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**Performance Improvement of Direct Torque Controlled
Permanent Magnet Synchronous Motor Drives**

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SCHOOL OF ELECTRICAL & ELECTRONIC ENGINEERING

2019

Performance Improvement of Direct Torque Controlled Permanent Magnet Synchronous Motor Drives

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A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirement for the degree of
Doctor of Philosophy

2019

Statement of Originality

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Authorship Attribution Statement

This thesis contains material from 1 paper published in the following peer-reviewed journal(s) / from papers accepted at conferences in which I am listed as an author.

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The contributions of the co-authors are as follows:

- Dr Zhang Xinan provided the initial project direction and edited the manuscript drafts.
- I prepared the manuscript drafts. The manuscript was reviewed and revised by Dr Zhang Xinan and Prof. Wang Youyi.
- I conducted all the simulation studies and data presenting.

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Abstract

Permanent Magnet Synchronous Motors (PMSM) have been gaining ever-increasing popularity in many applications thanks to their advantages such as long life, high torque and power density, excellent efficiency, etc. The applications of PMSM could be found in almost every industry such as home appliances, robotics, flywheels, industrial automation, electric vehicles, so on and so forth. Various control methods have been proposed for PMSM to attain satisfying performance with different application needs. For example, V/f control with stabilization loops could be used to control fan-type applications where dynamic response is not a must but efficiency and quietness matter. Field Oriented Control (FOC) is the dominant PMSM control method that achieve good dynamic response, smoothness and energy efficiency. Direct Torque Control (DTC) excels FOC in simpler control structure and supreme dynamic response but suffers from variable switching frequency and severe torque ripples. These defects generally lead to poor torque and flux regulation, high current harmonics and acoustic noise.

This thesis studies the performance improvement of direct torque controlled PMSM drives with more emphasis on the Surface-mounted PMSM(SPMSM), which is widely used due to its cost-effectiveness. Several sub-topics are covered in this thesis. In the first place, torque ripple reduction is investigated. Emerging techniques to address the torque ripple problem of DTC can be classified into three groups: (1) Model Predictive Torque Control (MPTC) which select an optimal voltage vector(VV) and apply it on the next full control cycle (one VV); (2) Duty cycle based MPTC which finds an optimal active voltage vector and zero voltage vector (two VVs) for the next control cycle and (3), Space Vector Modulation (SVM) MPTC which essentially uses two active vectors and one zero vectors for the next control cycle (three VVs). All the methods belong to the Finite Set MPC category. This thesis contributes an improved duty cycle based MPTC which finds the optimal duty cycle by a PI controller. It obtains excellent torque and flux regulation with fixed switching frequency.

In the second place, to further enhance the performance of PMSM drives, predictive

deadbeat control is proposed in this thesis. Compared to the existing work, this method is implemented in the stationary reference frame and has a simplified nature. It produces very smooth torque and flux responses using SVM. Moreover, the usage of prediction technique compensates for the one cycle digital delay and thus, is desirable for a digital control system.

Thirdly, fast and stable speed control is generally preferred in a wide speed range for high-speed PMSM drives, such as the flywheel storage system. Field Weakening (FW) techniques are generally used to extend the speed range to allow extra energy storage. Most of the existing FW methods for DTFC based drives have constrained constant power speed range (CPSR) due to the limited usage of DC-link voltage. Moreover, many of these FW approaches are heavily dependent on machine parameters. This can lead to degraded performance of drives or even control instability under some circumstances. To extend the CPSR and simultaneously enhance the parameter robustness of FW operation, this thesis proposes a new robust FW algorithm for DTFC based SPMSM drives. The proposed algorithm achieves extended CPSR and reduced parameter dependency.

It's worth mentioning that the previously described methods are based on using encoder as speed and rotor position feedback. Various sensor-less observers have been proposed in the literature to save the cost of such sensors and to improve the drives' reliability. As direct torque control inherently requires flux observers, it is straightforward to extend the flux observers into position and speed feedback. This thesis studied three stator flux observers and studied their performance and compensator gain setting method. The best performance dq frame flux observer is selected and being applied for all the DTC methods discussed in this thesis.

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List of Symbols

$\mathbf{v}_s$	Stator voltage vector with $\mathbf{v}_s = [u_\alpha; u_\beta]$ or $[u_d; u_q]$ or $[u_x; u_y]$ depends on the reference frame, same for other vectors $\mathbf{i}_s$, λ_s and λ_r
$\mathbf{i}_s$	Stator current vector
λ_s	Stator flux vector
λ_r	Rotor flux vector
λ_s	Magnitude of stator flux vector
$ \lambda_s $	Same as λ_s , magnitude of stator flux vector
λ_r	Magnitude of the rotor flux vector
λ_{pm}	Same as λ_r , magnitude of rotor flux vector
λ_f	In chapter 3, the flux linkage caused by virtual field winding In other chapters: Same as λ_r , magnitude of rotor flux vector
R_s	Motor phase resistance
L_d, L_q	d and q axis inductance
L_s	Motor phase winding self-inductance
θ_{re}, θ_e	Rotor d-axis electrical angle
ω_{re}, ω_e	Rotor d-axis electrical radius speed
T_e	Electrical torque generated by motor
T_L	Load torque
J	Moment of inertial
B	Viscous coefficient
P	Number of pole pairs
$\hat{}$	Estimated variable
δ	Torque angle

List of Acronyms

AC	Alternative Current
AFPM	Axial Flux Permanent Magnet
BLDC	Brushless Direct Current
CB	Current model Based
CPSR	Constant Power and Speed Range
DB	Deadbeat
DC	Direct Current
DTC	Direct Torque Control
DTFC	Direct Torque and Flux Control
EMF	Electro-Motive Force
FESS	Flywheel Energy Storage System
FOC	Field Oriented Control
FS-MPTC	Finite Set Model Predictive Torque Control
FW	Field Weakening or Flux Weakening
IPMSM	Interior or Inserted PMSM
MMF	Magneto-Motive Force
MPTC	Model Predictive Torque Control
MTPA	Maximum Torque Per Ampere
MTPF	Maximum Torque Per Flux
PM	Permanent Magnet
PMSM	Permanent Magnet Synchronous Machine or Motor
PTC	Predictive Torque Control
PWM	Pulse Width Modulation
SPMSM	Surface-mounted PMSM
SVM	Space Vector Modulation
SVPWM	Space Vector Pulse Width Modulation
THD	Total Harmonic Distortion
V/f	Voltage over frequency
VB	Voltage model Based

CHAPTER 1 Introduction

1.1 Background

1.1.1 PMSM Control and Applications

Features and Classification of PMSM

Permanent Magnet Synchronous Motors (PMSM) belongs to the AC synchronous motor in the AC machine family, as shown in Figure 1.1. It possesses many advantages as compared with other types of machines. Firstly, if compared with DC motor, life can last for tens of years as there's no brush and commutator short life problem. Secondly, as compared with the winded field machine and AC induction machine, the rotor filed is provided by a permanent magnet, thus the efficiency is much higher. In addition, as there's no field winding, the size can be much more compact, making it high in power density, suitable for those applications where size and weight are a concern. Thirdly, the PMSM could be made in a variety forms, such as Surface-mounted PMSM(SPMSM) or Interior or Inserted PMSM (IPMSM), radial flux type or axial direction type, linear type or rotational type, inside rotor or external rotor, etc. making it very flexible and suits a large variety of applications. Lastly, if compared with BLDC, the only difference is that the BLDC features trapezoidal shape back-EMF while the PMSM produces sinusoidal back-EMF. Thus, the torque ripple of PMSM is much smaller under sinusoidal voltage/current control methods such as FOC.

Figure 1.1 Classification of electrical machine family [1]

PMSM Application Overview

Due to these excellent advantages of PMSM, applications of PMSM could be found in a wide range of applications. For example, in industrial servo drive applications[2, 3], PMSM servo drives are found to drive industrial robot, machine tools, that require precision position/speed control, as well as rigid torque-speed curve to reject external disturbance. In transportation, electric vehicles[4] and ship propulsion[5] favor PMSMs due to the high efficiency and high power and torque density; In energy generation and storage applications, PMSM can be used for wind turbines[6, 7], flywheels[8-11], etc.

In recent years, in BLDC dominated applications such as wash machines, compressors in air-conditioners and fridges, fans and purifiers, there's a trend to shape the back-EMF into sinusoidal one, making them a PMSM. This is driven by ever demanding requirement for supreme product efficiency, smooth torque, and low acoustics noise.

PMSM Control Overview

The control of PMSM shares a lot of similarity with other types of AC machines, especially the AC induction machine. In fact, the three typical control methods, ie. The V/f scalar control, the field-oriented control (FOC) and the direct torque control

(DTC), are first been applied on the AC induction machine, and subsequently being transplanted to the PMSM machine. Figure 1.2 gives an overview of the classical ac machine methods based on [12] with updates on the DTC part.

Figure 1.2 Classification of classic PMSM control methods [12]

Scalar control, or V/f control, is an open-loop control method that is widely used by induction machine drives. The typical applications are those do not require fast dynamic responses, such as air-conditioners, conveyor belts, ventilating system, etc. The control of PMSM using V/f control is attracting some research interests due to the ever-increasing demand for motor efficiency. However, V/f for PMSM is inherently unstable and the machine can easily lose synchronization in the high-speed region or under abrupt load change [13-15]. In addition, since the motor current is not directly controlled to achieve maximum torque per ampere (MTPA), the efficiency is not guaranteed to be optimum. A substantial amount of research work has been carried out the efficiency and stability of PMSM drives using V/f control[13, 14, 16-19].

Self-control relies on an absolute position sensor (such as hall sensor or absolute encoder) to measure the rotor flux position, and then do phase commutation based on the position angle. It is more widely known as the six-step commutation control in BLDC control literature[12]. Since PMSM only differs BLDC by the shape of the back-emf, this method can be readily applied to PMSM. However, it is better to output a sinusoidal voltage/current so that the torque and acoustics noise can be reduced. The disadvantages include, an absolute position sensor is required, and the phase advance angle needs to be tuned. [20, 21] .

Field Oriented Control (FOC) is currently the most widely adopted control method for AC drives. It was first introduced in the 1970s by F. Blaschke [22] and K. Hasse

[23] to control induction machines. The general idea of FOC is to decompose the motor current into torque producing component (on q -axis) and flux-producing component (on d -axis), mimicking the dc motor control principle. That is, the torque and flux is controlled indirectly by I_d and I_q current vector components. The core of the FOC scheme are coordinate transformation blocks which allow calculation of field-oriented current components by using $\alpha\beta/dq$ transformation (Park transformation), and reference voltage vector $V_{s\alpha}$ $V_{s\beta}$ by using $dq/\alpha\beta$ transformation (inverse Park transformation) [23]. To know the field orientation, either position sensors such as encoders or resolves, or sensor-less observers, are required to know the rotor angle and perform Park and inverse Park transform.

In the mid-1980s, an alternative high-performance control method, known as Direct Torque and flux Control (DTC or DTFC, but in this thesis, DTFC refers to DTC with SVM), or Direct Self-Control (DSC) was introduced by Takahashi and Noguchi [24] and Depenbrock [25] for induction machines. In the late 1990s, the DTC has been extended to use with the PMSM machine in the stator reference frame, as reported by C. French [26] and L. Zhong [27]. The idea of DTC is to directly control the flux and torque in the stationary reference frame, without using coordinate transformation. Two hysteresis controllers are used to control the estimated torque and flux within a tolerance band of the reference torque and flux. There's no PWM modulation. One of the 7 vectors of the inverter is selected to control the increase/decrease of the torque and flux, depends on which sector the voltage vector is traveling in.

Compared with FOC, the classic DTC method possesses several advantages: 1) very fast torque and flux response, 2) simple structure, no coordinates transform and PI controllers 3), no precise rotor information is required, thus it is inherently sensor-less; 4), less parameter dependency. [28]

On the other hand, the classic DTC suffers several drawbacks, such as (1), large torque and flux ripple, (2) variable switching frequency. These defects generally lead to poor torque and flux regulation, high current harmonics and acoustic noise. In recent years, numerous research efforts have been done in the literature to overcome these limitations for AC drives in general and good results have been achieved, such

as Space Vector Modulation based DTC (SVM-DTC) [29, 30] (i.e. DTFC), Predictive Torque Control (PTC)[31, 32] or model predictive direct torque control (MPTC) [33]. These can be considered as improved DTC and a more detailed review will be given in the literature review part in Section 2.4.

1.1.2 Flywheel Energy Storage System

One of the typical applications of PMSMs is the Flywheel Energy Storage System (FESS). FESS is an ancient energy storage system where the kinetic energy is stored in a rotating flywheel. The energy being stored is proportional to the inertia of the flywheel and to the square of rotation speed, that is, $E = \frac{1}{2}J\omega^2$. A modern FESS consists of a flywheel, motor/generator, power electronics, and mechanical supporting system. As it usually spins very fast to store more energy in a smaller size, the flywheel is enclosed in a strong vacuum or helium chamber for both safety and loss reduction purposes.

