Topology and Mathematical Modelling of PWM Rectifiers

PWM is a kind of control whose full name is pulse width modulation technique. It is a technique that controls the on or off of the switching devices so that it can get a series of output pulses with the same amplitude, and utilize these pulses in place of a sine wave or a desired waveform.

The PWM rectifier is applied with PWM control mode, and it is a new type high frequency rectifier circuit composed of full-controlled devices. This chapter introduces the basic principles and the main circuit topology of PWM rectifier, and gives the mathematical models of different coordinate systems aimed to simplify the design of the control system.

3.1 The Structure and Theory of the PWM Rectifier

3.1.1 The Overview of PWM Rectifier Theory

The PWM rectifier utilizes full-controlled power switching devices, which is a controllable device from both AC side and DC side, and it is no more the traditional conception of the ordinary AC/ DC converter. It can control the power factor of the power grid side, like unity power factor control, and make the electrical power flow bidirectionally. When the PWM rectifier absorbs energy from the grid, which means it runs in the rectifier mode, while when the PWM rectifier pours energy into the grid, which means it runs in the inverter mode. The so-called unity power factor means that when running in PWM rectifier operation mode, its grid-side voltage and current are in the same direction (positive resistance characteristic). When the PWM rectifier runs in the inverter mode, its grid-side voltage and current are in reverse directions (negative resistance characteristic). Since the PWM rectifier grid-side current and power factor can be controlled, it is also applied in the Static Var Generator (SVG) [31] and Active Power Filter (APF) [32] that are non-rectification applications.

Therefore, the PWM rectifier is actually a four-quadrant operation device. First, the basic principle is demonstrated from the model of PWM rectifier circuit in Figure 3.1.

The PWM rectifier circuit consists of AC and DC circuits, and the power electronic switch bridge. The AC circuit includes the electromotive force e and AC side filter inductor L , etc. The DC circuit includes a load resistor R_L and the load electromotive force e_L , etc. The power electronic switch bridge can be composed of voltage or current bridges.

Figure 3.1 The PWM Rectifier Model Circuit

If the loss of the power electronic switch bridge can be neglected, based on principle of power balance between the AC and DC loop, the following can be achieved:

$$IU = i_{dc}v_{dc} \quad (3.1)$$

where, U and I ----- the model circuit's AC side voltage and current,

v_{dc} and i_{dc} ----- the model circuit's DC side voltage and current.

The (3.1) shows that by controlling the AC side of the PWM rectifier will achieve the goal of controlling the DC side. Therefore, from the AC side of the model circuit, the PWM rectifier's operation status and control principle can be analyzed.

On the steady-state condition, the AC side's vector diagram of the PWM rectifier's operation in four quadrants is shown in Figure 3.2.

To simplify the analysis, for the model of PWM rectifier circuit, it considers only the fundamental component of the AC side and neglects the PWM harmonics as well as the input reactor resistance value, assuming that the system is stable and at ideal state. Vector relationship analysis in Figure 3.2: taking the AC side of the grid EMF E as a reference, if it is assumed that the AC side input current size $|I|$ is unchanged, the inductor voltage vector

$|V_L| = \omega L |I|$ will be constant. Therefore, the running track of the PWM rectifier's AC side voltage vector $\mathbf{V}$ endpoint is a circle with the radius of $|V_L|$. So the operation state and power factor of the PWM rectifier is determined by the angle of the three-phase power source synthesized vector $\mathbf{E}$ and current vector $\mathbf{I}$. Therefore, by controlling the AC side voltage vector $\mathbf{V}$, the goals can be achieved that the PWM rectifier works in four-quadrant operation and the power factor is adjustable.

Figure 3.2 The PWM Rectifier's AC Side Steady State Vector Diagram

(a) Pure Inductance Characteristic Operation (b) Positive Resistance Characteristic Operation

(c) Pure Capacity Characteristic Operation (d) Negative Resistance Characteristic Operation

$\mathbf{E}$ ——Vector of the AC grid EMF

$\mathbf{V}$ —— Vector of the AC side voltage

$\mathbf{V}_L$ —— Vector of the AC side inductance voltage

$\mathbf{I}$ —— Vector of the AC side current

As Figure 3.2(a) shown, when the voltage vector $\mathbf{V}$ endpoint runs the circular locus point A, the current vector $\mathbf{I}$ lags the EMF vector $\mathbf{E}$ by 90° , which means the grid-side of PWM rectifier demonstrates the pure inductance characteristic. With respect to Figure 3.2(b), when the voltage vector $\mathbf{V}$ endpoint runs around the circular track to the point B, the current vector $\mathbf{I}$ and EMF vector $\mathbf{E}$ are parallel and in the same direction, so the grid-side of PWM rectifier exhibits the positive resistance characteristic, and when the PWM rectifier operates at the point B, the unity power factor can be realized. As for Figure 3.2(c), when the voltage

vector $\mathbf{V}$ endpoint runs to the circular locus point C, the current vector $\mathbf{I}$ leads EMF vector $\mathbf{E}$ by 90° , so the PWM rectifier's grid side exhibits pure capacity characteristic. When it comes to Figure 3.2(d), the voltage vector $\mathbf{V}$ endpoint runs around the circular track to the point D, and the current vector $\mathbf{I}$ and EMF vector $\mathbf{E}$ are parallel but in reverse direction, so the grid side of the PWM rectifier shows negative resistance characteristic. In other words, when the PWM rectifier operates at point D, the unity power factor inverter control is achieved.

From the above analysis, it can be learned that A, B, C, and D are the specific points at the operation circular track. So generally analyzed, the PWM rectifier's four quadrant operation rules are as followed.

Figure 3.3 General Diagram of the PWM Rectifier's AC Side Steady State Vector

(a) Converter Operation (b) Converter Operation (c) Inverter Operation (d) Inverter Operation

1. When the voltage vector $\mathbf{V}$ endpoint moves at the trajectory of circular arc AB , as shown in Figure 3.3(a), current vector $\mathbf{I}$ lags the EMF vector $\mathbf{E}$ and the angle is an acute angle, so the PWM rectifier absorbs active power and inductive reactive power from the grid. At this time the PWM rectifier is running at the rectification state and the power is transported through the PWM rectifier to the DC load from the grid.
2. When the voltage vector $\mathbf{V}$ endpoint moves at the trajectory of circular arc BC , as shown in Figure 3.3(b), the current vector $\mathbf{I}$ leads the EMF vector $\mathbf{E}$ and the angle is an acute angle, and the PWM rectifier absorbs active power and capacitive reactive power from

the grid. At this time the PWM rectifier is also running at the rectification state, which means power is conveyed through the PWM rectifier from the grid to DC load.

3. When the voltage vector V endpoint moves at the track of circular arc CD , as shown in Figure 3.3(c), the current vector I leads the EMF vector E and the angle is obtuse, thus the PWM rectifier transmits active power and capacitive reactive power to the grid. In this case, the PWM rectifier is running as an inverter, and power is poured back into the grid from the DC side through the PWM rectifier.
4. When the voltage vector V endpoint moves at the track of circular arc DA , as shown in Figure 3.3(d), the current vector I lags the EMF vector E and the angle is obtuse, and the PWM rectifier delivers active power and inductive reactive power to the grid. At this situation, the PWM Rectifier is running in the inverter status, in which power is delivered back to the grid through the PWM rectifier from the DC. When the voltage vector V endpoint runs back to the point A, the current vector is perpendicular to the voltage vector again, and the circuit works as the inductive load output, which means it completes the four-quadrant operation.

In summary, in order to achieve the four-quadrant PWM rectifier operation, controlling the AC voltage vector V is a good way. By controlling the AC voltage vector, the PWM rectifier can work in any position in the four quadrants.

3.1.2 The Topology of Voltage Source PWM Rectifier

The DC side of voltage source PWM rectifier (VSR) is applied with big capacity to store energy, and it serves as a low resistance voltage source.

(1) The Structure of Single Phase VSR Circuit

As shown in Figure 3.4, (a) and (b) are the main circuit structures of the single-phase half-bridge VSR and the single-phase full-bridge VSR respectively. They have the same structure of the AC side circuit, in which the input inductor is to filter the grid side's current harmonics.

As it is shown in Figure 3.4(a), the single-phase half-bridge VSR circuit, and there are two power switches in one bridge arm, while the other arm has the two capacitors in series. The two capacitors in series are used as the energy storage capacitors of the DC side as well. The single-phase full-bridge VSR circuit structure that is shown in Figure 3.4(b) has four power switches which constitute the "H" bridge shaped configuration. And each of the VSR

main circuit's power switches is in reverse parallel with a freewheeling diode, which is used to buffer the reactive power in the rectification process.

Figure 3.4 The Topologies of Single-Phase Half-Bridge and Full-Bridge

By comparing the structures of the two circuits, it can be learned that the single-phase half-bridge VSR circuit structure is relatively simpler, and its number of power switch is half of that in full-bridge VSR circuit. Therefore the single-phase half-bridge VSR has a lower cost, commonly used in low-cost and low-power applications. But on the condition that the circuit parameters of the two VSR's AC side are same, the half-bridge circuit power switch is exposed to a DC voltage that is twice of the full bridge circuit's, therefore the requirements for the breakdown voltage of half-bridge circuit's power switches are relatively high. In addition, the half-bridge circuit also requires the introduction of capacitive equalizing control so that the half-bridge circuit's capacitance midpoint electric potential is essentially the same, so a single-phase half-bridge VSR control is much more complex.

The single-phase full-bridge VSR PWM is analysed as followed.

Figure 3.4(b) has shown the topology structure of the single-phase full-bridge VSR main circuit, since there are two switch modes for each bridge arm, which are the upper arm on and the lower arm on, the single-phase full-bridge VSR has $2^2=4$ types of switch mode. To make a further research in these 4 types of switch mode and main circuit topology, it can be learned that the AC side voltage $v(t)$ is jumping between 0 and V_{dc} or 0 and $-V_{dc}$. Table 3.1 has shown the relations between switches' states and AC side's voltage $v(t)$.

Since state 3 and state 4 lead the single-phase full bridge's AC side voltage to be zero, they are named as zero modes. When the grid side instantaneous current direction is different,

the same switch state has different current loops. Take $i(t) > 0$ for example, Figure 3.5 depicts the current loops of switch state 1 and switch state 4 when the condition is $i(t) > 0$.

Table 3.1 Single-Phase Full Bridge VSR Switches' States

Switches' States	1	2	3	4
Devices On	V_1 (VD_1) V_4 (VD_4)	V_2 (VD_2) V_3 (VD_3)	V_1 (VD_1) VD_3 (V_3)	V_2 (VD_2) VD_4 (V_4)
AC side voltage $v(t)$	V_{dc}	$-V_{dc}$	0	0

Figure 3.5 Current Loops of Single-Phase Full Bridge VSR Different Switch States When $i(t) > 0$

When further analysing the current loops of single-phase full bridge on condition that the grid side current $i(t) > 0$, it is easily achieved that the current loops shown in Figure 3.5 are equivalent to the boost circuit shown in Figure 3.6.

Figure 3.6 Structure of the Boost Circuit

Since VSR has the boost circuit structure, the voltage source PWM rectifier is also known as the boost type PWM Rectifier. The VSR's transformation from AC side to DC side exhibits the boost converter's voltage step-up characteristics. Thus, when the PWM rectifier operates normally, the output DC voltage value is greater than the input AC voltage peak value, namely $V_{dc} > U_m$.

The AC side's inductor voltage can be described as:

$$V_L = L \frac{di(t)}{dt} = e(t) - kv_{dc} \quad (3.2)$$

in which, $k=1, 0$ or -1 .

When $k=1$, the inductor voltage is negative, and the input AC current $i(t)$ decreases. When $k=-1$, the inductor voltage is positive, and the input AC current $i(t)$ increases. When $k=0$, the input AC current trends is determined by the grid EMF $e(t)$. Thus, full control of the input current is realized [33].

When the prerequisite condition $V_{dc} > U_m$ is not met, for example, when it is just started, the input AC current is not controlled, so at this time the DC capacitor will be charged through the diode until its voltage reaches the peak value of the grid voltage.

(2) The Main Circuits' Structures of Three-Phase Half-Bridge and Full-Bridge VSRs

Figure 3.7 shows the main circuit's structure of the three-phase half-bridge VSR and the main circuit's structure of the three-phase full-bridge VSR separately.

Figure 3.7(a) is the topology of three-phase half-bridge VSR. Three-phase half-bridge VSR grid side is three-phase symmetrically connected without neutral line, and the VSR installs six power switches, the same as single-phase VSR does, every power switch in reverse parallel with a freewheeling diode in order to feedback reactive power. Currently, the three-phase half-bridge VSR is the most widely used three-phase PWM rectifier and it is also the study object in this thesis. Generally speaking, the so-called three-phase bridge circuit refers to the three-phase half-bridge circuit.

Three-phase half-bridge VSR is more suitable for three-phase balanced grid system. When the grid side of the three-phase power is asymmetrical, it will make a three-phase half-

bridge VSR control performance deteriorated, or even worse will cause the rectifier circuit failed. To overcome this disadvantage, the three-phase full-bridge VSR can be applied and its topology is shown in Figure 3.7(b). The DC bus of three-phase full-bridge VSR is in connection with three independently controllable single-phase full-bridge VSRs, and is connected to the three-phase four-line grid via a transformer. Therefore, the three-phase full-bridge circuit is actually a combination of three independently controllable single-phase full-bridge VSRs.

