Electrical driven winch control for on board deck machinery

Wu, Xiaofeng

2016

Wu, X. (2016). Electrical driven winch control for on board deck machinery. Master's thesis, Nanyang Technological University, Singapore.

<https://hdl.handle.net/10356/69186>

<https://doi.org/10.32657/10356/69186>

**Electrical Driven Winch Control
For
On Board Deck Machinery**

Wu Xiao Feng

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Master of Engineering

MARCH 2016

Acknowledgement

I would like to express my sincere gratitude to all those who ever gave me a lot of help during the writing of this thesis.

My deepest gratitude goes first and foremost to Professor Wang Youyi, my supervisor, for his constant encouragement and guidance. He has guided me in every stage of writing this thesis. This thesis could not have become its present form without his enlightening and consistent instructions.

Secondly, I would like to deliver my heartfelt gratitude to Dr. Shawn Liu, who gave me this chance to do this thesis. During these times, he gave me a lot of instructions on directions of this thesis and also provided some important advises on this thesis.

I am also greatly indebted to the researcher at Epsil@ntu: Dr. Niu Li, who have instructed and helped me a lot in the past several months. He always welcomed me to consult questions at anytime and gave me very useful suggestions on my work.

Finally my thanks would go to my beloved family and friends for their great confidence and loving considerations in me all through these years. I also owe my sincere gratitude to my fellow classmates and my friends who gave me their time and help in listening to me and helping me work out my problems during the difficult times.

Contents

I. INTRODUCTION	5
1.1 Background and motivations	5
1.1.1 Background of winch system	5
1.1.2 Advantages of Permanent Magnetic Synchronous Motor (PMSM).....	6
1.1.3 Comparison of different motor control methods	7
1.2 Objectives	9
1.3 Contributions	9
1.4 Organization.....	10
1.5 Summary	11
II. ANALYSIS OF PMSM MATHEMATICAL MODEL.....	12
2.1 Mathematical model of PMSM	12
2.2 Choice of PMSM for winch system.....	14
2.2.1 Cause of ripple torque	15
2.2.2 Selection of PMSM.....	16
2.3 Summary	16
III. THE DIRECT TORQUE CONTROL SYSTEM OF PMSM	17
3.1 Basic theory of DTC.....	17
3.2 Space vector control	19
3.3 Parameter acquisition model.....	21
3.4 Hysteresis control of flux and torque	22
3.5 Switch table of traditional DTC	23
3.6 System diagram of traditional DTC	24
3.7 Summary	25
IV. MODIFIED DTC METHOD TO REDUCE TORQUE RIPPLE	26
4.1 Torque ripple's damage to on board work	26
4.2 Analysis of torque ripple in DTC.....	26
4.3 12 Space vectors' control.....	27
4.4 Improvement and switch table of 12-sector DTC	29
4.5 Summary	29
V. SENSORLESS CONTROL OF WINCH DRIVE SYSTEM.....	30
5.1 Categories of sensorless control methods.....	30
5.2 Rotating HF voltage signal injection method for low-speed region.....	31
5.2.1 Theory of rotating hf voltage signal injection	31
5.2.2 HF injection in a 6/12-sector DTC drive	32
5.3 Back-EMF integral method for mid/high-speed region.....	34
5.4 Summary	35
VI. SIMULATION AND ANALYSIS OF RESULTS	36

6.1 Simulation	36
6.1.1 Diagram and schematic of simulation	36
6.1.2 Functions in C block (Controller).....	41
6.1.3 Simulation of 12-sector DTC	43
6.2 Results and analysis.....	45
6.3 Summary	50
VII. CONCLUSION AND RECOMMENDATIONS.....	51
7.1 Conclusion.....	51
7.2 Recommendations for further research	52
7.2.1 Improvement of anti-disturbance capability	52
7.2.2 Switch algorithm for combined methods	53
7.2.3 Hardware Test	53
References.....	57

Summary: Winches are widely used on board ship's deck machinery for anchor handling and mooring. The marine winch market was dominated by hydraulic systems and there is a trend to be replaced by more efficient electrical drive system. For both anchoring and mooring control, full load torque is required through the entire speed range including zero speed operation. This poses a challenge for the inverter driven motor control, especially when there is no speed or position sensor. This thesis aims to develop advanced control algorithms of permanent magnet (PM) machine including direct torque control (DTC) and sensorless control for marine winch application. In this thesis, direct torque control, as a new type of high-performance ac driving system after vector control technique, is mainly concerned. Furthermore, since the traditional switch-table based DTC ac motor drive is usually affected by the torque ripple problem and steady-state error problem in torque, this thesis introduced a modified 12-sector DTC control method. The subsystems of traditional DTC system and modified DTC system for permanent magnet synchronous machine (PMSM) are modeled and designed in PSIM software. Then the simulation models of the above two control strategies are also established. Simulation research is conducted and the results are corresponding with theory analysis: the improved system has good dynamic and static performance; less flux and torque ripple especially in the low speed range. The second part of this thesis is to realize sensorless control of DTC control system. Considered the actual situation on board, different methods for high-speed range and low-speed range are discussed. The rotating high frequency (*hf*) voltage signal injection as low-speed sensorless control method is good for its simplicity and better tracking accuracy, while the back-EMF integral as high-speed method is suitable for its mature industrial applications. And the *hf* injection method used in this thesis is designed to compatible with the DTC method applied in this thesis, which successfully extended the speed range without sacrificing the simplicity of DTC method.

Key word: Direct torque control (DTC), permanent magnet synchronous machine (PMSM), low speed, torque ripple, and sensorless control.

I. INTRODUCTION

1.1 Background and motivations

1.1.1 Background of winch system

A winch is a mechanical device that is used to pull in (wind up) or let out (wind out) or otherwise adjust the "tension" of a rope or wire rope (also called "cable" or "wire cable"). Its simplest form consists of a spool and an attached hand crank. In its larger forms, winches stand at the heart of machines as diverse as elevators, steam shovels and tow trucks. The spool can also be called the winch drum. More elaborate designs of winches have gear assemblies and can be powered by pneumatic, hydraulic, electric or internal combustion drives, where some may include a mechanical brake and/or a solenoid brake or ratchet and pawl device that prevents it from unwinding unless the pawl is retracted [1].

Figure 1.1 Deck machinery winches for anchoring and mooring [1]

Deck machinery winches for mooring and anchoring are the basic equipment on all marine vessels, as shown in Figure 1.1. Winching operations are very challenging since an anchor line or slack mooring with nearly zero tension at the beginning has to deal with torque rapidly racing to great levels in the next just a few seconds. And then it needs to keep the high output torque during variable speed ranges in the operation. To properly operate and secure high-mass marine vessels moving on the water surface, it is very important to have precise and 100% reliable control of the huge torque forces at work. [1][2]

Traditionally marine winches have been controlled by hydraulic systems, which have excellent dampening properties that, until recently, electrical controls had a hard time matching. However, hydraulics is often messy and complicated due to the hydraulic oil, pumps, piping and possible leaks.

Electrical three-speed winch motors with contactors have existed already for many years, as shown in Figure 1.2. It has already had electrical VSD (variable speed drive) solutions with basic functionality for winch operations. Although these old-generation electrical driven solutions offered improved control and positive environmental benefits, they were not widely accepted, because the DOL (Direct on Line) motors with 3 motor windings for low-middle-high speeds and tension control via load sensors had limited success, while the early VSD design which included a PLC with winch application has also been proved unsuccessful [3].

Figure 1.2 Electrical driven winches [3]

Drawbacks of this old generation included costly motors with three windings, high noise, a high starting current, and the requirement for electrical cabinets with contactors for each set of windings in the motor. Additionally, external control via old control cards generated questions about support, maintenance and spare parts. In other words, the external PLCs, encoders and load sensors equipped on the early VSD winches made the systems complicated and sometimes unreliable. The result has been that the traditional hydraulic systems have been maintained by approximately 70% of the worldwide marine vessels so far, which is understandable considering the fact that winch builders and ship-owners are extremely careful with their multimillion-dollar vessels and cargoes. This clearly is not an easy decision for them to apply new things before they are well proven with years of experience on both the hardware and software. And considering the total cost approximately 10 thousand dollars of the entire system, unless there are some irresistible reasons, otherwise the businessmen won't accept any changes to their moneymaker. [3][4][5]

But the change is now happening because new electronic technology has given the electrical option more precise and reliable control. Furthermore, environmental concern has also catalyzed the changeover. Hydraulic winches are becoming more controversial in the maritime business due to potential oil leaks in pristine areas such as the Barents Sea. So the electrically driven winches are nowadays often selected on board instead of hydraulic control.

In marine industry, the electrical drive improves reliability and enables more precise rope control and reduces operating noise over hydraulic drive. No hydraulics means no hydraulic fluid concerns, less parts, reduced installation space and reduced maintenance needs [6].

1.1.2 Advantages of Permanent Magnetic Synchronous Motor (PMSM)

As the rapid development of motor speed control theory, power electronics, microelectronics technology and the permanent magnetic materials, such frequency control of PMSM has entered a stage of in-depth research and extensive applications. PMSM has simple structure, high torque inertia ratio, high energy density, high efficiency, good maintenance and many other advantages, so it is used widely in Computer Numerical Control (CNC) machine tools, aerospace, marine industry and many other electric traction fields.

Figure 1.3 Induction motor to PMSM [6]

Compared with conventional induction motors showed in Figure 1.3, permanent magnet synchronous motor has the following advantages in marine condition:

- Significant energy-saving effect for limited on board resources. Firstly, permanent magnet synchronous motor replaces electrical excitation by a permanent magnet, so there is no excitation loss. Secondly, PMSM has no rotor windings, so there is no copper loss on the rotor. Thirdly, because stator and rotor are synchronous, there is no iron loss in the rotor core. Therefore, PMSM is more efficient than electric excited synchronous and asynchronous motors. Furthermore, PMSM does not require lagging exciting current that draws from the AC source, thus saving reactive power and increasing the power factor greatly [7].
- Highly durable construction to difficult weather conditions and heavy-duty operations on board. Permanent magnet synchronous motor has hard mechanical properties, so it has strong endurance to load changes caused by the environment disturbance on motor torque.
- Low wear shaft requirement on board. Permanent magnet synchronous motor has low inertia since there is no cage rotor.
- Simple for on board staff to control. Motor speed and power frequency are always keeping the accurate synchronization relationship, which means motor speed can be controlled by adjusting power frequency. So the control of PMSM is easier than induction motor.

The advantages mentioned above make permanent magnet synchronous motor widely used in modern industrial applications.

1.1.3 Comparison of different motor control methods

Since the late 1980s, varieties of AC variable speed technologies have become mature with high operational reliability due to the rapid development of converter technology, control theory and means of control. The main control strategies for permanent magnet synchronous motors have the following three categories, variable voltage variable frequency control (VVVF), vector control (VC) and direct torque control (DTC).

- Variable voltage variable frequency control (VVVF): Control variables of VVVF control strategy are the external variables of motor, namely voltage and frequency. The reference voltage and

frequency are the inputs of VVVF regulator, then the inverter produces a sinusoidal alternating voltage applied to the motor stator windings to make motor run at the specified voltage and reference frequency. This control method does not need to introduce any speed, position, voltage or current feedback signal from the motor, so it is an open-loop control, which is characterized by easy to implement and inexpensive. Further, wear and tear of the equipment can also be less during start/stop of the motor by using VVVF motor drive. Since the system does not require the introduction of speed, position, or any other feedback signal, it cannot work in precisely control condition, thereby this control method cannot obtain optimal electromagnetic torque. And there is only one regulator to achieve the modulation of input voltage and flux, thus resulting in the reduction of communication between input voltages, frequency signals, the motor final torque and speed reaction, so that the motor response becomes slow. This drive system is generally applicable to fan, pump and situations do not need precise control. [8][9]

- Vector control (VC): Vector control, also called field-oriented control (FOC), is a variable-frequency drive (VFD) control method in which the stator currents of a three-phase AC electric motor are identified as two orthogonal components that can be visualized with a vector. The basic idea of vector control technology is to imitate the control of DC motor, namely mimic the magnetic field orientation process. In the control system, the electromagnetic torque and the stator flux are proportional to the q-axis and d-axis current generated by the permanent magnet motor. And because of this positive relationship, control the cross-axis current generated by the permanent magnets, thus control the motor electromagnetic torque, which realize the speed control of motor. The benefits of vector control system include good torque response, accurate speed control, can achieve full load at zero speed, and similar operating characteristics to DC motor. However, in order to obtain high performance of torque and speed control, rotor position feedback device is required, which will increase the system hardware costs. Additionally, the regulator also reduces the communication between the input and output signals of the motor. [10][11] Therefore, compared with a DC motor drive, this control system is very complex due to the requirement of rotating coordinate transformation and counter-rotating coordinate transformation though the entire system seems simple. In fact, it is difficult to observe rotor flux accurately, and the system greatly influenced by the characteristics of the motor parameters, so that the actual effect of vector control is difficult to achieve the theoretical analysis.
- Direct torque control (DTC): The thought of direct torque control is to achieve electromagnetic torque control by directly controlling the stator flux of the motor. Unlike the vector control method, which controls torque by controlling the current, direct torque control observes the stator flux and torque of motor in the stator coordinate system. Then obtain flux and torque control signals produced by two hysteresis comparators whose inputs are the observed and reference value of flux and torque. Finally, calculate the position of stator flux in stator coordinate system based on the flux and torque control signals, then select the appropriate space vector voltage from switch table to control the stator flux direction, thereby control the torque. So basically DTC is one method used in variable frequency drives to control the torque (and thus finally the speed) of three-phase AC electric motors. [12][13] Compared with vector control method, its main advantage is that it abandons the complex decoupling idea, instead, to control the flux and torque directly, and replaces the rotor flux orientation by stator flux orientation, so that it avoids obtaining those motor parameters that cannot easily be determined. This control method does not require rotor position information or rotor parameters, and eliminates the need of complex rotating coordinate transformation. The calculation of the stator flux is only relative with the stator resistance which is easy to measure, therefore it is easier to estimate the flux more precisely and

the influence of motor parameters change is smaller. In addition, direct torque control method determines the voltage vector directly by the bias of output torque and stator flux. Compared with the previous governor theory, it has the more direct mean of control, the simpler calculation process, the faster torque response and the better dynamic performance. Accordingly, direct torque control has aroused widespread attention and rapid development, and now it has the most widely prospects for application among varieties of AC speed control methods.

The above comparison of three different control methods shows that the direct torque control method is the best suitable way for winch control because of its precise control, simplicity and good low-speed performance.

1.2 Objectives

From the above introduction, the objectives of this thesis can be set as below:

- Establish a direct torque control (DTC) system for permanent magnetic synchronous motor (PMSM) working in entire speed region.
- Improve the performance of DTC system by solving the torque ripple problem existed in the traditional DTC method.
- Realize sensorless control of DTC system in entire speed region without sacrificing the simplicity of DTC system.

And according to the on board requirement, the electrical driven system should have the following features: working with full load in entire speed region, simple structure for easy control and fast dynamic response, and energy saving due to the limited on board resources.

1.3 Contributions

According to the objectives mentioned in Section 1.2, the main contribution of this thesis is the establishment of an entire Sensorless DTC system that is suitable for the electrical driven winch application. The features of this system are mentioned in the Section 1.2, and to achieve those goals, following contributions in the control system have been made to meet the requirement of marine condition:

- To realize the motor working with full load in entire speed region, the key point is the performance of motor with high load at extreme-low speed and even 0 speed. There are very little examples in this kind of condition, because the low-speed systems are usually designed specifically for its application or with mechanical devices to help at low or 0 speed, as the elevator as the most decent example. But in winch application, there is no mechanical support, which means whether the braking or speed maintenance at low speed is all relied on motor itself. So based on the fundamental knowledge of PMSM and control strategies, this thesis analyzed how to choose the proper PMSM for the winch system in Chapter II and explained why DTC can achieve full torque in full speed range in Chapter III.