The core component of FESS is the electric motor/generator that used to drive the flywheel. Various different electric machine types can be used, for example, induction machine (IM), Switched Reluctance Machine (SRM) and Synchronous Reluctance Machine (SynRM), Brushless DC Motor (BLDC), Permanent Magnet Synchronous Machine (PMSM), Homopolar machine (HM), etc. Among them, PMSM is the most common choice, for its wide speed range, high efficiency and power density, smooth operation, as well as versatile configurations are possible. For example, the flux can be radial direction as most normal PMSMs do, or it can also be axial flux, which can make the machine much compact, offering even higher efficiency, higher power and torque density than the radial flux counterparts.

The typical control requirements for FESS includes:

- Fast charge and discharge speed. FESS is typically used for a short period of energy storage, similar to super-capacitors. The discharge speed must be fast enough when needed. And it should also be capable of fast charging to absorb transient energy from load.
- Low torque ripple. Excessive torque ripple would cause vibration and

acoustics noise which jeopardizes the system life and the usage environment.

- Robust and reliable. As FESS features dozens of years of service life and low maintenance, the controller must be robust and reliable to match these features.
- Wide speed range. This means more energy can be stored going for higher speed, and higher discharge depth going for lower speed.

1.2 Scope of this thesis

This thesis focus on the PMSM control for energy storage applications, and the DTC and its improvements will be of the interest of this thesis. A dual airgap Axial Flux PM (AFPM) machine prototype for FESS was designed and constructed in the lab by [34]. Figure 1.3 illustrates the main components of the machine and the pictures of the machine assembly can be found in Figure 1.4. This machine will be used as an evaluation platform for validating various control algorithms. This machine can be treated as two surface-mounted PMSMs connected in series and can be driven by a single power electronics inverter. It is also cable of controlling the axial displacement for reducing the bearing supporting force. A more detailed introduction and modeling of this machine will be given in Chapter 3.

Figure 1.3 Main components of the dual-airgap AFPM machine in a simplified FESS structure [8]

Figure 1.4 Picture of prototype dual-airgap AFPM machine [8].(a). the rotor and stator assembly. (b). rotor magnet assembly.

It is worth highlighting that, although the evaluation platform is based on this special machine, the proposed methods for the torque and flux control could be applied to any other types of SPMSM as well.

1.3 Objective and Contributions

Given that the FESS requires fast dynamic response as well as high reliability, and that the direct torque control offers such characteristics, it is motivated to explore the application of improved direct torque control for the flywheel energy storage applications. In addition, as the dual airgap AFPM flywheel motor is unique, it would be interesting to test the motor with various control strategies, especially the proposed control methods. The objective of this thesis could be summarized as follows:

- (1) One major drawback of DTC is large torque ripple, especially if the switching frequency is low. However, for large power flywheel applications, it is always desirable to keep the switching frequency low so that the power loss could be lower. Excessive torque ripple also means poor efficiency of energy storage, as it produces vibration and acoustics, and increases the mechanical loss and stress. This study aims to reduce the torque ripple on the family of the DTC control method and its variants.

- (2) For flywheel applications, it is always important that it could have a wider speed range. Wide speed range would mean more energy can be stored on the flywheel. Field weakening is usually adopted to extend the speed range of PM machines at given dc bus voltage. This study aims to propose an improved field weakening strategy in conjunction with DTFC.
- (3) It is desirable for a flywheel system to last for decades with no or minimal maintenance. Reliability is another essential feature of a flywheel. PM machine usually requires rotor position feedback to do commutation and speed calculation. However, having position feedback sensors not only increases the cost but also compromises the reliability when the sensor is faulty. This study aims to study the performance of the observer-based sensor-less method and its application to the dual airgap AFPM machine.

With all the aforementioned objectives, investigations are carried out to explore better control performance for the flywheel motor under study. The dual airgap AFPM flywheel is employed to study the torque and flux control, leaving the axial levitation control untouched. The contributions of this study can be summarized as follows:

- (1) This study proposes a simplified deadbeat model predictive direct torque control strategy for PMSM drives with reduced torque ripple and constant switching frequency. Different from most of the existing approaches, the proposed method calculates the reference voltage vector in the stationary reference frame based on the deadbeat technique. Moreover, space vector modulation is directly employed to synthesize the reference voltage vector which eliminates the complexity of voltage vector selection in conventional PTC.
- (2) This thesis presents a duty cycle based finite-set model predictive torque control for PMSM drives. The proposed method contributes to significantly reduce torque ripples experienced in conventional FS-MPTC by using two voltage vectors(VVs) in one sampling cycle. The duty ratio active VV is determined based on the PI controller, which equivalently adjusts the weighting between torque and flux control in FS-MPTC.
- (3) This thesis explores the direct torque and flux control (DTFC) with SVM for

PMSM drives. To ensure a wide speed range, field weakening (FW) methods for DTFC are generally used. This thesis proposes a new robust FW algorithm for DTFC based SPMSM drives. The proposed algorithm achieves an extended speed region and reduces parameter dependency.

- (4) This thesis conducts further study on the dual airgap AFPM flywheel motor and demonstrates that various above-mentioned control methods could be applied to that motor.

1.4 Organization of the Thesis

There are seven chapters and three appendices, which have been organized as follows:

Chapter 1 briefly introduces the background of the study. The objectives and major contributions are also stated.

Chapter 2 presents the most often used PMSM reference frames and the corresponding models in those frames. The AFPM for the FESS model is also introduced in this chapter. A fundamental literature review related to the research topics of this thesis is also given in this chapter.

Chapter 3 introduces DTFC for SPMSM control. As the feedback for torque and flux are key to DTFC, stator flux observers are evaluated together with DTFC. All the subsequent chapters will be also based on the same evaluated stator flux observer.

Chapter 4 extends the space vector modulation based direct torque and flux wheel control (SVM-DTFC) to field weakening region. A new field weakening control approach is proposed to expend the constant power speed range. This chapter first introduces conventional field weakening operation and MPTA trajectories for SPMSM. Then the proposed SVM-DTFC based field weakening method is given. To verify the validity of the proposed algorithm, simulation results are provided.

Chapter 5 presents performance improvement work over another SVM based predictive torque control (DB-SVM-PTC) strategy. This chapter first gives an overview of the proposed strategy and the proposed system block diagram, followed

by a detailed derivation of the voltage vectors with a 2-step ahead prediction on torque and flux. Then the proposed strategy is first being validated through Simulink simulation using the dual airgap AFPM machine parameters, and then by experiment using the actual dual airgap AFPM flywheel motor.

Chapter 6 works on performance improvement using finite set model predictive torque control (FS-MPTC). A duty cycle based FS-MPTC is presented in this chapter. This chapter first introduces the mathematic models being used and the conventional FS-MPTC, followed by the proposed duty cycle FS-MPTC control algorithm. Comparisons on both steady-state and dynamic performance between classic FS-MPTC and duty FS-MPTC are carried out by simulation. In addition, a comparison between the proposed duty cycle FS-MPTC and a deadbeat-based duty-cycle PTC has also been done to show that duty cycle MPTC does not have the flux-deregulation problem.

Chapter 7 summarizes the key conclusion of this thesis. Meanwhile, some potential research topics related to this thesis are recommended for future works.

The structure of this thesis is illustrated by the block diagram as shown in Figure 1.5.

Figure 1.5 Structure of this thesis

CHAPTER 2 Fundamentals and Literature Review

2.1 Introduction

This chapter first briefly introduced the fundamentals of PMSM control in Section 2.1, including space vector concept, reference frames, PMSM models in three most often used reference frames: $\alpha\beta$ frame, dq frame, and xy frame. Then in Section 2.3, the dual airgap flywheel motor model was presented and one can find that the model is equivalent to an SPMSM. In Section 2.4, a brief review of the improvement of DTC control for PMSM, including SVM-DTC, DB-DTC, Finite Control Set Model Predictive Torque control (FCS-MPTC, or simply MPTC), duty cycle MPTC, etc. Review of Flux Weakening methods are introduced in Section 2.5. Finally, Section 2.6 reviewed the stator flux observer which is essential for direct torque and flux control.

2.2 PMSM Reference Frames and Model

For 3-phase AC electric machine modeling and control, a 3-phase variable, such as voltage, is often expressed in a combined vector form $\mathbf{v}_s$ given by [1]:

$$\mathbf{v}_s = \frac{2}{3}(v_a + \mathbf{a}v_b + \mathbf{a}^2v_c) \quad (2.1)$$

where $\mathbf{a} = e^{j\frac{2\pi}{3}}$. v_a , v_b and v_c instantaneous scalar value in phase a, b, c axis as shown in Figure 2.1(a). Same is true for the current vector $\mathbf{i}_s$, stator flux linkage vector $\boldsymbol{\lambda}_s$, etc. The combined vector is more convenient expressed in two-axis coordinate $\alpha\beta$ frame by expanding $e^{j\frac{2\pi}{3}} = \cos\left(\frac{2\pi}{3}\right) + j\sin\left(\frac{2\pi}{3}\right)$, and this conversion is often called Clarke Transform.

$$\begin{aligned} v_\alpha &= v_a \\ v_\beta &= (v_a + 2v_b)/\sqrt{3} \end{aligned} \quad (2.2)$$

The term $2/3$ is to make sure the combined vector has the same amplitude as a phase-

amplitude.

Figure 2.1 (a).definition of space vector and (b) commonly used reference frame

Using the vector definition method, the voltage vector $\mathbf{v}_s$, current vector $\mathbf{i}_s$, rotor flux vector λ_r and stator flux vector λ_s in the $\alpha\beta$ frame are illustrated in Figure 2.1(b). The $\alpha\beta$ frame is a static frame, with α axis aligned with phase a. Viewing from $\alpha\beta$ frame, all the vectors are rotating. As PMSM is synchronous motor, the vectors are rotating at same (average) speed ω_e .

In addition to the $\alpha\beta$ frame, there are two most often used rotating reference frames: dq reference frame and xy reference frame (or $f\tau$ frame in some literature, for example [35]). The dq frame is called the rotor flux reference frame, with the d-axis aligned with rotor permanent flux linkage. The xy frame is called the stator flux reference frame, with the x-axis aligned with stator flux linkage. The angle between $\alpha\beta$ frame and dq frame is called rotor electrical angle θ_e . The angle between dq and xy frame is called torque angle δ .

It is more convenient to do FOC current control in dq frame and to do DTFC (i.e. DTC) in xy frame. To convert a vector from $\alpha\beta$ frame to dq frame, just need to

multiply the vector with $e^{-j\theta_e}$. Still using the voltage vector as an example, after multiplication, it becomes:

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} \cos\theta_e & \sin\theta_e \\ -\sin\theta_e & \cos\theta_e \end{bmatrix} \begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} \quad (2.3)$$

This process is also known as Park Transform.

To further transform a vector from dq frame to xy frame, simply multiply with $e^{-j\delta}$:

$$\begin{bmatrix} v_x \\ v_y \end{bmatrix} = \begin{bmatrix} \cos\delta & \sin\delta \\ -\sin\delta & \cos\delta \end{bmatrix} \begin{bmatrix} v_d \\ v_q \end{bmatrix} \quad (2.4)$$

The more detailed description of various transforms could be found in Appendix C and also in some good reference books, such as [1] and [36].

2.2.1 Model in the stator reference frame

The electrical model of a PMSM in the stator reference frame (abc or $\alpha\beta$ frame) can be expressed as:

$$\mathbf{v}_s = R_s \mathbf{i}_s + \frac{d\boldsymbol{\lambda}_s}{dt} \quad (2.5)$$

$$\boldsymbol{\lambda}_s = \boldsymbol{\lambda}_a + \boldsymbol{\lambda}_r \quad (2.6)$$

$$T_e = \frac{3}{2} P (\boldsymbol{\lambda}_s \times \mathbf{i}_s) \quad (2.7)$$

The cross product is calculated by augmenting the two-component vector into a three-dimensional vector using a zero in the third row. For example, in the dq frame, $\boldsymbol{\lambda}_s$ and $\mathbf{i}_s$ becomes $[\lambda_{s\alpha} \ \lambda_{s\beta} \ 0]^T$ and $[i_{s\alpha} \ i_{s\beta} \ 0]^T$, respectively. Then the result $\boldsymbol{\lambda}_s \times \mathbf{i}_s$ takes the third item from the resulting vector $[0 \ 0 \ \lambda_{s\alpha} i_{s\beta} - \lambda_{s\beta} i_{s\alpha}]^T$ [36].

Note that the flux linkage is a sum of the armature flux linkage $\boldsymbol{\lambda}_a = L_s \mathbf{i}_s$ and flux linkage caused by permanent on the rotor $\boldsymbol{\lambda}_r = \lambda_{pm} e^{j\theta_e}$.