Figure 3.7 The Topologies of Three-Phase VSRs

Since the three single-phase full-bridge VSRs are independently controlled with respect to the three-phase alternate currents, when the grid is not balanced, it will not seriously affect the control performance of the PWM rectifier. But the three-phase full-bridge VSR uses 12 power switches, twice of that in the three-phase half-bridge circuit, so it has a higher cost, which means it is less applied in industry.

Since the conventional three-phase VSR's topology is a three-phase half-bridge configuration, this section will analyse the PWM process of the three-phase half-bridge VSR. The following three-phase VSR refers to the three-phase half-bridge VSR.

Compared with the single-phase full-bridge VSR, the three-phase VSR's main circuit has one more bridge arm. The three-phase bridge arms are applied with the symmetrical three-phase sine wave modulation signals that have the same amplitude and frequency while the phase difference is 120° . Since the two power switches of the same bridge arm cannot be turned on at the same time, every phase arm of the three-phase VSR has two switch modes, which are the upper arm turned on or the lower arm turned on. The three-phase VSR has $2^3 = 8$ kinds of switch modes, and can be expressed by unipolar binary logic switching function s_j ($j = a, b, c$) as:

$$s_j = \begin{cases} 1 & V_j \text{ and } VD_j \text{ are on; } V'_j \text{ and } VD'_j \text{ are off} \\ 0 & V_j \text{ and } VD_j \text{ are off; } V'_j \text{ and } VD'_j \text{ are on} \end{cases} \quad (j = a, b, c) \quad (3.3)$$

where, V_j and VD_j ($j = a, b, c$) represents the upper arm's power switch and freewheeling diode, while V'_j and VD'_j ($j = a, b, c$) represents the lower arm's power switch and freewheeling diode. The three-phase VSR's 8 kinds of switch modes are demonstrated in Table 3.2.

In Table 3.2, the switch mode 7 and 8 cause the VSR's AC side three-phase line voltage to be zero, so they are named as "zero mode". Normally, the "zero mode" is chosen by the principle that it generates the least number of power switch's on and off. Because different grid side current's instantaneous directions correspond to different current loops, it is more complex compared with the single-phase VSR current loop. Let's assume the three-phase grid currents $i_a > 0$, $i_b < 0$ and $i_c > 0$. Figure 3.8 shows the current loops corresponding to the PWM control eight kinds of switch modes when the grid currents $i_a > 0$, $i_b < 0$ and $i_c > 0$.

Table 3.2 Three-Phase VSR PWM Switch Mode

Switch Mode	1	2	3	4
Device On	$V_a(VD_a)$	$V'_a(VD'_a)$	$V_a(VD_a)$	$V'_a(VD'_a)$
	$V'_b(VD'_b)$	$V_b(VD_b)$	$V_b(VD_b)$	$V'_b(VD'_b)$
	$V'_c(VD'_c)$	$V'_c(VD'_c)$	$V'_c(VD'_c)$	$V_c(VD_c)$
Switch Function	001	010	011	100

Switch Mode	5	6	7	8
Device On	$V_a(VD_a)$	$V'_a(VD'_a)$	$V_a(VD_a)$	$V'_a(VD'_a)$
	$V'_b(VD'_b)$	$V_b(VD_b)$	$V_b(VD_b)$	$V'_b(VD'_b)$
	$V_c(VD_c)$	$V_c(VD_c)$	$V_c(VD_c)$	$V'_c(VD'_c)$
Switch Function	101	110	111	000

The analytical method for other current direction combinations is similar, so there is no need to describe it again in this thesis.

(a) Switch Mode 1

(b) Switch Mode 2

(c) Switch Mode 3

(d) Switch Mode 4

(e) Switch Mode 5

(f) Switch Mode 6

(g) Switch Mode 7

(h) Switch Mode 8

Figure 3.8 Current Loops of Three-Phase VSR PWM Different Switch Modes

3.2 Three-Phase Mathematical Model

In the late 1980s, the scholars proposed the PWM rectifier continuous discrete dynamic mathematical model and control strategy based on coordinate transformation. This section will take the three-phase voltage source PWM rectifier as study object and introduce the three-phase voltage source PWM rectifier's mathematical model [34].

3.2.1 VSR Mathematical Model in the Three-Phase Stationary a-b-c Coordinate

The three-phase VSR's general mathematical model is based on a three-phase VSR topology, which is in the three-phase stationary a-b-c coordinate, and it can be established that a general mathematical description of VSR via utilizing circuit basic laws (Kirchhoff's voltage and current laws). Three-phase VSR topology is shown in Figure 3.9.

The symbols in Figure 3.9 represent the following meanings as:

- (1) e_a, e_b and e_c ----- Three-phase power source EMF instantaneous values.
- (2) i_a, i_b and i_c ----- Rectifier's input three-phase current instantaneous values.
- (3) V_{aN}, V_{bN} and V_{cN} ----- Rectifier's grid side input voltage instantaneous values.

(4) L and R ----- Grid side filter reactor's inductance value; Equivalent resistance value of inductance parasitic resistance as well as power switch loss resistance.

(5) C and R_L ----- DC side electrolytic capacitors; Load resistance.

(6) V_{dc} ----- DC side voltage.

(7) U_N ----- Zero potential reference voltage.

Figure 3.9 The Three-Phase VSR Main Circuit

To simplify the analysis, it usually can make the following assumptions:

- (1) Power grid EMFs e_a, e_b and e_c are symmetrical three-phase sine wave EMFs;
- (2) The filter reactor is linear and the saturation is neglected;
- (3) The power switches' loss and AC inductors as well as grid side resistance is equivalent to be represented by resistance R ;
- (4) The load is a pure resistance R_L and the system is running in the rectification state.

The three-phase power grid EMFs are expressed as:

$$\begin{cases} e_a = U_m \cos \omega t \\ e_b = U_m \cos \left(\omega t - \frac{2}{3} \pi \right) \\ e_c = U_m \cos \left(\omega t + \frac{2}{3} \pi \right) \end{cases} \quad (3.4)$$

in which, ω is the AC power source angular frequency.

Firstly, the unipolar binary logic switch function of three-phase PWM rectify bridge is defined as:

$$s_k = \begin{cases} 1 & \text{upper arm is on; lower arm is off.} \\ 0 & \text{upper arm is off; lower arm is on.} \end{cases} \quad (k = a, b, c) \quad (3.5)$$

Based on the Kirchhoff's Voltage Law (KVL), the three-phase VSR's AC side loop equation in the a-b-c coordinate is:

$$L \frac{d}{dt} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} e_a \\ e_b \\ e_c \end{bmatrix} - R \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} - \begin{bmatrix} v_{aN} + v_{NO} \\ v_{bN} + v_{NO} \\ v_{cN} + v_{NO} \end{bmatrix} \quad (3.6)$$

When V_a is turned on while V_a' is turned off, $s_a = 1$ and $v_{aN} = v_{dc}$; when V_a is turned off while V_a' is turned on, $s_a = 0$ and $v_{aN} = 0$. Since $v_{aN} = v_{dc} s_a$, (3.6) can be transformed into:

$$L \frac{d}{dt} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} e_a \\ e_b \\ e_c \end{bmatrix} - R \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} - \begin{bmatrix} v_{dc} s_a + v_{NO} \\ v_{dc} s_a + v_{NO} \\ v_{dc} s_a + v_{NO} \end{bmatrix} \quad (3.7)$$

Based on the three-phase balance relation of voltage source inverter's bridges, the following equations can be achieved:

$$e_a + e_b + e_c = 0 \quad i_a + i_b + i_c = 0 \quad (3.8)$$

From (3.7) and (3.8), the following equation can be derived [35]:

$$v_{NO} = -\frac{v_{dc}}{3} \sum_{k=a,b,c} s_k \quad (3.9)$$

In addition, the DC side capacitor's positive node is applied by Kirchhoff's Current Law (KCL) and the following is achieved:

$$C \frac{dv_{dc}}{dt} = i_a s_a + i_b s_b + i_c s_c - \frac{v_{dc}}{R_L} \quad (3.10)$$

Doing the calculation with equations from (3.7) to (3.10) as well as considering the introduction of state variable X as $X = [i_a, i_b, i_c, v_{dc}]^T$, it can be achieved that the state equation of the three-phase VSR's general mathematical model in the form of unipolar binary logic switch function:

$$Z \dot{X} = AX + BE \quad (3.11)$$

in which,

$$A = \begin{bmatrix} -R & 0 & 0 & -(s_a - \frac{1}{3} \sum_{k=a,b,c} s_k) \\ 0 & -R & 0 & -(s_b - \frac{1}{3} \sum_{k=a,b,c} s_k) \\ 0 & 0 & -R & -(s_c - \frac{1}{3} \sum_{k=a,b,c} s_k) \\ s_a & s_b & s_c & -1/R_L \end{bmatrix} \quad (3.12)$$

$$Z = \begin{bmatrix} L & 0 & 0 & 0 \\ 0 & L & 0 & 0 \\ 0 & 0 & L & 0 \\ 0 & 0 & 0 & C \end{bmatrix} \quad (3.13)$$

$$B = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1/R_L \end{bmatrix} \quad (3.14)$$

$$E = [e_a, e_b, e_c, e_L]^T \quad (3.15)$$

3.2.2 The VSR Mathematical Model in the Two-Phase Stationary α - β Coordinate

From (3.5) to (3.10), it is not difficult to find out that, in the three-phase stationary a-b-c coordinate, the three-phase VSR switch function model is:

$$\begin{cases} C \frac{dv_{dc}}{dt} = \sum_{k=a,b,c} i_k s_k - i_L \\ L \frac{di_k}{dt} + Ri_k = e_k - v_{dc} \left(s_k - \frac{1}{3} \sum_{j=a,b,c} s_j \right) \\ \sum_{k=a,b,c} e_k = \sum_{k=a,b,c} i_k = 0 \end{cases} \quad (3.16)$$

in which, s_k ----- the unipolar binary logic switch function (k=a,b,c).

i_L ----- VSR DC side load current.

The three-phase PWM rectifier's mathematical model in the three-phase stationary coordinate system is concluded based on Kirchhoff's circuit laws, and the physical meaning is clear and intuitive as well as easy to be understood. However, in the AC side, there are a number of nonlinear time-varying AC variables among couplings exist, which is a huge disadvantage for designing the control system. Through coordinate transformation, it can be realized that converting the PWM rectifier's mathematical model in the three-phase stationary a-b-c coordinate into a mathematical model in the two-phase perpendicularly stationary α - β coordinate, or a mathematical model in the synchronous rotating d-q coordinate which synchronizes with the grid's fundamental frequency [36]. This will help transform the time-varying AC variables into rotating DC variables that are easy to be controlled. Thereby, the decoupling control on AC side is achieved.

There are two kinds of coordinate transformations: "equal power" transformation and "equal amount" transformation. In this thesis, the "equal power" transformation principle is applied. In order to be in the two-phase synchronous rotating d-q coordinate system, it should firstly transform the three-phase stationary a-b-c coordinate into a two-phase perpendicularly stationary α - β coordinate system.

Assuming that axis α coincides with axis a , axis β leads axis a by phase angle 90° . The position relationship between the two coordinate systems is shown in Figure 3.10.

The common vector X is chosen to represent the symmetrical three-phase AC variables (voltage or current), and it considers the three-phase AC variables' frequency to be ω .

Figure 3.10 Coordinate (a, b, c) and Coordinate (α , β)

For the common vector X , the components in the two coordinates have the following relations:

$$\begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \quad (3.17)$$

or

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & 0 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} \quad (3.18)$$

In (3.17) and (3.18), the following conditions are for the three-phase VSR:

$$\begin{cases} x_k \in \{e_k, i_k, s_k\} (k = a, b, c) \\ x_l \in \{e_l, i_l, s_l\} (l = \alpha, \beta) \end{cases}$$

Substituting (3.8) into (3.6), it can be achieved that the three-phase voltage source rectifier's mathematical model in the two-phase stationary α - β coordinate system:

$$\begin{cases} L \frac{di_\alpha}{dt} = e_\alpha - Ri_\alpha - V_{dc}s_\alpha \\ L \frac{di_\beta}{dt} = e_\beta - Ri_\beta - V_{dc}s_\beta \\ C \frac{dV_{dc}}{dt} = i_\alpha s_\alpha + i_\beta s_\beta - \dot{i}_L \end{cases} \quad (3.19)$$

The differential equations can be transformed into a diagram, in which the inputs are e_α , e_β , s_α , s_β , outputs are V_{dc} , and intermediate variables are i_α , i_β , i_{dc} , i_L .

Applying Laplace transformation with (3.19), it can be achieved that:

$$\begin{cases} sLI_\alpha(s) = e_\alpha(s) - RI_\alpha(s) - V_{dc}(s)s_\alpha \\ sLI_\beta(s) = e_\beta(s) - RI_\beta(s) - V_{dc}(s)s_\beta \\ sCV_{dc}(s) = I_\alpha(s)s_\alpha + I_\beta(s)s_\beta - I_L(s) \end{cases} \quad (3.20)$$

Converting (3.20) into a diagram, it can be shown in Figure 3.11.