- In the traditional DTC, the existence of torque ripple problem does not meet the energy-saving and safety requirement for on board work. In order to minimize the torque ripple problem, a new 12-sector DTC was developed. In 12-sector DTC, the original 6 space vectors had been successfully extended to 12 space vectors, so that the control process would be smoother, namely, less unreasonable torque ripple. In Chapter IV, the idea of this new method would be explained in detail about why and how it can reduce torque ripple.
- Since there are no possibilities for the implement of mechanical sensors on board, sensorless control is required in winch system. Once the sensorless control was introduced into the system, the system would become very complicated. In order to keep the system structure as simple as possible, the switch table based DTC were adopted instead of SVPWM based DTC in this thesis. However, the most effective sensorless control methods in low-speed region nowadays require the injection of high frequency voltage or current, but without PWM module means that the most common and simple way of high frequency carrier injection cannot be used. So Chapter V illustrated a unique way of high frequency carrier injection, which successfully used the existed space vectors to create the high frequency voltage signal in traditional or 12-sector DTC. In this way the simplicity of DTC would be kept.

So generally speaking, the contribution of this thesis is the establishment of sensorless DTC system with several analysis and new ideas adopted based on the main objective to make the system applicable for the winch application. Finally in Chapter VI, all the theories and analysis will be verified in simulation, while some further research recommendations will be given in the last chapter.

1.4 Organization

This thesis consists three parts to introduce the electrical driven winch system from background to the future prospects. There are three parts as the main parts of winch system this thesis mainly focuses on: the motor (PMSM), main control algorithm (DTC), and sensorless control. Each of them will be discussed in separate chapters at length.

Part I (Chapters I, II and III) presents the background and the basic knowledge of winch system, PMSM and DTC. Chapter I provides a general introduction of this thesis. The first part of this chapter contains information on the history of winch system, the comparison between different motors and control theories. And base on these information, explains why choose PMSM and DTC as the main parts of electrical winch system. The rest parts of Chapter I introduce the objectives, contributions and organization of this thesis. Chapter II mainly deals with PMSM on theoretical base. The mathematical model of PMSM is discussed firstly in order to build a strong basement of all rest contents in this thesis. And the second part of this chapter offers some suggestions on the choice of PMSM for winch system, based on the discussion of ripple torque that may be generated only by hardware. Chapter III introduces DTC system in detail including the basic theory and each part of the DTC system. This chapter explained DTC theory firstly, and based on the theory the rest sections of Chapter III illustrate each part of DTC system minutely including space vectors, parameters' acquisition, hysteresis control and switch table. And at the end of this chapter, a completely system diagram of DTC is given, which shows the connections between each parts and the operation way of the entire system clearly.

Part II (Chapters IV and V) introduces the main control algorithms of DTC and sensorless control adopted in the electrical driven winch system. Chapter IV introduces a modified DTC method based on the traditional DTC mentioned in Chapter III. As discussed in the first section, torque ripple can do a lot of

damage to the on board work. Then the second section analyzes the reason and how much the torque ripple can be in DTC. So based on the analysis, the rest parts introduces a modified DTC to reduce the torque ripple by extending the effective space vector numbers to 12. And a theoretical proof that the new method truly reduces the torque ripple to a great extent is also given in this chapter. Chapter V provides the sensorless control strategies for a full speed range system. So the different methods for the different speed ranges will be adopted. Both methods are discussed in detail in this chapter, especially the low-speed method. Not only because it is a little complicated, but also it needs to coordinate with the DTC algorithm used in this system while the high-speed method is much simpler and effective.

Part III (Chapters VI and VII) gives the result and prospect of this thesis. Chapter VI shows the simulation of the whole system. Every function part of the system along with the system diagram is explained in detail. The result and analysis are also given in the second section of this chapter. Chapter VII concludes the thesis, summarizes Parts I, II and III, and discusses the future opportunities for applying electrical winch system.

1.5 Summary

In this chapter, firstly introduced the history and current situations of winch control system, then compared induction motor and permanent magnet synchronous motor, and finally compared three different control methods. In conclusion, electrical driven winch system is replacing traditional hydraulic driven system nowadays and the permanent magnet synchronous motor with direct torque control is the best solution to it. After the background introduction, the objectives, contributions and organization of this thesis are given to show the entire structure of the thesis. So the following chapters will gradually introduce every aspect of the whole system according to the organization introduced in Section 1.4. In chapter II, the mathematical model of PMSM will be discussed as the beginning. It is very necessary since the mathematical model will be a very strong theory basement for all other discussions next in this thesis.

II. ANALYSIS OF PMSM MATHEMATICAL MODEL

Chapter I introduced some background knowledge and structure of this thesis. So from this chapter, each part of the electrical winch system will be discussed in detail. Since the PMSM is chosen to be the motor of this system, its mathematical model, which is the base of all control strategies, should be the first thing to concern. Furthermore, for DTC as the main control method, the flux and torque expressions in different coordinate systems are the main targets in the analysis.

2.1 Mathematical model of PMSM

In the analysis of the mathematical model of permanent magnet synchronous motor, in order to simplify the analysis, [14][15] introduced a way to make the following assumptions:

- Stator winding is star connection and winding current is symmetrical three-phase sine wave.
- Stator magnetic field is sinusoidal distribution and irrespective of harmonic saturation.
- Ignore the eddy current and hysteresis losses of the motor.

At the basis of the above assumptions, when establishing permanent magnet synchronous motor mathematical model in different coordinate systems, the relationships between various coordinate systems are shown in Figure 2.1.

Figure 2.1 Coordinate systems of PMSM

In Figure 2.1, A-axis of ABC coordinate system coincides with α -axis of the $\alpha\beta$ coordinate system. DQ coordinate system rotates at the rotor angular velocity ω_r and d-axis points the direction of the rotor flux ψ_f . XY coordinate system rotates at the stator flux rotation angular velocity ω_e and x-axis points the stator flux's direction. Angle between d-axis and A-axis is θ_r . The angle between stator ψ_s and rotor flux vector ψ_f , δ , is the motor power angle. [16][17]

In the three-phase stator stationary ABC coordinate system, ignoring the stator windings, the voltage and flux vector can be expressed as (2-1) and (2-2):

$$\begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} = \begin{bmatrix} R_a & 0 & 0 \\ 0 & R_b & 0 \\ 0 & 0 & R_c \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + p \begin{bmatrix} \psi_a \\ \psi_b \\ \psi_c \end{bmatrix} \quad (2-1)$$

In (2-1), p is the differential operator, $p = d/dt$. Since the three-phase stator windings are completely symmetrical, the resistance three-phase stator windings are equal, $R_a = R_b = R_c$. ψ_a, ψ_b, ψ_c are the total flux of each phase winding:

$$\begin{bmatrix} \psi_a \\ \psi_b \\ \psi_c \end{bmatrix} = \begin{bmatrix} L_a & M_{ab} & M_{ac} \\ M_{ba} & L_b & M_{bc} \\ M_{ca} & M_{cb} & L_c \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \begin{bmatrix} \psi_{ra}(\theta_r) \\ \psi_{rb}(\theta_r) \\ \psi_{rc}(\theta_r) \end{bmatrix} \quad (2-2)$$

In (2-2), $\theta_r = w_r t$, w_r is the rotor angular velocity. $\psi_{ra}(\theta_r), \psi_{rb}(\theta_r), \psi_{rc}(\theta_r)$ are the rotor flux linkage generated in the ABC three-phase windings. The self-inductance of three-phase stator windings $L_a = L_b = L_c = L_s$. The mutual inductance of three-phase stator windings $M_{ab} = M_{ac} = M_{ba} = M_{bc} = M_{ca} = M_{cb} = M_s$ [18].

Rotor flux distribution in the air gap of is the sinusoidal type, so the rotor flux linkage in the phase windings can be expressed as (2-3) [19]:

$$\begin{bmatrix} \psi_{ra}(\theta_r) \\ \psi_{rb}(\theta_r) \\ \psi_{rc}(\theta_r) \end{bmatrix} = \psi_f \begin{bmatrix} \cos(\theta_r) \\ \cos(\theta_r - 2\pi/3) \\ \cos(\theta_r + 2\pi/3) \end{bmatrix} \quad (2-3)$$

Because the current through the three-phase winding is symmetrical, it gives that $i_a + i_b + i_c = 0$. Substituting this condition into (2-1) and let $L = L_s - M_s$, gives that:

$$\begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} = \begin{bmatrix} R_s + pL & 0 & 0 \\ 0 & R_s + pL & 0 \\ 0 & 0 & R_s + pL \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} - w_r \psi_f \begin{bmatrix} \cos(\theta_r) \\ \cos(\theta_r - 2\pi/3) \\ \cos(\theta_r + 2\pi/3) \end{bmatrix} \quad (2-4)$$

Establishing mathematical model in dq coordinate system is very convenient for analyzing the stability and dynamic performance of PMSM in the control processes. The voltage loop equation of PMSM in the dq coordinate system is showed in (2-5) [20]:

$$\begin{bmatrix} u_d \\ u_q \end{bmatrix} = \begin{bmatrix} R_s + pL_d & -L_q w_r \\ -L_d w_r & R_s + pL_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + w_r \psi_f \begin{bmatrix} 0 \\ 1 \end{bmatrix} \quad (2-5)$$

Assuming that the motor is linear and parameters don't change with temperature or any other environment conditions. Ignoring the hysteresis and eddy current loss. And with no damper winding rotor, (2-6) represents the stator flux equation in dq coordinates:

$$\begin{bmatrix} \psi_d \\ \psi_q \end{bmatrix} = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \psi_f \begin{bmatrix} 1 \\ 0 \end{bmatrix} \quad (2-6)$$

And the electromagnetic torque equation is [21]:

$$T_e = 1.5n_p(\psi_d i_q - \psi_q i_d) \quad (2-7)$$

The equation of motion is [22]:

$$T_e - T_L = J \frac{dw_r}{dt} + B w_r \quad (2-8)$$

In (2-8), T_L is the load torque. J is the inertia of the motor and B is the damping factor.

According to the coordinate transformation theory, transform the PMSM model in the dq coordinate to $\alpha\beta$ coordinate system to obtain the voltage equation in $\alpha\beta$ coordinates system from (2-5):

$$\begin{bmatrix} u_\alpha \\ u_\beta \end{bmatrix} = \begin{bmatrix} R_s + pL_d & (L_d - L_q)w_r \\ -(L_d - L_q)w_r & R_s + pL_q \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + [(L_d - L_q)(w_r i_d - i_q) + w_e \psi_f] \begin{bmatrix} -\sin(\theta_r) \\ \cos(\theta_r) \end{bmatrix} \quad (2-9)$$

Perform coordinate transformation for (2-6) to obtain the flux equation in the $\alpha\beta$ coordinate system:

$$\begin{bmatrix} \psi_\alpha \\ \psi_\beta \end{bmatrix} = L_q \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + [(L_d - L_q)i_d + \psi_f] \begin{bmatrix} \cos(\theta_r) \\ \sin(\theta_r) \end{bmatrix} \quad (2-10)$$

In the $\alpha\beta$ coordinate system, ψ_α and ψ_β can also be obtained by integrating the back-EMF:

$$\psi_\alpha = \int (u_\alpha - R_s i_\alpha) dt \quad (2-11)$$

$$\psi_\beta = \int (u_\beta - R_s i_\beta) dt \quad (2-12)$$

And the electromagnetic torque equation is:

$$T_e = 1.5n_p(\psi_\alpha i_\beta - \psi_\beta i_\alpha) \quad (2-13)$$

Until here, the information on mathematical model of PMSM is enough for the further DTC discussion, since the equations (2-11), (2-12) and (2-13) are the most important expressions of PMSM in DTC method. After the establishment of mathematical model, in order to select the most proper PMSM for the system, Section 2.2 will analysis the ripple torque in the hardware model of PMSM.

2.2 Choice of PMSM for winch system

As mentioned in abstract and Chapter I, PMSM often works at very low speed in winch system, so the performance at low speed range is mainly concerned in the choice of PMSM, in other words, torque ripple has to be reduced as much as possible. To generate a constant torque, the induced electromotive force (EMF) and the current of PMSM should be a sine wave without harmonic disturbances [23]. But in the actual motor, the magnetic field distribution of permanent magnet rotor or the spatial distribution of stator winding is not ideal sine wave, in addition, the frequency devices as power supply to the stator though uses the fast current tracking control technology, which can possible track sine wave, but still inevitably contain high harmonics in stator current and EMF [24]. The harmonic torque generated due to EMF and stator current waveform distortion is called ripple torque.

2.2.1 Cause of ripple torque

Following is the quantitative analysis of field-oriented control ripple torque of PMSM. Assuming:

- The magnetic circuit is not saturated and ignoring the impact of cross-axis armature reaction.
- Not considering the damping effect of the permanent steel of rotor and the rotor core.
- Symmetrical three-phase stator windings, Y-type connection with non-neutral, no 3 times or multiples of 3 times harmonics in stator current and the stator current is also free of even harmonics.

A-phase current and induced electromotive force can be expressed as follows:

$$i_A(t) = I_{m1}\sin(\omega t) + I_{m5}\sin(5\omega t) + I_{m7}\sin(7\omega t) + \dots \quad (2-14)$$

$$u_A(t) = E_{m1}\sin(\omega t) + E_{m3}\sin(3\omega t) + E_{m5}\sin(5\omega t) + E_{m7}\sin(7\omega t) + \dots \quad (2-15)$$

where ω is the fundamental angular velocity and electrical angular velocity of rotor.

So the electromagnetic power of A-phase is:

$$u_A(t)i_A(t) = P_0 + P_2\cos(2\omega t) + P_4\cos(4\omega t) + P_6\cos(6\omega t) + \dots \quad (2-16)$$

Similarly, the electromagnetic power of B-phase and C-phase:

$$u_B(t)i_B(t) = P_0 + P_2\cos(2\omega t - 120^\circ) + P_4\cos(4\omega t - 120^\circ) + P_6\cos(6\omega t - 120^\circ) + \dots \quad (2-17)$$

$$u_C(t)i_C(t) = P_0 + P_2\cos(2\omega t - 240^\circ) + P_4\cos(4\omega t - 240^\circ) + P_6\cos(6\omega t - 240^\circ) + \dots \quad (2-18)$$

So the electromagnetic torque is

$$T_{em} = [u_A(t)i_A(t) + u_B(t)i_B(t) + u_C(t)i_C(t)]/\Omega \quad (2-19)$$

Ω is the mechanical angular velocity of rotor. Substitute (2-16), (2-17) and (2-18) into (2-19):

$$T_{em}(t) = T_0 + T_6\cos(6\omega t) + T_{12}\cos(12\omega t) + T_{18}\cos(18\omega t) + \dots \quad (2-20)$$

in which,

$$T_6 = 1.5[I_{m1}(E_{m7} - E_{m5}) + I_{m5}(E_{m11} - E_{m1}) + I_{m7}(E_{m1} + E_{m13}) + I_{m11}(E_{m5} + E_{m17}) + \dots]/\Omega$$

$$T_{12} = 1.5[I_{m1}(E_{m13} - E_{m11}) + I_{m5}(E_{m17} - E_{m7}) + I_{m7}(E_{m19} - E_{m5}) + I_{m11}(E_{m23} - E_{m1}) + \dots]/\Omega$$

$$T_{18} = 1.5[I_{m1}(E_{m19} - E_{m17}) + I_{m5}(E_{m23} - E_{m13}) + I_{m7}(E_{m25} - E_{m11}) + I_{m11}(E_{m29} - E_{m7}) + \dots]/\Omega$$

Written in matrix form:

$$\begin{bmatrix} T_0 \\ T_6 \\ T_{12} \\ T_{18} \end{bmatrix} = \frac{3}{2\Omega} \begin{bmatrix} E_{m1} & E_{m5} & E_{m7} & E_{m11} \\ E_{m7} - E_{m5} & E_{m11} - E_{m1} & E_{m1} + E_{m13} & E_{m5} + E_{m17} \\ E_{m13} - E_{m11} & E_{m17} - E_{m7} & E_{m19} - E_{m5} & E_{m23} - E_{m1} \\ E_{m19} - E_{m17} & E_{m23} - E_{m13} & E_{m25} - E_{m11} & E_{m29} - E_{m7} \end{bmatrix} \begin{bmatrix} I_{m1} \\ I_{m5} \\ I_{m7} \\ I_{m11} \end{bmatrix} \quad (2-21)$$

From the above analysis, the interaction between same times of induced electromotive force harmonics and current harmonics generates the average torque. The interaction between different frequency EMF harmonics and current harmonics generates the ripple torque whose ripple frequency is six times to the fundamental frequency. The amplitude of each ripple torque is related to the waveform distortion of current and EMF.