So, if expressed in $\alpha\beta$ component, the SPMSM model becomes:

$$\begin{aligned} v_{s\alpha} &= R_s i_{s\alpha} + \frac{d\lambda_{s\alpha}}{dt} \\ v_{s\beta} &= R_s i_{s\beta} + \frac{d\lambda_{s\beta}}{dt} \end{aligned} \quad (2.8)$$

$$\begin{aligned} \lambda_{s\alpha} &= L_s i_{s\alpha} + \lambda_{pm} \cos\theta_e \\ \lambda_{s\beta} &= L_s i_{s\beta} + \lambda_{pm} \sin\theta_e \end{aligned} \quad (2.9)$$

$$T_e = \frac{3}{2} P (\lambda_{s\alpha} i_{s\beta} - \lambda_{s\beta} i_{s\alpha}) \quad (2.10)$$

And the mechanical model is given by:

$$T_e = J \frac{d\omega_m}{dt} + B \omega_m + T_L \quad (2.11)$$

where T_L is load torque, J is the moment of inertial, B is viscous coefficient, ω_m is mechanical angular speed. The mechanical speed relates to rotor electrical speed by a number of pole pairs P :

$$\omega_m = P \omega_e \quad (2.12)$$

Note that the mechanical model is universal regardless of the reference frame. So the mechanical equation is neglected in other reference frames.

2.2.2 PMSM in dq reference frame

The model for both IPMSM and SPMSM in dq reference frame is:

$$v_d = R_s i_d + L_d \frac{di_d}{dt} - \omega_e L_q i_q \quad (2.13)$$

$$v_q = R_s i_q + L_q \frac{di_q}{dt} + \omega_e L_q i_q + \omega_e \lambda_{pm} \quad (2.14)$$

$$\lambda_{sd} = L_d i_d + \lambda_{pm} \quad (2.15)$$

$$\lambda_{sq} = L_q i_q$$

$$T_e = \frac{3}{2} P (\lambda_{sd} i_q - \lambda_{sq} i_d) \quad (2.16)$$

$$T_e = \frac{3}{2} P (\lambda_{pm} i_q + (L_d - L_q) i_d i_q) \quad (2.17)$$

When applying for SPMSM, just need to replace L_d and L_q with L_s , since $L_d = L_q = L_s$. Note that for simplicity, the dq components of $\mathbf{v}_s$ are just expressed as v_d and v_q , instead of v_{sd} and v_{sq} . Same for i_d and i_q .

2.2.3 PMSM in xy reference frame

The model for both IPMSM and SPMSM in xy reference frame is given by [28]:

$$V_x = R_s i_x + \frac{d\lambda_s}{dt} \quad (2.18)$$

$$V_y = R_s i_y + \omega_e \lambda_s + \frac{d\delta}{dt} \lambda_s \quad (2.19)$$

$$T_e = \frac{3P}{2L_s} \lambda_s \lambda_{pm} \sin\delta \quad (2.20)$$

where δ is torque angle. λ_s is the amplitude of stator flux linkage vector λ_s and λ_{pm} is the amplitude of the permanent magnet flux linkage vector λ_r . This model is widely used in DTC control of PMSM.

2.3 Dual airgap AFPM flywheel motor model

The design and model of this dual airgap AFPM flywheel motor have been introduced in [8, 11, 37] and in the thesis [34]. For the reader's convenience, the model is rederived in this section.

The dual airgap AFPM flywheel machine consists of two halves, namely the upper half and the lower half. Each upper or lower half has its own stator and rotor. The upper and lower windings can be connected in series so that only one 3-phase power converter is required. The principle of the dual airgap AFPM machine is illustrated in Figure 2.2. The upper half winding and rotor permanent magnet produces an electromagnetic torque T_1 and the lower half winding and rotor permanent magnet produces an electromagnetic torque T_2 . The total torque produces by the machine will be a sum of the two: $T_e = T_1 + T_2$.

In this AFPM machine, the torque is used to control the rotation of the flywheel. The torque equation will be given after the electrical model is derived.

Figure 2.2 Illustration of the dual-airgap AFPM machine principle

Figure 2.3 Three-phase (a,b,c) and rotating (d,q) windings of a 2-pole AFPM machine

As the permanent magnets are mounted on the surface of the disk-type rotor, this machine can be analyzed as a non-salient PM machine. The model of the dual-airgap AFPM machine is firstly obtained based on the coordinate axes of a 2-pole AFPM machine, and then be extended to a $2P$ -pole machine. The three-phase windings and dq -winding layout of the 2-pole AFPM machine are defined in Figure 2.3. The

rotating d - and q -axis windings are in quadrature of each other and their magnetic axes are aligned with the d - and q -axis of the rotor. The d -axis of the rotor is chosen to be aligned with the center lines of the magnets and the q -axis is between the magnets.

For either of the two halves of this machine, assuming the permanent magnets on the rotor are represented by the magnetomotive force of $N_f i_f$ and ignoring the leakage inductance, the flux matrix of the machine half can be expressed as:

$$\lambda = (\lambda_f \quad \lambda_a \quad \lambda_b \quad \lambda_c)^T = \begin{pmatrix} L_{ff} & M_{fa} & M_{fb} & M_{fc} \\ M_{af} & L_{as} & M_{ab} & M_{ac} \\ M_{bf} & M_{ba} & L_{bs} & M_{bc} \\ M_{cf} & M_{ca} & M_{cb} & L_{cs} \end{pmatrix} \begin{pmatrix} i_f \\ i_a \\ i_b \\ i_c \end{pmatrix} \quad (2.21)$$

where:

$$L_{as} = L_{bs} = L_{cs} = L_s = \frac{\mu_0 \pi (R_o^2 - R_i^2) N_s^2}{8P^2 g}$$

$$L_{ff} = L_f = \frac{\mu_0 \pi (R_o^2 - R_i^2) N_f^2}{8P^2 g}$$

$$M_{af} = M_{fa} = L_m \cos \theta$$

$$M_{bf} = M_{fb} = L_m \cos (\theta - 2\pi/3)$$

$$M_{cf} = M_{fc} = L_m \cos (\theta + 2\pi/3)$$

And the PM to winding mutual inductance is given by:

$$L_m = \frac{\mu_0 \pi (R_o^2 - R_i^2) N_s N_f}{8P^2 g} \quad (2.22)$$

$$N_s = k_{w1} N_1 \quad (2.23)$$

N_1 is the number of turns in each phase of the stator windings and k_{w1} is the winding factor.

The magnetomotive force $N_f i_f$ is given by:

$$N_f i_f = 2P \frac{B_r l_m}{\mu_r \mu_0} \quad (2.24)$$

where B_r and μ_r are the remanent flux density and relative permeability of the PM respectively, l_m denotes the magnet length and P is the number of pole pairs. μ_0 is the magnetic permeability of the air with the value of $4\pi \times 10^{-7}$ H/m. The design parameters of this machine are listed in Table 2.1.

By using the Park's transform from abc frame to dq frame, the flux linkage $\lambda = (\lambda_f \ \lambda_a \ \lambda_b \ \lambda_c)^T$ in (3.1) can be transformed into:

$$\begin{bmatrix} \lambda_f \\ \lambda_d \\ \lambda_q \end{bmatrix} = \begin{pmatrix} L_f & \frac{3}{2}L_m & 0 \\ L_m & \frac{3}{2}L_s & 0 \\ 0 & 0 & \frac{3}{2}L_s \end{pmatrix} \begin{bmatrix} i_f \\ i_d \\ i_q \end{bmatrix} \quad (2.25)$$

Table 2.1 The dual airgap AFPM machine design parameters

Item & Symbol	Value & Unit
Number of pole pair, P	2
Equivalent upper air-gap g10	2.877 mm
Equivalent lower air-gap g20	2.763 mm
Magnet type	NdFeB
Remanent flux density, Br	1.25T
Stator outer diameter Ro	45mm
Stator inner diameter, Ri	26mm
PM length of upper rotor, lm1	2.0 mm
PM length of lower rotor, lm2	1.87 mm
Number of slots	24
Number of slots per pole per phase	2
Number of phases	3
Skew angle	18.7 degree
Flywheel rim diameter,	12.4 cm
Number of stator winding turns N1 and N2	208 turns each

Ignoring the leakage inductance, the direct axis inductance L_d and the quadrature axis inductance L_q are obtained as:

$$L_d = L_q = \frac{3}{2} L_s = \frac{3\mu_0\pi(R_o^2 - R_i^2)N_s^2}{16P^2g} \quad (2.26)$$

The amplitude of the flux induced by the permanent magnets in the stator phase is:

$$\lambda_{pm} = L_m i_f = \frac{\mu_0\pi(R_o^2 - R_i^2)N_s N_f}{8P^2g} i_f \quad (2.27)$$

Finally, in the d - q reference frame, the stator flux linkage can be expressed as:

$$\begin{aligned} \lambda_d &= L_d i_d + \lambda_{pm} \\ \lambda_q &= L_q i_q \end{aligned} \quad (2.28)$$

which is the same as the ordinary radial flux PM machine model in the d - q frame.

The electromagnetic torque can be expressed as:

$$T_e = \frac{3}{2} P (\lambda_d i_q - \lambda_q i_d) \quad (2.29)$$

As $L_d = L_q$, the torque equation can be further simplified as:

$$T_e = \frac{3}{2} P \lambda_{pm} i_q \quad (2.30)$$

The voltage equation can be derived as:

$$\begin{aligned} v_d &= R_s i_d + L_d \frac{di_d}{dt} - \omega_e L_q i_q \\ v_q &= R_s i_q + L_q \frac{di_q}{dt} + \omega_e L_d i_d + \omega_e \lambda_{pm} \end{aligned} \quad (2.31)$$

The axial force is derived by using the co-energy method and is given by:

$$\begin{aligned} F &= -\frac{\partial W_{co}}{\partial g} = -\frac{\partial \left[\frac{1}{2} (i_f \ i_d \ i_q) (\lambda_f \ \lambda_d \ \lambda_q)^T \right]}{\partial g} \\ &= \frac{\mu_0 \pi (R_o^2 - R_i^2)}{16 P^2 g^2} \left[(N_f i_f)^2 + \frac{5}{2} N_s N_f i_d i_f + \frac{3}{2} N_s^2 (i_d^2 + i_q^2) \right] \end{aligned} \quad (2.32)$$

Note that in all the above equations the effective air gap is represented by g . It should be replaced by $g_1 = g_{10} - z$ and $g_2 = g_{20} + z$ for the upper half and lower half respectively. g_{10} and g_{20} are design parameters representing the effective airgap length at the equilibrium point. Also, note that z is much smaller than g_{10} and g_{20} (μm scale versus mm scale).

The force and torque produced by the individual half of the machine can be calculated by substituting the corresponding parameters for the half-machine with subscript 1 and 2 respectively.

As the machine is controlled by connecting the upper and lower winding in series, the current will be same ($i_{d1} = i_{d2}, i_{q1} = i_{q2}$), the following equations can be established:

$$R_{s12} = R_{s1} + R_{s2} \quad (2.33)$$

$$L_{s12} = L_{s1} + L_{s2} \quad (2.34)$$

$$L_{d12} = L_{d1} + L_{d2} \quad (2.35)$$

$$L_{q12} = L_{q1} + L_{q2} \quad (2.36)$$

$$\lambda_{pm12} = \lambda_{pm1} + \lambda_{pm2} \quad (2.37)$$

$$T_{e12} = T_1 + T_2 = \frac{3}{2} P \lambda_{pm12} i_q \quad (2.38)$$

$$F_{12} = F_1 + (-F_2) = F_1 - F_2 \quad (2.39)$$

Then, the machine can be considered as a single SPMSM, and as the torque is independent of i_d current, the rotation of the machine can be controlled the same way as SPMSM. In the rest part of this thesis, if not stated explicitly, T_e is to represent the total electrical torque of full machine (instead of using T_{e12}), and same for the machine parameters, $R_s, L_s, L_d, L_q, \lambda_{pm}$, all refer to full machine-level parameters.

2.4 Improvements in DTC methods

Direct torque control (DTC) has been widely used in modern electrical drives due to its simplicity and fast dynamic response [27, 38]. Nonetheless, classical DTC is associated with two problems known as high torque ripples and variable switching frequency. These defects generally lead to poor torque and flux regulation, high current harmonics and a certain level of acoustic noise at low speeds [39-41]]. A substantial amount of research work has been carried out to overcome the aforementioned two problems. To alleviate or overcome these issues, many innovative research proposals have been put forth with varying degrees of success. Notably, the duty ratio control (D-DTC) [42], predictive torque control (P-DTC) [43], constant switching frequency-based DTC (CSF-DTC) [44] and the space vector modulation based DTC (SVM-DTC) [45-47] or DTFC are the most commonly used methods.

Finite Set-Model Predictive Torque Control (FS-MPTC, or simply MPTC) is considered a significant improvement over the classic DTC as it can predict the future steps and dynamic select the optimal voltage vector to satisfy certain cost function.