Figure 3.11 The Diagram of the Rectifier's Model in Two-Phase Stationary α - β Coordinate

3.2.3 The Mathematical Model of VSR in the Two-Phase Synchronous Rotating d-q Coordinate System

After the 3s/2s transformation, the general vector X is rotating with angular frequency ω in a two-phase perpendicular stationary α - β coordinate system. Then it has to do the second type transformation, which transforms the general vector X into the synchronous rotating d-q coordinate system with angular frequency ω . In this case, the vector X is relatively stationary

with respect to the d-q coordinate system, and the components in the d-q coordinate axes are stationary DC components. Based on the instantaneous reactive power theory, when describing the three-phase power and in order to simplify the analysis, it usually makes the d axis's direction in the two-phase rotating d-q coordinate system the same with that of the grid voltage vector E , namely d-axis is coaxial with the vector E . The d-axis component represents the active component, and the q-axis component represents the reactive component. In addition, the angle between the d-axis and α -axis can be set to be φ_0 on the initial conditions, which is shown in Figure 3.12.

Figure 3.12 The Coordinate Transformation

The components of general vector X in the d-q coordinate and α - β coordinate meet the following relations:

$$\begin{bmatrix} x_d \\ x_q \end{bmatrix} = \begin{bmatrix} \cos \psi_s & \sin \psi_s \\ -\sin \psi_s & \cos \psi_s \end{bmatrix} \begin{bmatrix} x_\alpha \\ x_\beta \end{bmatrix} \quad (3.21)$$

in which, $\psi_s = \omega t + \varphi_0$, ω is the power grid's angular frequency.

Substituting (3.21) into the three-phase VSR general mathematical model's (3.16), the three-phase VSR d-q mathematical model is obtained:

$$\begin{cases} L \frac{di_d}{dt} - \omega L i_q + R i_d = e_d - v_{dc} s_d \\ L \frac{di_q}{dt} + \omega L i_d + R i_q = e_q - v_{dc} s_q \\ C \frac{dv_{dc}}{dt} = i_q s_q + i_d s_d - i_L \end{cases} \quad (3.22)$$

in which, x_d and x_q are the d and q components separately of x_k , ($k=a,b,c$) ($x=i, e, s$).

Defining $V_d = V_{dc} S_d$ ----- the component on the axis d of three-phase VSR's AC side bridge arm voltage vector V_{dc} .

Defining $V_q = V_{dc} S_q$ ----- the component on the axis q of three-phase VSR's AC side bridge arm voltage vector V_{dc} .

Then it can be achieved as:

$$\begin{cases} L \frac{di_d}{dt} - \omega L i_q + R i_d = e_d - v_d \\ L \frac{di_q}{dt} + \omega L i_d + R i_q = e_q - v_q \end{cases} \quad (3.23)$$

The DC side is

$$C \frac{dv_{dc}}{dt} = i_q s_q + i_d s_d - \frac{V_{dc}}{R_L} \quad (3.24)$$

The rectifier's mathematical model in the d-q synchronous rotating coordinate can be demonstrated in a diagram as Figure 3.13.

Figure 3.13 The Diagram of the Rectifier's Mathematical Model in the d-q Synchronous Rotating Coordinate

3.3 Chapter Summary

In this chapter, firstly an overview of the basic principles of the PWM rectifier is presented. Secondly, it demonstrates and analyses the PWM rectifier topology. Finally, it describes the process of establishing the PWM rectifier's mathematical model in detail. This

chapter has made the foundation of the new control strategy which is known as power ripple minimization.

In the next chapter, it will present the DPC PWM rectifier based on the power ripple minimization at length. It is a continuous study on AC/DC converter in one generator and one motor marine ship's propulsion system after Chapter 3.

Chapter 4

The DPC PWM Rectifier Based on the Power Ripple Minimization

Based on the proposed DPC power ripple minimization control strategy in a control period, a zero vector and a non-zero vector are applied, and at the same time it needs to optimize the duration time of the non-zero vector.[37] The specific implementation includes two parts: vector selection and vector acting time, which are specifically demonstrated below.

4.1 Vector Selection Based on the Instantaneous Power Theory

The vector selection table in this thesis is based on the instantaneous power theory and the power model. Firstly, according to the instantaneous power theory, this section analyzes the effects of various non-zero voltage vectors on active power and reactive power, and then selects the appropriate non-zero vector to regulate active and reactive powers; while the zero vector is selected based on the principle of the minimum number of switching times between selected non-zero vector and zero one.

4.1.1 The Instantaneous Power Theory

(1) The instantaneous power which is calculated in the three-phase a-b-c coordinate[38]:

In the three -phase a-b-c coordinate,
the instantaneous voltage vector V_{abc} :

$$V_{abc} = [v_a \quad v_b \quad v_c]^T \quad (4.1)$$

$$|V_{abc}| = \sqrt{v_a^2 + v_b^2 + v_c^2} \quad (4.2)$$

and the instantaneous current vector I_{abc} :

$$I_{abc} = [i_a \quad i_b \quad i_c]^T \quad (4.3)$$

$$|I_{abc}| = \sqrt{i_a^2 + i_b^2 + i_c^2} \quad (4.4)$$

The instantaneous voltage vector V_{abc} and current vector I_{abc} in three-phase a-b-c coordinate are shown in Figure 4.1.

Figure 4.1 The Instantaneous Voltage and Current Vectors in the Three-phase a-b-c Coordinate

The instantaneous active power is defined as: the scalar product of the instantaneous phase current vector and the instantaneous phase voltage vector, which is expressed as:

$$p = V_{abc} \bullet I_{abc} = v_a i_a + v_b i_b + v_c i_c = |V_{abc}| |I_{abc}| \cos \varphi = |V_{abc}| |i_d| \quad (4.5)$$

The instantaneous reactive power is defined as: the vector product of the instantaneous phase current vector and the instantaneous phase voltage vector, which is expressed as:

$$q = V_{abc} \times I_{abc} = v_a^* i_a + v_b^* i_b + v_c^* i_c = |V_{abc}| |I_{abc}| \sin \varphi = |V_{abc}| |i_q| \quad (4.6)$$

The instantaneous power factor is defined as:

$$\lambda = \frac{p}{\sqrt{p^2 + q^2}} \quad (4.7)$$

which can also be expressed as $\lambda = \cos \varphi$, in which φ is the angle between the voltage vector and current vector in Figure 4.1.

According to (4.6) and (4.7), the instantaneous active power p and the instantaneous reactive power q in vector form in the a-b-c stationary coordinate can be expressed as following:

$$\begin{bmatrix} p \\ q \end{bmatrix} = \begin{bmatrix} v_a & v_b & v_c \\ v_a^* & v_b^* & v_c^* \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (4.8)$$

in which v_a^*, v_b^*, v_c^* can be achieved from (4.9) :

$$\begin{bmatrix} v_a^* \\ v_b^* \\ v_c^* \end{bmatrix} = \frac{1}{\sqrt{3}} \begin{bmatrix} v_b - v_c \\ v_c - v_a \\ v_a - v_b \end{bmatrix} = \frac{1}{\sqrt{3}} \begin{bmatrix} v_{bc} \\ v_{ca} \\ v_{ab} \end{bmatrix} \quad (4.9)$$

(2) The instantaneous power calculated in the stationary two-phase α - β coordinate

If the transformation matrix from stationary three-phase a-b-c coordinate to stationary two-phase α - β coordinate based on the equal power transformation theory is defined as $T_{abc/\alpha\beta}$, then:

$$T_{abc/\alpha\beta} = \sqrt{\frac{2}{3}} \begin{pmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{pmatrix} \quad (4.10)$$

The instantaneous voltage and current vectors $V_{\alpha\beta}$ 、 $I_{\alpha\beta}$ in the stationary two-phased α - β coordinate are expressed as:

$$V_{\alpha\beta} = \begin{bmatrix} v_{\alpha} \\ v_{\beta} \\ 0 \end{bmatrix} \quad I_{\alpha\beta} = \begin{bmatrix} i_{\alpha} \\ i_{\beta} \\ 0 \end{bmatrix} \quad (4.11)$$

Then, the power equations in the stationary two-phased α - β coordinate are:

$$p = V_{\alpha\beta} \bullet I_{\alpha\beta} = v_{\alpha} i_{\alpha} + v_{\beta} i_{\beta} \quad (4.12)$$

$$q = |V_{\alpha\beta} \times I_{\alpha\beta}| = |(v_{\alpha} i_{\beta} - v_{\beta} i_{\alpha})| \quad (4.13)$$

In conclusion, the instantaneous power can be expressed in matrix as:

$$\begin{bmatrix} p \\ q \end{bmatrix} = \begin{pmatrix} v_{\alpha} & v_{\beta} \\ -v_{\beta} & v_{\alpha} \end{pmatrix} \begin{bmatrix} i_{\alpha} \\ i_{\beta} \end{bmatrix} \quad (4.14)$$

(3)The instantaneous power calculated in the rotating coordinate.

If the transformation matrix from stationary three-phase a-b-c coordinate to rotating two-phase d-q coordinate is defined as $T_{abc/dq}$, then:

$$T_{abc/dq} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos \omega t & \cos\left(\omega t - \frac{2}{3}\pi\right) & \cos\left(\omega t + \frac{2}{3}\pi\right) \\ \sin \omega t & \sin\left(\omega t - \frac{2}{3}\pi\right) & \sin\left(\omega t + \frac{2}{3}\pi\right) \end{bmatrix} \quad (4.15)$$

The instantaneous voltage and current vectors V_{dq} and I_{dq} in the rotating two-phased d-q coordinate are expressed as:

$$V_{dq} = \begin{bmatrix} v_d \\ v_q \end{bmatrix} \quad I_{dq} = \begin{bmatrix} i_d \\ i_q \end{bmatrix} \quad (4.16)$$

So the instantaneous power calculated in the d-q coordinate can be expressed as:

$$p = V_{dq} \bullet I_{dq} = v_d i_d + v_q i_q \quad (4.17)$$

$$q = |V_{dq} \times I_{dq}| = v_q i_d - v_d i_q \quad (4.18)$$

After the $T_{abc/dq}$ transformation, it can achieve:

$$\begin{cases} v_d = \sqrt{3/2} U_m \\ v_q = 0 \end{cases} \quad (4.19)$$

in which U_m is the peak value of the input AC voltage.

Thus, the instantaneous active power and reactive power can be simplified as:

$$\begin{cases} p = \sqrt{\frac{3}{2}} U_m i_d \\ q = -\sqrt{\frac{3}{2}} U_m i_q \end{cases} \quad (4.20)$$

According to the instantaneous power theory, the value of the instantaneous power can be estimated in different coordinate systems. Although in different coordinates, the expressions of instantaneous power are different, the values are the same. So you can select the appropriate coordinate system to calculate the instantaneous power, which increases the flexibility of the control system design.

4.1.2 The Rectifier Power Model

The power equations of active power and reactive power in VSR are expressed as [39]:

$$\begin{cases} L \frac{dp}{dt} = 1.5 U_m^2 - R p - \omega L q - \sqrt{\frac{3}{2}} U_m s_d v_{dc} \\ L \frac{dq}{dt} = -R q + \omega L p + \sqrt{\frac{3}{2}} U_m s_q v_{dc} \end{cases} \quad (4.21)$$

Since $p_{rd} = \sqrt{\frac{3}{2}} U_m u_{rd}$ and $q_{rq} = \sqrt{\frac{3}{2}} U_m u_{rq}$, then (4.21) can be transformed into:

$$\begin{cases} L \frac{dp}{dt} = 1.5 U_m^2 - R p - \omega L q - p_{rd} \\ L \frac{dq}{dt} = -R q + \omega L p + q_{rq} \end{cases} \quad (4.22)$$

If neglecting the resistances on the AC source side, namely R in (4.22), it can be

obtained that the change rates of active power p and reactive power q as following:

$$\begin{cases} \frac{dp}{dt} = (1.5U_m^2 - \omega Lq - p_{rd})/L \\ \frac{dq}{dt} = (\omega Lp + q_{rq})/L \end{cases} \quad (4.23)$$

4.1.3 The Divided Sectors

Three-phase PWM rectifier can output 8 kinds of switching states, which is in accordance with eight voltage space vectors, namely $V_0, V_1, \dots, V_7$, in which $V_1, \dots, V_6$ are non-zero vectors and V_0, V_7 are zero vectors. The value of voltage vector can be decided by s_a, s_b, s_c and V_{dc} , namely $s_a s_b s_c = 000 \sim 111$ corresponding to $V_0 \sim V_7$ [$V_0(000), V_1(100), V_2(110), V_3(010), V_4(011), V_5(001), V_6(101), V_7(111)$]. The status of the switch is not only decided by the output of power hysteresis comparator, but also the spatial sector position of the grid voltage vector. Thus DPC control method requires the real-time location of the grid electromotive force vector space sector, to determine the sector where the present grid electromotive force vector is.

In this thesis, the division method of 6 sectors is applied. The six non-zero voltage space vectors of the three-phase PWM rectifier output divide the $\alpha\beta$ plane into six separate vector sectors, namely sector I to sector VI, which can be shown in Figure 4.2. This division method forms a regular hexagon, which consists of six non-zero voltage space vectors ($V_k, k=1,2,\dots,6$) and two zero voltage space vectors (V_0, V_7).

Figure 4.2 Sectors of the DPC PWM Rectifier

4.1.4 The None-zero Vectors' Effects on the Power

Because each voltage vector ($V_0 \sim V_7$) of the PWM rectifier has different effects on the instantaneous active power and reactive power, it is necessary to regulate the output of instantaneous active power and reactive power by choosing the optimal voltage vector. In the vector selection table, the required voltage vector V_r is determined by the hysteresis comparator output and the sector of grid voltage vector.