When the system is working in high-speed region, ripple torque may be absorbed by the rotor inertia. But when the system runs at a very low speed, it will make the rotor speed fluctuate and seriously affect the speed stability. So we should pay full attention to minimize ripple torque when we design and choose PMSM for low-speed usage.

2.2.2 Selection of PMSM

In order to reduce the ripple torque, the harmonics of induced electromotive force and stator current should be reduced as much as possible. So we should note the following in the selection of PMSM for low-speed work:

- Select the number of slots per pole per phase as big as possible to weaken the higher harmonics of induced electromotive force.
- Select the appropriate pitch factor of stator winding to weaken the EMF and MMF harmonics.
- As far as possible to choose the right shape and size of magnetic pole, so that the waveform of rotor excitation magnetic field is as sinusoidal distribution.

2.3 Summary

In this chapter, Section 2.1 illustrates the analysis of mathematical model of PMSM, especially some important flux and torque expressions in DTC, which makes a strong base for the introduction of control methods afterwards. And based on the analysis in Section 2.1, Section 2.2 analyzes the ripple torque which will be an important factor when PMSM works at a low-speed range. Therefore, the analysis in Section 2.2 offers some suggestions in the selection of PMSM for low-speed work. After the analysis of PMSM's model, on the strong theoretical basement, the following chapter will introduce the DTC of PMSM as the main control algorithm of the winch system.

III. THE DIRECT TORQUE CONTROL SYSTEM OF PMSM

The introduction of mathematical model of PMSM in Chapter II built a very good basement for the discussion of DTC in this chapter. As mentioned in Chapter I, the features of simple structure, full speed range operation and fast dynamic response make DTC the most suitable control strategies for electrical driven winch system. In this chapter, in order to explain how exactly DTC works, the basic theory and each function part of DTC system will be introduced in detail.

3.1 Basic theory of DTC

The thought of direct torque control is to achieve electromagnetic torque control by directly controlling the stator flux of the motor. Direct torque control observes the stator flux and torque of motor in the stator coordinate system. Then obtain flux and torque control signals produced by two hysteresis comparators whose inputs are the observed and reference value of flux and torque. Finally, calculate the position of stator flux in stator coordinate system based on the flux and torque control signals, then select the appropriate space vector voltage from switch table to control the stator flux direction, thereby control the torque. In a word, direct torque control is one method used in variable frequency drives to control the torque and thus finally the speed. [25][26][27]

Formula (2-7) shows that the electromagnetic torque of permanent magnet synchronous motor depends on the stator current of dq-axis, and it has not reached completely decoupled. To this end, there is the need to perform coordinate transformation decoupling in order to obtain the torque equation of permanent magnet synchronous motor in the xy coordinate system, in which x-axis is the stator flux ψ_s while y-axis is the rotor flux ψ_r [28]. The relation of xy coordinate system and dq coordinate system is shown in Figure 2.1. The conversion formula between the xy coordinate system and dq coordinate system is shown as (3-1) below [29]:

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} \cos\delta & -\sin\delta \\ \sin\delta & \cos\delta \end{bmatrix} \begin{bmatrix} i_x \\ i_y \end{bmatrix} \quad (3-1)$$

Formula (3-1) is also applicable to voltage and flux vectors, so the flux equation in xy coordinate system is:

$$\begin{bmatrix} \psi_x \\ \psi_y \end{bmatrix} = \begin{bmatrix} L_d \cos^2\delta + L_q \sin^2\delta & -L_d \sin\delta \cos\delta + L_q \sin\delta \cos\delta \\ -L_d \sin\delta \cos\delta + L_q \sin\delta \cos\delta & L_d \sin^2\delta + L_q \cos^2\delta \end{bmatrix} \begin{bmatrix} i_x \\ i_y \end{bmatrix} + \psi_f \begin{bmatrix} \cos\delta \\ -\sin\delta \end{bmatrix} \quad (3-2)$$

Since $\psi_y = 0$, i_x can be obtained from (3-2):

$$i_x = \frac{2\psi_f \sin\delta - [(L_d - L_q) \cos 2\delta + (L_d + L_q)] i_y}{(L_d - L_q) \sin 2\delta} \quad (3-3)$$

And because $\psi_x = \psi_s$, i_y can be obtained from (3-3):

$$i_y = \frac{1}{2L_d L_q} [2\psi_f L_q \sin\delta - |\psi_s| (L_q - L_d) \sin 2\delta] \quad (3-4)$$

Based on (2-6) and (2-7), replacing i_d, i_q by i_x, i_y through coordinate transformation to obtain the electromagnetic torque equation in xy coordinate system:

$$T_e = \frac{3n_p|\psi_s|}{4L_dL_q} [2\psi_f L_q \sin\delta - |\psi_s|(L_q - L_d)\sin 2\delta] \quad (3-5)$$

Formula (3-5) shows the torque of the motor can be divided into two parts. The first part is the electromagnetic torque of the motor, which is generated by cross-axis armature reaction while the second part is the reluctance torque generated by the motor salient structure including salient pole and non-salient pole two different types [30].

For non-salient pole permanent magnet synchronous motors, since $L_d = L_q = L_s$, the electromagnetic torque equation can be expressed as [31]:

$$T_e = \frac{3n_p}{2L_s} |\psi_s| \psi_f \sin\delta = \frac{3n_p}{2L_s} |\psi_s| \psi_f \sin(w_{sr}t + \delta_0) \quad (3-6)$$

In (3-6), w_{sr} is the rotation angular velocity of stator flux relative to the rotor flux. δ_0 is the initial value of the previous torque angle changing moment. From (3-6), when the stator flux amplitude keeps constant and the torque angle changes from -90° to 90° , the motor torque increases with the torque angle increases. And when the torque angle is 90° , the maximum torque is reached. Calculate derivative respect to time of both sides of the (3-6) to obtain the growth of torque at time $t = 0$ as (3-7):

$$\frac{dT_e}{dt} = \frac{3n_p}{2L_s} |\psi_s| \psi_f w_{sr} \cos\delta \quad (3-7)$$

When the torque angle varied within the range of $-90^\circ \sim 90^\circ$ (motor's stable working area, in which $0 \sim 90^\circ$ is the electric status), right side of (3-7) is always positive, indicating that the torque is increasing with the increase of the torque angle.

For salient pole permanent magnet synchronous motors, $L_d \neq L_q$, doing the same calculation of derivative of (3-5):

$$\begin{aligned} \frac{dT_e}{dt} \Big|_{t=0} &= \frac{3n_p|\psi_s|}{4L_dL_q} [2\psi_f L_q w_{sr} \cos\delta - 2|\psi_s|(L_q - L_d)w_{sr} \cos 2\delta] \\ &= \frac{3n_p|\psi_s|}{4L_d} [2\psi_f \cos\delta - 2|\psi_s|(1 - 1/\rho) \cos 2\delta] w_{sr} \end{aligned} \quad (3-8)$$

In (3-8), $\rho = L_q/L_d$ is called salient factor. As (3-8) indicated, salient factor has a great influence on the change of torque and torque angle of the motor. When $\rho < 1$, (3-8) is always positive so that the motor is at electric status, which means that the relationship between torque change and torque angle change is the same as the non-salient pole PMSM. When $\rho > 1$, quadrature-axis inductance is greater than the direct-axis inductance, so the reluctance torque is negative, therefore the overall trend of the motor torque and torque angle may be inconsistent because of the influence of reluctance torque. In order to ensure the consistency of both total torque and torque angle changing, the growth of torque needs to meet the formula below [32]:

$$\frac{dT_e}{dt}|_{\delta=0} \geq 0 \quad (3-9)$$

The following condition can be deduced from (3-8) and (3-9):

$$|\psi_s| \leq \frac{L_q}{L_q - L_d} \psi_f = \frac{\rho}{\rho - 1} \psi_f \quad (3-10)$$

So when using direct torque control of permanent magnet synchronous motors at $\rho > 1$ condition, the given value of the stator flux amplitude is required to meet the conditional expression (3-10), in order to ensure the consistency of motor output torque, which is to ensure the motor torque can be controlled effectively by controlling the torque angle [33].

When the stator flux maintains at a constant value, the torque of the motor changes due to the torque angle varies. Under normal circumstances, the electrical time constant is much smaller than the mechanical time constant, so change the rotational speed of the stator flux is easier than rotor flux. Thus changing the angle of the torque can be achieved by varying the rotation speed and direction of the stator flux. In summary, the basic idea of direct torque control theory of permanent magnet synchronous motor is to hold the motor flux amplitude constant under the condition that motor torque and torque angle change at the same direction, and control the angle between stator and rotor flux to control motor torque. What's more, given the close relationship between torque and torque angle, to regulate the torque quickly can be realized by changing the torque angle rapidly. [34][35]

The discussion above illustrates the core idea of DTC clearly. Then from Section 3.2, each function part of the DTC system will be introduced separately, and in Section 3.6, the complete system diagram will be given to show that how these parts working together to realize the direct torque control of PMSM.

3.2 Space vector control

Stator voltage space vector u_s is defined as follow [36]:

$$u_s = \frac{2}{3} (u_a + u_b e^{j2\pi/3} + u_c e^{j4\pi/3}) \quad (3-11)$$

Figure 3.1 Three-phase voltage-source inverter [37]

In (3-11), u_a , u_b , u_c are the instantaneous phase voltage of three-phase power supply. When a voltage source inverter shown in Figure 3.1 powers the permanent magnet synchronous motor, the voltages (u_a , u_b , u_c) are entirely determined by the three switching states S_a , S_b and S_c . If $S_a = 1$, then $u_a = U_{dc}$,

otherwise $u_a = 0$, which is similar to u_b and u_c . So there are six non-zero voltage vectors: $U_1(100)$, $U_2(110)$, $U_3(010)$, $U_4(011)$, $U_5(001)$, $U_6(101)$ and two zero-voltage vectors: $U_0(000)$, $U_7(111)$. Six non-zero voltage vectors are different from each other at 60° , which are shown in Figure 3.2, and the inverter output voltage can also be represented by (3-12) [38]:

$$\begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_a \\ S_b \\ S_c \end{bmatrix} U_{dc} \quad (3-12)$$

Figure 3.2 Space vector voltages of three-phase inverter [39]

According to the unified theory of motor, the relationship between stator flux and the input voltage is:

$$\psi_s = \int (u_s - R_s i_s) dt \quad (3-13)$$

Ignoring the resistance gives that:

$$\psi_s = \int u_s dt \quad (3-14)$$

The discretization of (3-14) is:

$$\psi_s = u_s t + \psi_{s0} \quad (3-15)$$

where ψ_{s0} is the initial value of flux. This suggests that the stator flux vector ψ_s will follow the direction of voltage vector u_s at the speed proportional to the input voltage to move. So control the input voltage vector u_s of permanent magnet synchronous motor can realize precise control of the amplitude, rotation direction and speed of stator flux.

To facilitate the selected voltage vector to control the amplitude of stator flux, the voltage vector plane is divided into six parts shown as Figure 3.3. In each sector, select one of the two adjacent voltage vectors to control the amplitude of the flux ψ_s . For example, when the flux at U_1 range and at counterclockwise rotation, vector voltage application of U_2 can increase the amplitude of the flux, on contrary, U_3 can reduce the amplitude of the flux. Thus, select the appropriate voltage vector can control the amplitude of the flux to ensure the flux changes in the certain range and direction. [40][41]

Figure 3.3 The control circle of stator flux

In direct torque control strategy, the amplitude of the stator flux maintains a certain value by the above-mentioned voltage vector control. Then control the angle δ between stator and rotor flux to change the amplitude of torque, and the angle δ can be adjusted by controlling the voltage space vector. The rotation speed of the rotor flux remaining unchanged, therefore the angle δ can also be adjusted by adjusting the stator flux instantaneous rotational speed.

Assuming that the motor's rotor is rotating counterclockwise, if the actual torque is less than the given value, select the voltage vector that makes the stator flux rotate counterclockwise, thus the angle δ increases so the actual torque increase. Once the actual torque is higher than the given value, select the voltage vector that makes the stator flux rotate at the opposite direction, which results in the lower angle of δ . In this way of selecting voltage vector, the stator flux rotates as expected.

The above-mentioned way of control is widely realized by two hysteresis controllers based on Bang-Bang control theory in traditional direct torque control with the estimation of flux and torque.

3.3 Parameter acquisition model

In the permanent magnet synchronous motor direct torque control system, the amplitude and angle of stator flux are required. Therefore we need to estimate the amplitude and the components of stator flux in $\alpha\beta$ coordinate system. The stator flux cannot be directly detected, but can be acquired by calculation of the stator voltage, stator current and stator resistance. And there are two ways to get the stator flux by calculation:

- 1) The voltage model: The stator flux can be calculated from (2-11) and (2-12). This model is generally used for flux estimation because this method adopted for the stator winding back EMF integral operation, which is the simplest and most commonly used way. But when the motor is running at low speed, the back EMF is low and the role of $R_s i_s$ can not be ignored, since it can offset the back EMF, which will cause drift and error accumulation between the calculated value and actual value. In such condition, only close-loop method, for example Extended Kalman Filter (EKF), which is a non-linear adaptive filter in finding the best estimate from the noisy data, can be used. [42][43] However, due to less motor parameters required and no need of the motor speed information, in this way, voltage model usually used for calculating flux amplitude in simple structure permanent

magnet synchronous motor direct torque control system. When the motor speed is high, the stator voltage drop $R_s i_s$ is far less compared to the back EMF, even can be ignored, so this model can calculate the amplitude of stator flux when motor runs at high speed [44].

- 2) The current model: The stator flux can be calculated from (2-6). In (2-6), both equations are represented in dq coordinate system, so the rotation angle of rotor θ is required to perform coordinate transformation of current. And in the absence of the rotor position sensor, the rotor position needs to be estimated. In fact, for realizing the hysteresis control of torque and stator flux, the inverter switching frequency must be very high, which requires the estimation of stator flux to be completed in the same time order at least. This can be done in voltage model, but is difficult to realize in the current model, because it not only requires the measurement of the rotor position, but also the communications between the rotor position sensor and the motor control module, furthermore, the integral of the voltage model itself has filtering properties, but in the current model it includes all current harmonics, which can result in a huge decline of estimation accuracy. [45][46] For all these reasons, the two are unlikely to complete the stator flux estimation in the same time order. In addition, the accuracy of estimated flux in current model is also depends on the current parameters of the model whether is consistent with the actual value, so there is also a need for testing the relevant parameters.

Similarly, we must firstly know the feedback value of the electromagnetic torque to achieve the electromagnetic torque control in the permanent magnet synchronous motor direct torque control system. For the current model technology, to measure electromagnetic torque directly is very difficult, so it requires the use of indirect method to acquire the electromagnetic torque. The most common way is to use (2-13) to calculate electromagnetic torque feedback value, namely the estimation value.

3.4 Hysteresis control of flux and torque

The control of flux and torque are based on the estimation of flux and torque and realized by two hysteresis controllers, namely hysteresis comparators. The operation principle of hysteresis comparator is: when the error of current value and the given value is within the tolerance range of the hysteresis comparator, the comparator's output remains unchanged, but once exceeded this range, the hysteresis comparator will give the corresponding value [47]. The hysteresis torque and flux hysteresis control principle are shown in Figure 3.4 and Figure 3.5.

Figure 3.4 Torque hysteresis comparator

Figure 3.5 Flux hysteresis comparator

For torque hysteresis comparator, T_e^* is the reference torque value while T_e is the current actual electromagnetic torque value. There are three critical values of torque hysteresis comparator, respectively, ΔT_e , $-\Delta T_e$ and 0. When $T_e^* - T_e > \Delta T_e$, which means the estimated torque value is smaller than the reference torque value, the electromagnetic torque needs to be increased. On contrary, when $T_e^* - T_e < -\Delta T_e$, which means the estimated torque value is bigger than the reference torque value, the electromagnetic torque needs to be decreased. When $-\Delta T_e < T_e^* - T_e < \Delta T_e$, which means the error between estimated value and reference value is within the tolerance range of the system, there is no need to make any adjustment. So the value of ΔT_e should be determined very carefully since either too small or too large will increase the instability of the system. [48][49]

For flux hysteresis comparator, similarly, $|\psi_s^*|$ is the reference flux value while $|\psi_s|$ is the current actual flux value. And there are also three critical values of flux hysteresis comparator, respectively, $\Delta\psi$, $-\Delta\psi$ and 0. When $|\psi_s^*| - |\psi_s| > \Delta\psi$, which means the estimated flux value is smaller than the reference flux value, the flux needs to be increased. On contrary, when $|\psi_s^*| - |\psi_s| < -\Delta\psi$, which means the estimated flux value is bigger than the reference flux value, the flux needs to be decreased.