At the same time, it maintains the advantages of fast response and less rotor position dependency. It also overcomes the one cycle delay in digital control through prediction [48, 49]. The basic idea of these approaches is to select one optimal VV to minimize the torque and flux errors in the next sampling cycle. To realize online optimization, the effects of different VVs are iteratively calculated and compared. This implies high computational burden since the prediction and cost function evaluation must be done for all seven VVs, including six active vectors and one zero vector, for a typical two-level DC/AC inverter fed drives. Although efforts are made to reduce the computational load of the controller by using selective VVs in MPTC [50, 51]. Iterative calculation is still unavoidable because of the existence of several selective VVs in each sampling cycle. Furthermore, for MPTC, the inverter switching frequency is still variable and the mitigation of torque and flux ripples is limited by the application of only one VV in an entire sampling cycle [42].

In order to reduce torque and flux ripples whilst maintaining constant switching frequency, researchers proposed space vector modulation (SVM) based deadbeat control for AC machine drives [29, 46, 52, 53]. These methods are implemented in the rotary d-q reference frame which requires coordinate transformation and decoupling of the cross-coupling terms. This indicates increased algorithm complexity. Moreover, the majority of them did not employ predictive torque control (PTC) [29, 46, 53]. As a consequence, the time delay caused by digital control can not be compensated and the improvement of drive performance is limited. Recently, a deadbeat MPTC approach with discrete SVM is proposed for PMSM drives [52]. In this approach, the deadbeat technique is employed to determine the reference VV in the rotary d-q frame. Then, three discretized virtual VVs that are adjacent to the reference VV are selected as candidates and are iteratively calculated in the online optimization process. By numerically comparing their corresponding cost functions, one optimal virtual VV is chosen and subsequently synthesized by using SVM. Despite its advantages, this approach only approximates the reference VV by using discretized virtual VVs. Consequently, its contribution to torque and flux ripple reduction heavily depends on the accuracy of such approximations, which can be unsatisfying when all three virtual VVs are not very close to the reference VV. Also, its computational burden is still high because three VVs have to be evaluated

iteratively in online optimization and coordination transformation is required by deadbeat control.

As the SVM based methods make uses of 3 VVs, the torque ripple as significantly reduced. However, for high power applications, it is desirable to keep the voltage vector low in order to reduce switching loss. To achieve a balance between reducing torque ripple and switching loss, recently, the duty-cycle based MPTC is proposed in the literature. Compared with conventional MPTC methods, one active VV and one zero VV is used within one sampling period. The time ratio between the period for active VV and the sampling time is called duty ratio. As it is sampling-based, it also resolved the variable switching frequency problem of conventional MPTC methods. The method to find the duty ratio varies from one to another.

Table 2.2 summarizes the features and advantages of the above-reviewed methods.

Table 2.2 Comparison of DTC and MPTC Methods

Methods	Classic DTC; Classic MPTC	DTFC; Deadbeat-SVM MPTC	Duty MPTC
Voltage Vector (VV)	1 VV	3 VVs	2 VVs
Torque Ripple	High	Low	Medium
Switching Freq	Variable	Fixed	Fixed

2.5 Field Weakening Methods

Field weakening (or Flux Weakening, both are abbreviated as FW) is one of the basic functions that modern ac drive should have [54]. The purpose of field weakening is to extend the motor speed beyond base speed until available inverter voltage and current limit is reached. As many applications do not require maximum torque at high speed, a trade-off between torque and speed can be made, as long as the maximum available torque satisfies the application need. This is very useful to ensure wide speed operation of electric drives.

The discussion and derivation of FW trajectories are based on the steady-state model

of PMSM, most conveniently in the dq frame. The max voltage V_{max} is usually limited by dc-link voltage and PWM modulation methods, while the max current I_{max} is limited by either inverter max current or motor max current, whichever is lower. Both refer to amplitude. Hence in dq frame, the constraints are given by:

$$v_d^2 + v_q^2 \leq v_{max}^2 \quad (2.40)$$

$$i_d^2 + i_q^2 \leq i_{max}^2 \quad (2.41)$$

The steady-state PMSM model is obtained by letting $di/dt = 0$ from the dq reference frame model. Then the voltage limit constraint defines an ellipse (for IPMSM) or a circle (SPMSM) with the size inverse proportional to motor speed. The detailed analytical expression for the ellipse or circle can be found in [1]. In addition, the base speed at which FW begins and the maximum speed all depends on machine parameters. For example, for an SPMSM, with $\lambda_{pm} \geq L_s i_{max}$ (called finite speed system, which is the case for most SPMSM), after neglecting the stator resistance, the voltage constraint circle, the base, and maximum speed is given by[1]:

$$i_q^2 + (i_d + \frac{\lambda_{pm}}{L_s})^2 \leq \frac{v_{max}^2}{\omega_e^2 L_s^2} \quad (2.42)$$

$$\omega_{base} = \frac{V_{max}}{\sqrt{\lambda_{pm}^2 + (L_s i_{max})^2}} \quad (2.43)$$

$$\omega_{max} = \frac{V_{max}}{\lambda_{pm} - L_s i_{max}} \quad (2.44)$$

In general, there are various existing field-weakening solutions and they can be roughly divided into feedforward compensation, feedback compensation and a combination of the two [1, 54, 55]. The feedforward compensation adjusts the torque limit based on the FW curve, it is an open-loop method and highly depends on the machine parameter. The feedback compensation method usually employs a PI controller to adjust the i_d reference and i_q limit. The input of the PI controller is the difference between the max voltage and the current regulator output voltage amplitude. This method is robust to parameter variations, but the tuning of the PI controller is by trial and error. The dynamic performance might also be affected.

As the field weakening is expressed in terms of current i_d and i_q , it is straight forward

to be applied with FOC control methods. The majority of the FW method discussed in the literature belongs to this group, for example, [54, 56, 57]. For DTC drives, the control variable is torque and stator flux. The current limiting cannot be directly imposed. Instead, the current limiting is achieved indirectly by imposing torque limit. In addition, as the DTC for PMSM is usually done in stator flux frame (xy frame), the current and voltage limit is expressed with xy components, making it very different from FOC controlled drives. For example, the voltage constraints, torque constraints, and stator flux constraints are given by [35]:

$$|V| = \sqrt{V_x^2 + V_y^2} \leq V_{max}^2 \quad (2.45)$$

$$I_y \leq \sqrt{I_{max}^2 - I_x^2} \quad (2.46)$$

$$T_e \leq \frac{3}{2} P \lambda_s \sqrt{I_{max}^2 - I_x^2} \quad (2.47)$$

$$\lambda_s \leq \frac{\sqrt{V_{max}^2 - (R_s I_x)^2} - R_s I_x}{\omega_e} \quad (2.48)$$

Similar work of FW with DTC of PMSM could be found in [8, 35, 58]. A comparison between FW using FOC and DTC was conducted by [59]. General speaking, FW using DTC in xy frame has several advantages, such as robust to parameter variations, simplicity of calculation and more stable in control [59].

2.6 Stator Flux Observer

The motor stator flux observer not only estimates speed and angle for Park transformation and speed closed-loop control but also provides estimated torque and flux which are essential for direct torque and flux control. It is well known that voltage model-based flux observers are more accurate at high speed, while current model-based flux observers are advantages at low speed. Hence a large number of state-of-art drives utilize closed-loop flux-observers which combines the voltage model and current model for a wide speed range of operation.

The combined model was initially introduced for induction machine drives [60] and employed extensively in recent works on PMSM drives [61],[62],[63],[64],[8, 35, 65,

66], synchronous reluctance drives [67] and induction machine as well [68],[69].

Observer accuracy and dynamic response are of paramount importance to DTC drives. The parameter sensitivity can be analyzed by using FRF as shown in [60]. [68] summarized various combination method and unified them using the so-called observer characteristics functions. It is shown in [68] that the conventional Gopinath-style model possesses a non-linear flux trajectory when transiting between CB and VB. Improvement of making them linear are conducted in [68] and [35], at the cost of increased complexity.

The combined flux observers have significantly improved the sensitivity to dc offset, motor parameter variations, sensor noise, inverter nonlinearities, etc. At a very low-speed range, the stator flux observer provides less accurate stator flux and torque estimation due to motor parameter variations. [65] developed a (disturbance input decoupling) DID stator flux observer which uses disturbance information from a current observer to significantly reduce flux estimation error due to magnetic saturation and PM flux linkage variations for IPMSMs. Stator resistance is the most sensitive parameter at low speeds. [64] incorporated a novel and simple stator resistance estimator into the sensor-less drive to compensate for the effects of stator resistance variation for IPMSM and [8] presented a similar for an SPMSM drive with an even simpler observer structure.

2.7 Summary

In this chapter, the PMSM model in three most often used reference frames are introduced: stationary $\alpha\beta$ frame, rotor flux dq frame, and stator flux xy frame. This thesis is based on these models and reference frames. Specifically, Chapter 3 will introduce the DTFC which is based on frame xy , but the stator flux observers will be based on stationary $\alpha\beta$ frame. Chapter 4 will continue the DTFC with Field Weakening which will still be based on the xy frame. Chapter 5 and Chapter 6 are all based on the stationary reference frame, demonstrating the simplicity of the MPTC methods.

A literature review pertaining to the research topic of this thesis is also conducted in

this chapter. SVM-DTFC, Deadbeat PTC, and FS-MPTC are the main improvements to the DTC and this thesis made contribute based on these methods. The FW method in the i_d - i_q plane which is convenient for FOC needs to be transformed into the flux-torque plane for the DTC based methods. The analysis method for combined flux observer using characteristic function is utilized by this thesis in Chapter 3.2.

CHAPTER 3 Sensor-less Flux and Torque Observer Study for the Direct Torque and Flux Controlled Surface Mounted Permanent Magnet Flywheel Motor

3.1 Introduction

In Chapter 2, the special dual airgap AFPM flywheel motor has been modeled to be an equivalent surface-mounted permanent motor (SPMSM) for the speed and torque control. As an energy storage device, the flywheel FESS is required to provide high burst power with fast dynamic response to meet the stringent demands of future smart grids where the penetration of renewable generation is high. Furthermore, high reliability and robustness are necessary to ensure the proper function of FESS. In practice, all these requirements of FESS are directly imposed on the drives of its SPMSM.

In recent years, numerous control approaches have been proposed to enhance the performance of SPMSM drives. Unfortunately, many of them are based on field-oriented control (FOC) [10, 70, 71], which heavily relies on the knowledge of the rotor position. In contrast, direct torque control (DTC) offers fast dynamic response but much less dependency on rotor position and machine parameters. Direct torque and flux control (DTFC) is one of the most accepted variants of DTC. It employs space vector modulation (SVM) to produce very smooth torque and flux responses while maintaining the main features of classical DTC. The overall block diagram of DTFC is illustrated in Figure 3.1.

The structure of DTFC is very similar to that of FOC. It uses two PI regulators to regulate torque and flux separately in the xy reference frame. The torque reference comes from the external speed control output and the flux reference is from the MPTA curve. The regulator output voltage V_x and V_y are converted into $\alpha\beta$ frame and modulated by an SVPWM. It effectively reduced the torque ripple and making the switching frequency constant than conventional DTC.

$$T_e = \frac{3P}{2L_s} \lambda_s \lambda_{pm} \sin \delta \quad (3.3)$$

From (3.1), neglecting the resistor voltage drop, the stator flux could be directly regulated by x -axis voltage V_x , as given by

$$V_x \approx \frac{d\lambda_s}{dt} \quad (3.4)$$

Differentiating (3.3) with time, and assumes that the stator flux amplitude is being tightly regulated, the change of torque is proportional to the torque angle change as shown in (3.5):

$$\frac{dT_e}{dt} = \frac{3P}{2L_s} \lambda_s \lambda_{pm} \cos \delta \frac{d\delta}{dt} \quad (3.5)$$

The assumption is usually true because DTC gives preference to the flux control.

Neglecting the resistance drop in (3.2) and integrate (3.5) into (3.2), the y -component voltage becomes:

$$V_y \approx \frac{1}{K} \frac{dT_e}{dt} + \omega_e \lambda_s \quad (3.6)$$

where $K = \frac{3P}{2L_s} \lambda_{pm} \cos \delta$.

From the above analysis, the stator flux and torque could be regulated by two PI regulators separately as shown in Figure 3.2 and Figure 3.3 respectively. $R_s i_x$ and $R_s i_y$ could be treated as a disturbance and can be compensated by the PI regulator. $\omega_e \lambda_s$ can also be treated as disturbance, it could be added in as feedforward term for faster response.

The output of the two PI controllers V_x and V_y are then transformed back to the $\alpha\beta$ reference frame by (4.5) and modulated by SVPWM.

Figure 3.2 Stator Flux PI Controller using V_x

Figure 3.3 Torque PI Controller using V_x

$$\begin{bmatrix} V_\alpha \\ V_\beta \end{bmatrix} = \begin{bmatrix} \cos\theta_s & -\sin\theta_s \\ \sin\theta_s & \cos\theta_s \end{bmatrix} \begin{bmatrix} V_x \\ V_y \end{bmatrix} \quad (3.7)$$

In (3.7), the stator flux angle could be calculated by the estimated stator flux observer:

$$\hat{\theta}_s = \tan^{-1}\left(\frac{\hat{\lambda}_{s\beta}}{\hat{\lambda}_{s\alpha}}\right) \quad (3.8)$$

The stator flux observer is then being covered by the next section.