Based on the VSR mathematical model and neglecting the R , the following can be achieved:

$$I = I(0) + \frac{1}{L} \int_0^t (E - V_r) dt \quad (4.24)$$

in which V_r is the output voltage vector of the VSR.

Assuming the acting time of voltage vector is T , and current vector at the time before zero is I , at the same time the reference current vector is I_{ref} , so it can transform the formula (4.24) into:

$$I_{ref} = I + \frac{1}{L} \int_0^T (E - V_r) dt = I + \frac{T}{L} (E - V_r) \quad (4.25)$$

The (4.25) shows that in order to make the current vector I track the reference current vector I_{ref} , it has to make the VSR output the appropriate voltage vector V_r , which can force the current vector I track the reference current vector I_{ref} in the direction of vector $E - V_r$.

Assuming that the grid electromotive vector is in the VI sector, which can be shown in Figure 4.3, I_{ref} is the given current vector corresponding to p_{ref} , q_{ref} . If I is less than I_{ref} , then $p_{ref} > p$ and $q_{ref} < q$. In this case, it should select vector V_5 (001) that is the output of VSR, so that the current vector I can approach the given one I_{ref} at the direction of vector $E - V_r$, which will lead to p approaching p_{ref} and q approaching q_{ref} .

In the perspective of the relationships between each voltage vector, it can make a further analysis of the non-zero vectors' effects on instantaneous active and reactive power. As is known to all, d-axis current is active current component, and q-axis current is reactive current component. Thus, it can set the d-axis in the d-q rotating coordinate system to coincide with the position of the vector E . The established coordinate is shown in Figure 4.3.

Figure 4.3 Voltage Vector Tracing and Selection of VSR Space Vector V_r

Assuming that the non-zero vector is V_r (r is the vector rank number) in the control period, then the projection's length and direction of $E - V_r$ on the d-q coordinate determine the increments' value and direction of the system's active and reactive power. Based on the active power equation in (4.20), it can be learned that the projection of $E - V_r$ on the d axis determines the active power's increment's value and its direction is the same with the increment's; while from the reactive power equation in (4.20), it can be derived that the projection of $E - V_r$ on the q axis determines the reactive power's increment's value and its direction is opposite with the increment's.

Taking the effects of vectors V_1 、 V_3 、 V_5 、 V_6 as examples, Figure 4.4 shows their effects on instantaneous power separately. The projection's direction of $E - V_1$ on the d axis is negative, which means the system's incremental active power is decreasing; while the projection's direction of $E - V_1$ on the q axis is negative, which means the system's incremental reactive power is increasing. The projection's direction of $E - V_3$ on the d axis is positive, which means the system's incremental active power is increasing; while the projection's direction of $E - V_3$ on the q axis is negative, which means the system's incremental reactive power is increasing. The projection's direction of $E - V_5$ on the d axis is positive, which means the system's incremental active power is increasing; while the projection's direction of $E - V_5$ on the q axis is positive, which means the system's incremental reactive power is decreasing. The projection's direction of $E - V_6$ on the d axis is

negative, which means the system's incremental active power is decreasing; while the projection's direction of $E - V_6$ on the q axis is positive, which means the system's incremental reactive power is decreasing.

Figure 4.4 Schematic Diagram of the Non-zero Vectors' Effects

In conclusion, when the grid electromotive vector E is in the sector VI, non-zero vectors' effects on the power are:

- (1) Vector V_6 can decrease active and reactive power, and the acting time determines the decrements.
- (2) Vector V_3 can increase active and reactive power, and the acting time determines the increments.
- (3) Vector V_1 can decrease active power and increase reactive power, and the acting time determines the decrement and increment.
- (4) Vector V_5 can increase active power and decrease reactive power, and the acting time determines the increment and decrement.

The above conclusions are based on the fact that the grid electromotive vector E is in the sector VI, so the conclusions can be generalized by assuming that the grid electromotive vector E is in the sector k and the followings can be summarized:

- (1) Vector V_k can decrease active and reactive power, and the acting time determines the decrements.

(2) Vector V_{k+3} can increase active and reactive power, and the acting time determines the increments.

(3) Vector V_{k+1} can decrease active power and increase reactive power, and the acting time determines the decrement and increment.

(4) Vector V_{k-1} can increase active power and decrease reactive power, and the acting time determines the increment and decrement.

In the above vectors, k is applied by cycle counting method, namely $k=\text{mod}(k,6)$.

4.1.5 The Vector Selection Table

It is intensively analyzed that the non-zero voltage vectors' effects on the active and reactive power previously. Based on the conclusion in Section 4.1.2, it can come up with the vector selection table shown in Table 4.1.

Table 4.1 Vector Selection Table

Δp	Δq	Sector Number k
Increase	Increase	V_{k+3}
Increase	Decrease	V_{k-1}
Decrease	Increase	V_{k+1}
Decrease	Decrease	V_k

In Table 4.1, $\Delta p=p-p_{ref}$, $\Delta q=-q_{ref}$, which are representing the signals of power hysteresis comparator output errors. $k=\text{mod}(k,6)$.

The non-zero vector can be selected according to Table 4.1, And in order to reduce the switching frequency and switching losses, which will improve the system's efficiency, the zero vector is selected with the principle of minimum number of switching between it and the selected non-zero vector. Because active power and reactive power are decided by the sector number k and power hysteresis comparator's output, by which the PWM rectifier can refer to the vector selection Table 4.1 to select V_r ($s_a s_b s_c$) and realize real-time adjustment. V_r has the ability to adjust the active and reactive power at the same time. Therefore, the vector selection table is the core of the DPC control system.

4.2 The Vector Acting Time Based on Power Ripple Optimization

4.2.1 The Building of Power Ripple Optimization's Objective Function

The building of the power ripple optimization's objective function is on the basis of the theory that the root mean square (RMS) value of the errors between the active and reactive power and the reference ones in a controlled period is the minimum. In this part, it will calculate the acting time of the non-zero voltage vector in a controlled period in order to achieve minimum power ripple, which is based on the three-phase PWM power model.

Assuming that the active and reactive power at time k are P_k and Q_k separately, then the non-zero vector's and zero vector's corresponding active power change rates are s_{np} and s_{zp} respectively; while the reactive power change rates are s_{nq} and s_{zq} respectively. The control goal is to minimize the RMS value of the errors between the active and reactive power and the reference ones in the next controlled period. In other words, it needs to derive the non-zero vector's optimization acting time t_{op} from making the following function $F_{ripple}(t_{op})$ output the least value:

$$F_{ripple}(t_{op}) = \frac{1}{t_{sp}} \int_0^{t_{op}} (P_k + s_{np}t - P_{ref})^2 dt + \frac{1}{t_{sp}} \int_{t_{op}}^{t_s} (P_k + s_{np}t_{op} + s_{zp}(t - t_{op}) - P_{ref})^2 dt \\ + \frac{1}{t_{sp}} \int_0^{t_{op}} (Q_k + s_{nq}t - Q_{ref})^2 dt + \frac{1}{t_{sp}} \int_{t_{op}}^{t_s} (Q_k + s_{nq}t_{op} + s_{zq}(t - t_{op}) - Q_{ref})^2 dt \quad (4.26)$$

in which P_{ref} , Q_{ref} are reference values of active and reactive power; t_{sp} is controlled sampling time period.

By calculating the derivative of $F_{ripple}(t_{op})$ with respect to t_{op} , namely:

$$\frac{dF_{ripple}(t_{op})}{dt_{op}} = 0 \quad (4.27)$$

the non-zero vector's optimization acting time t_{op} can be achieved.

4.2.2 The Result of Power Ripple Optimization's Objective Function

It can get the solution result t_{op} by numerous solving processes (the details are shown in Appendix A.1) of (4.27) and the result is shown as following:

$$t_{op} = \frac{2(P_{ref} - P_k)(s_{np} - s_{zp}) + 2(Q_{ref} - Q_k)(s_{nq} - s_{zq})}{2(s_{np}^2 + s_{zp}^2) + s_{nq}^2 + s_{zq}^2 - 3(s_{np}s_{zp} + s_{nq}s_{zq})} \\ + \frac{s_{zp}(s_{zp} - s_{np}) + s_{zq}(s_{zq} - s_{nq})}{2(s_{np}^2 + s_{zp}^2) + s_{nq}^2 + s_{zq}^2 - 3(s_{np}s_{zp} + s_{nq}s_{zq})} t_s \quad (4.28)$$

in which, s_{np} , s_{zp} , s_{nq} and s_{zq} can be achieved from (4.23). If the optimization acting time t_{op}

achieved from (4.28) is negative or larger than sampling period time t_{sp} , this is suggesting that the system is not becoming stable and in this case $t_{op}=t_{sp}$.

4.3 The Switch Table

The traditional DPC system usually utilizes the divided plane with 12 sectors.[40] This divided plane is achieved by dividing the α - β plane evenly via the VSR's output 6 non-zero vectors and 2 zero vectors and their bisectors, and then it can get the independent $\theta_1 \sim \theta_{12}$. This plane is shown in Figure 4.5(b).

(a) Three-phase VSR Voltage Space Vectors

(b) α - β Plane with 12 Sectors

Figure 4.5 The Traditional 12 Sectors' Division Method

Table 4.2 Switch Table

s_p	s_q	$s_a \quad s_b \quad s_c$											
		θ_1	θ_2	θ_3	θ_4	θ_5	θ_6	θ_7	θ_8	θ_9	θ_{10}	θ_{11}	θ_{12}
0	0	101	100	100	110	110	010	010	011	011	001	001	101
0	1	100	110	110	010	010	011	011	001	001	101	101	100
1	0	001	001	101	101	100	100	110	110	010	010	011	011
1	1	010	010	011	011	001	001	101	101	100	100	110	110

The sector divider calculates the power grid electromotive E 's phase angel θ by the equation $\theta = \arctan(e_\alpha/e_\beta)$, which e_α 、 e_β are known. Then by applying the phase angel range equation $(n-2)\pi/6 \leq \theta_n \leq (n-1)\pi/6$ ($n=1, 2, \dots, 12$), it can determine the sector in which power grid electromotive E locates. For example, $\theta = \arctan(e_\alpha/e_\beta) = -30^\circ \sim 0^\circ$, which indicates the voltage space vector E locates in the first sector θ_1 .

The PWM rectifier can select the switch state from the switch table shown in Table 4.2, in which voltage space vector $\mathbf{E}$'s sector θ and the corresponding power hysteresis comparator output s_p and s_q determine the switch state. The switch state equals the output voltage space vector and thus controls the instantaneous active and reactive power.

4.4 The Simulations and Analysis

4.4.1 The Marine Ship's Generator and PWM Rectifier Simulations

Based on the power ripple minimization strategy, the PWM rectifier DPC has been promoted into a better level. The Simulation has proved it.

Figure 4.6 shows the schematic diagram of PWM rectifier DPC based on power ripple minimization. It is applied in the marine ship's propulsion system which just includes one generator and the PWM rectifier, also known as the AC/DC rectifier part.

The system and control parameters are displayed in Table 4.3. The two simulation models share the same parameters in Table 4.3, while they have two different control methods within them.

The reference values of active power in the two models are the same and so the reference active power value can be calculated by $P_{ref} = \frac{U_{dc}^2}{R_L} = \frac{(700V)^2}{46\Omega} = 10.65kW$, and reference reactive power value is set to be $Q_{ref} = 0Var$. They are the target values for the two models' active power and reactive power.

Table 4.3 System and Control Parameters

Parameters	Signs	Values
Line resistance	R	0.001 Ω
Line inductance	L	0.0067 H
DC-Bus capacitor	C	0.0047 F
Load resistance	R_L	46 Ω
Line-line voltage(RMS)	U_N	381 V
Line voltage frequency	f	50 Hz
DC voltage	U_{dc}	700 V
Sampling period	t_{sp}	5 μs
Active power	P	10.45 kW
Reactive power	Q	0 Var

Figure 4.6 The Schematic Diagram of PWM Rectifier DPC Based On Power Ripple Minimization

Figure 4.7 The Vector Selection Model

Figure 4.7 shows the vector selection model, which is used by both the PWM rectifier DPC based on power ripple minimization and the traditional PWM rectifier DPC. While the power ripple minimization strategy has a unique model that is named as vector time model and it is shown in Figure 4.8

Figure 4.8 The Vector Time Model

Figure 4.9 Simulation Model of the PWM Rectifier DPC Based on the Power Ripple Minimization

Figure 4.10 Simulation Model of the Traditional PWM Rectifier DPC

The generator and PWM rectifier DPC based on the power ripple minimization

model is shown in Figure 4.9. U_{dc} is the rectifier's output DC voltage. P and Q are the active power and reactive power, respectively. I is the power grid current. While Figure 4.10 shows the simulation model of the traditional PWM rectifier DPC.

Figure 4.11 shows the power grid's instantaneous active power P waveform in the power ripple minimization strategy model. It fluctuates at the value of 10.65kW. Figure 4.12 shows the reactive power Q waveform and it fluctuates at the value of 0 Var. By comparison, the traditional voltage oriented DPC PWM rectifier model's active and reactive power P_o and Q_o waveforms are shown in Figure 4.13 and Figure 4.14.