Under the joint action of the torque and flux hysteresis comparator, the moving trail of stator flux should be as Figure 3.3.

3.5 Switch table of traditional DTC

Based on the outputs of two hysteresis controllers, direct torque control can be realized by the switch table which only contains motion vector. Let τ represents the torque status value. When $\tau = 1$, the electromagnetic torque needs to be increased, on contrary, the electromagnetic torque needs to be decreased when $\tau = -1$. Similarly, let ϕ represents the flux status value. When $\phi = 1$, the stator flux needs to be increased and when $\phi = -1$, the stator flux needs to be decreased. The switch table is showed as Table 3.1.

Table 3.1 The switch table based on only motion vectors [50]

ϕ	τ	θ_1	θ_2	θ_3	θ_4	θ_5	θ_6
1	1	$U_2(110)$	$U_3(010)$	$U_4(011)$	$U_5(001)$	$U_6(101)$	$U_1(100)$
	-1	$U_6(101)$	$U_1(100)$	$U_2(110)$	$U_3(010)$	$U_4(011)$	$U_5(001)$
-1	1	$U_3(010)$	$U_4(011)$	$U_5(001)$	$U_6(101)$	$U_1(100)$	$U_2(110)$
	-1	$U_1(100)$	$U_2(110)$	$U_3(010)$	$U_4(011)$	$U_5(001)$	$U_6(101)$

where θ_i shows the space sector of the stator flux, and the value of ϕ and τ are determined by (3-16) and (3-17) [50][51]:

$$\tau(k) = \begin{cases} 1 & T_e^* - T_e > \Delta T_e \\ \tau(k-1) & -\Delta T_e < T_e^* - T_e < \Delta T_e \\ -1 & T_e^* - T_e < -\Delta T_e \end{cases} \quad (3-16)$$

$$\phi(k) = \begin{cases} 1 & |\psi_s^*| - |\psi_s| > \Delta\psi \\ \phi(k-1) & -\Delta\psi < |\psi_s^*| - |\psi_s| < \Delta\psi \\ -1 & |\psi_s^*| - |\psi_s| < -\Delta\psi \end{cases} \quad (3-17)$$

where $\tau(k-1)$ and $\phi(k-1)$ are the status value of torque and flux in the previous control cycle.

As known that in the synchronous motor direct torque control system, the function of zero voltage vectors is to maintain the current torque constantly. Depending on the selected motion vectors and the zero voltage vectors, synchronous motor's torque in switch table can be in three states: increasing, hold and reduction. Based on the role of the zero voltage vectors, there is a new control method, which makes the torque hysteresis comparator become a three-level comparator. When the error between estimated value and reference value is within the tolerance range of the system, select zero voltage vectors to maintain the torque output to reduce torque ripple, thereby improving motor steady-state performance. The torque status value is determined by (3-18) [52], when $\tau = 0$, the system needs to maintain the torque output. The new switch table is shown as Table 3.2.

$$\tau(k) = \begin{cases} 1 & T_e^* - T_e > \Delta T_e \\ 0 & -\Delta T_e < T_e^* - T_e < \Delta T_e \\ -1 & T_e^* - T_e < -\Delta T_e \end{cases} \quad (3-18)$$

Table 3.2 The switch table adopted zero voltage vectors [52][53]

ϕ	τ	θ_1	θ_2	θ_3	θ_4	θ_5	θ_6
1	1	$U_2(110)$	$U_3(010)$	$U_4(011)$	$U_5(001)$	$U_6(101)$	$U_1(100)$
	0	$U_7(111)$	$U_0(000)$	$U_7(111)$	$U_0(000)$	$U_7(111)$	$U_0(000)$
	-1	$U_6(101)$	$U_1(100)$	$U_2(110)$	$U_3(010)$	$U_4(011)$	$U_5(001)$
-1	1	$U_3(010)$	$U_4(011)$	$U_5(001)$	$U_6(101)$	$U_1(100)$	$U_2(110)$
	0	$U_0(000)$	$U_7(111)$	$U_0(000)$	$U_7(111)$	$U_0(000)$	$U_7(111)$
	-1	$U_1(100)$	$U_2(110)$	$U_3(010)$	$U_4(011)$	$U_5(001)$	$U_6(101)$

Development principles of new switching table is similar to the conventional switching table, but one thing should be noted that when $\tau = 0$, the output must be the zero voltage vector. This is because the core of direct torque control is the control of torque, namely torque is the most important. And flux is the second, which means it will not affect the results if do not control flux in one or two cycles, so output zero voltage vector to smooth the torque fluctuations when $\tau = 0$.

3.6 System diagram of traditional DTC

Direct torque control system can be obtained through the above discussion and PMSM schematic is shown in Figure 3.6, which consists of a permanent magnet synchronous motors, inverters, flux and torque calculation and sector judgment module, speed sensors, switches table and regulators module

etc. The working principle and control process is as follows. Detect the inverter's output three-phase current and DC bus voltage firstly. Calculate the stator flux by coordinate transformation and system control law secondly. After that calculate the instantaneous torque of the motor according to the calculated flux and measured current. Then based on stator flux of $\alpha\beta$ -axis to determine the location of its sector θ . The speed regulator is used to determine the torque reference value based on the deviation of the speed reference value and the actual speed value compared to the feedback torque value, then transfer the deviation through the hysteresis comparator to obtain the torque control signal τ . After that the motor speed can be obtained by optical encoder, which can also be estimated by the rotational speed of the stator flux without speed sensor. Compared the reference stator flux value and the actual value to obtain the flux control signal ϕ by transferring the deviation through the same comparison hysteresis comparator. Finally, select voltage vector by the three control signals (θ , τ , ϕ) via the switch table to determine the most appropriate switch status, so that the inverter can drive PMSM. It is the traditional DTC control flow which has already been proved to have high efficiency and low losses in PMSM control. [54][55][56]

Figure 3.6 Traditional Direct Torque Control Schematic of PMSM

In general, traditional direct torque control is suitable for the PMSM control in most situations, however, the torque ripple existed in traditional DTC still can not be ignored especially in low-speed range. In low-speed working area, space vector needs more precise control, which makes 6-sector division seems a little rough. So a 12-sector direct torque control method is developed, which makes control more precisely not only in high-speed region but also in low-speed region.

3.7 Summary

In this chapter, the theory and system of traditional direct torque control method is introduced in detail from basic theory to each function part. Generally speaking, as the name suggests, the core idea of DTC is to control the torque directly, and it realized by controlling the stator flux. Though the traditional DTC is applicable for most situations, the torque ripple problem in DTC still can be a potential threat to the on board work. So in the next chapter, a modified DTC method to reduce torque ripple will be discussed.

IV. MODIFIED DTC METHOD TO REDUCE TORQUE RIPPLE

Different from the ripple torque caused by the hardware mentioned in Section 2.2, the torque ripple discussed in this chapter is caused by the DTC itself. They are not same but not conflict, which means they can both cause damage to the system. Studies have shown that in motor direct torque control system, in some cases, there will be switching table fail phenomenon, which means the practical effect of the voltage vector to torque is contrary to the expectations, which usually happens in the situation of different changing trend of torque and torque angle, and results in unreasonable torque ripple [57][58]. Furthermore, a more precise control method is needed when the motor works at low speed. In order to reduce the torque ripple of the DTC system, a new 12-sector DTC method is developed based on the traditional DTC method.

4.1 Torque ripple's damage to on board work

Torque ripple is the phenomenon that motor's torque fluctuates when it is running. Torque ripple will definitely influence the stability of speed and power of motor, especially when it is larger than the tolerance range. For on board work, winch often drags heavy load and works at low speed range, so the stability of motor is very important. Once the amplitude of torque fluctuation is too large to keep motor working at a stable speed, it will cause the following problems:

- Torque ripples have bad effects on the synchronous motors if they cause speed oscillations which contradict with the advantage of the synchronous motors being have constant speed applications. Also, if these ripples have frequency coinciding with the natural mechanical frequency of the system, severe torsional oscillation will occur, and the shaft may be damaged.
- The energy consumption of motor will increase. Since the very limited on board resources, unnecessary energy lose is the first thing we need to avoid.
- On board staff cannot locate the position of anchor correctly, namely the reliability of the system will be influenced. It can be very dangerous at sea if the sailor cannot set a proper speed for winch because of a wrong anchor position.

Above all, the objective is to minimize the torque ripple while not sacrifice the simplicity of DTC. In the following sections, a modified DTC to reduce torque ripple will be introduced. Firstly, the cause and extent of torque ripple in traditional DTC will be analyzed in Section 4.2.

4.2 Analysis of torque ripple in DTC

According to (3-7), when the change of the magnitude of stator flux and torque angle is consistent, torque and torque angle change consistently, on the contrary, when the change of the magnitude of stator flux and torque angle is conflict, there may be the situation that torque and torque angle change inconsistently.

Take Section I (θ_1) and U_6 in Figure 3.3 as the example. Stator flux rotates counterclockwise and ignores the change of rotor flux. From Table 3.2, the function of U_6 in Section I is to decrease the torque.

Figure 4.1 the effect of U_6 on stator flux in Section I

where ψ_s is the initial stator flux, ψ_s' is the stator flux after one sample time, ψ_r is the rotor flux, δ is the torque angle and θ is the angle between voltage vector and the initial edge of sector I. From Figure 4.1, it follows that

$$\frac{\psi_s'}{\psi_s} = \frac{\sin(\theta + 60^\circ + \lambda)}{\sin(\theta + 60^\circ)} \quad (4-1)$$

So the torque comparison between before and after is

$$\frac{T_e'}{T_e} = \frac{\sin(\theta + 60^\circ + \lambda)\sin(\delta - \lambda)}{\sin(\theta + 60^\circ)\sin(\delta)} \quad (4-2)$$

Considering $T_e' - T_e$ in (4-2) gives that

$$\sin(\theta + 60^\circ + \lambda)\sin(\delta - \lambda) - \sin(\theta + 60^\circ)\sin(\delta) = \sin(\delta - \lambda - \theta - 60^\circ)\sin(\lambda) \quad (4-3)$$

Since $\sin(\lambda) > 0$ and $\theta \in [-30^\circ, 30^\circ]$, from (4-3) we can know that if $\delta - \lambda < 60^\circ$, torque decreases; if $\delta - \lambda > 60^\circ$, the torque increases when stator flux (θ) in the range $[-30^\circ, \delta - \lambda - 60^\circ]$ and the torque decreases in the range $[\delta - \lambda - 60^\circ, 30^\circ]$. So $[-30^\circ, \delta - \lambda - 60^\circ]$ is the range where the expected and actual effect of U_6 may cause conflict. And similar conclusion can be drawn for the rest voltage vectors. So, in any sections, delimit $-60^\circ, -30^\circ, 0^\circ$ are the three flux-increase with torque-angle-decrease vectors, and $120^\circ, 150^\circ, 180^\circ$ are the three flux-decrease with torque-angle-increase vectors. They can all cause this kind of conflict in torque control and this is also the reason why torque ripple exists in DTC method.

So if we can narrow the possible conflict range, then the torque ripple will be controlled in an acceptable area.

4.3 12 Space vectors' control

According to (3-7), control the torque angle can control the torque with the consistent amplitude of flux (including stator flux and rotor flux) [59]. If there are more voltage vectors available in flux space sectors, the effect of torque control will be improved. So the traditional 6 space sectors are divided further into 12 space sectors shown in Figure 4.2. The increased 6 effective space voltage vectors generated by adopting two-phase conducted way of inverter. Therefore, the control strategy of 12 sectors increases

the available voltage vector to the maximum extent while does not change the premise of the conventional voltage source inverter circuit topology.

Figure 4.2 12 output voltage vectors of voltage source inverter

Under the three-phase and two-phase mixed conduction mode, the output 12 effective voltage vectors of voltage source inverter are shown in Table 4.1.

Table 4.1 12 output voltage vectors of voltage source inverter

	S_a	S_b	S_c	Amplitude	Phase
U_1	1	0	0	$2U_{dc}/3$	0
U_2	1	-1	0	$U_{dc}/2$	$\pi/6$
U_3	1	1	0	$2U_{dc}/3$	$\pi/3$
U_4	-1	1	0	$U_{dc}/2$	$\pi/2$
U_5	0	1	0	$2U_{dc}/3$	$2\pi/3$
U_6	0	1	-1	$U_{dc}/2$	$5\pi/6$
U_7	0	1	1	$2U_{dc}/3$	π
U_8	0	-1	1	$U_{dc}/2$	$7\pi/6$
U_9	0	0	1	$2U_{dc}/3$	$4\pi/3$
U_{10}	-1	0	1	$U_{dc}/2$	$3\pi/2$
U_{11}	1	0	1	$2U_{dc}/3$	$5\pi/3$
U_{12}	1	0	-1	$U_{dc}/2$	$11\pi/6$

S_a, S_b, S_c , respectively, represents three-phase bridge inverter output, where 1 means upper-bridge is turned on so the lower-bridge is shutdown, while 0 represents the opposite of 1, and -1 indicates that both the upper and lower bridges are switched off. U_{dc} is the DC bus voltage. Phase is the angle between voltage vector and α -axis.

Table 4.1 also shows that the amplitude of the new 6 voltage vectors (U_{even}) is less than the amplitude of the old 6 voltage vectors (U_{odd}) in traditional DTC. Direct torque control method using hysteresis comparator to qualitatively control the amplitude of torque and stator flux. The phase angle of the voltage vector is to determine the increase or decrease of stator flux and torque, while the role of the amplitude is to decide how much the fluctuations on the torque and stator flux, so the amplitude of the voltage vector change does not affect feasibility of direct torque control. Therefore, there is no need to enhance the voltage vector amplitude in two-phase conduction mode to the same as the three-phase conduction mode.

4.4 Improvement and switch table of 12-sector DTC

After the modified 12-sector DTC adopted, we do the similar analysis as in Section 4.1, Section I ($[-30^\circ, 30^\circ]$) in traditional DTC has been divided into Section I ($[0^\circ, 30^\circ]$) and Section XII ($[-30^\circ, 0^\circ]$) in the 12-sector DTC. Table 4.2 shows the comparison of conflict range in the two kinds of DTC.

Table 4.2 Comparison of conflict section

	Section	Conflict range
6-sector DTC	Section I ($[-30^\circ, 30^\circ]$)	$\theta \in [-30^\circ, \delta - \lambda - 60^\circ]$
12-sector DTC	Section I ($[0^\circ, 30^\circ]$)	$\theta \in [0^\circ, \delta - \lambda - 60^\circ]$
	Section XII ($[-30^\circ, 0^\circ]$)	$\theta \in [-30^\circ, \delta - \lambda - 90^\circ]$

From Table 4.2 we can easily get that the possible conflict can be avoided in the range $[\delta - \lambda - 90^\circ, 0^\circ]$ when $\delta - \lambda > 60^\circ$. So the effect of 12-sector DTC can minimize the conflict range to maximum 50% compared to traditional DTC.

As mentioned above, there are three similar function vectors, -60° , -30° and 0° vector. With the same analysis, the conflict range of -30° and 0° are $[0^\circ, \delta - \lambda - 30^\circ]$ and $[0^\circ, \delta - \lambda]$ in section I of 12 sectors. So choose the -60° vector when $\phi = 1$ and $\tau = 0$ and the 120° vector when $\phi = 0$ and $\tau = 1$ to reduce the range of unreasonable torque ripple to the maximum extent, thereby reducing the torque ripple. The new switch table is developed as Table 4.3. In addition, different from the studies so far, zero voltage vectors $U_0(000)$ and $U_{13}(111)$ are also used in this thesis.