3.3 Stator flux observer

Stator flux, angle, electrical torque, and rotor speeds are the key feedbacks for Direct Torque Control and motor drive speed control. If stator flux is estimated, the rotor flux, torque, speed, and rotor angle can be calculated from it together with the motor current and parameters, as given by (3.9) to (3.14).

$$\hat{T}_e = \frac{3}{2}P(\hat{\lambda}_{s\alpha}i_{s\beta} - \hat{\lambda}_{s\beta}i_{s\alpha}) \quad (3.9)$$

$$\hat{\lambda}_s = \sqrt{\hat{\lambda}_{s\alpha}^2 + \hat{\lambda}_{s\beta}^2} \quad (3.10)$$

$$\hat{\lambda}_{r\alpha} = \hat{\lambda}_{s\alpha} - L_s i_{s\alpha} \quad (3.11)$$

$$\hat{\lambda}_{r\beta} = \hat{\lambda}_{s\beta} - L_s i_{s\beta} \quad (3.12)$$

$$\hat{\theta}_{re} = \tan^{-1} \left(\frac{\hat{\lambda}_{r\beta}}{\hat{\lambda}_{r\alpha}} \right) \quad (3.13)$$

$$\hat{\omega}_{re} = \frac{d}{dt} \hat{\theta}_{re} \quad (3.14)$$

Now the key is to estimate the stator flux. Various stator flux observers are available in the literature. For this thesis study, the three most common stator flux observes are selected for evaluation and they are introduced below.

3.3.1 Stator Flux Observers for Evaluation Study

A. Voltage Model (Observer 1)

From the SPMSM model in (2.5), the stator flux vector can be calculated by integration using (3.15), The voltage integration based method in (3.15) is known as voltage model.

$$\hat{\lambda}_s = \int (\mathbf{v}_s - R_s \mathbf{i}_s) dt + \hat{\lambda}_{s0} \quad (3.15)$$

where $\hat{\lambda}_{s0}$ is the initial stator flux position (i.e. initial rotor flux position as the initial current is zero).

In practical implementation, low pass filter with a very low cut off frequency ω_c is usually employed to avoid pure integrator saturation, as given by (3.16).

$$\hat{\lambda}_s = (\mathbf{u}_s - R_s \mathbf{i}_s) \frac{\omega_c}{s + \omega_c} + \hat{\lambda}_{s0} \quad (3.16)$$

The resultant block diagram for the voltage model is depicted in Figure 3.4:

Figure 3.4 Structure of stator flux observer 1

B. Voltage Model in $\alpha\beta$ frame with PI compensator (Observer 2)

The pure voltage model suffers from integration drifting problem due to measurement noise and DC-offset, the original stator flux estimator in (3.15) is added with a compensation voltage V_{comp} :

$$\hat{\lambda}_s = \int (\mathbf{v}_s - R_s \mathbf{i}_s + \mathbf{V}_{comp}) dt + \hat{\lambda}_{s0} \quad (3.17)$$

V_{comp} is used to compensate for various errors in the stator flux estimation, such as inverter non-linearity, dead time, stator resistance small variations, integration DC-offset, and measurement noise.

$$\mathbf{V}_{comp} = \left(K_p + \frac{K_i}{s} \right) (\mathbf{i}_s - \hat{\mathbf{i}}_s) \quad (3.18)$$

The current again is estimated from the estimated stator flux using (3.19)

$$\hat{\mathbf{i}}_s = (\hat{\lambda}_s - \hat{\lambda}_r) / L_s \quad (3.19)$$

where the rotor flux estimation is reconstructed from the estimated rotor angle $\hat{\theta}_{re}$:

$$\hat{\lambda}_r = \lambda_{pm} \begin{bmatrix} \cos(\hat{\theta}_{re}) \\ \sin(\hat{\theta}_{re}) \end{bmatrix} \quad (3.20)$$

The structure of this observer is depicted in Figure 3.5, from which it is seen that only motor input voltage $\mathbf{v}_{\alpha\beta}$ and output current $\mathbf{i}_{\alpha\beta}$ in the $\alpha\beta$ frame is required. There's

no involvement in speed adaptation and coordinates transformation. It's a very simple structure.

It's worth pointing out that, at very low speed, the stator voltage is very small and the accuracy of the stator flux estimation is limited by the stator resistance variation, inverter nonlinearities, and measurement noise. An R_s estimation method together with this observer is introduced in [18]. For the application of the flywheel, the machine usually operates not at very low speed, thus the resistance change is not considered in this thesis.

Figure 3.5 Structure of stator flux observer 2

C. Voltage and Current Model in dq frame with PI compensator (Observer 3)

This type of observer can be found in [72] and [28], it is also known as Gopinath flux observer. The block diagram for this observer can be seen in Figure 3.6.

Figure 3.6 Stator Flux Observer 3

Comparing Figure 3.5 with Figure 3.6, it is seen that the two observers share a similar structure, the only difference is that the current model in the left side is calculated in the dq frame, based on the dq frame flux model in (2.15). Or equivalently, the current observer is derived from the stator flux model in dq frame as shown in (3.21):

$$\hat{\mathbf{i}}_s = T(\theta_{re})^{-1}(T(\theta_e)\hat{\lambda}_{s_{\alpha\beta}} - \hat{\lambda}_{r_{dq}})/L_{dq} \quad (3.21)$$

where $T(\theta_e)$ and $T(\theta_{re})^{-1}$ are Park and inverse-Park transformation as given in Appendix C, and $\hat{\lambda}_{r_{dq}}$ is the vector $[\lambda_{pm}; 0]$ in dq frame.

The transformation and inverse transformation lead to slightly higher complexity than the second observer which is only in a stationary frame. However, this transformation makes both the stator flux and observed angle being involved in the current model of stator flux, whereas the second observer only used the observed angle.

3.3.2 Explanation with characteristic function

The stator flux observer 3 is a combination of both voltage model which is good at high speed and a current model that is good at low speed. The two models compensate each other and able to achieve a wide range of operations. There is numerous theoretical analysis in the literature for observer 3, such as in [61] and [60]. This thesis took a slightly different approach. Since observer 3 and observer 2 share a similar structure, observer 2 is used for the theoretical analysis. Observer 2 could also be viewed in terms of combined observer theory.

If (3.20) is treated as $\hat{\lambda}_{r_{cm}}$, the diagram in Figure 3.5 could be redrawn as shown in Figure 3.7.

The stator flux could be expressed as (3.22):

$$\hat{\lambda}_s = \hat{\lambda}_{s_{vm}} + \frac{1}{s} \mathbf{V}_{com} \quad (3.22)$$

And the compensation voltage is written as (3.23):

$$V_{com} = \frac{k_p s + k_i}{s} (i_s - \hat{i}_s) \quad (3.23)$$

Figure 3.7 Redrawn stator flux observer showing voltage model and current model

In which the estimated current is derived from:

$$\hat{i}_s = (\hat{\lambda}_s - \hat{\lambda}_{r\_cm}) / L_s \quad (3.24)$$

$$\hat{\lambda}_{r\_cm} = \lambda_{pm} e^{j\hat{\theta}_e} \quad (3.25)$$

Substitute (3.24) and (3.25) into (3.23), V_{com} becomes

$$V_{com} = \frac{k_p s + k_i}{s} \cdot \frac{L_s i_s - (\hat{\lambda}_s - \hat{\lambda}_{r\_cm})}{L_s} \quad (3.26)$$

Let $\hat{\lambda}_{s\_cm} = \hat{\lambda}_{r\_cm} + L_s i_s$, (3.26) can be rewritten as:

$$V_{com} = \frac{k_p s + k_i}{s} \cdot \frac{\hat{\lambda}_{s\_cm} - \hat{\lambda}_s}{L_s} \quad (3.27)$$

Finally, substituting (3.27) into (3.23), and with some arrangement, it turns into a combined voltage model and current model in (3.28):

$$\hat{\lambda}_s = \frac{s^2}{s^2 + \frac{1}{L_s}(k_p s + k_i)} \hat{\lambda}_{s\_vm} + \frac{\frac{1}{L_s}(k_p s + k_i)}{s^2 + \frac{1}{L_s}(k_p s + k_i)} \hat{\lambda}_{s\_cm} \quad (3.28)$$

$$\hat{\lambda}_s = F(s) \hat{\lambda}_{s\_vm} + (1 - F(s)) \hat{\lambda}_{s\_cm} \quad (3.29)$$

The function $F(s)$ is called characteristics equation, given by:

$$F(s) = \frac{s^2}{s^2 + \frac{1}{L_s}(k_p s + k_i)} = \frac{s^2}{s^2 + K_p s + K_i} \quad (3.30)$$

Where K_p and K_i are related to the PI parameters according to (3.31):

$$\begin{aligned} K_p &= \frac{k_p}{L_s} \\ K_i &= \frac{k_i}{L_s} \end{aligned} \quad (3.31)$$

To set the K_p and K_i parameters in the characteristic function, one possible way is to define the transition regions, as given by:

$$\begin{aligned} K_p &= \omega_1 + \omega_2 \\ K_i &= \omega_1 \omega_2 \end{aligned} \quad (3.32)$$

where $\omega_1 < \omega_2$ are motor speed. ω_1 and ω_2 determines the transition region between the current model and voltage model. Current model dominates when speed below ω_1 and voltage model dominates when speed above ω_2 . For this method, it can achieve good results if set $\omega_1 \leq 0.1$ and set ω_2 to a value that the voltage mode model works reliably.

Another possible way to set the PI parameters is based on the following:

$$K_p = 2\xi\omega_0 \quad (3.33)$$

$$K_i = \omega_0^2$$

where ξ is damping ratio and ω_0 is crossover frequency. According to control theory, it's a good practice to set $\xi=1$. The crossover frequency should be carefully selected according to the specific applications.

For example, if the transition region-based method is called method 1, and the crossover frequency-based method is called method 2, a bode plot for comparison can be plotted for voltage mode characteristic function as shown in Figure 3.5(a), and for current mode characteristic function as shown in Figure 3.5(b), respectively.

Figure 3.8 Bode plot of characteristic function for (a) voltage model $F(s)$ and (b) current model $1-F(s)$ for both method 1 and method 2

The bode plot is based on the following values:

- Method 1: $\omega_1 = 0.1$ rad/s, $\omega_2 = 10$ rad/s

- Method 2: $\xi = 1$ and $\omega_0 = 10$ rad/s

Based on the above value, it is seen that the transition region-based method yields better results in terms of earlier entering pure voltage mode.

3.4 Simulation Study

The DTFC drive model for the SPMSM control system is built-in Matlab/Simulink environment as shown in Appendix B. All the three stator flux observers are included in the model. Simulation step length is set at 0.5 us, while the control loop and PWM switching time is set at 10 kHz. The machine parameters listed in Table 3.1 are used for simulation. The observer sampling time is the same as the torque and flux control loop. A low pass filter with a 2ms time constant is used for speed filtering.

Table 3.1 SPMSM parameters for simulation

Rated torque, T	2.85 Nm
Magnet flux linkage, λ_{pm}	0.1905 V·s/rad
Phase resistance, R_s	4.67 Ω
dq -axes inductance, $L_d = L_q$	26.8 mH
Number of pole pairs, P	2
DC bus voltage	150 V
Rated phase current	5 A
No-load speed	1950 rpm
Inertial J	1e-4 kg·m ²
Viscous coefficient B	1e-4 Nm·s/rad

To compare the performance of stator flux observers 1 to 3, the exact same simulation conditions are used. The Flux and Torque PI controllers for the DTFC are listed in Table 3.3. They are designed according to the closed-loop response from Figure 3.2 and Figure 3.3, with tighter control for the Flux. The Observer 1 only has a single cut-off frequency $\omega_c = 0.02$, which cannot be too high. Observer 2 and 3 have the

same PI parameters. The values are roughly derived from $\omega_1 = 0.05$ and $\omega_2=37$ as given by (3.32) and (3.33).

The simulations use step torque response. A 0.2 Nm load torque and reference torque are applied to the model from the beginning and the reference torque steps from 0.2 Nm to 0.3 Nm at 0.1 s. This torque drives the flywheel motor from zero speed to max speed which is around 1920rpm in the 0.2Nm load condition. The estimated speed from stator flux observers is compared with the actual speed (or idea speed) given by the model output.

Table 3.2 Control parameter settings for simulation

Observers	Observer Parameters	Flux Controller PI Parameters	Torque Controller PI parameters
Observer 1	$\omega_c = 0.02$		
Observer 2	$k_{p\_obs} = 1$ $k_{i\_obs} = 0.05$	$k_{p\_l} = 460$ $k_{i\_l} = 6$	$k_{p\_T} = 80$ $k_{i\_T} = 2$
Observer 3	$k_{p\_obs} = 1$ $k_{i\_obs} = 0.05$		

Figure 3.9 presents the simulation results using Observer 1. It is seen that there are large ripples on all of the torque, flux, and speed, especially when the rpm goes higher. Note that when the rpm reaches base speed around 1920rpm, there are large distortions on flux and torque, this is normal phenonium and can also be observed on other observers.