Figure 4.11 Power Ripple Minimization Strategy Model's Instantaneous Active Power P Waveform

Figure 4.12 Power Ripple Minimization Strategy Model's Instantaneous Reactive Power Q Waveform

Figure 4.13 Traditional Voltage Oriented DPC Model's Instantaneous Active Power P_o Waveform

Figure 4.14 Traditional Voltage Oriented DPC Model's Instantaneous Reactive Power Q_o Waveform

The (4.29) is introduced, which is to calculate the power ripple in 0.1 s and N is the data number in 0.1s.

$$\begin{cases} P_{ripple} = \sqrt{\frac{1}{N} \sum_{i=1}^N (P_i - \frac{1}{N} \sum_{i=1}^N P_i)^2} \\ Q_{ripple} = \sqrt{\frac{1}{N} \sum_{i=1}^N (Q_i - \frac{1}{N} \sum_{i=1}^N Q_i)^2} \end{cases} \quad (4.29)$$

The results are shown in the Table 4.4.

Table 4.4 Comparisons of Simulation Results of Two Methods at Steady State Performance

	Power Ripple Minimization	Traditional DPC
P_{ripple}	163.32	181.54
Q_{ripple}	288.01	485.20

From Table 4.4, it can conclude that the active power ripple and reactive power ripple has decreased by 11.16% and 68.47% respectively, which means the power ripple minimization strategy has made a significant performance in stabilizing the power grid's power ripple.

The current harmonics are analyzed by the FFT analysis tool, and Figure 4.15 shows the simulation results of THD (Total Harmonic Distortion) analysis obtained from power ripple minimization DPC. By comparison, Figure 4.16 shows the simulation results of THD analysis obtained from traditional voltage oriented DPC. It can be learned that the power

ripple minimization method has made the grid current THD dropped from 4.93% to 3.70% and the harmonics are mainly at the frequencies: $f = (6n \pm 1) \times 50 \text{ Hz}$, ($n = 1, 2, \dots$).

(a) Grid Current Waveform

(b) Grid Current THD Analysis

Figure 4.15 Simulation Results of THD Analysis Obtained From Power Ripple Minimization DPC

(a) Grid Current Waveform

(b) Grid Current THD Analysis

Figure 4.16 Simulation Results of THD Analysis Obtained From Traditional Voltage Oriented DPC

4.4.2 The Analysis

The simulation results prove the power ripple minimization strategy can restrain the power ripple of the marine ship power grid and also decrease the current harmonics, which will enhance the power system steady state performance significantly.

Although the PWM rectifier DPC based on the power ripple minimization is only applied in this one generator and one motor marine ship's propulsion system, it will show a great deal of significant performance in the multi-generators and multi-motors power system in the marine ship.

4.5 Chapter Summary

This chapter demonstrates the design of the power ripple minimization PWM rectifier controller in detail. The vector selection based on the instantaneous power theory and the calculation of vector acting time are the two major research objects, which determine the power ripple minimization PWM rectifier controller.

The simulation results have verified that the power ripple minimization control strategy indeed diminishes the power ripple of the power system in the one generator and one motor marine ship's propulsion system. And the harmonics of the currents are also suppressed.

Next chapter will focus on the electric motor control problems of the one generator and one motor marine ship's propulsion system.

Chapter 5

DC Motor Double Closed Loop Control System

In this chapter, a DC motor works as the propulsion motor of the marine ship and a double closed loop control strategy [41], in which a Fuzzy-PID controller is applied, is also introduced to handle the marine ship's propulsion under the condition of load fluctuations.

5.1 The Double Closed Loop System Operation Principle

The speed and current double closed loop control system of the DC motor is the most typical control system, and its operation principle schematic is shown in Figure 5.1.

Figure 5.1 The Operation Principle Schematic of DC Motor Double Closed Loop Control System

ASR is the speed regulator; ACR is the current regulator; TG is the tacho-generator; TA is current transformer; GT is the trigger; V is the rectifier; M is the DC motor armature; U_n^* is the speed given voltage; U_n is the feedback speed given voltage; U_i^* is the current given voltage; U_i is the feedback current given voltage.

The characteristic of the double closed loop DC motor speed control system is that the motor's speed and current are controlled respectively by two independent regulators. The output of the speed regulator is the given input value of the current regulator, therefore the current loop can adjust the armature current along with the deviation value of the rotor speed.

When the rotor speed value is smaller than the given speed value, the ASR will produce a bigger output value which is the input of the ACR, and then current regulator will increase output armature current, finally leading that the rotor achieves acceleration and the speed increases. This process will continue work until the feedback rotor speed equivalent to the given speed value. It will work vice versa.

5.2 The Dynamic Model of the Double Closed Loop System

A few assumptions should be made before the dynamic model of the DC motor is set. So the assumptions are:

- 1) The offset of the motor is good enough;
- 2) The armature reaction, eddy current effect and the magnetic hysteresis effect are all neglected;
- 3) The excitation current is constant.

Hence, it can get the mathematical equation and the motion equation of the DC motor respectively:

$$U_d = Ri_d + L \frac{di_d}{dt} + E \quad (5.1)$$

$$T_e - T_L = \frac{GD^2}{375} \frac{dn}{dt} \quad (5.2)$$

In U_d is the armature voltage; R , L and i_d are the armature circuit's resistor, inductance and current. E is the opposing electromagnetic force and it can be calculated by $E = C_e \phi n$. T_e and T_L are the motor's electromagnetic torque and the load torque respectively, and T_e can be calculated by $T_e = C_m \phi I_d$. GD^2 is the moment of inertia value which has been transformed to the motor's rotor side.

And therefore, it can achieve the transfer function from (5.1) and (5.2) as:

$$\frac{I_d(s)}{U_d(s) - E(s)} = \frac{1/R}{T_l s + 1} \quad (5.3)$$

$$\frac{E(s)}{I_d(s) - I_{dl}(s)} = \frac{R}{T_m s} \quad (5.4)$$

where $T_l = L/R$ is electromagnetic time constant of the armature circuit(s); $I_{dl} = T_l/C_m$ is load current (A); T_m is electromechanical time constant of the marine ship's electric motor system.

Because $n = \frac{E}{C_e \phi}$, it can get the dynamic structure of the DC motor shown in Figure 5.2.

Figure 5.2 The Dynamic Structure Schematic of the DC Motor

5.3 The Design of the Regulators

5.3.1 The Design of Current Regulator

In Figure 5.2, there are cross connections between feedback opposing electromotive force E and the feedback current effect U_i , which makes the design very complicated. In most cases, the electromagnetic time constant T_l is much smaller than the electromechanical time constant T_m , so the speed's changing is much slower than the current's changing. Then E is a slower changing disturbance value for the current loop, so it can assume that the opposing electromotive force E is unchangeable, also can be expressed as $\Delta E \approx 0$, during the current's instantaneous changing time interval. So the appropriate equivalent structure schematic is shown in Figure 5.3.

Hence, it can neglect the feedback opposing electromotive force loop's effect on the current loop on the condition that:

$$\omega_{ci} \geq \sqrt{\frac{1}{T_m T_l}} \quad (5.5)$$

where, ω_{ci} is cut-off frequency of the current loop's open-loop frequency characteristics.

If transferring the given filtering loop and the feedback filtering loop into the whole loop, as well as making the given input as $U_i^*(s)/\beta$, then the current loop becomes the unit negative feedback system. This is shown in Figure 5.4.

Finally, since T_{oi} and T_s are both much smaller than T_l , so they can be treated as one single inertia element and the time constant is:

$$T_{\sum i} = T_s + T_{oi} \quad (5.6)$$

And the ultimate equivalent structure is show in Figure 5.5. The equivalence's condition is:

$$\omega_{ci} \leq 3 \sqrt{\frac{1}{T_s T_{oi}}} \quad (5.7)$$

Figure 5.3 The Equivalent Structure Schematic of the Current Loop

Figure 5.4 The Current Loop Structure Schematic

Figure 5.5 The Ultimate Equivalent Structure Schematic of the Current Loop

The current loop is chosen as the typical system which is featuring in the following function. And the transfer function is:

$$W_{ACR}(s) = \frac{K_i(\tau_i s + 1)}{\tau_i s} \quad (5.8)$$

where, K_i is the proportional coefficient of the current regulator; τ_i is leading time constant.

In order to offset the regulator zero point and the controlled object's time constant polar point, it usually chooses:

$$\tau_i = T_l \quad (5.9)$$

$$K_i = \frac{K_i K_s \beta}{\tau_i R} \quad (5.10)$$

The current regulator's parameters are K_i and τ_i , in which τ_i has been determined, so proportional coefficient K_i is the only one to be confirmed. Generally speaking, the overshoot value is expected to be $\delta_i \leq 5\%$, so it can refer to the chart and get: $\xi = 0.707$ and $K_i T_{\Sigma i} = 0.5$.

$$K_i = \omega_{ci} = \frac{1}{2T_{\Sigma i}} \quad (5.11)$$

$$K_i = \frac{T_l R}{2K_s \beta T_{\Sigma i}} = \frac{R}{2K_s \beta} \left(\frac{T_l}{T_{\Sigma i}} \right) \quad (5.12)$$

5.3.2 The Design of Speed Regulator

The current loop's closed-loop transfer function $W_{cli}(s)$ can be derived from Figure 5.4 and expressed as:

$$\begin{aligned}
W_{cli}(s) &= \frac{I_d(s)}{U_i^*(s)/\beta} \\
&= \frac{\frac{K_I}{s(T_{\sum i}s+1)}}{1 + \frac{K_I}{s(T_{\sum i}s+1)}} \\
&= \frac{1}{\frac{T_{\sum i}}{K_I}s^2 + \frac{1}{K_I}s + 1}
\end{aligned} \tag{5.13}$$

Ignoring the high order term, the $W_{cli}(s)$ can be simplified as

$$W_{cli}(s) \approx \frac{1}{\frac{1}{K_I}s + 1} \tag{5.14}$$

on condition that:

$$W_{cn} \leq \frac{1}{3} \sqrt{\frac{K_I}{T_{\sum i}}} \tag{5.15}$$

in which, W_{cn} is the cut-off frequency of the speed loop's open loop frequency characteristic.

The current loop can be transformed as following in the speed loop:

$$\frac{I_d(s)}{U_i^*(s)} = \frac{W_{cli}(s)}{\beta} \approx \frac{\frac{1}{\beta}}{\frac{1}{K_I}s + 1} \tag{5.16}$$

It can replace the current loop in Figure 5.2 with the ultimate equivalent current loop in Figure 5.5. Then the whole dynamic structure of the speed regulator is shown in Figure 5.6.

Figure 5.6 The Dynamic Structure Schematic of the Speed Regulator

It can move the speed given filtering box and the speed feedback filtering box into the loop and should also change the given input value into $U_n^*(s)/\alpha$, as well as merging the time constants $1/K_I$ and T_{on} as one single constant $T_{\Sigma n}$, then:

$$T_{\Sigma n} = \frac{1}{K_I} + T_{on} \quad (5.17)$$

Then the speed loop can be simplified as shown in Figure 5.7.

The ASR transfer function is:

$$W_{ASR} = \frac{K_n(\tau_n s + 1)}{\tau_n s} \quad (5.18)$$

where, K_n is the proportional coefficient of the speed regulator; τ_n is the leading time constant.

Therefore, the open loop transfer function of the speed regulator system is:

$$W_n(s) = \frac{K_n \alpha R (\tau_n s + 1)}{\tau_n \beta C_e \phi T_m s^2 (T_{\Sigma n} s + 1)} \quad (5.19)$$

Making the open loop gain of the speed loop as:

$$K_N = \frac{K_n \alpha R}{\tau_n \beta C_e T_m} \quad (5.20)$$

So finally,

$$W_n(s) = \frac{K_N (\tau_n s + 1)}{s^2 (T_{\Sigma n} s + 1)} \quad (5.21)$$

Figure 5.7 The Simplified Schematic of the Speed Loop

The ultimate equivalent structure of the speed regulator system is shown in Figure 5.8.

Figure 5.8 The Ultimate Equivalent Structure Schematic of the Speed Loop

5.4 The Fuzzy-PID Controller

5.4.1 The Theory of PID Control

PID control is a control strategy which is a combination of proportional, integral and derivative. The PID controller is the application of the PID control method. In the industrial field, PID controller has shown magnificent performance in most of the cases. Thus, the PID controller is a conventional and common controller in the industrial sector.

In the analog control system, it is proportional, integral and derivative that are the most favorable control methods. The structure diagram of analog PID control is shown in Figure 5.9.

PID controller is a kind of linear controllers. The difference of system given value $R(t)$ and output value $Y(t)$ determines the system's output error value $E(t)$.

$$E(t) = R(t) - Y(t) \quad (5.22)$$

The PID control strategy can be expressed as following:

$$u(t) = K_p \left[E(t) + \frac{1}{T_i} \int_0^t E(\tau) d\tau + T_d \frac{d}{dt} E(t) \right] \quad (5.23)$$

With the Laplace transformation, it can transform (5.23) into:

$$G(s) = \frac{U(s)}{E(s)} = K_p \left(1 + \frac{1}{T_i s} + T_d s \right) \quad (5.24)$$

Figure 5.9 The Structure Diagram of the PID Controller

In (5.24), K_p is the coefficient of the proportion, T_i is the integration time constant and T_d is the differential time constant.

In conclusion, the proportional, integral and derivative have the following effects respectively.

(1) The proportional control block: it is proportional to the error $E(t)$, and when there is an error $E(t)$ input, the controller will have an instant output which diminishes the error.