Table 4.3 Improved 12-sector DTC switch table

ϕ	τ	θ_1	θ_2	θ_3	θ_4	θ_5	θ_6	θ_7	θ_8	θ_9	θ_{10}	θ_{11}	θ_{12}
I	I	$U_3(110)$	$U_5(010)$	$U_5(010)$	$U_7(011)$	$U_7(011)$	$U_9(001)$	$U_9(001)$	$U_{11}(101)$	$U_{11}(101)$	$U_1(100)$	$U_1(100)$	$U_3(110)$
	0	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$	$U_0(000)$
	-I	$U_{11}(101)$	$U_{12}(10-1)$	$U_1(100)$	$U_2(1-10)$	$U_3(110)$	$U_4(-110)$	$U_5(010)$	$U_6(01-1)$	$U_7(011)$	$U_8(0-11)$	$U_9(001)$	$U_{10}(-101)$
-I	I	$U_5(010)$	$U_6(01-1)$	$U_7(011)$	$U_8(0-11)$	$U_9(001)$	$U_{10}(-101)$	$U_{11}(101)$	$U_{12}(10-1)$	$U_1(100)$	$U_2(1-10)$	$U_3(110)$	$U_4(-110)$
	0	$U_0(000)$	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$	$U_0(000)$	$U_{13}(111)$
	-I	$U_9(001)$	$U_{11}(101)$	$U_{11}(101)$	$U_1(100)$	$U_1(100)$	$U_3(110)$	$U_3(110)$	$U_5(010)$	$U_5(010)$	$U_7(011)$	$U_7(011)$	$U_9(001)$

After using 12-sector DTC method, the torque ripple will be reduced to some extent compared to the traditional direct torque control theoretically. In the following simulation part, we can see that the torque ripple is reduced obviously by using the modified DTC method.

4.5 Summary

In this chapter, firstly, the influences of torque ripple on motor and on board work are discussed. Then the cause of torque ripple in DTC is analyzed, which make a strong foundation of the improvement work. After that, based on the analysis and traditional DTC, the space vectors are extended to 12 to smooth the switch process. And it is proved to be effective to reduce maximum 50% torque ripple. Finally choose the proper vector to build the new switch table. In conclusion, Chapters III and IV introduced every aspect of DTC, and the 12-sector DTC discussed in this chapter theoretically can be the main control algorithm of winch system. So in the next chapter, the last part of winch system, the sensorless control of PMSM that especially designed for DTC method and marine condition will be introduced.

V. SENSORLESS CONTROL OF WINCH DRIVE SYSTEM

After the introduction of PMSM and DTC in Chapters II, III and IV, the last part of electrical driven winch system is the sensorless control. Considered the marine condition, it is hard to set up speed or position sensors on board, so the sensorless control for the system is very necessary. In the direct torque control method introduced in the previous parts, position sensor is not used, so the sensorless control methods mentioned in this part means control methods without speed sensor. Speed feedback determines the reference torque value, one of the three control signals. In order to acquire the precise speed value of PMSM in winch system working in whole speed range, we will separate high-speed region and low-speed region, which means we will adopt different strategies according to different speed ranges. And both sensorless control methods must be compatible with the DTC method, in other words, not compromise the simplicity of DTC method.

5.1 Categories of sensorless control methods

To achieve sensorless control of permanent magnet synchronous motors, information on the rotor speed is extracted from measured stator currents and voltages at motor terminals. Open-loop estimators or closed-loop observers are used for motor position estimation. By exploiting the non-ideal property of motor to acquire position information, sensorless control is extended to low speed range even including zero [60].

Permanent magnet synchronous motor sensorless control methods can be roughly divided into three categories. One is open-loop calculation based on the ideal model of motor, another is close-loop calculation based on the various observation models, and the last one is the algorithm based on the non-ideal characteristics of the motor as the typical representative high frequency (*hf*) injection method.

The open-loop algorithm is based on the mathematical model of the motor, such as direct calculation method, estimation method based on the inductance change and back-EMF integration method. The calculating process in such methods is simple, and can get more accurate estimation of rotor position under the premise of precise motor parameters.

The principle of open-loop calculation method is simple and has fast dynamic response. But such methods are highly vulnerable to changes in parameters, so these algorithms are unstable. With the development of automatic control technology, domestic and foreign scholars have studied various observation-based closed-loop algorithms, which are widely used currently as extended Kalman filter, sliding mode observer, model reference adaptive algorithms and other adaptive algorithms [61]. The observation accuracy and system stability have been greatly improved. However, all of these applied to high-speed operation sensorless control technologies are directly or indirectly based on the extracted position signal from the back EMF, then estimate the rotor position by calculating the induced electromotive force from the motor's voltage equation. Since the back EMF is proportional to the magnitude and speed, when the speed is very low or zero, the signal to noise ratio of counter electromotive force is small, combined with other confounding factors, these methods cannot accurately estimate the rotor speed and position in low-speed region.

To solve the speed observation problem at low speed, Matthew J. Corley and R.D. Lorenz proposed high frequency (*hf*) signal injection method of permanent magnet synchronous motor for rotor position estimation at low and zero speed [62], which utilize the salient features of salient pole machine itself or

the salient characteristics performance in a non-salient pole machine under the high-frequency signal to estimate the rotor position and speed. Nowadays, it becomes the most effective way of permanent magnet synchronous motors running at low speed without mechanical sensors. So the high frequency signal injection method is mainly concerned in this thesis considered the low-speed requirement.

High frequency injection method obtains the rotor position and speed by injecting high-frequency voltage or current into motor to detect the corresponding voltage or current. Depending on the injected signal, high-frequency injection methods can be divided into the three categories in [63]: rotating high frequency voltage signal injection, rotating high frequency current signal injection and fluctuating high frequency voltage signal injection. Considered the complexity of implementation, current injection is not a possible way to realize on board. The rotating high frequency injection method has been studied and used by many people but the other one is rarely mentioned in the same situation especially combined with DTC. What's more, in [64], it compared the rest two high frequency injection methods and drew the conclusion that both methods can be effective to detect the spatial position of the rotor over the entire speed range, but the rotating high frequency voltage signal injection method system is simpler with higher tracking accuracy, better static and dynamic performance and the most important, can realize perfectly with the traditional or 12-sector DTC. So the rotating high frequency voltage signal injection method will be the prior choice of the winch system.

5.2 Rotating *HF* voltage signal injection method for low-speed region

5.2.1 Theory of rotating *hf* voltage signal injection

In rotating *hf* voltage signal injection method, the high frequency rotating carrier is injected into the $\alpha\beta$ coordinate system. The most used way of injection is to superpose a high frequency voltage carrier on the fundamental excitation, so that the position information of rotor can be acquired from the resulting high frequency current components. And in order to reduce the limitations in the use of band-pass filters, the injected voltage carrier's frequency, ω_i , should be high enough to make sure it can be spectrally separated from the fundamental excitation sufficiently [65]. One kind of the injected high frequency carrier is as below:

$$\begin{bmatrix} u_{\alpha i} \\ u_{\beta i} \end{bmatrix} = V_i \begin{bmatrix} -\sin(\omega_i t) \\ \cos(\omega_i t) \end{bmatrix} \quad (5-1)$$

where V_i is the amplitude of voltage and ω_i represents the rotating frequency.

The voltage model of a salient pole PMSM discussed in chapter II can be written as follows, using $L = (L_d + L_q)/2$ and $\Delta L = (L_d - L_q)/2$, where θ_r is the rotor position:

$$\begin{bmatrix} u_\alpha \\ u_\beta \end{bmatrix} = R_s \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} L - \Delta L \cos(2\theta_r) & -\Delta L \sin(2\theta_r) \\ -\Delta L \sin(2\theta_r) & L + \Delta L \cos(2\theta_r) \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \psi_f \frac{d}{dt} \begin{bmatrix} \cos(\theta_r) \\ \sin(\theta_r) \end{bmatrix} \quad (5-2)$$

The carrier frequency mostly used is between 500Hz to 2000Hz, then the stator inductance will dominate the stator resistance, as $\omega_i L_s \gg R_s$, thus (5-2) can be approximately to:

$$\begin{bmatrix} u_{\alpha i} \\ u_{\beta i} \end{bmatrix} \approx \frac{d}{dt} \begin{bmatrix} L - \Delta L \cos(2\theta_r) & -\Delta L \sin(2\theta_r) \\ -\Delta L \sin(2\theta_r) & L + \Delta L \cos(2\theta_r) \end{bmatrix} \begin{bmatrix} i_{\alpha i} \\ i_{\beta i} \end{bmatrix} \quad (5-3)$$

From (5-1) to (5-3), the resulting high frequency current is as follows:

$$\begin{bmatrix} i_{\alpha i} \\ i_{\beta i} \end{bmatrix} = \frac{V_i}{\omega_i L_d L_q} \begin{bmatrix} L \cos(\omega_i t) + \Delta L \cos(2\theta_r - \omega_i t) \\ L \sin(\omega_i t) + \Delta L \sin(2\theta_r - \omega_i t) \end{bmatrix} \quad (5-4)$$

Equation (5-4) shows that both negative sequences components where the information of rotor position is contained and the positive sequence components are both included in the resulting high frequency current. To extract the information of position angle, a band-pass filter is firstly used to separate the high frequency current components from the fundamental excitation, and then the rotor position signal, θ_r , can be demodulated by the synchronous filter [66]. Figure 5.1 shows the resulting *hf* current demodulation block diagram:

Figure 5.1 The demodulation block diagram

From Figure 5.1, after the band-pass filter, using Park transformation with rotating angle $-\theta_i$ ($[T_{(-\theta_i)}]$) to transform the high frequency current $i_{\alpha\beta i}$ to a reference coordinate system rotating synchronously with the injected signal, so then the components of high frequency carrier become DC with the negative sequence component which is located at $-2\omega_i$, shown as (5-5) where $\theta_i = \int \omega_i t$:

$$\begin{bmatrix} i_{dp} \\ i_{qp} \end{bmatrix} = [T_{(-\theta_i)}] \cdot \begin{bmatrix} i_{\alpha i} \\ i_{\beta i} \end{bmatrix} = \frac{V_i}{\omega_i L_d L_q} \begin{bmatrix} L + \Delta L \cos(2\theta_r - 2\omega_i t) \\ L + \Delta L \sin(2\theta_r - 2\omega_i t) \end{bmatrix} \quad (5-5)$$

Then the high-pass filter is used to block the DC component. Then using the similar Park transformation ($[T_{(2\theta_i)}]$) to transform the resulting signal to a coordinate system which is synchronous rotating as the negative sequence, so that rotor position signal can be obtained as (5-6):

$$\begin{bmatrix} i_{dn} \\ i_{qn} \end{bmatrix} = [T_{(2\theta_i)}] \cdot \begin{bmatrix} i_{dp} \\ i_{qp} \end{bmatrix} = \frac{V_i}{\omega_i L_d L_q} \begin{bmatrix} \Delta L \cos(2\theta_r) \\ \Delta L \sin(2\theta_r) \end{bmatrix} \quad (5-6)$$

Finally simply use the arctan function to obtain the rotor angle information:

$$\theta_r = \frac{1}{2} \arctan\left(\frac{i_{qn}}{i_{dn}}\right) \quad (5-7)$$

5.2.2 HF injection in a 6/12-sector DTC drive

Simplicity and fast dynamic response are the biggest features of DTC system. However, the high frequency voltage injection is quite hard to realize because the torque and flux errors are converted to the switching signals directly by switch table. The use of SVPWM can solve this kind of problem but also will sacrifice the simple structure of DTC drive. Since the on board drive ought to be as simple as possible, a new approach needs to be found.

Considering the switch table is the only determination of switches' status, the only way to inject a rotating high frequency voltage carrier is to modify the DTC vector pattern determined by the switch table directly. Since the 12-sector DTC has the similar vector pattern as the traditional DTC, we take the 6-sector DTC as the example.

A very short period of time T_{inj} in a complete DTC sampling period T_s ($T_{inj} \ll T_s$) is used to inject the high frequency voltage carrier shown in Figure 5.2:

Figure 5.2 Modified DTC pattern

In modified DTC pattern showed in Figure 5.2, $V_i, V_j (V_k, V_l)$ are the original voltage vectors applied according to the DTC switch table, while V_n is the injected *hf* voltage carrier ($1 \leq n \leq 6(12)$), which means the 6 or 12 vectors are consecutively injected in every DTC sampling period to superpose the high frequency voltage carrier on the base voltage vectors. Figure 5.3 shows the superposition process where ω_e is the stator flux rotating angular velocity.

Figure 5.3 Superposition of high frequency carrier on the voltage vectors

What's more, the *hf* carrier injection does not need to be done in every sampling period, which means the injection can happen in alternate periods according to the frequency requirement. So it's very easy to change both frequency and amplitude of the injected carrier. The amplitude depends on T_{inj} , while the frequency f_{ic} decided by the repetition rate R of each injected vector shown in Figure 5.2. The repetition rate R represents the times one vector injected, for example, if $R = 2$, the sequence of injected voltage vectors at the end of each sample period could be $(V_1, V_1, V_2, V_2, V_3, V_3 \dots)$. And since the vectors

are injected sequentially at each sample period T_s , the injected carrier's frequency can be obtained as (5-7):

$$f_{ic} = \frac{1}{6 \cdot T_s \cdot R} \left(\text{or } \frac{1}{12 \cdot T_s \cdot R} \right) \quad (5-8)$$

where $R \geq 1$, and 6 or 12 are depended on which of the two DTC methods is adopted. And since the DTC sampling frequency should be high enough, the frequency of injected carrier will not be restricted.

After the implement of the hf carrier injection strategy mentioned in Section 5.2.2, then using the common knowledge of high-frequency voltage injection theory introduced in Section 5.2.1, the high frequency voltage injection method for sensorless control of winch drive system in low-speed range can be fulfilled without sacrificing the simplicity of DTC method.

So the above-mentioned hf voltage injection method uses the existed voltage vectors to the maximum extent to solve the hf carrier injection problem without the help of PWM. In this way, the low-speed sensorless control method can be successfully applied. In Section 5.3, the high-speed sensorless control method will be discussed. Since the Back-EMF integral method is already widely used in industrial application, it is much simpler than the low-speed strategy.

5.3 Back-EMF integral method for mid/high-speed region

Compared to the high-frequency injection, Back-EMF integral is much easier to realize. Since PMSM is a synchronous motor, when the motor is at stable operation, the stator and rotor flux rotates synchronously, and the angle difference between them is the load torque angle, so the rotor position can get by calculating the stator flux phase angle directly [67]. The stator flux can be obtained by the integration of counter electromotive force using voltage equation:

$$\psi_d = \int (u_d - R_s i_d) dt \quad (5-9)$$

$$\psi_q = \int (u_q - R_s i_q) dt \quad (5-10)$$

So the stator flux angle is:

$$\theta_r = \arctan(\psi_q / \psi_d) \quad (5-11)$$

And in order to suppress the zero drift, a feedback adjustment coefficient will be introduced into the structure.

Though this is an open-loop calculation method, it can still keep good performance at high speed because the motor's parameters don't change easily due to the temperature or magnetic saturation in high-speed region. And considered the low-speed strategy has already brought a lot of complexity into the system, the Back-EMF integral method is very proper for its simple structure and good dynamic performance.

5.4 Summary

This chapter introduces the sensorless control strategies of electrical driven winch system. At first, the different sensorless control strategies are compared to choose the most proper two for the entire speed range application. And after selected high-frequency voltage injection method as low-speed strategy, in Section 5.2, modified vector pattern makes the high frequency injection in switch table based DTC applicable and do not compromise its simplicity at meantime. Finally, Back-EMF integral method introduced in Section 5.3 has already proved to become a simple and mature way of sensorless control in mid or high-speed region. With the end of this chapter, all the three main parts of the electrical driven winch system have been introduced theoretically. In order to verify all theories discussed in the previous chapters, Chapter VI will show the simulation of the overall system and the analysis of the simulation results.

VI. SIMULATION AND ANALYSIS OF RESULTS

According to the direct torque control system introduced in the Chapters III and IV and the sensorless control methods illustrated in chapter V, the simulation model of traditional and modified sensorless direct torque control of permanent magnet synchronous motor system is established with the help of the PSIM software. PSIM is a simulation software specifically designed for power electronics and motor drives. PSIM provides a powerful simulation environment for power electronics, analog and digital control, magnetics, and motor drive system studies with fast simulation and friendly user interface [68].