The simulation results using Observer 2 and Observer 3 are presented in Figure 3.10 and Figure 3.11, respectively. Comparing the first subplot of Figure 3.10 (a) and Figure 3.11 (a), it is seen that Observer 3 provides a straighter ramp up and reaches the base speed faster. Comparing the flux and flux error, it is seen that Observer 3 performs much better than Observer 2 as well.

Although Observer 2 is not giving as good result as Observer 3, it still performs better than Observer 1, which can be observed by comparing Figure 3.10 with Figure 3.9.

Figure 3.9 Simulation of SPMSM Sensor-less DTFC Torque Step with Flux Observer 1. (a) Speed, Flux and Torque; (b) Error of Speed, Flux and Torque

Figure 3.10 Simulation of SPMSM Sensor-less DTFC Torque Step with Flux Observer 2. (a) Speed, Flux and Torque; (b) Error of Speed, Flux and Torque

Figure 3.10 and Figure 3.11 also demonstrate excellent torque and flux tracking capability for the DTFC control below base speed. As explained above, the large oscillation toward the end of the simulation is due to the motor speed reaches base speed and torque and flux regulation are limited. They will be addressed by the Field Weakening method in Chapter 4.

Through simulation, this thesis concludes that Observer 3 is the best performer and thus is selected for all the subsequent chapters, despite it's slightly more complex than the other two Observers.

Figure 3.11 Simulation of SPMSM Sensor-less DTFC Torque Step with Flux Observer 3. (a) Speed, Flux and Torque; (b) Error of Speed, Flux and Torque

3.5 Conclusions

This chapter introduced the direct torque and flux control (DTFC) for the flywheel SPMSM. As torque and flux observers are key to the success of DTFC, three different stator flux observers are reviewed and evaluated. Through a simulation study, it is seen that stator flux observer 3 – the combined voltage and current model in dq frame

with PI compensator gives the best performance.

In the next chapter, an improved field weakening method with DTFC of PMSM will be proposed. The simulation is built upon the dual airgap AFPM machine model and the stator flux observer 3 will be used by the rest of the chapters.

CHAPTER 4 A new Field Weakening Control approach for a Direct Torque and Flux Controlled Surface Permanent Magnet Motor

4.1 Introduction

In Chapter 3, the sensor-less direct torque and flux control (DTFC) of SPMSM flywheel was introduced. This chapter is still based on the sensor-less DTFC control, with the focus moving to the flux weakening region.

To guarantee excellent performance of DTFC in high-speed region, the field-weakening (FW) operation is indispensable. In theory, FW operation starts when the motor speed exceeds its base speed, before which the maximum torque per ampere (MTPA) trajectory dominates to minimize the copper loss [6]. When the motor speed is high, change-over from MTPA to FW operation is necessary since the back electromotive force (emf) of the motor goes up with the increase of speed. This indicates that a smaller portion of DC-link voltage is available for torque regulation. Plenty of research papers have been published to investigate the FW operations of direct torque controlled electric machines [73-79]. Unfortunately, many of them focus on the field-weakening operation of direct torque controlled induction motors and thus are not readily applicable to the PMSMs [73, 77, 79]. In [74], a new FW scheme is proposed to extend the virtual signal injection control of IPMSM into the high-speed region. Excellent dynamic response and good parameter robustness are attained by the proposed scheme. Nonetheless, the method is designed for field-oriented controlled IPMSM drives. Hence, it is not suitable for DTC based electrical drives. A linear FW control of IPMSM is initialized in [75] to provide both torque and current regulation in a wide speed range. It simplifies the calculation of working points for FW operation to a large extent while obtaining a smooth speed transition in FW areas. Nevertheless, detailed machine model and d,q-axes currents are used in the method, leading to high parameter and rotor angular position dependency. In [78], a novel torque control method of IPMSM in the FW region is proposed to realize improved

DC-link voltage utilization. It makes use of the voltage angle control with d-axis current feedback. Due to the employment of rotor reference frame coordinate transformation, voltage vector selection scheme and parameter dependent voltage/current vector calculations, this method is associated with high computational complexity and reduced parameter robustness.

The FW schemes for direct torque and flux controlled PMSM drives in [45] and [47] compute the base speed and FW trajectory are purely based on the machine model. Consequently, they are heavily dependent on machine parameters. Recently, a self-learning direct flux vector control algorithm is proposed for PMSM in both constant torque and FW regions [76]. It is reported to be insensitive to machine parameter variations, producing smooth torque response and achieving a good dynamic response. Despite its advantageous features, this method needs a rotor reference frame coordinate transformation and a complex self-learning controller. This increases its dependence on the knowledge of accurate rotor angular position as well as imposing a higher computational burden.

To extend the Constant Power and Speed Range (CPSR) and simultaneously enhance the parameter robustness of DTFC based PMSM drives, this thesis proposes a new FW algorithm that dynamically adjusts reference flux and torque values, using SPMSM for case study. The proposed algorithm employs a simple PI controller and thus features at very low computational complexity. Moreover, the entire control system incorporating the proposed FW algorithm is implemented in the stationary reference frame, which eliminates coordinate transformation. This reduces the rotor position dependency of DTFC. The proposed methods could be extended to IPMSM as well.

The rest of this chapter is organized as follows: The trajectories of MTPA and conventional FW of SPMSM are given in Section 4.2. Then, the proposed DTFC based FW algorithm is presented in Section 4.3. To verify the validity of the proposed algorithm, simulation results are provided in Section 4.4. Finally, a conclusion is drawn in Section 4.5.

4.2 MTPA and Field Weakening Trajectories for SPMSM

In high-performance AC machine drives, two trajectories are commonly used to achieve high efficiency in a wide speed range. They are known as the MTPA and FW trajectories.

4.3.1 Maximum Torque Per Ampere Trajectory

For SPMSM drives, the MPTA is simply $i_d=0$. However, for DTFC, the control variable is torque and flux, which requires the MTPA trajectory to be expressed as $\lambda_s - T_e$ curve.

By defining β as the angle between the stator current vector $\mathbf{i}_s$ and the q-axis of rotor reference frame, dq-axes currents can be expressed as (4.1), where I_s represents the amplitude of stator current. In addition, the electromagnetic torque of a PMSM is given in (4.2) with notation P denoting the number of pole pairs.

$$\begin{aligned} i_d &= -I_s \sin \beta \\ i_q &= I_s \cos \beta \end{aligned} \quad (4.1)$$

$$T_e = \frac{3}{2} P \lambda_{pm} i_q \quad (4.2)$$

$$T_e = \frac{3}{2} P \lambda_{pm} I_s \cos \beta \quad (4.3)$$

Substituting (4.1) into (4.2), the motor torque can be re-derived as (4.3) with respect to I_s . From (4.3), it is observable that the maximum torque can be achieved by controlling I_s and β in an appropriate way. If the derivative of (4.3) is made equal to zero, the condition for MTPA can be developed. This condition is shown in (4.4). In principle, the MTPA trajectory can be realized if the reference trajectory of the flux controller is following the relation in (4.4). The transformation of relation (4.4) to the torque-flux plane is usually implemented numerically by using a look-up-table that provides point-to-point mapping of torque and stator flux amplitude.

$$\lambda_s = \sqrt{\left(\frac{2L_s T_r}{3P \lambda_{pm}}\right)^2 + \lambda_{pm}^2} \quad (4.4)$$

where T_r is torque reference which is the output of the speed PI controller.

4.2.2 Field Weakening Trajectory

When the rotor speed of SPMSM goes above base speed, the MTPA trajectory will not be traversed. Consequently, the FW trajectory must be pursued to avoid the demand for reference phase voltage that exceeds the maximum inverter limit. It is pointed out that the FW trajectory can be briefly divided into two regions, i.e. Region I and Region II dominated by the current limit and voltage limit conditions, respectively.

- **Region I:** In this region, the transition from MTPA to FW begins when the demanded torque starts exceeding the torque limit defined by the maximum available DC-link voltage. This transition is autonomously and smoothly realized by comparing the reference stator flux governed by (4.5) with its counterpart determined by (4.4) and selecting the minimum value. The notation V_{max} represents the maximum phase voltage of SPMSM confined by the available DC-link voltage.

$$|\lambda_s^*| = \frac{V_{max}}{\omega_{re}} \quad (4.5)$$

In general, when the SVM of DTFC operates in linear modulation region, V_{max} is given by (4.6); whereas it is increased to (4.7) in the average sense when the SVM of DTFC operates in the six-step over-modulation region, which is pushing the utilization of DC-link voltage to its upper limit. Here, I_{max} denotes the maximum phase current. By comparing (4.6) and (4.7), it is obvious that the latter increases the utilization rate of DC-link voltage, resulting in a larger average driving torque and subsequently a higher speed limit for SPMSM. Hence, (4.7) is desirable from the perspectives of producing faster dynamic responses and enhanced energy storage capacity for FESS. It needs to be pointed out that, in order to produce an average phase voltage of (4.7), equation (4.8) is usually used. The detailed reason will be explained in the next section.

$$V_{max} = \frac{V_{dc}}{\sqrt{3}} - I_{max}R_s \quad (4.6)$$

$$V_{max} = \frac{2V_{dc}}{\pi} - I_{max}R_s \quad (4.7)$$

$$V_{max} = \frac{2}{3}V_{dc} - I_{max}R_s \quad (4.8)$$

- **Region II:** When the SPMSM reaches very high speed, it enters the deep FW in Region II. In this region, a new torque limit in (4.9) is imposed by the maximum allowable motor/inverter current. Physically, by steering the torque along its limit in (4.9), the torque angle is tracking its limit of 90° for SPMSM. If parameter variation occurs, such as the change in value of L_s caused by magnetic saturation. It can result in an over-estimated torque limit. In this case, instability of drives can be triggered because the actual torque is demanded to track a reference value that is beyond its physical limit in Region II. To ensure the stability of SPMSM drives in the high-speed region, which is of vital importance for FESS, the improvement of parameter robustness of FW is necessary.

$$T_{e(limit)} = \frac{3}{2}P\lambda_{pm} \sqrt{I_{max}^2 - \left(\frac{\lambda_s^2 - \lambda_{pm}^2 - L_s^2 I_{max}^2}{2\lambda_{pm}L_s} \right)^2} \quad (4.9)$$

From the description above, maximized usage of DC-link voltage and parameter robustness of FW are two main concerns for FESS. The realization of these two goals can contribute to significantly improve the performance of FESS. This motivates the proposal of the new FW algorithm in the next section.

4.3 Proposed Field Weakening Algorithm

Based on the previous discussion, the robustness of conventional DTFC based FW is unsatisfying. To overcome this issue, a new FW algorithm is proposed in this section. The proposed FW algorithm contributes to deliver the maximum allowable speed for SPMSM in FESS and control robustness improvement via two simple PI controllers. They can be defined as the reference flux adjustment PI regulator and the torque limit adjustment PI regulator. Their mechanisms are explained in sub-section A and B.

A. Reference Flux Adjustment Scheme

As stated above, the utilization of DC-link voltage must be maximized to produce the maximum phase voltage. In this manner, the highest possible speed, as well as the fastest dynamic response, can be obtained in SPMSM drives. Nonetheless, the direct substitution of (4.7) into (4.5) will cause control failure since (4.7) is only achievable in the average sense instead of the instantaneous sense. To be specific, (4.7) cannot be substituted into (4.5) for instantaneous reference flux computation. Instead, the value of V_{max} needs to be dynamically adjusted to make it approximately equivalent to (4.7) on average. Such an adjustment is realized by the proposed algorithm that is conceptually shown in Figure 4.1. It is visible that equation (4.8) is used in both the input of adjustment PI controller and the base of flux adjustment in (4.10).

Figure 4.1 Block diagram of the proposed reference flux adjustment scheme

Physically, (4.8) represents the maximum transient phase voltage that is achievable only at the six vertices of the hexagonal voltage limit in voltage vector space. Therefore, it can be used as the base of reference flux computation. Decrement (i.e. negative $\Delta |\lambda_s^*|$) will be made on this base when the maximum available transient phase voltage is exceeded as shown in (4.10) and in the right part of Figure 4.1. Similarly, an increment will be added to this base when more phase voltage can be used. Hence, the final reference stator flux $|\lambda_s^*|_{adjusted}$ is dynamically adjusted. To find out the actual maximum allowable phase voltage, the comparison of (4.8) and the output of the PI torque regulator, i.e. v_y , is performed as shown in the left part of Figure 4.1. This comparison checks if the torque PI controller output v_y saturates caused by an unrealizably high reference flux. Then, based on the level of saturation, the reference flux will be rapidly reduced until torque control is de-saturated. This is

the point that the maximum allowable phase voltage is used.

$$|\lambda_s^*|_{adjusted} = \frac{V_{max}}{\omega_{re}} + \Delta|\lambda_s^*| \quad (4.10)$$

B. Torque Limit Adjustment Scheme

It is mentioned in Section 4.2 that the torque limit of Region II in FW operation is sensitive to machine parameter variations. Therefore, to enhance its robustness against parameter changes, an auto-adjustment algorithm for torque limit $T_{e(limit)}$ is proposed as shown in Figure 4.2.