(2) The integral control block: it is used to eliminate the steady state error and make the system approach the expected operation state. The integration time constant T_i determines the integral control's performance, which will have a significant effect when T_i is a small value, and will have a little effect when T_i is a large value.

(3) The derivative control block: it reflects the change speed rate of error $E(t)$. The system will bring in a correct signal when the change speed is getting high, which will quicken the system's control speed and reduce the regulation time.

The algorithm of the PID controller is simple and robust, and it has no static error in the steady state. Therefore, 90% of the controllers in the industrial sector are applied with the PID controller, which has dominated the controller field.

With the improvements of industrial automation, more complex control systems are emerging. These complex systems have the following features: big lagging, nonlinear, time-varying, parameters unknown and random interference, etc. These factors make it very difficult to establish the mathematical models, which is the fundamental of the PID control.

Thus, the disadvantages of the PID control are getting obvious. So the new control strategies which meet industrial production quality control requirements are to be introduced.

5.4.2 The Theory of Fuzzy Control

The fuzzy control technology [42] is an advanced smart control strategy, which combines the fuzzy mathematics that works as the fundamental theory, and the computer as well as the classical control technologies. The fuzzy control technology is one of the most widely applied smart control technologies and shares a lot of commons with the artificial intelligence. They both have the three factors of inference, learning and associating.

The fuzzy control takes the same structure as the conventional digital control system, and the fuzzy control's diagram is shown in Figure 5.10.[43]

Figure 5.10 Diagram of the Fuzzy Control System

The main difference between the fuzzy control strategy and the conventional control strategy is the controller. The inner structure of the fuzzy controller is shown in Figure 5.11.

The fuzzy controller's operation theory can be described as: the system achieves the controlled variable by sampling, then does the calculation with the given value and gets an error signal. After the A/D transformation, an accurate digital number is achieved and it should be fuzzified with the fuzzy controller. Then the fuzzified value is used to act on the controlled object, and usually a professional operator can decide to operate based on his or her abundant experience. A few rules have been summarized by the operators' and experts' knowledge and experience to make it applied easily in practical situation. The fuzzified value utilizes these rules and makes fuzzy inference. Then by de-fuzzifying and D/A processing, finally the value which can act on the controlled object is obtained.

Figure 5.11 Schematic of Fuzzy Control

5.4.3 The Design of Fuzzy-PID Controller

Figure 5.12 Design Diagram of Fuzzy Logic Controller

Figure 5.12 shows a two-dimension fuzzy controller, whose inputs are error $e(t)$ and error change rate $ec(t)$, and its output is u . And their fuzzy sets are E , EC and U respectively.[44-47]

The fuzzy subset and the domain of E , EC and U are expressed as following:

the fuzzy subset is $\{NB, NM, NS, ZO, PS, PM, PB\}$;

and the domain is $\{-4, -3, -2, -1, 0, 1, 2, 3, 4\}$.

What's more, the membership function's types of the E and EC are set as Gauss function distribution curve, and the membership function's type of U is set as Triangle function distribution curve. And the membership functions of E , EC and U are shown in Figure 5.13.

The fuzzy control rules are summarized based on previous experience and knowledge. It is shown in Table 5.1.

Table 5.1 Fuzzy Control Rules

U		E						
		NB	NM	NS	ZO	PS	PM	PB
EC	NB	NB	NB	NM	NM	NS	ZO	PS
	NM	NB	NB	NM	NS	ZO	PS	PM
	NS	NB	NM	NS	NS	ZO	PS	PM
	ZO	NB	NM	NS	ZO	PS	PM	PB
	PS	NB	NS	ZO	PS	PS	PM	PB
	PM	NB	NS	ZO	PS	PM	PM	PB
	PB	NB	NS	PS	PM	PM	PB	PB

From Table 5.1, it can be derived that the fuzzy inference as the format of if ... then...

If E is NB and EC is NB, then U is NB;

If E is NM and EC is NB, then U is NB;

If E is NS and EC is NB, then U is NM;

...

...

If E is PB and EC is PB, then U is PB;

In total, there are 49 phrases in Table 5.1.

5.5 The Simulink Simulations

The simulation model of the DC motor double closed loop control system is demonstrated in Figure 5.13. Figure 5.14 is the conventional PID method, while Figure 5.15 shows the Fuzzy-PID control method. The same parameters are given to the two methods' simulations as following described. Due to the confidential policy of Rolls-Royce Corporate @NTU Lab, the detailed parameters can't be disclosed.

The given rotor speed is 800 rpm, and the load changes at the time $t=3s$ from $50 \text{ N}\cdot\text{m}$ to $150 \text{ N}\cdot\text{m}$. The simulation time span is 6 seconds. Therefore, it can achieve the two methods' rotor speed waveforms shown in Figure 5.16.

Figure 5.13 Simulation Model of the DC Motor Double Closed Loop Control System

Figure 5.14 Conventional PID Control Method

Figure 5.15 Fuzzy-PID Control Method

Figure 5.16 The Motor's Rotor Speed Waveform

From Figure 5.16, it can be derived that the Fuzzy-PID method can significantly make the overshoot down and nearly eliminate it. When the load changes at time $t=3s$, the PID control method experiences a fast speed drop, while the Fuzzy-PID control method make the rotor speed approximately stabilized at the target speed.

5.6 Chapter Summary

This chapter mainly presents the DC motor control strategy in the one generator and one motor marine ship's propulsion system. Based on the simulation results, the Fuzzy-PID controller has shown its superiority over the conventional PID control in the DC motor double closed loop control system. It is important that the marine ship cruises at a stable speed on the sea where there are many ocean currents and tides affecting the propeller.

The initial goal that the propeller speed is stable is realized. However, there are still some existing problems shown in the simulation. Taken as an example, the Fuzzy-PID control shows great performance in stabilizing the rotor's speed in transient period. While at steady state, the motor speed has little pulses or ripples.

Next chapter will focus on design of an energy storage system which can stabilize the DC bus voltage between the AC/DC converter and the DC/AC inverter. It also can store and release power energy when it is necessary.

Chapter 6

The Energy Storage System Based on a Reverse Parallel Boost-Buck Voltage Stabilizer

Because the marine ship's load fluctuates with time and it will have a backlash effect on the generator system, and it causes the voltage on the DC bus unstable. In order to diminish the DC Voltage ripple as well as save the energy, an energy storage system based on a reverse parallel boost-buck voltage stabilizer in which a super capacitor works as the energy storage device [48, 49] is designed. The block diagram is illustrated in Figure 6.1.

6.1 The Design of a Reverse Parallel Boost-Buck Voltage Stabilizer

As for the voltage disturbance in the one generator and one motor marine ship's propulsion system, this chapter introduces an energy storage system based on a reverse parallel boost-buck voltage stabilizer and it also works as a passage between the marine ship's system and the energy storage. The energy storage system's operation theory is: when the marine ship's load is increasing and it will lead to the decrease of the DC link voltage, the super capacitor energy storage will release the energy within it through the reverse parallel Boost-Bust system and the DC link voltage will be boosted to the rated level; while when the marine ship's load is decreasing and it will make the DC link voltage increase, the super capacitor energy storage will absorb the energy from the DC bus through the reverse parallel boost-bust system and the DC link voltage will be depressed to the rated level.

Figure 6.1 The Energy System in the One Generator and One Motor Marine Ship's Propulsion System

This chapter establishes the Boost-Buck circuit state space equation. For the different operating conditions of DC bus voltage, a reverse parallel dual-passage Boost-Buck circuits'

system with PI closed loop control strategy is designed. Finally, the Simulink simulation is utilized to test the energy storage system. Figure 6.2 shows the diagram of the marine ship's super capacitor energy storage system based on the reverse parallel Boost-Buck circuits.

Figure 6.2 The Diagram of Super Capacitor Energy Storage System Based on the Reverse Parallel Boost-Buck Circuits

6.2 Super Capacitor Energy Storage System

The super capacitor is a new type of energy storage device in addition to the traditional capacitor and the chargeable battery. When it compares with the traditional capacitor, the super capacitor possesses a larger capacity, a higher energy unit and a wider range of operation temperature as well as a longer lifespan. When it comes to compare with the battery, the super capacitor holds much more power and is environment-friendly. Therefore, the super capacitor is an efficient and environmentally friendly energy storage device, and it is very suitable for the application in the marine ship's power system to replace the batteries which have the shortest life expectancy in the power system. The super capacitors used in the marine ship's power system will be of crucial importance to the whole marine ship's power grid, and will solve some of the problems caused by other energy storage devices.

The diagram of the marine ship's power system based on the energy storage system in which a super capacitor works as the storage unit is shown in Figure 6.3. [50] The DC bus works as a stable DC voltage station. When the DC bus voltage decreases, the super capacitor energy storage will release its energy via DC/DC converter into the DC bus until the bus's voltage returns to its rated value. While, when the DC bus voltage rises, the bus's energy will pour into super capacitor energy storage via DC/DC converter until the bus's voltage reduces to its rated value.

Because of the load changing, in order to maintain the stable operation of the marine ship's power system, this thesis selects the relatively high-performance super capacitor working as the power energy storage core body. However, the voltage of the super capacitor energy storage changes fast during capacitor's charging and discharging operations. Thus, the design of a bi-directional DC/DC converter, in which a two-passage Boost-Buck circuit system with a wide working range operates, play a crucial role in keeping the DC bus voltage stable and constant[51-53].

Figure 6.3 Diagram of the Super Capacitor Energy Storage System in the Marine Ship's Power System

6.3 The Voltage Stabilizer Based on the Reverse Parallel Boost-Buck Circuits

In order to achieve the energy can flow bi-directionally between the marine ship's DC bus and the super capacitor, two Boost-Buck circuits which are connected by the means of reverse parallel. It can be shown in Figure 6.2.

6.3.1 The Operation Principle and State Space Equations of the Boost-Buck Circuit

Figure 6.4 The Diagram of the Boost-Buck Circuit

The Boost-Buck chopper circuit is also named as the reverse polarity chopper circuit, whose output voltage U_o can be higher as well as lower than the input voltage U_d , and it is a DC converter with a single none-isolated thyristor.

The main circuit of the Boost-Buck is composed of the thyristor, diode, inductor and capacitor, the same components as the Boost circuit and Buck circuit have. The diagram of the Boost-Buck circuit is shown in Figure 6.4. What makes the Boost-Buck circuit different from the Boost and Buck chopper circuits is that the output voltage polarity is reversed compared with the input voltage polarity.

The equivalent schematic diagrams of the Boost-Buck circuit when it is under operation state are shown in Figure 6.5.

Figure 6.5 has shown the equivalent schematic diagrams when the VT is in the states of on and off. From Figure 6.5(a), it can be learned that the DC voltage source charges the energy storage component inductor L when the VT is on, the current flowing in the inductor L is increasing gradually. By comparison, when the VT turns off, the inductor L discharges and the current flowing in the inductor L is decreasing gradually.

When $t=0$, VT is turned on, the DC voltage source U_d is totally applied on the boost

(a) The Equivalent Schematic Diagram When VT on

(b) The Equivalent Schematic Diagram When VT off

Figure 6.5 The Equivalent Schematic Diagrams of the Boost-Buck Circuit

inductor L and the inductor current i_L is increasing linearly. At this time, the diode D closes and the load is powered by filter capacitor C .

$$L \frac{di_L}{dt} = U_d \quad (6.1)$$

At the very moment of $t = t_{on}$, i_L reaches the maximum value $i_{L\max}$. During the time of VT is on, the $\Delta i_{L(+)}$ is:

$$\Delta i_{L(+)} = \frac{U_d}{L} \times T_{on} = \frac{U_d}{L} \times D_y \times T_s \quad (6.2)$$

When $t = t_{on}$, VT will be turned off, i_L will be freewheeling through diode D . The energy stored in the inductor L will flow into the load resistor R and the power accumulator

capacitor C , and charge the capacitor. At this moment, the voltage applied on the inductor L is $-U_o$. Thus, the i_L will decrease linearly.

$$L \frac{di_L}{dt} = -U_o \quad (6.3)$$

When $t = t_s$, i_L will reach its minimum value $i_{L\min}$. During the period of the VT off, the decrement of i_L , namely $\Delta i_{L(-)}$ is:

$$\Delta i_{L(-)} = \frac{U_o}{L} \times (T_s - T_{on}) = \frac{U_o}{L} \times (1 - D_y) \times T_s \quad (6.4)$$

When $t = t_s$, VT will be turned on again, and the circuit will operate into another periodic cycle.

From the above analysis, it can be learned that the inductor L serves as an energy storage and energy transfer device, when VT is on, L will absorb and store energy, and the load resistor R will be supplied with power by the capacitor C . When VT is cut off, the load resistor R will be powered by the inductor L .

When the circuit operates at the stable state, the increment $\Delta i_{L(+)}$ of the inductor L current i_L during the time when VT is on, is equal to the decrement $\Delta i_{L(-)}$ of the inductor L current i_L during the time when VT is off. Based on (6.2) and (6.4), it can achieve the relation equation between the input voltage and output voltage as:

$$\frac{U_o}{U_d} = -\frac{D_y}{1 - D_y} \quad (6.5)$$

From (6.5), the conclusion can be drawn as: when $D_y = 0.5$, then $|U_o| = |U_d|$; if $D_y < 0.5$, then $|U_o| < |U_d|$; if $D_y > 0.5$, then $|U_o| > |U_d|$. Thus, the Boost-Buck chopper's output voltage can be higher as well as lower than the DC source's voltage.