6.1 Simulation

6.1.1 Diagram and schematic of simulation

According to Figure 3.6, the simulation system diagram is shown as Figure 6.1. DTC system includes PMSM module, inverter module, parameter detecting module, flux and torque estimation module, hysteresis control module and switch table module etc. Among them, the last three modules are realized by code in C-block function. The sensorless control part has speed estimation module and speed control module.

Figure 6.1 Diagram of simulation model

According to Figure 6.1, the simulation schematic is established as Figure 6.2, and each part of the system is marked clearly in red blocks.

From Figure 6.2, there are seven modules in the simulation schematic: the main circuit module, the PMSM module, the stator voltage acquire module, the rotor position estimation module, the speed control module, the flux estimation module and the C block controller module. Their components and the functions will be introduced in detail as following.

- 1) The main circuit module is the hardware drive circuit, so it consists of the DC power, inductor and capacitor for filtering, six MOSFETs as three-phase inverter.

Figure 6.3 Main circuit

- 2) The PMSM module includes the permanent magnet synchronous motor, mechanical load, speed sensor and position sensor (no use in this simulation, just for testing).

Figure 6.4 PMSM module

- 3) The stator voltage acquire module is realized by calculating the stator voltage according to (3-12), where S_a , S_b , S_c are the output signal of switch table:

Figure 6.5 Stator voltage acquire module

- 4) The rotor position estimation module is the core part of sensorless control. The left part of Figure 6.6 is the *hf* carrier injection part. S1 to S3 are the switch table value and HF1 to HF3 are the *hf* injection switch status value. They are both determined in C block controller. So according to the vector pattern in Figure 5.2, the unique way of *hf* carrier injection is realized by the combination of several logical gates, which are used to determine whether to adopt DTC switch table value or *hf* injection switch status value to the inverter. The synchronous filter and back-EMF integral on the right part of the figure are just realized by using the corresponding function blocks in the simulation software according to the theory diagram in Figure 5.1 and the calculation formula (5-10).

Figure 6.6 Rotor position estimation module

- 5) The speed control module, as shown in Figure 6.7, consists of the speed estimation part and the PI part. The speed estimation part is just using the differentiator to the rotor position information got in the rotor position estimation module to calculate the actual speed of motor. Then a PI regulator is adopted to eliminate the error between the actual speed and the reference speed in the PI part. Ultimately, the control of speed is realized by controlling torque in DTC, so the output of the PI regulator is set as the reference value of torque. The parameters of PI regulator are adjusted manually according to the results of a lot of experiment tests.

Figure 6.7 Speed control module

- 6) The flux estimation module is established according to (2-11) and (2-12), as shown in Figure 6.8:

Figure 6.8 Flux estimation module

- 7) Then the outputs of the above-mentioned modules: i_α , i_β , u_α , u_β , ψ_α , ψ_β , U_{dc} , w , θ (position) and T_e * all enters the controller, a C-function block. The rest modules are all realized in this C block by code. The output includes the switch value of DTC and hf injection and some testing signals. The controller C block is shown in Figure 6.9.

Figure 6.9 Controller C block

Since the C block controller has some of the core control algorithms, the code will be introduced separately in Section 6.1.2 in detail.

6.1.2 Functions in C block (Controller)

The variable definition of the controller including the reference value of stator flux, the parameters of motor, and some variables used in DTC algorithm like voltages, currents, flux, torque and so on as shown in Figure 6.10:

```
#define FSET 1.5
#define Ld 3.16e-2
#define Lq 3.16e-2
#define Rs 2.38
#define PI 3.1415926535

int flagf;
int flagt;
double dUa,dUb,dUdc;
double dIa,dIb,dId,dIq;
double dFa,dFb,dFs;
double dTe,TSET;
double dDeltaT,dDeltaF;
int sector = 0;
int vector = 0;
```

Figure 6.10 Variable definitions

Figure 6.11 shows the calculation of flux based on (6-1) and the torque calculation based on (2-13).

$$\psi_s = \sqrt{\psi_{s\alpha}^2 + \psi_{s\beta}^2} \quad (6-1)$$

```
// calc fai
dFs = sqrt(dFa * dFa + dFb * dFb);

//calc Te
dTe = 1.5*2*(dFa*dIb - dFb*dIa);
```

Figure 6.11 Flux and torque calculation

There are three status value (θ , τ , ϕ) need to be determined as mentioned in Chapter III. τ and ϕ are determined by the flux and torque hysteresis control showed as Figure 6.12 while the sector (θ) judgment is realized by using the space relationship between ψ_s , $\psi_{s\alpha}$ and $\psi_{s\beta}$ as Figure 6.13.

```
//calc flux and torque status value
dDeltaT = TSET - dTe;
dDeltaF = FSET - dFs;
if(dDeltaF > 0.01) flagf = 1;
if(dDeltaF < -0.01) flagf = -1;
if(dDeltaT > 0.1) flagt = 1;
if(dDeltaT < -0.1) flagt = -1;
if(dDeltaT >= -0.1 && dDeltaT <= 0.1) flagt = 0;
```

Figure 6.12 Calculation of τ and ϕ

```

// Judge sector
if (dFa >= 0)
{
    if( dFb > (dFs/2)) sector = 2;
    if( -dFb > (dFs/2)) sector = 6;
    if( dFb <= (dFs/2) && dFb >= -(dFs/2)) sector = 1;
}
else
{
    if( dFb > (dFs/2)) sector = 3;
    if( -dFb > (dFs/2)) sector = 5;
    if( dFb <= (dFs/2) && dFb >= -(dFs/2)) sector = 4;
}

```

Figure 6.13 Calculation of θ

The role of switch table is to determine which space vector will be used for maintaining rotation of motor. In this thesis, the switch table with zero voltage vectors is used and Figure 6.14 shows one of the situation when $\tau = 1$ and $\phi = 1$. The rest 5 situations are similar, so no more repeat here.

```

//judge vector
if( flagf == 1 && flagt == 1)
{
    switch(sector)
    {
        case 1 : vector = 2; break;
        case 2 : vector = 3; break;
        case 3 : vector = 4; break;
        case 4 : vector = 5; break;
        case 5 : vector = 6; break;
        case 6 : vector = 1; break;}
}

```

Figure 6.14 Switch table (section I)

After the appropriate space vector is selected, the final step is to realize switch table shown as Figure 6.15. The value of out [] represents the switch value in the switch table.

```

//Modulation
switch(vector)
{
    case 0 : out[0]=0; out[1]=0; out[2]=0;break;
    case 1 : out[0]=1; out[1]=0; out[2]=0;break;
    case 2 : out[0]=1; out[1]=1; out[2]=0;break;
    case 3 : out[0]=0; out[1]=1; out[2]=0;break;
    case 4 : out[0]=0; out[1]=1; out[2]=1;break;
    case 5 : out[0]=0; out[1]=0; out[2]=1;break;
    case 6 : out[0]=1; out[1]=0; out[2]=1;break;
    case 7 : out[0]=1; out[1]=1; out[2]=1;break;
}

```

Figure 6.15 Modulation module

The output signals will directly control the switches of upper-bridge MOSFETs and control the switches of lower-bridge MOSFETs by passing a NOT gate.

And additionally, the output switch values of hf injection are also generated at the end of the code list as Figure 6.16, where k is the repetition rate.

```

//HF injection
switch(i)
{
    case 1 : out[8]=1; out[9]=0; out[10]=0;break;
    case 2 : out[8]=1; out[9]=1; out[10]=0;break;
    case 3 : out[8]=0; out[9]=1; out[10]=0;break;
    case 4 : out[8]=0; out[9]=1; out[10]=1;break;
    case 5 : out[8]=0; out[9]=0; out[10]=1;break;
    case 6 : out[8]=1; out[9]=0; out[10]=1;break;
} k++;
if(k=4) {k=0;i++;}
if(i=7) {i=1;}

```

Figure 6.16 *hf* voltage injection

The code of C block controller presented in this section along with other function parts introduced in Section 6.1.1 together build the simulation of traditional DTC system. The traditional DTC is the basement of the entire system and each part of the system can be seen clearly in the simulation schematic, so it is introduced at the beginning of this chapter. Base on the discussion in the traditional DTC simulation, Section 6.1.3 will introduce the differences on the modified 12-sector DTC.

6.1.3 Simulation of 12-sector DTC

The simulation schematic of 12-sector DTC is similar with the traditional DTC of PMSM. The biggest change in schematic is that there are no more NOT gates to simplify the control of MOSFETs because of the change of switch table. And since the calculation way of stator voltage changes, three coefficients are introduced into (3-12) to help the calculation. Compared to Figure 6.5, the three switch values (S_a , S_b , S_c) used to calculate the voltage have already been adjusted by the coefficients ($C1$ to $C3$) before the calculation in Figure 6.17, so the function module showed in Figure 6.5 will still work in this schematic. But most of the differences existed in θ determination module, switch table module, stator voltages acquire module and modulation module, are in C block controller. Figure 6.17 shows the main changes of schematic outside the C block.

Figure 6.17 Changes in 12-sector schematic

The new θ determination module showed as Figure 6.18 is changed because the space has been divided into 12 sectors, but is still determined according to the space relationship between ψ_s , $\psi_{s\alpha}$ and $\psi_{s\beta}$.

```

// Judge sector
if (dFa >= 0 && dFb >= 0)
{
    if( dFb < (dFs/2)) sector = 1;
    else if( dFa < (dFs/2)) sector = 3;
    else sector = 2;
}
if (dFa >= 0 && dFb < 0)
{
    if( -dFb < (dFs/2)) sector = 12;
    else if( dFa < (dFs/2)) sector = 10;
    else sector = 11;
}
if (dFa < 0 && dFb >= 0)
{
    if( dFb < (dFs/2)) sector = 6;
    else if( -dFa < (dFs/2)) sector = 4;
    else sector = 5;
}
if (dFa < 0 && dFb < 0)
{
    if( -dFb < (dFs/2)) sector = 7;
    else if( -dFa < (dFs/2)) sector = 9;
    else sector = 8;
}

```

Figure 6.18 θ determination module

The new switch table, as shown in Figure 6.19, in the situation of $\tau = 1$ and $\phi = 1$:

```

//judge vector
if( flagf == 1 && flagt == 1)
{
    switch(sector)
    {
        case 1 : vector = 3; break;
        case 2 : vector = 5; break;
        case 3 : vector = 5; break;
        case 4 : vector = 7; break;
        case 5 : vector = 7; break;
        case 6 : vector = 9; break;
        case 7 : vector = 9; break;
        case 8 : vector = 11; break;
        case 9 : vector = 11; break;
        case 10 : vector = 1; break;
        case 11 : vector = 1; break;
        case 12 : vector = 3; break;}
}

```

Figure 6.19 12-sector switch table

The voltage acquirement in this control method is not that simple as the traditional way, because the voltage cannot be calculated simply by (3-12) in the two-phase conduct mode. So the adjustment coefficients are introduced into the calculation based on the (3-12), and these adjustment coefficients are integrated into the modulation module as output [6-8] shown in Figure 6.20. The coefficients are equal to zero when the inverter is three-phase conducted since (3-12) works. When only two phases conducted, the coefficients need to satisfy some conditions to make (3-12) work. Assume the three coefficients are x, y and z , so $S_a = S_1 + x$, $S_b = S_2 + y$ and $S_c = S_3 + z$. Then (3-12) turns into (6-2):

$$\begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ -1 & 2 & -1 \\ -1 & -1 & 2 \end{bmatrix} \begin{bmatrix} S_a + x \\ S_b + y \\ S_c + z \end{bmatrix} U_{dc} \quad (6-2)$$

Then apply two-phase conditions to (6-2) and the conclusion can be drawn after solve the equation: the coefficients of "1" and "0" are equal, for example, if apply $u_{12}(10-1)$, then $x = y$, meanwhile, the coefficient of "-1" (z) needs to satisfy $x - z = -1.5$. And this conclusion can be testified in every two-phase condition. So in order to simplify the calculation, the coefficients (0, 0, 1.5) are adopted in this thesis.

Additionally, for modulation module, since there is a two-phase conduction mode, the NOT gate cannot be used for simplifying code anymore, which means each switch have its own control signal directly from the C block.

```
//Modulation
switch(vector)
{
  case 0 : out[0]=0; out[1]=1; out[2]=0;out[3]=1; out[4]=0; out[5]=1;out[6]=0; out[7]=0; out[8]=0;break;
  case 1 : out[0]=1; out[1]=0; out[2]=0;out[3]=1; out[4]=0; out[5]=1;out[6]=0; out[7]=0; out[8]=0;break;
  case 2 : out[0]=1; out[1]=0; out[2]=0;out[3]=0; out[4]=0; out[5]=1;out[6]=0; out[7]=1.5; out[8]=0;break;
  case 3 : out[0]=1; out[1]=0; out[2]=1;out[3]=0; out[4]=0; out[5]=1;out[6]=0; out[7]=0; out[8]=0;break;
  case 4 : out[0]=0; out[1]=0; out[2]=1;out[3]=0; out[4]=0; out[5]=1;out[6]=1.5; out[7]=0; out[8]=0;break;
  case 5 : out[0]=0; out[1]=1; out[2]=1;out[3]=0; out[4]=0; out[5]=1;out[6]=0; out[7]=0; out[8]=0;break;
  case 6 : out[0]=0; out[1]=1; out[2]=1;out[3]=0; out[4]=0; out[5]=0;out[6]=0; out[7]=0; out[8]=1.5;break;
  case 7 : out[0]=0; out[1]=1; out[2]=1;out[3]=0; out[4]=1; out[5]=0;out[6]=0; out[7]=0; out[8]=0;break;
  case 8 : out[0]=0; out[1]=1; out[2]=0;out[3]=0; out[4]=1; out[5]=0;out[6]=0; out[7]=1.5; out[8]=0;break;
  case 9 : out[0]=0; out[1]=1; out[2]=0;out[3]=1; out[4]=1; out[5]=0;out[6]=0; out[7]=0; out[8]=0;break;
  case 10 : out[0]=0; out[1]=0; out[2]=0;out[3]=1; out[4]=1; out[5]=0;out[6]=1.5; out[7]=0; out[8]=0;break;
  case 11 : out[0]=1; out[1]=0; out[2]=0;out[3]=1; out[4]=1; out[5]=0;out[6]=1; out[7]=0; out[8]=0;break;
  case 12 : out[0]=1; out[1]=0; out[2]=0;out[3]=1; out[4]=0; out[5]=0;out[6]=0; out[7]=0; out[8]=1.5;break;
  case 13 : out[0]=1; out[1]=0; out[2]=1;out[3]=0; out[4]=1; out[5]=0;out[6]=0; out[7]=0; out[8]=0;break;
}
```

Figure 6.20 New switch table module with adjustment coefficients

Above all, every aspect of the simulation, from the whole system schematic to each function modules, has been presented minutely in Section 6.1, so in the next section, the results of simulation will be discussed and analyzed.

6.2 Results and analysis

The parameters of the PMSM model: power rating $P_n = 1200W$, speed rating $n = 2000r/min$, torque rating $T_n = 9550 * P_n / n = 5.73N \cdot m$, number of Poles $P = 4$ ($n_p = 2$), inductance $L_d = L_q = 31.6mH$, stator resistance $R_s = 2.38\Omega$, peak line-to-line back EMF constant is $98.67V/krpm$, moment of Inertia $J = 0.00179$.

The control parameters: the total simulation time is $0.9s$, reference value of stator flux $\psi_s^* = 1.5Wb$, start speed is $90rad/s$ and speed changes to $5rad/s$ when $t = 0.6s$, initial load torque is $1N \cdot m$ then changes to $5N \cdot m$ when $t = 0.3s$.

Sensorless control parameters: DTC sampling time $T_z = 5\mu s$, h_f voltage injection time $T_i = 0.25\mu s$, and repetition rate $k = 4$.

Under such simulation circumstance, we can clearly see the performances of two DTC methods and sensorless control methods in the entire speed range with light or full load of the motor.

The speed (rad/s) response curve, torque ($N \cdot m$) response curve, torque ripple ($N \cdot m$) curve, stator flux (Wb) curve and rotor position ($^\circ$) estimation curve are shown as following.