Figure 4.2 Block diagram of the proposed torque limit adjustment scheme

The fundamental logic of such an adjustment scheme is to properly reduce the torque limit $T_{e(limit)}$ when it exceeds the corresponding physical limit caused by variations in machine parameters. Over-estimated $T_{e(limit)}$ is compensated by torque angle δ control in Figure 4.2. In principle, by keeping $\delta \leq 90^\circ$, the stability of FW in **Region II** is guaranteed. Moreover, $\delta = \theta_s - \hat{\theta}_{re}$ (θ_s and $\hat{\theta}_{re}$ denotes the stator flux angle and estimated rotor electrical angle) is independent of machine parameters. Thus, the error $90^\circ - \delta$ can be used directly to compensate for inaccuracy of $T_{e(limit)}$. This is done through the usage of Figure 4.2 and (4.11).

$$T_{e(limit)_adj} = T_{e(limit)} + \Delta T_{e(limit)} \quad (4.11)$$

By incorporating the proposed FW algorithm with torque limit adjustment into DTFC, the overall control system block diagram can be obtained as shown in Figure 4.3.

4.4 Simulation Study

To verify the effectiveness of the proposed robust DTFC based FW algorithm, simulations are carried out in MATLAB/SIMULINK environment as presented in Appendix B. The parameters of dual airgap AFPM SPMSM in Chapter 3 are used in the simulations as shown in Table 3.1. Noticeably, the sampling frequency of the controller is set to be 10 kHz. Simulation results are presented and compared with their counterparts obtained using the conventional DTFC based FW.

Firstly, simulations of step speed response under both the DTFC with conventional FW approach and the proposed FW algorithm are carried. They are shown in Figure 4.4, Figure 4.4 for the conventional approach and in Figure 4.5 for the proposed approach.

Figure 4.4 Conventional FW algorithm: Simulation of SPMSM step speed response from 600rpm to 4000rpm using (4.5) and (4.7)

Figure 4.4 presents the situation of acceleration failure because of the substitution of (4.7) into (4.5) in conventional FW. The direct employment of (4.7) results in a reference flux that is too high for FW operation. Subsequently, improper FW

trajectory is utilized and the acceleration fails. Instead, when (4.6) is substituted into (4.5) in conventional FW, the applied voltage vectors are confined in linear modulation range of SVM. This gives a conservative FW trajectory, but the acceleration from 600rpm to 4000rpm is successful as depicted by subplot 1 of Figure 4.4.

Figure 4.5 Conventional FW algorithm: Simulation of SPMSM step speed response from 600rpm to 4000rpm with the conventional FW algorithm using (4.5) and (4.6)

It is noteworthy that, despite the success in acceleration, the speed response is not smooth. This is caused by the fluctuations in flux and torque responses as illustrated in subplots 2 and 3 of Figure 4.4, respectively. The fundamental reason for such a phenomenon is the lower average phase voltage governed by (4.6), which produces lower flux and torque development capability. This gives fluctuations in flux and torque when DTFC is tuned to provide fast acceleration responses demanding large average torque. Although the fine-tuning of DTFC can eliminate these fluctuations, it will simultaneously slow down the acceleration of SPMSM. In comparison, the proposed method obtains faster acceleration response with smooth flux and torque responses under the same condition as shown in Figure 4.6. By comparing subplot 1 of Figure 4.4 and Figure 4.6, it is observed that the acceleration is nearly 20% faster using the proposed algorithm. Furthermore, from subplot 3 in both of the figures, it

is seen that the average torque output of the proposed algorithm is larger and smoother.

Figure 4.6 Proposed FW algorithm: Simulation of SPMSM step speed response from 600rpm to 4000rpm using (4.10) and (4.8)

Secondly, owing to its capability of applying the maximum phase voltage, the proposed algorithm is able to deliver the maximum shaft power and subsequently drives the SPMSM to reach the highest speed. This is demonstrated in Figure 4.7 and Figure 4.8. By comparing the subplot 1 of these two figures, it is clearly seen that the maximum speed obtained by the conventional DTFC based FW is below 5000rpm. However, the proposed FW algorithm can drive the SPMSM to reach 5500rpm, demonstrating its superior performance. Additionally, the speed response of the proposed FW is faster because of its higher average output torque seen from the comparison of subplot 3 in Figure 4.7 and Figure 4.8.

Figure 4.7 Conventional FW algorithm: Simulation of SPMSM step speed response from 600rpm to 500rpm with the conventional FW using (4.5) and (4.6)

Figure 4.8 Proposed FW algorithm: Simulation of SPMSM step speed response from 600rpm to 5500rpm using (4.10) and (4.8)

Figure 4.9 Conventional FW algorithm: Simulation of SPMSM step speed response from 600rpm to 4000rpm with +20% detuning in L_s

Figure 4.10 Proposed FW algorithm: Simulation of SPMSM step speed response from 600rpm to 4000rpm with +20% detuning in L_s

Another advantage of the proposed FW algorithm is its strong robustness against machine parameter variations. This is verified through the robustness tests shown in Figure 4.9 to Figure 4.12. On one hand, in Figure 4.9 and Figure 4.10, step speed test from 600rpm to 4000rpm is carried out with +20% detuning in L_s . It simulates the effect of L_s decrease caused by magnetic saturation, which results in detuning in the value of L_s used in the control algorithm. By comparing subplot 1 of the two figures, it is seen that speed control fails in the conventional FW with detuned L_s . The speed of SPMSM is clamped and oscillates around 3000rpm due to the fact that + detuning in L_s leads to an over-estimated $T_{e(limit)}$, which is above the actual torque production capability. As a result, the torque control collapses as shown in subplot 3 of Figure 4.9. This is the reason for its speed control failure. In contrast, the proposed FW properly reduces the over-estimated $T_{e(limit)}$ back to its actual physical limit through the torque limit adjustment scheme. Consequently, successful torque and speed control are maintained as depicted by subplots 3 and 1 in Figure 4.10.

On the other hand, step speed test from 600rpm to 4000rpm is carried out with -20% detuning in R_s is also carried out. The results are illustrated in Figure 4.11 and Figure 4.12. This test illustrates the situation of stator resistance R_s increase caused by temperature rising in practical FESS applications. Since the increase in R_s is not detected by the controller, the value of R_s used in the algorithm is detuned. By examining the speed, flux and torque responses in subplots 1, 2 and 3 of Figure 4.11, it is observed that the SPMSM runs out of control with conventional FW due to the detuning in R_s . In addition, the speed estimation using the stator flux observer in Chapter 3 is also greatly deteriorated when the entire drives system loses control. This is catastrophic for the FESS. Unlikely, the excellent drives performance is maintained by the proposed FW with the same -20% detuning in R_s as illustrated by Figure 4.12. This is achieved by the auto-adjustment scheme for reference flux computation.

Figure 4.11 Conventional FW algorithm: Simulation of SPMSM step speed response from 600rpm to 4000rpm with -20% detuning in R_s

Figure 4.12 Proposed FW algorithm: Simulation of SPMSM step speed response from 600rpm to 4000rpm with -20% detuning in R_s

4.5 Conclusion

This chapter proposes a new field-weakening algorithm for direct torque and flux controlled (DTFC) surface-mounted permanent magnet synchronous machine drives. It contributes to enhancing the parameter robustness of field-weakening operations in high-speed drives. Moreover, it effectively extends the constant power speed range of SPMSM drives, which is very important for a flywheel-based energy storage system. Smooth torque and flux responses, as well as constant switching frequency, are obtained due to the employment of space vector modulation in DTFC.

In the subsequent chapters, predictive torque and flux control methods will be discussed, with a focus on torque ripple reduction. The FW methods proposed in this chapter could be potentially extended into the predictive control based methods since they all use flux and torque as a control reference.

CHAPTER 5 Deadbeat Predictive Torque Control with Space Vector Modulation

5.1 Introduction

In this chapter, in order to further diminish the ripples of torque and flux and simultaneously reduce the computational burden while maintaining a constant switching frequency, this chapter proposes a simplified PTC strategy for PMSM drives employing SVM. In the proposed strategy, the reference VV is firstly determined by the deadbeat technique in the stationary reference frame. Then, SVM is directly utilized to synthesize the desired reference VV. Consequently, no iterative calculation is needed and the reference VV is analytically synthesized rather than being approximated. Due to the fact that one cycle implementation delay [80] is compensated by PTC and three VVs are used in each sampling cycle, excellent torque and flux regulation can be achieved. The block diagram of the proposed control strategy is depicted in Figure 5.1.

Figure 5.1 Block diagram of the proposed simplified deadbeat PTC with SVM

5.2 Basic Principle of Torque Prediction

The mathematical model in $\alpha\beta$ frame (2.5) to (2.7) in Chapter 2 can be rewritten as:

$$L_q \frac{d\mathbf{i}_s}{dt} = -R_s \mathbf{i}_s + \mathbf{j}\omega_{re}(\boldsymbol{\lambda}_s - L_q \mathbf{i}_s) + \mathbf{v}_s \quad (5.1)$$

$$\frac{d\boldsymbol{\lambda}_s}{dt} = \mathbf{v}_s - R_s \mathbf{i}_s \quad (5.2)$$

$$T_e = \frac{3}{2} P(\boldsymbol{\lambda}_s \times \mathbf{i}_s) \quad (5.3)$$

It is noted that both the stator flux and the electromagnetic torque can be observed from measured stator currents and applied voltages. To avoid integral saturation caused by sensor imperfections, low pass filters can be used to replace the integrator in (5.2) [81, 82]. The design of flux and torque observer has been studied in Chapter 3 hence not repeated here.

By discretizing (5.1)-(5.3) via forward Euler method and re-writing them in scalar forms, equations (5.4)-(5.6) can be derived.

$$\begin{aligned} \begin{bmatrix} i_{s\alpha}(k+1) \\ i_{s\beta}(k+1) \end{bmatrix} &= \begin{bmatrix} 1 - \frac{R_s}{L_q} T_s & -\omega_{re} T_s \\ \omega_{re} T_s & 1 - \frac{R_s}{L_q} T_s \end{bmatrix} \begin{bmatrix} i_{s\alpha}(k) \\ i_{s\beta}(k) \end{bmatrix} \\ &+ \begin{bmatrix} 0 & \frac{\omega_{re}}{L_q} T_s \\ -\frac{\omega_{re}}{L_q} T_s & 0 \end{bmatrix} \begin{bmatrix} \lambda_{s\alpha}(k) \\ \lambda_{s\beta}(k) \end{bmatrix} + \frac{T_s}{L_q} \begin{bmatrix} v_{s\alpha}(k) \\ v_{s\beta}(k) \end{bmatrix} \end{aligned} \quad (5.4)$$

$$\begin{bmatrix} \lambda_{s\alpha}(k+1) \\ \lambda_{s\beta}(k+1) \end{bmatrix} = \begin{bmatrix} \lambda_{s\alpha}(k) \\ \lambda_{s\beta}(k) \end{bmatrix} + T_s \left(\begin{bmatrix} v_{s\alpha}(k) \\ v_{s\beta}(k) \end{bmatrix} - R_s \begin{bmatrix} i_{s\alpha}(k) \\ i_{s\beta}(k) \end{bmatrix} \right) \quad (5.5)$$

$$T_e(k+1) = \frac{3}{2} P \left(\lambda_{s\alpha}(k+1) i_{s\beta}(k+1) - \lambda_{s\beta}(k+1) i_{s\alpha}(k+1) \right) \quad (5.6)$$

where notations k and T_s represent the k -th sampling instant and the controller sampling period, respectively.

In real-time digital control, the presence of one-cycle delay usually leads to increased torque and flux ripples [42]. To compensate for this effect, the torque and flux should be predicted for one more step, i.e. the prediction of $\boldsymbol{\lambda}_s(k+2)$ and $T_e(k+2)$ are needed. This can be done by following the flow chart in Figure 5.2.

Figure 5.2 Flow chart of predictive torque control.

5.3 Simplified deadbeat PTC with SVM

In PTC, the reference torque T_e^* and flux amplitude $|\lambda_s^*|$ are generally determined by speed regulator and the maximum torque per ampere trajectory, respectively, as illustrated by the block diagram in Figure 5.3.

Figure 5.3 Reference torque and flux amplitude calculation.