In Figure 6.2, the parallel Boost-Buck circuit I shares the same structure with the parallel Boost-Buck circuit II, so it only needs to analyze one of the Boost-Buck circuits. According to the detailed analysis of the above Boost-Buck chopper circuit, it can be easily learned that when the fully-controlled device IGBT is in the on state, the power source will

supply the inductor L and store energy in it through the IGBT, while Capacitor C maintains the output voltage substantially constant and supplies the load with power. Then the IGBT is turned off, the energy stored in the inductor L flows into the load. Thus, the polarity of the output voltage is opposite to the polarity of the input voltage, and the input and output voltage relationship can be expressed in the following equation:

$$U_o = -\frac{D}{1-D}U_i \quad (6.6)$$

in which, U_i is the input voltage of the circuit; U_o is the output voltage of the circuit; D is the duty cycle of IGBT. The minus sign represents the polarity reverse between input voltage and output voltage.

From the above analysis, it can be concluded that changing the duty cycle D of the IGBT will make the output voltage of the Boost-Buck circuit higher or lower than the input voltage. When $0 < D < \frac{1}{2}$, the circuit's output voltage is lower than the input voltage; when $\frac{1}{2} < D < 1$, the circuit voltage is higher than the input voltage; if all the device losses are neglected, when $D = \frac{1}{2}$, the output voltage of the circuit will be equal to the input voltage.

Base on the operation principle of the Boost-Buck circuit, it can achieve the state equations under the two types of the switch mode.

When the fully-controlled device IGBT is on, the state equation is shown:

$$\frac{d}{dt} \begin{pmatrix} I_L \\ U_o \end{pmatrix} = \begin{pmatrix} 0 & 0 \\ 0 & -\frac{1}{RC} \end{pmatrix} \begin{pmatrix} I_L \\ U_o \end{pmatrix} + \begin{pmatrix} \frac{1}{L} \\ 0 \end{pmatrix} U_i \quad (6.7)$$

When the fully-controlled device IGBT is off, the state equation is shown:

$$\frac{d}{dt} \begin{pmatrix} I_L \\ U_o \end{pmatrix} = \begin{pmatrix} 0 & \frac{1}{L} \\ -\frac{1}{C} & -\frac{1}{RC} \end{pmatrix} \begin{pmatrix} I_L \\ U_o \end{pmatrix} \quad (6.8)$$

When merging (6.7) and (6.8), the state space equation of the Boost-Buck circuit can be achieved:

$$\frac{d}{dt} \begin{pmatrix} I_L \\ U_o \end{pmatrix} = \begin{pmatrix} 0 & \frac{1-D}{L} \\ -\frac{1-D}{C} & \frac{1-2D}{RC} \end{pmatrix} \begin{pmatrix} I_L \\ U_o \end{pmatrix} + \begin{pmatrix} \frac{D}{L} \\ 0 \end{pmatrix} U_i \quad (6.9)$$

In (6.7), (6.8) and (6.9), I_L is the inductor current value, L is the inductor's inductance value, C is the capacitor's capacitance value, D is the duty cycle of the IGBT and R is the load resistor's resistance value.

6.3.2 The Control Strategy of the Reverse Parallel Boost-Buck Circuits

In order to improve voltage stability of the marine ship's power system, the DC bus voltage is considered as the control objective. When the marine ship's DC bus output voltage U_o is below the rated value, controlling the parallel Boost-Buck I to release the energy and transfer it to the DC bus to increase its voltage, the control block diagram of the PI closed-loop control strategy is demonstrated in Figure 6.6. Comparing the detected DC bus voltage U_o with the reference voltage value U_{ref} , it can obtain the error signal e , which serves as the input of the PI regulator. So PI regulator's output value and the terminal voltage value of the super capacitor energy storage unit are calculated by mathematical algorithm, and then proceed to limiting amplitude loop, thus the duty cycle D is achieved. The duty cycle D is compared with the triangular wave, the PWM that drives the parallel Boost-Buck I is generated, so the objective of controlling the DC bus voltage is obtained.

Figure 6.6 The Control Block Diagram of the PI Closed-Loop Control Strategy When U_o Below Rated Value

When the marine ship's DC bus output voltage U_o is above the rated value, controlling the parallel Boost-Buck II to release the redundant energy and transfer it to the super capacitor energy storage unit to decrease the DC bus voltage, the control block diagram of the PI closed-loop control strategy is displayed in Figure 6.7. Comparing the detected DC bus voltage U_o with the reference voltage value U_{ref} , it can obtain the error signal e' , which serves as the input of the PI regulator. So the PI regulator's output value and the terminal voltage value of the super capacitor energy storage unit are calculated by mathematical algorithm, and then proceed to limiting amplitude loop, thus the duty cycle D' is achieved. The duty cycle D' is compared with the triangular wave, the PWM that drives the parallel Boost-Buck II is generated, once again the objective of controlling the DC bus voltage is obtained. U_c' is the charging voltage of the super capacitor energy storage unit.

Figure 6.7 The Control Block Diagram of the PI Closed-Loop Control Strategy When U_o Above Rated Value

6.3.3 The Design of the Boost-Buck Circuit's Components

Firstly, it should calculate and select the inductor L . The inductor loss is the most difficult loss to be eliminated, which means an inductor a large inductance will also have a large resistance. The power loss which the inductor consumes under large current will be considerable. Therefore, in the case that satisfying the system requirements, it usually selects an inductor with a small inductance. The inductor is generally composed of the windings and the magnetic core, and it plays a role in transmit power, storing energy and filtering harmonics, which determines the amplitude of the input current's high-frequency ripple current. In order to calculate the minimum inductance value which can be applicable to the entire operation condition, it is has to be guaranteed that even all the parameters in the most unfavorable combination group, the selected inductance still will keep the ripple current

within a specific range. For the boost converter, the boost inductor can be calculated by the following equation:

$$L = \frac{U_m}{\Delta I} DT_s \quad (6.10)$$

Therefore, the condition for the most unfavorable combination group is: the input voltage and the duty cycle reach the highest value respectively, and the ripple current is at the minimum value. According to the reference material, the inductor ripple current's amplitude is generally 10% to 20% of the average current's value. When calculating the ripple current, it usually chooses 20% of the average current value as the ripple current's amplitude, namely:

$$\Delta I = 0.2I_{\max} = 0.2 \frac{\sqrt{2}P_{out}}{U_{in}} \quad (6.11)$$

When U_L is selected the maximum value, the ΔI will be chosen the minimum value. When the input voltage is the minimum value and the output voltage reaches the maximum value, the duty cycle gets the maximum value as:

$$D_{\max} = 1 - \frac{U_{in\min}}{U_{o\max}} \quad (6.12)$$

Base on the above rules, the working frequency of the converter's control gate signal is 20 kHz ($T_s = 10\mu s$). Thus, it can be achieved as $L \geq 4.38mH$ by doing the calculation.

Secondly, it is to select the filtering capacitor C . When the thyristor VT is on, the capacitor powers the load and the current is $-I_o$. When the thyristor VT is off, the induced EMF in the inductor L forces the diode D turned on, and it not only supplies the load with power but also complements the capacitor's electric charges which are released when the VT is on.

When the VT is on, the voltage drop on the capacitor C is:

$$\Delta U(-) = \frac{I_o t_{on}}{C} \quad (6.13)$$

When the VT is off, the voltage increment on the capacitor C is:

$$\Delta U(+) = \frac{1}{C} \int_0^{t_{off}} I_c dt \quad (6.14)$$

When the circuit operates in the steady state, $\Delta U(+) = \Delta U(-)$, so it can get:

$$C = \frac{I_o \cdot t_{on}}{\Delta U} = \frac{U_o \cdot D \cdot T}{\Delta U \cdot R} \quad (6.15)$$

In order to satisfy the requirement of the output ripple voltage's relative value, the filtering capacitor can be determined by:

$$C \geq \frac{U_o \cdot D \cdot T}{\Delta U \cdot R} = \frac{U_o \cdot D}{\Delta U \cdot R \cdot f} \quad (6.16)$$

in which, ΔU_o is the output voltage's pulsation value and it is about 1 V. On the worst condition that D and U_o are the maximum values respectively, it can be achieved as $C \geq 13.4 \mu F$ after doing the calculation

For the withstand voltage characteristics of the capacitor, it usually takes the rule that withstanding voltage value is twice of the operating voltage value.

Finally it comes to select the power thyristors and diodes. The thyristors' and output diodes' rated current should be larger than the maximum peak inductor current, and the rated voltages should be larger than the input voltage. This system adopts the power MOSFET, which depends on the major charge carriers' conduction, and there is no storage effect. So current rise and fall are very fast, and the storage time is short. It also has low switching loss, which means it can operate at a high switching frequency.

Similarly, the output diode is selected by the fast recovery rule. The diode selection principle is mainly based on the comprehensive consideration of a few parameters such as the forward voltage drop, reverse voltage drop and switching speed. The higher forward voltage drop, the greater the power loss. Also long time of the diode's turn-off time will affect the conversion efficiency and may produce an instant high voltage, damaging the components. The last requirement the diode must have is a high reverse voltage.

6.4 The Simulation of Energy Storage System

Figure 6.8 shows the simulation model of one generator and one motor marine ship's propulsion system without the energy storage system, while Figure 6.9(a) shows the simulation model of one generator and one motor marine ship's propulsion system with the energy storage system based on the reverse parallel Boost-Buck circuits, which is shown in Figure 6.9(b).

Because it involves the confidential policy of Rolls-Royce Corporate @NTU Lab, the detailed parameters are not disclosed.

The motor load is changing from 0 to 3 seconds, and all the other parameters are the same for the two systems. The target value of DC bus voltage is 600 V.

The DC bus voltage U_0 simulation waveforms of the two systems are shown in Figure 6.10.

Figure 6.8 Simulation Model of Marine Ship's Propulsion System without the Energy Storage System

(a) The Marine Ship's Propulsion System with the Energy Storage System

(b) The Reverse Parallel Boost-Buck Circuits

Figure 6.9 Simulation Model of Marine Ship's Propulsion System with the Energy Storage System Based on the Reverse Parallel Boost-Buck Circuits

Figure 6.10 shows the load fluctuations of the marine ship's propulsion system.

Figure 6.11(a) shows the DC bus voltage U_0 of the marine ship's propulsion system without the energy storage system.

While Figure 6.11(b) illustrates the DC bus voltage U_0 of the marine ship's propulsion system with the energy storage system based on the reverse parallel Boost-Buck circuits.

Figure 6.10 The Diagram of Load Fluctuations

(a) DC Bus Voltage U_0 of the Marine Ship's Propulsion System without the Energy Storage System

(b) DC Bus Voltage U_0 of the Marine Ship's Propulsion System with the Energy Storage System

Figure 6.11 DC Bus Voltage U_0 Simulation Waveforms of the Two Systems

By comparing the two systems' DC bus voltage U_0 , it can be obviously learned that the U_0 of the marine ship's propulsion system without the energy storage system has an unstable waveform due to the motor load changes. This unstable DC voltage will pour a great of harmonics into the power grid which will counteract the performance of the PWM rectifier DPC based on the power ripple minimization demonstrated in Chapter 4.

While the DC bus voltage U_0 of the marine ship's propulsion system with the energy storage system based on the reverse parallel Boost-Buck circuits is approximately stable at 600 V, which means it can provide a perfect power source for the next stage inverters and converters. Also, it eliminates the motor load change's bad influence on the power grid and plays as a crucial stabilizer as well as an energy storage unit in the mariner ship's power system.

6.5 Chapter Summary

In this chapter, the energy storage system based on the reverse parallel Boost-Buck circuits is demonstrated in detail. It not only serves as a DC voltage stabilizer, but also absorbs or releases energy when it is essential. The simulation results have proven the energy storage system to be a great device that guarantees the stable operation of the power system on the marine ship.

The energy storage system and the PWM rectifier DPC based on the power ripple minimization can work together to make the one generator and one motor marine ship's power system stable and normal.

Chapter 7

Conclusions and Recommendations

7.1 Conclusions

In this thesis, a DPC PWM rectifier based on the power ripple minimization in the marine ship's propulsion system is proposed. And it is the core of the one motor and one generator marine ship's propulsion system, since it can ensure a stable power voltage and minimize the harmonics in the power grid.

In Chapter 2, the one generator and one motor marine ship propulsion system model is set up. For the purpose of pertinence, the whole system simulation test is divided into two parts which are AC/DC part and DC motor control part.

In Chapter 3 and 4, it is the power ripple that dominates the focus of the study. The power ripple minimization strategy is illustrated in detail and the simulations have proven its advantages over the traditional DPC strategy and it will supply the marine ship with good quality of power. At steady state, the marine ship power grid active power ripple and reactive power ripple have been decreased by 11.16% and 68.47% respectively which will diminish the power loss and extend power electronics' as well as other the electric appliances' lifespans. In addition, the current THD drops from 4.93% to 3.70% and decreased by 24.95%. This guarantees the quality of power supply.