Figure 6.21 Speed response curve in traditional DTC

Figure 6.22 Speed response curve in 12-sector DTC

From Figures 6.21 and 6.22, both systems started with torque $= 1N \cdot m$, then came to steady state ($v = 90rad/s$) after around 0.1s. When $t = 0.3s$, load torque changed from $1N \cdot m$ to $5N \cdot m$, both systems performed good anti-jamming capability to torque changes. When $t = 0.6s$, reference speed changed drastically from $90rad/s$ to $5rad/s$, both systems reached steady after less than 0.1s and worked well in low speed range. Generally speaking, both systems show good speed responses in the full speed range.

At first sight, compared Figures 6.21 and 6.22, the conventional DTC shows better adjusting ability than 12-sector DTC when noticed the adjusting time of response to reference speed change. The conventional DTC can reach steady stage around 0.05s before 12-sector DTC method when speed changes. However, in the real marine application, 0.05s lead in the response time is almost negligible. So the performance at steady stage is the most concern since the torque ripple problem will make

differences in the performance at steady working stage. From these two figures, there is only a little differences can be seen because the figures presented in a large range of speed. But if we check them carefully, at steady stage, the speed ripples in conventional DTC are quite obvious while there are nearly no ripples in 12-sector DTC. Figures 6.21 and 6.22 are just showed to testify that this system with either DTC method can have good speed response and work well in entire speed region with either light load or heavy load. The comparison of torque ripple will be discussed in detail later in this section.

Following are the torque responses of the systems, which are similar to the situations in the speed responses:

Figure 6.23 Torque response curve in traditional DTC

Figure 6.24 Torque response curve in 12-sector DTC

From Figures 6.23 and 6.24, both systems shows good torque control ability with light or full load in entire speed region. And similar to the speed response comparison mentioned above, the torque response time to speed change in traditional direct torque control method is around 0.05s less than the 12-sector direct torque control method, which is also negligible in the real applications. But again, there are torque ripples can be found in the steady time of traditional direct torque control method if we have a careful look while there are scarcely torques ripple existed in the 12-sector direct torque control method.

The above discussions on speed and torque response prove that the system can meet the requirement of full load in full speed range on the whole. But as mentioned above in the discussion of speed response, the adjust time usually very short and the machine works at stable status for most time, so the torque ripple problem discussed in Chapter IV can not be ignored in the DTC part. In order to have a clear comparison of torque ripple in the two DTC methods, we take the time range 0.4s to 0.6s to see their differences at high speed as Figure 6.25 and 0.7s to 0.9s to see the low speed range as Figure 6.26.

Figure 6.25 Torque ripple at high speed

From Figure 6.25, the maximum fluctuation value of torque is $0.4 \text{ N} \cdot \text{m}$ in 6-sector DTC and $0.2 \text{ N} \cdot \text{m}$ in 12-sector DTC when the load torque is $5 \text{ N} \cdot \text{m}$. Generally speaking, the torque ripple has reduced 50% when apply 12-sector DTC and the error ratio is less than 5%, which is an acceptable range in marine application. The situations in low speed range can be more significant as Figure 6.26.

Figure 6.26(a) Torque ripple at low speed (6-sector)

Figure 6.26(b) Torque ripple at low speed (12-sector)

Figure 6.26 shows the torque performance at extreme low speed (5rad/s). Surprisingly, the 12-sector DTC shows an even better effect in reducing torque ripple compared to the high-speed region. The reduction amplitude of torque ripple can even reach 83.3% in some range of low-speed time. Besides the impressed performance in reducing torque ripple, we can also notice that the fluctuating frequency of the torque has also been decreased obviously, which means the 12-sector DTC offers a more stable working performance at low speed. This is very meaningful for the winch application since there is quite a lot low-speed time in winch's work.

Above all, the results in Figures 6.25 and 6.26 not only show that the analysis in chapter V is reasonable, which the new control method has significantly reduced torque ripple level, but also prove that the 12-sector DTC is very proper for winch application for its remarkable low-speed performance.

Figure 6.27 Stator flux curve in traditional DTC

Figure 6.28 Stator flux curve in 12-sector DTC

Since the torque control realized by controlling the stator flux in DTC, the moving trace of stator flux is also what we are interested in. Figures 6.27 and 6.28 show the control circles of the stator flux in two DTC methods. Referring to related theory in Chapter III, both of them are as good as what we expected.

Figure 6.29 Rotor position estimation and the real position

Figure 6.29 shows the result of sensorless control. Before 0.6s, the back-EMF integral method shows a great effect for tracking the rotor's position in high-speed range as it has proved in most industry application. When a sudden decline happens in operation speed at 0.6s, the *hf* injection method has also shown a good ability in tracking the actual position of rotor especially after the adjusting time. So in general, the performance of the combined sensorless control methods can meet the requirement in both high-speed range and low-speed range.

6.3 Summary

This chapter shows the results and analysis of traditional DTC and modified DTC simulation with sensorless control. From the simulation we can see that traditional DTC has a fair performance whether in high-speed range or low-speed range, but the torque ripple cannot be ignored since the fluctuation will be magnified, which can become serious in low-speed region. In modified DTC method, the torque ripple has been greatly reduced. Though the system response to speed change is a bit late compared to the traditional DTC, which can be negligible in the real work, 12-sector DTC is proved to be very suitable for whole-speed range work especially for low-speed region. And the sensorless control methods are also proved to be trustworthy in the entire speed range. Since the whole content of the electrical driven winch system have been introduced completely, Chapter VII will do some conclusion to this thesis and offers some future research recommendations.

VII. CONCLUSION AND RECOMMENDATIONS

The previous parts of this thesis mainly introduce the basic theory of direct torque control of PMSM and its modified version as well as the sensorless control. Current works are all based on the marine condition to make it simple, good dynamic performance and energy saving. In this part, some conclusions drew from the simulation result and some future research recommendations will be illustrated.

7.1 Conclusion

In this thesis, according to the requirement of electrical driven winch system, following works have been done:

- Introduced the background of winch system and compared different motor type with different control methods to explain why DTC system of PMSM is the best choice currently.
- Illustrated the mathematical model of PMSM and the basic theory of DTC system.
- Proposed 12-sector DTC as a modified version of traditional DTC to greatly reduce torque ripple problem.
- Combined back-EMF integral and high frequency voltage injection as the sensorless control method. And the way of hf carrier signal injection is specifically designed for switch-table based DTC in order not to compromise the simplicity of DTC system.
- Finished the simulation of PMSM DTC system, and analyzed the result of simulation.

From the works have been done mentioned above, the following conclusions can be drawn:

- The electrical driven winch system based on the direct torque control methods can achieve full torque in entire speed range, which is the most basic requirement for winch application. Figure 6.21 to 6.24 showed that overall, the speed and output torque of PMSM was stable both in high-speed range and extreme low-speed range with full torque.
- The electrical driven winch system has inherited the features of simplicity, short response time to speed change, good anti-jamming capability to the torque variation and good control of flux circle from DTC system. These good characters can be seen clearly from the figures in Chapter VI make the winch system reliable and efficient. As discussed in Chapter I, considering the limited resources on board, simplicity is one of the most important requirements for winch system because it will influence its efficiency, dynamic performance, energy consumption and so on. So it is easy to see that every modification made on the DTC system including the sensorless control in this thesis has one precondition that is to keep this system as simple as possible. Finally, the system successfully satisfied its requirement and could still keep these advantages in DTC.
- 12-sector direct torque control method makes the electrical driven winch system more reliable and stable, especially effective in low-speed range, since it solves the torque ripple problem existed in traditional DTC to a very good extent. From the comparison of traditional DTC and 12-sector DTC in

Figures 6.25 and 6.26, the output torque is much more stable especially at low speed after using more voltage vectors. Though applying SVPWM can solve the torque ripple problem thoroughly in theory, however, it will obey the rules of simplicity. Now that the 12-sector DTC has already controlled the torque ripple in a very low and acceptable range, the simpler structure makes 12-sector DTC here the best choice for the winch system.

- The combined way of back-EMF integral and high frequency voltage injection can be successfully applied in the DTC system and perfectly realize sensorless control in entire speed region without compromising the simplicity of DTC system. Generally speaking, the *hf* injection method will bring great complexity to the system and need the help of PWM to inject *hf* carrier. But the clever use of space voltage vectors in Chapter V not only successfully injected *hf* carrier but also made the switch table based DTC used in this thesis still applicable for the winch system. After solving this problem, the rest contents in sensorless control part are very mature technologies nowadays.

In conclusion, the sensorless DTC system of PMSM established in this thesis is a very good choice for the electrical driven system for its simplicity, fast dynamic response and energy saving features besides the basic functions according to the requirements of winch application. But there are still many things need further research, which will be discussed in Section 7.2 to make the electrical driven winch system better and better.

7.2 Recommendations for further research

Though the DTC system introduced in this thesis is very proper for the marine choice, but when considered the severe sea condition or other situations, we can still see many potentials in this system. Following are some recommendations for further research in this area.

7.2.1 Improvement of anti-disturbance capability

In real marine condition, the disturbance can be any type and hard to simulate in software. So in the real situation, whether PID controller can be trusted is an unknown question. The main advantage of the classical PID controller is not required to know the exact model of the controlled object, whose system error elimination control strategy is based on the error between the target and actual values. So far PID controller is the most commonly used control method in AC servo control system. But while PID controllers are applicable to many control problems, and often perform satisfactorily without any improvements or only coarse tuning, they can perform poorly in some applications, and do not in general provide optimal control [69]. The classical PID controller has shown some defects with the development of digital processing techniques:

- The way error acquired is not very reasonable.
- Ideal differentiator physically is difficult to achieve.
- “Weight sum” method is not the best.
- The feedback of Integral has many side effects [70].

So base on the classical PID controller, we can use more stable adjusting controlling like the Active Disturbances Rejection Controller (ADRC) and so on. Compared with the classic PID controller, advanced adjusting controller usually has the following advantages:

- The transition process is arranged suitably to make the system response faster and no overshoot.
- Extract differential signals reasonably and overcome the differential signal distortion exists in PID controller.
- Use nonlinear combination of error signal.
- Estimate the disturbance inside and outside the system, and generates a compensation component in the control, in order to compensate for impact on system performance caused by the internal and external disturbances [71].

As a result, the operator does not need a precise analytical description of the system, as one can assume the unknown parts of dynamics as the internal disturbance in the plant. Robustness and the adaptive ability of this method make it an interesting solution in scenarios where the full knowledge of the system is not available[72]. This kind of controller not only control the output of the system, but also control the differential output of each stage, which results in the state of entire system well controlled and compensation for the disturbance. So the performance of the control system has been significantly improved in terms of stability and robustness. So the replacement of PID controller is a future research option.

7.2.2 Switch algorithm for combined methods

Since we adopt different methods in different speed ranges, when the speed variation is severe, how to switch them smoothly is also a problem needs to solve. Current studies shows that the weighted algorithm is a possible solution, which the key it to determine the best switch range of different methods, namely the maximum and minimum limits of adopting the different methods [73]. Generally, the lower speed limit n_{min} of the switch range should higher than the lowest speed limit of high-speed method, while the higher speed limit n_{max} of the switch range should lower than the highest speed limit of low-speed method. When the estimation value of speed $\tilde{n}$ is lower than the lower limit n_{min} , the estimation of speed will be obtained by low-speed method $\tilde{w}_{low}$. Similarly, the high-speed method will be adopted to get $\tilde{w}_{high}$ when the estimation value of speed $\tilde{n}$ is higher than the higher limit n_{max} . Then the final the estimation value of speed $\tilde{w} = \lambda_1 \tilde{w}_{low} + \lambda_2 \tilde{w}_{high}$, where λ is the weighted coefficient.

The weighted algorithm theoretically smooth the switching process and is not complicated to realize. However, the biggest problem of this algorithm is when the estimated speed value of different methods are conflict in the switch section, it has no such ability to figure out the wrong one, in other words, this algorithm is kind of vulnerable. So if we want to develop a stronger algorithm, further research is also necessary.

7.2.3 Hardware Test

Unfortunately, due to the limited time and resources, the hardware test hasn't been done in this thesis. But as a design for the marine application, the system introduced in this thesis still needs the hardware

test to check its performance in the real application. With the development of MCU and Power Electronics, the Numerical Control Technology has been widely used in the motor control area. Since the PMSM is a high-level, multi-variable and strong coupling nonlinear system, the high-performance control strategies need a lot of real-time computing in order to ensure the real-time control of the system [74]. So a high-performance DSP should be adopted as the core of the controller.

The hardware system includes the hardware part and software part. The hardware system based on the high-performance DSP controller mainly contains CPU control board, main circuit, detecting circuit, protection circuit and so on. The hardware system diagram is designed as Figure 7.1.

Figure 7.1 Hardware system diagram

Following will discuss each part of the hardware system separately in detail:

- DSP chip is a kind of microprocessor that particularly suitable for digital signal processing. The most successful DSP chips are made by TI Company. Among them, TMS320C2000 series, which is designed to meet a wide range of digital motor control applications, is one of the most suitable DSP controllers for the hardware system. Take TMS320F2812 as an example in this series, it is a 32-bit DSP and its working frequency can reach up to 150MHz [75]. It also has the features which are useful for the hardware system including up to 128K Flash Memory with 18K SRAM, two 12-bit ADCs with 16 channels and up to 56 independent programmable General Purpose I/O (GPIO) pins etc. [76] Generally speaking, the popular DSP chips nowadays all have been optimized for controller application and integration, which make the application developer much easier, faster and more effective in their design process. The final choice of the DSP chip should be a very careful decision after considering like cost and some other factors.
- Main circuit of the system includes rectifier circuit, filter circuit and the inverter circuit showed in Figure 7.1. Rectifier circuit is suggested to adopt three-phase uncontrolled rectifier module since there is no need to change the DC bus voltage. The capacitance of filter capacitor should be as high as possible theoretically, however, the volume and price of the capacitor are also very important in

choosing. Then considering the limit in capacity and withstand voltage of the electrolytic capacitor, a useful way is usually to connect two small capacitors in series. The Power devices in inverter circuit use the IPM (Intelligent Power Modules). They are a combination of power semiconductors like IGBTs, MOSFETs and diodes with peripheral components like driver circuits [77]. In general, IPM module not only integrated power switching device like IGBT with its corresponding drive circuits, but also has the over-voltage, over-current and over-temperature fault detection circuit, which can sent the detected fault signal into the DSP for interrupt handling [78]. IPM is very suitable for the inverter used to drive motor because of its high reliability and convenience.

- The detecting circuit is used to detect stator current and DC bus voltage. The main point here is to design the proper operational amplifier and voltage follower according to the parameters of sensors and AD modules in DSP, to let the detected value within the input range of AD.
- The isolation and protection circuit is used to protect the low-voltage circuits from high-voltage circuits when faults happen. As discussed above in point 2, IPM has offered fault signals when fault happens, so use an optical coupler (OC), which is an optical component which allows the transmission of light in only one direction [79], to isolate all the signals between IPM and DSP and create an interrupt in DSP to block the signals from DSP to IPM when a fault signal received, then all the Power devices in IPM will shut off to prevent the fault from beginning.

Above discussed are the main parts in the hardware system, while there are also some other circuits like the DSP clock circuit, reset circuit and so on. They are either easy to design just following the datasheet or common used in a lot of applications, so there is no further discussion on them.

The software design will be much easier since the C block controller used in the simulation software is quite similar to the real DSP programming. After integrating other function parts, the procedure of the code in the C block controller will be the same as the procedure of the main interrupt, which is the most core part of program in DSP. And the other interrupt is the protection interrupt that has already mentioned above. The widely used procedure of main program in DSP for motor drive is showed as Figure 7.2.

Figure 7.2 Procedure of the main program [80]

So above is the discussion on the hardware and software design of the hardware test for the electrical driven winch system. Generally speaking, the discussions in this section are qualitative. The quantitative design, which needs to consider all kinds of parameters and situations, can only be done in the real work.

References

- [1]. Electrical Colleges, Electric Winches, From the World Wide Web: <http://www.wiml.org/219-electric-winches.html>.
- [2]. Lincoln J M, Lucas D L, McKibbin R W, et al., Reducing commercial fishing deck hazards with engineering solutions for winch design, *Journal of safety research*, 2008, 39(2): 231-235.
- [3]. ABB website, Today's electrical deck winch control solutions deserve a deeper look, From the World Wide Web: <http://new.abb.com/drives/segments/winches/winch-control-solutions>.
- [4]. Cooper G, Ghassemieh E., Risk assessment of patient handling with ambulance stretcher systems (ramp/(winch), easi-loader, tail-lift) using biomechanical failure criteria, *Medical engineering & physics*, 2007, 29(7): 775-787.
- [5]. Henao H, Fatemi S M J R, Capolino G A, et al., Wire rope fault detection in a hoisting winch system by motor torque and current signature analysis, *IEEE Transactions on Industrial Electronics*, 2011, 58(5): 1727-1736.
- [6]. Kawai H and Tasaka Y., Evaluation of the energy-saving performance of the PMSM drive system, *Power Electronics Conference (IPEC)*, 2010 International. IEEE, 2010: 1605-1608.
- [7]. Liu H, Li S., Speed control for PMSM servo system using predictive functional control and extended state observer, *IEEE Transactions on Industrial Electronics*, 2012, 59(2): 1171-1183.
- [8]. Vas P., *Vector control of AC machines*, Oxford University Press, USA, 1990.
- [9]. Li S, Haskew T A, Swatloski R P, et al., Optimal and direct-current vector control of direct-driven PMSG wind turbines, *IEEE Transactions on power electronics*, 2012, 27(5): 2325-2337.
- [10]. Takken W, Knols B G J., Malaria vector control: current and future strategies, *Trends in parasitology*, 2009, 25(3): 101-104.
- [11]. Vaez-Zadeh S, Jalali E., Combined vector control and direct torque control method for high performance induction motor drives, *Energy conversion and management*, 2007, 48(12): 3095-3101.
- [12]. Takahashi I and Noguchi T., A new quick-response and high-efficiency control strategy of an induction motor, *Industry Applications*, *IEEE Transactions on*, 1986 (5): 820-827.
- [13]. Kowalski C T, Lis J, Orłowska-Kowalska T., FPGA implementation of DTC control method for the induction motor drive, *The International Conference on Computer as a Tool*. IEEE, 2007: 1916-1921.
- [14]. Sun Y W., Development and application of the modern PMSM and AC-servo system, *Small & Special Electrical Machines*, 2010, 38(3): 71-73.
- [15]. Mohamed Y A R I., Design and implementation of a robust current-control scheme for a PMSM vector drive with a simple adaptive disturbance observer, *IEEE transactions on industrial electronics*, 2007, 54(4): 1981-1988.
- [16]. Liu H, Li S., Speed control for PMSM servo system using predictive functional control and extended state observer, *IEEE Transactions on Industrial Electronics*, 2012, 59(2): 1171-1183.
- [17]. Nahid-Mobarakeh B, Meibody-Tabar F, Sargos F M., Back EMF estimation-based sensorless control of PMSM: Robustness with respect to measurement errors and inverter irregularities, *IEEE Transactions on Industry Applications*, 2007, 43(2): 485-494.
- [18]. Caponio A, Cascella G L, Neri F, et al., A fast adaptive memetic algorithm for online and offline control design of PMSM drives, *IEEE Transactions on Systems, Man, and Cybernetics, Part B (Cybernetics)*, 2007, 37(1): 28-41.
- [19]. Ortega C, Arias A, Caruana C, et al., Improved waveform quality in the direct torque control of matrix-converter-fed PMSM drives, *IEEE Transactions on Industrial Electronics*, 2010, 57(6): 2101-2110.

- [20]. Akrad A, Hilairat M, Diallo D., Design of a fault-tolerant controller based on observers for a PMSM drive, *IEEE Transactions on Industrial Electronics*, 2011, 58(4): 1416-1427.
- [21]. Zhang X, Sun L, Zhao K, et al., Nonlinear speed control for PMSM system using sliding-mode control and disturbance compensation techniques, *IEEE Transactions on Power Electronics*, 2013, 28(3): 1358-1365.
- [22]. Choi H H, Vu N T T, Jung J W., Digital implementation of an adaptive speed regulator for a PMSM, *IEEE Transactions on Power Electronics*, 2011, 26(1): 3-8.
- [23]. Jia H P, Sun D and He Y., The PMSM DTC based on variable structure sliding mode, *Proceedings of the Chinese Society of Electrical Engineering*. 2006, 26(20): 134-138.
- [24]. Sun D and Meng J., A single neuron PID controller based PMSM DTC drive system fed by fault tolerant 4-switch 3-phase inverter, *Industrial Electronics and Applications*, 2006 1ST IEEE Conference on. IEEE, 2006: 1-5.
- [25]. Lai Y S and Chen J H., A new approach to direct torque control of induction motor drives for constant inverter switching frequency and torque ripple reduction, *Energy Conversion*, *IEEE Transactions on*, 2001, 16(3): 220-227.
- [26]. Paponpen K, Konghirun M., An improved sliding mode observer for speed sensorless vector control drive of PMSM, *Power Electronics and Motion Control Conference*, 2006. IPEMC 2006. CES/IEEE 5th International. IEEE, 2006, 2: 1-5.
- [27]. Preindl M, Bolognani S., Model predictive direct speed control with finite control set of PMSM drive systems, *IEEE Transactions on Power Electronics*, 2013, 28(2): 1007-1015.
- [28]. Geyer T, Beccuti G A, Papafotiou G, et al., Model predictive direct torque control of permanent magnet synchronous motors, 2010 IEEE Energy Conversion Congress and Exposition. IEEE, 2010: 199-206.
- [29]. Metidji B, Taib N, Baghli L, et al., Low-cost direct torque control algorithm for induction motor without AC phase current sensors, *IEEE Transactions on Power Electronics*, 2012, 27(9): 4132-4139.
- [30]. Cheng Y W, Jia K X, Jun Y Y, et al., *Modern motor control technology*, 2009.
- [31]. Aktas M, Turkmenoglu V., Wavelet-based switching faults detection in direct torque control induction motor drives, *IET science, measurement & technology*, 2010, 4(6): 303-310.
- [32]. Ortega C, Arias A, Caruana C, et al., Improved waveform quality in the direct torque control of matrix-converter-fed PMSM drives, *IEEE Transactions on Industrial Electronics*, 2010, 57(6): 2101-2110.
- [33]. Arbi J, Ghorbal M J B, Slama-Belkhodja I, et al., Direct virtual torque control for doubly fed induction generator grid connection, *IEEE Transactions on Industrial Electronics*, 2009, 56(10): 4163-4173.
- [34]. Sun D, He Z, He Y, et al., Four-switch inverter fed PMSM DTC with SVM approach for fault tolerant operation, *Electric Machines & Drives Conference*, 2007. IEMDC'07. IEEE International. IEEE, 2007, 1: 295-299.
- [35]. Akrad A, Hilairat M, Diallo D., Design of a fault-tolerant controller based on observers for a PMSM drive, *IEEE Transactions on Industrial Electronics*, 2011, 58(4): 1416-1427.
- [36]. Zhang Z, Zhao Y, Qiao W, et al., A space-vector-modulated sensorless direct-torque control for direct-drive PMSG wind turbines, *IEEE Transactions on Industry Applications*, 2014, 50(4): 2331-2341.
- [37]. Sutikno T, Idris N R N, Jidin A, et al., An improved FPGA implementation of direct torque control for induction machines, *IEEE Transactions on Industrial Informatics*, 2013, 9(3): 1280-1290.
- [38]. Gui-xi J, Hui-chao Z, Hong-jun S., Simulation research on PMSM vector control system based on SVPWM, *Electrical and Control Engineering (ICECE)*, 2010 International Conference on. IEEE, 2010: 1936-1940.

- [39]. West N T, Lorenz R D., Implementation and evaluation of a stator and rotor flux linkage-based dead-beat, direct torque control of induction machines at the operational voltage limits, 2007 IEEE Industry Applications Annual Meeting. 2007.
- [40]. Patel C, Rajeevan P, Fast direct torque control of an open-end induction motor drive using 12-sided polygonal voltage space vectors, IEEE transactions on power electronics, 2012, 27(1): 400-410.
- [41]. Yu H S, Zhao K Y, Guo L, et al., Maximum torque per ampere control of PMSM based on port-controlled Hamiltonian theory, Zhongguo Dianji Gongcheng Xuebao(Proceedings of the Chinese Society of Electrical Engineering). 2006, 26(8): 82-87.
- [42]. Zhong L, Rahman M F, Hu W Y, et al., A direct torque controller for permanent magnet synchronous motor drives, Energy Conversion, IEEE Transactions on, 1999, 14(3): 637-642.
- [43]. Zaimeddine R, Undeland T., DTC control schemes for induction motor fed by three-level NPC-VSI using space vector modulation, SPEEDAM 2010. IEEE, 2010: 966-971.
- [44]. Lu Z, Sheng H, Hess H L, et al., The modeling and simulation of a permanent magnet synchronous motor with direct torque control based on Matlab/Simulink, Electric Machines and Drives, 2005 IEEE International Conference on. IEEE, 2005: 7 pp.-1156.
- [45]. Buja G S and Kazmierkowski M P., Direct torque control of PWM inverter-fed AC motors-a survey, Industrial Electronics, IEEE Transactions on, 2004, 51(4): 744-757.
- [46]. Ziane H, Retif J M, Rekioua T., Fixed-switching-frequency DTC control for PM synchronous machine with minimum torque ripples, Canadian Journal of Electrical and Computer Engineering, 2008, 33(3/4): 183-189.
- [47]. Khoucha F, Lagoun S M, Marouani K, et al., Hybrid cascaded H-bridge multilevel inverter motor drive DTC control for Electric Vehicles, Electrical Machines, 2008. ICEM 2008. 18th International Conference on. IEEE, 2008: 1-6.
- [48]. Liu Y, Zhu Z Q, Howe D., Commutation-torque-ripple minimization in direct-torque-controlled PM brushless DC drives, IEEE Transactions on Industry Applications, 2007, 43(4): 1012-1021.
- [49]. Jiang Z, Hu S, Cao W., A new fuzzy logic torque control scheme based on vector control and direct torque control for induction machine, Innovative Computing Information and Control, 2008. ICICIC'08. 3rd International Conference on. IEEE, 2008: 500-500.
- [50]. Yu W H, Cun T, Yi K G, et al., In-depth research on direct torque control of permanent magnet synchronous motor, IECON 02 [Industrial Electronics Society, IEEE 2002 28th Annual Conference of the]. IEEE, 2002, 2: 1060-1065.
- [51]. Payam A F, Hashemnia M N, Faiz J., Robust DTC control of doubly-fed induction machines based on input-output feedback linearization using recurrent neural networks, Journal of Power Electronics, 2011, 11(5): 719-725.
- [52]. LI Y and ZHU H., Overview of sensorless control of permanent magnet synchronous motors, Electric Drive, 2009, 9: 002.
- [53]. Chitra V, Ravichandran K S, Varadarajan R., Adaptive neuro fuzzy inference system based DTC control for matrix converter, Research Journal of Applied Sciences, Engineering and Technology, 2012, 4(8): 929-936.
- [54]. Han J., From PID to active disturbance rejection control, Industrial Electronics, IEEE transactions on, 2009, 56(3): 900-906.
- [55]. Moulay-Idriss C, Mohamed B., Application of the DTC control in the photovoltaic pumping system, Energy Conversion and Management, 2013, 65: 655-662.
- [56]. LI C, LI X., A modified stator flux observer of DTC control system, Electric Machines & Control Application, 2006, 6: 008.
- [57]. Corley M J and Lorenz R D., Rotor position and velocity estimation for a salient-pole permanent magnet synchronous machine at standstill and high speeds, Industry Applications, IEEE Transactions on, 1998, 34(4): 784-789.

- [58]. Lee K B, Blaabjerg F., Sensorless DTC-SVM for induction motor driven by a matrix converter using a parameter estimation strategy, *IEEE Transactions on Industrial Electronics*, 2008, 55(2): 512-521.
- [59]. Nian H, Quan Y and Li J., Rotor displacement sensorless control strategy for PM type bearingless motor based on the parameter identification, *Electrical Machines and Systems*, 2009. ICEMS 2009. International Conference on. IEEE, 2009: 1-5.
- [60]. Schrodli M, Hofer M and Staffler W., Sensorless control of PM synchronous motors in the whole speed range including standstill using a combined INFORM/EMF model, *Power Electronics and Motion Control Conference*, 2006. EPE-PEMC 2006. 12th International. IEEE, 2006: 1943-1949.
- [61]. Han J., From PID to active disturbance rejection control, *Industrial Electronics*, *IEEE transactions on*, 2009, 56(3): 900-906.
- [62]. Dawson D M, Hu J and Burg T C., *Nonlinear control of electric machinery*, Marcel Dekker, Inc., 1998.
- [63]. Zhang Y, Zhu J, Guo Y, et al., A sensorless DTC strategy of induction motor fed by three-level inverter based on discrete space vector modulation, *Power Engineering Conference*, 2009. AUPEC 2009. Australasian Universities. IEEE, 2009: 1-6.
- [64]. Jing Q H and Wen G Z., ADRC control for large time-delay systems, *Control and Decision*, 1999, 4.
- [65]. Nian X, Wang T, Wang J, et al., Adaptive stator resistance estimation method for speed sensorless DTC controlled IM drives, *APEC 07-Twenty-Second Annual IEEE Applied Power Electronics Conference and Exposition*. IEEE, 2007: 214-221.
- [66]. Abu-Rub H, Stando D, Kazmierkowski M P., Simple speed sensorless DTC-SVM scheme for induction motor drives, *Bulletin of the Polish Academy of Sciences: Technical Sciences*, 2013, 61(2): 301-307.
- [67]. Bian C, Ren S, Ma L., Sensorless DTC of super high-speed PMSM, *2007 IEEE International Conference on Automation and Logistics*. IEEE, 2007: 3060-3064.
- [68]. Manual P U., Powersim Inc. 2003.
- [69]. Åström K J, Hägglund T., *Advanced PID control*, ISA-The Instrumentation, Systems and Automation Society, 2006.
- [70]. Bolognani S, Zigliotto M and Zordan M., Extended-range PMSM sensorless speed drive based on stochastic filtering, *Power Electronics*, *IEEE Transactions on*, 2001, 16(1): 110-117.
- [71]. O'Dwyer A., *Handbook of PI and PID controller tuning rules*, London: Imperial college press, 2009.
- [72]. Li J, MA Y, LIU W, et al., Dynamic mode switch algorithm in multi-cell MU-MIMO systems, *Journal of Xidian University*, 2011, 5: 002.
- [73]. Bolognani S, Tubiana L and Zigliotto M., Extended Kalman filter tuning in sensorless PMSM drives, *Industry Applications*, *IEEE Transactions on*, 2003, 39(6): 1741-1747.
- [74]. Gu W., *Taught you how to learn DSP—Based on TMS320X281x*, 2011.
- [75]. TMS320x281x DSP. Analog-to-Digital Converter (ADC) reference guide, Texas Instruments, 2005.
- [76]. Infineon website., *Intelligent Power Modules (IPM).*, From the World Wide Web: <http://www.infineon.com/cms/en/product/power/motor-control-and-gate-driver-ics/intelligent-power-modules/channel.html?channel=db3a30431f848401011fa3657ad1349b>.
- [77]. Foo G, Rahman M F., Sensorless sliding-mode MTPA control of an IPM synchronous motor drive using a sliding-mode observer and HF signal injection, *IEEE Transactions on Industrial Electronics*, 2010, 57(4): 1270-1278.
- [78]. Cheok A D, Wang Z., DSP-based automated error-reducing flux-linkage-measurement method for switched reluctance motors, *IEEE Transactions on Instrumentation and Measurement*, 2007, 56(6): 2245-2253.
- [79]. Jezernik K, Korelič J, Horvat R., PMSM sliding mode FPGA-based control for torque ripple reduction, *IEEE Transactions on Power Electronics*, 2013, 28(7): 3549-3556.
- [80]. Swierczynski D, Kazmierkowski M P and Blaabjerg F., DSP based direct torque control of permanent magnet synchronous motor using space vector modulation (DTC-SVM), *Industrial Electronics*, 2002. ISIE 2002. Proceedings of the 2002 IEEE International Symposium on. IEEE, 2002, 3: 723-727.