In Chapter 5, it introduces a double closed loop control system into the DC motor control part. The simulation work has been done to test the control system's performance and the result indicates there is a huge space to be satisfied. Thus, a Fuzzy-PID controller is added into the DC motor double closed loop control system. From the simulations of the two control methods, namely PID and Fuzzy-PID, it can be observed clearly that the motor speed overshoot and undershoot are nearly eliminated in the Fuzzy-PID controller system. The setting and resetting times are also shortened. Therefore, the Fuzzy-PID control displays significant performance in marine ship's motor speed control, which means the marine ship can cruise under some extreme conditions like huge tides and waves.

In Chapter 6, an energy storage system is illustrated at length. It is based on the reverse parallel Boost-Buck circuits, which plays as a switch passage between the DC bus

and the energy storage unit, which is a super capacitor. When the motor load changes, the voltage will start to fluctuate and it will backwards affect the voltage outputs of the upper stage converters or inverters. So, the energy storage system is essential and useful to stop the voltage tremor caused by load changes from influencing on the generator. And the simulation waveforms of the DC bus voltage U_o in the two different systems have manifested the energy storage system's good performance in stabilizing DC bus voltage and diminish voltage tremor, which protects a stable power grid along with the Power Ripple Minimization PWM rectifier.

In summary, the one generator and one motor marine ship's propulsion system has been fully studied and the simulations have been conducted to verify the proposed new methods and designs. The simulation results turn to be good and rewarding. What is not perfect is that there are no physical experiments as a result of equipment shortage and limited time.

7.2 Recommendations for Future Research

Although this thesis has studied quite a few research problems and the solutions to them are tested to be satisfying. There are still some areas that remains problems, which are worthy of studying and working.

1. The capacity expansion of the marine ship's power system. This thesis only considers one generator and one motor system. More generators and more motors on the marine ship can be studied in the future research. Since the marine ship is getting larger and larger, it is necessary and practical to consider a few generators and many motors. The power system on the ship will bear much more responsibility of stable operation and power supply. Thus, the control of converters and inverters are still crucial and urgent.

More complicated load fluctuations should be studied in the marine ship's propulsion system in the future research. The model of load fluctuations needs to reflect the real situations when the marine ship is cruising on the sea.

2. The design of multi-level Fuzzy-PID controllers. Chapter 5 depicts the design of one Fuzzy-PID controller in the DC motor double closed loop control system. The simulation result is satisfying but not optimal. The multi-level Fuzzy-PID controllers should be designed to meet the utmost requirements.

The motor speed usually changes rapidly at the transient state, when the Fuzzy-PID can show its magnificent performance, so one Fuzzy-PID controller will not be suitable when the speed recovers from the transient state. Therefore, a few Fuzzy-PID controllers should be designed and arrayed by level. When the motor speed changes, the one that is most suitable for the requirement of the Fuzzy-PID controllers should start to work. All the Fuzzy-PID controllers should work under coordination.

3. AC energy storage systems should be introduced and implemented in the marine ship's power system. In order to make the power system stable when the load changes rapidly, it is a useful way of introducing AC storage system between the inverter and the AC motor. This will make the voltage tremor at terminal less possible to backwards affect the upper converters and inverters, as well as the operation of generators.

4. The physical experiments should be conducted in the future research. Although the simulations can reflect the proposed control methods to some degree, it is still quite necessary and important to carry out physical experiments when the equipment is available and time is abundant.

Above recommendations for future research are rather promising and practically useful. Although these research problems will cost a lot of work, the harvest will be tremendously rewarding and profitable.

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Appendix

A.1 The Derivation of Equation (4.28)

The Power Ripple Optimization's Objective Function is:

$$F_{ripple}(t_{op}) = \frac{1}{t_{sp}} \int_0^{t_{op}} (P_k + s_{np}t - P_{ref})^2 dt + \frac{1}{t_{sp}} \int_{t_{op}}^{t_s} (P_k + s_{np}t_{op} + s_{zp}(t - t_{op}) - P_{ref})^2 dt \\ + \frac{1}{t_{sp}} \int_0^{t_{op}} (Q_k + s_{nq}t - Q_{ref})^2 dt + \frac{1}{t_{sp}} \int_{t_{op}}^{t_s} (Q_k + s_{nq}t_{op} + s_{zq}(t - t_{op}) - Q_{ref})^2 dt \quad (1)$$

Based on the derivation formula for the integral, the Power Ripple Optimization's Objective Function can be divided into:

$$\text{Part 1:} \quad \frac{d}{dt_{op}} \left[\int_0^{t_{op}} (P_k + s_{np}t - P_{ref})^2 dt \right] = (P_k + s_{np}t_{op} - P_{ref})^2 \quad (2)$$

Part 2:

$$\frac{d}{dt_{op}} \int_{t_{op}}^{t_s} (P_k + s_{np}t_{op} + s_{zp}(t - t_{op}) - P_{ref})^2 dt \\ = \frac{d}{dt_{op}} \int_{t_{op}}^{t_s} \left[(P_k + s_{zp}t - P_{ref})^2 + 2(P_k + s_{zp}t - P_{ref})(s_{np} - s_{zp})t_{op} + (s_{np} - s_{zp})^2 t_{op}^2 \right] dt \quad (3)$$

$$\frac{d}{dt_{op}} \int_{t_{op}}^{t_s} (P_k + s_{zp}t - P_{ref})^2 dt = -(P_k + s_{zp}t_{op} - P_{ref})^2 \quad (4)$$

$$\frac{d}{dt_{op}} \int_{t_{op}}^{t_s} 2(P_k + s_{zp}t - P_{ref})(s_{np} - s_{zp})t_{op} dt \\ = -2(P_k + s_{zp}t_{op} - P_{ref})(s_{np} - s_{zp})t_{op} + \int_{t_{op}}^{t_s} \frac{\partial}{\partial t_{op}} \left[2(P_k + s_{zp}t - P_{ref})(s_{np} - s_{zp})t_{op} \right] dt \quad (5) \\ = -2(P_k + s_{zp}t_{op} - P_{ref})(s_{np} - s_{zp})t_{op} + 2 \int_{t_{op}}^{t_s} (P_k + s_{zp}t - P_{ref})(s_{np} - s_{zp}) dt$$

$$\frac{d}{dt_{op}} \int_{t_{op}}^{t_s} (s_{np} - s_{zp})^2 t_{op}^2 dt \\ = -t_{op}^2 (s_{np} - s_{zp})^2 + \int_{t_{op}}^{t_s} \frac{\partial}{\partial t_{op}} \left[(s_{np} - s_{zp})^2 t_{op}^2 \right] dt \quad (6) \\ = -t_{op}^2 (s_{np} - s_{zp})^2 + 2t_{op} (s_{np} - s_{zp})^2 (t_s - t_{op})$$

So based on the equations (2),(4),(5) and (6), it can be achieved as:

$$\begin{aligned}
& \frac{d}{dt_{op}} \int_0^{t_{op}} (P_k + s_{np}t - P_{ref})^2 dt + \frac{d}{dt_{op}} \int_{t_{op}}^{t_s} (P_k + s_{np}t_{op} + s_{zp}(t - t_{op}) - P_{ref})^2 dt \\
&= (P_k + s_{np}t_{op} - P_{ref})^2 - (P_k + s_{zp}t_{op} - P_{ref})^2 - 2(P_k + s_{zp}t_{op} - P_{ref})(s_{np} - s_{zp})t_{op} \\
&+ 2 \int_{t_{op}}^{t_s} (P_k + s_{zp}t - P_{ref})(s_{np} - s_{zp}) dt - t_{op}^2 (s_{np} - s_{zp})^2 + 2t_{op} (s_{np} - s_{zp})^2 (t_s - t_{op}) \\
&= 2 \int_{t_{op}}^{t_s} (P_k + s_{zp}t - P_{ref})(s_{np} - s_{zp}) dt + 2t_{op} (s_{np} - s_{zp})^2 (t_s - t_{op})
\end{aligned} \tag{7}$$

Also, equation (8) can also be achieved.

$$\begin{aligned}
& \frac{d}{dt_{op}} \int_0^{t_{op}} (Q_k + s_{nq}t - Q_{ref})^2 dt + \frac{d}{dt_{op}} \int_{t_{op}}^{t_s} (Q_k + s_{nq}t_{op} + s_{zq}(t - t_{op}) - Q_{ref})^2 dt \\
&= 2 \int_{t_{op}}^{t_s} (Q_k + s_{zq}t - Q_{ref})(s_{nq} - s_{zq}) dt + 2t_{op} (s_{nq} - s_{zq})^2 (t_s - t_{op})
\end{aligned} \tag{8}$$

Then, based on the equations (7) and (8), the derivative of $F_{ripple}(t_{op})$ with respect to t_{op} can be calculated:

$$\therefore \frac{dF_{ripple}(t_{op})}{dt_{op}} = 0 \tag{9}$$

$$\begin{aligned}
& \therefore \frac{d}{dt_{op}} \int_0^{t_{op}} (P_k + s_{np}t - P_{ref})^2 dt + \frac{d}{dt_{op}} \int_{t_{op}}^{t_s} (P_k + s_{np}t_{op} + s_{zp}(t - t_{op}) - P_{ref})^2 dt \\
&+ \frac{d}{dt_{op}} \int_0^{t_{op}} (Q_k + s_{nq}t - Q_{ref})^2 dt + \frac{d}{dt_{op}} \int_{t_{op}}^{t_s} (Q_k + s_{nq}t_{op} + s_{zq}(t - t_{op}) - Q_{ref})^2 dt \\
&= 0
\end{aligned} \tag{10}$$

The equation (10) is equivalent to:

$$\begin{aligned}
& 2 \int_{t_{op}}^{t_s} (P_k + s_{zp}t - P_{ref})(s_{np} - s_{zp}) dt + 2t_{op} (s_{np} - s_{zp})^2 (t_s - t_{op}) \\
&+ 2 \int_{t_{op}}^{t_s} (Q_k + s_{zq}t - Q_{ref})(s_{nq} - s_{zq}) dt + 2t_{op} (s_{nq} - s_{zq})^2 (t_s - t_{op}) \\
&= 0
\end{aligned} \tag{11}$$

Then, the equation (11) is to be calculated in order to achieve the non-zero vector's optimization acting time t_{op} in one control period, which can make the Power Ripple Optimization's Objective Function output the minimum value.

The calculation process of equation (11) is shown as following:

$$\begin{aligned}
& \int_{t_{op}}^{t_s} (P_k + s_{zp}t - P_{ref})(s_{np} - s_{zp})dt \\
&= \int_{t_{op}}^{t_s} (P_k - P_{ref})(s_{np} - s_{zp})dt + \int_{t_{op}}^{t_s} s_{zp}(s_{np} - s_{zp})tdt \\
&= (P_k - P_{ref})(s_{np} - s_{zp})(t_s - t_{op}) + \frac{1}{2}s_{zp}(s_{np} - s_{zp})(t_s - t_{op})^2
\end{aligned} \tag{12}$$

It can also be achieved that:

$$\begin{aligned}
& \int_{t_{op}}^{t_s} (Q_k + s_{zq}t - Q_{ref})(s_{nq} - s_{zq})dt \\
&= \int_{t_{op}}^{t_s} (Q_k - Q_{ref})(s_{nq} - s_{zq})dt + \int_{t_{op}}^{t_s} s_{zq}(s_{nq} - s_{zq})tdt \\
&= (Q_k - Q_{ref})(s_{nq} - s_{zq})(t_s - t_{op}) + \frac{1}{2}s_{zq}(s_{nq} - s_{zq})(t_s - t_{op})^2
\end{aligned} \tag{13}$$

Therefore equation (11) can be equivalent to:

$$\begin{aligned}
& 2(P_k - P_{ref})(s_{np} - s_{zp})(t_s - t_{op}) + s_{zp}(s_{np} - s_{zp})(t_s - t_{op})^2 + 2t_{op}(s_{np} - s_{zp})^2(t_s - t_{op}) \\
& + 2(Q_k - Q_{ref})(s_{nq} - s_{zq})(t_s - t_{op}) + s_{zq}(s_{nq} - s_{zq})(t_s - t_{op})^2 + 2t_{op}(s_{nq} - s_{zq})^2(t_s - t_{op}) \\
& = 0
\end{aligned} \tag{14}$$

So it can be concluded that $t_s - t_{op} = 0$ or

$$\begin{aligned}
& 2(P_k - P_{ref})(s_{np} - s_{zp}) + s_{zp}(s_{np} - s_{zp})t_s - s_{zp}(s_{np} - s_{zp})t_{op} + 2t_{op}(s_{np} - s_{zp})^2 \\
& + 2(Q_k - Q_{ref})(s_{nq} - s_{zq}) + s_{zq}(s_{nq} - s_{zq})t_s - s_{zq}(s_{nq} - s_{zq})t_{op} + 2t_{op}(s_{nq} - s_{zq})^2 \\
& = 0
\end{aligned} \tag{15}$$

After the calculation, it can get:

$$\begin{aligned}
t_{op} &= \frac{2(P_{ref} - P_k)(s_{np} - s_{zp}) + 2(Q_{ref} - Q_k)(s_{nq} - s_{zq})}{2(s_{np}^2 + s_{zp}^2) + s_{nq}^2 + s_{zq}^2 - 3(s_{np}s_{zp} + s_{nq}s_{zq})} \\
&+ \frac{s_{zp}(s_{zp} - s_{np}) + s_{zq}(s_{zq} - s_{nq})}{2(s_{np}^2 + s_{zp}^2) + s_{nq}^2 + s_{zq}^2 - 3(s_{np}s_{zp} + s_{nq}s_{zq})} t_s
\end{aligned} \tag{16}$$