Thus, based on the relations $T_e^* = T_e(k+2)$ and $|\lambda_s^*| = \sqrt{\lambda_{s\alpha}^2(k+2) + \lambda_{s\beta}^2(k+2)}$ imposed by deadbeat technique, equations (5.7) and (5.8) can be derived (see Appendix A for detailed derivation).

$$v_{s\beta}(k+1) = \mu_1 + \mu_2 v_{s\alpha}(k+1) \quad (5.7)$$

$$|\lambda_s^*|^2 = [M_1 + v_{s\alpha}(k+1)]^2 + [M_2 + v_{s\beta}(k+1)]^2 \quad (5.8)$$

where μ_1, μ_2 are constants given by

$$\mu_1 = \frac{\frac{2T_e^*}{3P} - M_1N_2 + M_2N_1}{\frac{M_1}{L_q} - N_1}, \quad \mu_2 = \frac{\frac{M_2}{L_q} - N_2}{\frac{M_1}{L_q} - N_1}$$

with M_1, M_2, N_1, N_2 expressed as

$$M_1 = \lambda_{s\alpha}(k+1) - T_s R_s i_{s\alpha}(k+1)$$

$$M_2 = \lambda_{s\beta}(k+1) - T_s R_s i_{s\beta}(k+1)$$

$$N_1 = \left(1 - \frac{T_s R_s}{L_q}\right) i_{s\alpha}(k+1) - (\omega_{re} T_s) i_{s\beta}(k+1) + \frac{\omega_{re} T_s}{L_q} \lambda_{s\beta}(k+1)$$

$$N_2 = (\omega_{re} T_s) i_{s\alpha}(k+1) + \left(1 - \frac{T_s R_s}{L_q}\right) i_{s\beta}(k+1) - \frac{\omega_{re} T_s}{L_q} \lambda_{s\alpha}(k+1)$$

Substituting (5.7) into (5.8) and re-arranging the terms, the voltage vector governing equation can be derived as

$$(1 + \mu_2^2) v_{s\alpha}^2(k+1) + 2(M_1 + M_2\mu_2 + \mu_1\mu_2) v_{s\alpha}(k+1) + (\mu_1^2 + 2M_2\mu_1 + M_1^2 + M_2^2 - |\lambda_s^*|^2) = 0 \quad (5.9)$$

Obviously, there are two solutions for $v_{s\beta}(k+1)$. It is noted that the smaller solution is generally chosen since the larger solution usually results in a voltage vector that exceeds the voltage limit of DC/AC inverter as illustrated by Figure 5.4. After determining the value of $v_{s\beta}(k+1)$, $v_{s\alpha}(k+1)$ can be calculated via (5.7). Consequently, the desired reference VV is determined. To synthesize this reference VV, SVM is employed in this chapter. In SVM, two adjacent active VVs are selected based on the angular position $\theta_v = \tan^{-1} \left(\frac{v_{s\beta}(k+1)}{v_{s\alpha}(k+1)} \right)$ as elaborated in Table 5.1. The time intervals associated with the two active VVs (t_1 and t_2) and zero VV (t_0) are calculated by (5.10)-(5.12), where θ_v^* is computed in Table I and modulation index m is given in (5.13).

$$t_1 = T_s m \frac{\sin\left(\frac{\pi}{3} - \theta_v^*\right)}{\sin\frac{\pi}{3}} \quad (5.10)$$

$$t_2 = T_s m \frac{\sin\theta_v^*}{\sin\frac{\pi}{3}} \quad (5.11)$$

$$t_0 = T_s - t_1 - t_2 \quad (5.12)$$

$$m = \frac{\sqrt{v_{s\alpha}^2(k+1) + v_{s\beta}^2(k+1)}}{\frac{2}{3}V_{dc}} \quad (5.13)$$

Figure 5.4 Possible voltage vectors and voltage limit imposed by inverter.

Table 5.1 Distribution of Active VV in SVM

θ_v	$(0, \frac{\pi}{3})$	$(\frac{\pi}{3}, \frac{2\pi}{3})$	$(\frac{2\pi}{3}, \pi)$	$(-\pi, -\frac{2\pi}{3})$	$(-\frac{2\pi}{3}, -\frac{\pi}{3})$	$(-\frac{\pi}{3}, 0)$
Sector	1	2	3	4	5	6
Active VVs	$V_1(100)$	$V_2(110)$	$V_3(010)$	$V_4(011)$	$V_5(001)$	$V_6(101)$
θ_v^*	θ_v	$\theta_v - \frac{\pi}{3}$	$\theta_v - \frac{2\pi}{3}$	$\theta_v + \pi$	$\theta_v + \frac{2\pi}{3}$	$\theta_v + \frac{\pi}{3}$

5.4 Simulation Results

To verify the effectiveness of the proposed method, a PMSM drive system is modeled in MATLAB/SIMULINK environment as presented in Appendix B. The same parameters of the dual airgap AFPM flywheel motor as given in Table 3.1 are used. The simulation results of the proposed method are compared with those obtained by classical MPTC. Noticeably, the controller sampling frequency is chosen to be 10kHz for both control approaches and one-cycle implementation delay is added in all simulations.

Firstly, reversed torque performance is tested at zero motor speed and the results of both control approaches are given in Figure 5.5. By comparing Figure 5.5 (a) and Figure 5.5 (b), it is seen that the torque ripples produced by the proposed method are maintained to be as low as 16.7% of the ripples of classical MPTC. Moreover, the response time of torque reversal is nearly the same by the two approaches. This means the proposed method contributes to significantly attenuate the torque ripples without degrading the dynamic response.

Figure 5.5 Reversed torque response of PMSM: (a) torque (up) and zoom in (below) of the classical MPTC (b) torque (up) and zoomed in (below) of the proposed method.

The performance of the PMSM speed step is shown in Figure 5.6, where the results obtained by the classical MPTC and the proposed method are illustrated in Figure 5.6

(a) and Figure 5.6 (b), respectively. From subplot 1 of Figure 5.6 (a) and (b), it can be observed that the response time of speed reversal for the afore-mentioned two methods is identical. Nonetheless, torque ripples of the proposed method are much smaller than those of the classical MPTC as shown by subplot 2, indicating the superiority of the proposed method.

Figure 5.6 Reversed speed response of PMSM: (a) classical MPTC speed and torque response (b) proposed MPTC speed and torque response

To further demonstrate the contribution of the proposed method on the reduction of flux ripples and stator current total harmonic distortion (THD), simulations are carried out and the results are given in Figure 5.7 and Figure 5.8. By comparing Figure 5.7 (a) and Figure 5.8 (a), one can see that the flux ripples and current harmonics produced by the proposed approach are much smaller than those of the classical MPTC. This implies a much lower THD, i.e. higher efficiency of PMSM drives.

The frequency spectrum and THD of stator current produced by the classical MPTC and the proposed method are shown in Figure 5.7 (b) and Figure 5.8 (b), respectively.

(a)

(b)

Figure 5.7 Steady-state stator flux and current response of the classical MPTC under 0.3Nm load at 1200 rpm speed: (a) subplot 1: amplitude of stator flux subplot 2: phase-a current; (b) frequency spectrum and THD of stator current

From these two figures, a comparison of the numerical values of stator current THD substantiate the statement that the proposed method is superior in terms of THD. Furthermore, from the frequency spectrums provided in these two figures, it is noticeable that a constant switching frequency of 10kHz is obtained by the proposed method compared to a variable switching frequency of the classical MPTC

(a)

(b)

Figure 5.8 Steady-state stator flux and current response of the proposed method under 0.3Nm load at 1200 rpm speed: (a) subplot 1: amplitude of stator flux, subplot 2: phase-a current; (b) frequency spectrum and THD of stator current.

5.5 Experiment Results

For further investigation and verification of the control system, the experimental platform has been built, as shown in Figure 5.9(a). Hardware implementation of the system is illustrated by functional blocks as shown in Figure 5.9(b). The dual airgap AFPM flywheel motor is used for the experiment. The parameters of the AFPM machine used in experiments are tabulated in Table 3.1. The upper and lower windings are connected in series and the three phases are connected in Y-connection. A 3-phase SiC MOSFET inverter from CREE Inc is employed for the power state. The control algorithm is implemented in the dSPACE DS1103 controller. The sampling and switching periods of the drive are set to be 100 μ s. The deadtime of the MOSFET is set at 2 μ s and the DC bus voltage is set at 100Vdc.

(a)

(b)

Figure 5.9 Experiment platform and the flywheel motor. (a) photo of the overall experiment setup; (b) system function block diagram

The system block diagram in Figure 5.1 is first simulated in Matlab/Simulink. Then the model is generated into code by dSPACE coder and downloaded into the DS1103 main CPU for running. The control variable could be viewed and modified on-the-fly via the dSPACE control desk software running on the PC. The PWM generation is done by a TI DSP inside the dSPACE board. The ADC sampling is using the 16-bit main ADC on the master board. Due to the synchronization issue, there's noise on the measured current, causing periodical switching noise on all the signals.

For this chapter's study, the optical encoder is not used. Instead, the stator flux observer introduced in Chapter 3.3 is used for stator flux estimation, as well as torque calculation and speed feedback.

Figure 5.10 Experiment Result of Classic MPTC: step from 600rpm to 1200rpm. (a) reference and actual speed (b)torque response (c) stator flux

To demonstrate the better performance of the proposed method against the classic MPTC method, the motor was first run at 600rpm and then step to 1200rpm at around 3 sec. Both are operating in a sensor-less way. Figure 5.10 shows the result of the classic MPTC method and Figure 5.11 shows the result of the proposed method. By comparing Figure 5.10 and Figure 5.11, one can see that the speed response of the two methods is similar. However, the classic MPTC produces a very large speed ripple, torque ripple, and flux ripple than that of the proposed method.

The zoom-in view of motor current at 1200rpm in the above two experiments as shown in Figure 5.12. Comparing Figure 5.12 (a) and (b), It is very obvious that the

proposed method produces much clean motor current than the classic MPTC. Thus, the effectiveness of the proposed method is fully demonstrated.

Figure 5.11 Experiment Result of Proposed MPTC: step from 600rpm to 1200rpm.

(a) reference and actual speed (b) torque response (c) stator flux

Figure 5.12 Experiment Result: Steady-state motor current at 1200rpm (a) classic MPTC; (b) proposed MPTC.

5.6 Conclusion

In this chapter, a simplified deadbeat predictive torque control strategy is proposed to improve the performance of PMSM drives with significantly reduced computational burden. SVM technique is employed to mitigate torque and flux ripples while obtaining constant switching frequency. Coordinate transformation is not needed by the proposed method. Compared to the conventional model predictive torque control, the proposed method reduces torque and ripples as well as current THD to a large extent without degrading dynamic responses.

SVM modulation essentially employs 3 VVs (2 active VVs and 1 zero VV) to reduce torque ripple for PTC. In the next chapter, 2 VVs (1 active and 1 zero VV) based method will be proposed for FS-MPTC.

CHAPTER 6 Duty Cycle Based Model Predictive Torque Control for PMSM

6.1 Introduction

Finite Set-Model Predictive Torque Control (FS-MPTC, or simply MPTC) is considered a significant improvement over the classic DTC as it can predict the future steps and dynamic select the optimal voltage vector to satisfy certain cost function. At the same time, it maintains the advantages of fast response and less rotor position dependency.

As only one VV is used on the entire sampling cycle, the conventional FS-MPTC still suffers relative high torque ripples despite the cost function includes the torque and flux error constraints. Increasing switching frequency is able to alleviate this problem, but the computation burden and switching loss also significantly increased. Some recent efforts have been made to mitigate the torque ripple of FS-MPTC, including offline torque ripple minimization [83], virtual voltage vector methods [52, 84], but all at the cost of significantly increased complexity. The idea of applying a duty cycle for the FS-MPTC was recently proposed by [85] for induction machines. It first predicts the effect of zero VV on torque control and compensates for the possible deficiency via a set of discretized new VVs. As the determination of the duty ratio is not based on torque or flux error, this approach implies a degradation on its accuracy, in addition to the increased computation burden.

To effectively address the torque ripple problem, this thesis proposes a method using two VVs (including one active and one zero VV) in each sampling cycle. The proposed method employs a PI controller to dynamically adjust the duty cycle of the active VV. It contributes to overcoming the flux-deregulation problem mentioned in [86]. Block diagram of the proposed method is illustrated in Figure 6.1. This method applies to both IPMSM and PMSM.

Figure 6.1 Block diagram of the proposed model predictive torque control.

6.2 Model and Conventional FS-MPTC

The prediction of torque and flux is based on the mathematical model of PMSM in the stationary ($\alpha - \beta$) reference frame as described in Chapter 2.1. Copied to here as:

$$\mathbf{v}_s = R_s \mathbf{i}_s + \frac{d\lambda_s}{dt} \quad (6.1)$$

$$T_e = \frac{3}{2} P (\lambda_s \times \mathbf{i}_s) \quad (6.2)$$

The conventional FS-MPTC first does torque and flux prediction based on predicted current, and then do iterative cost function minimization by comparing the numerical value of cost function with all the possible voltage vectors. The detail of these two steps as described in [86] are given below:

1) T_e and λ_s prediction

To estimate $T_e(k+1)$ and $\lambda_s(k+1)$, where $(k+1)$ represents the next sampling cycle, the value of $\lambda_s(k)$ must be obtained first. In DTC, the stator flux can be estimated by (6.3),

$$\lambda_s(k) = \lambda_s(k-1) + T_s(\mathbf{v}_s(k-1) - R_s \mathbf{i}_s(k-1)) \quad (6.3)$$

where T_s is the sampling period. Based on the measured stator current $\mathbf{i}_s(k)$ at the current sampling cycle and $\lambda_s(k)$, $\lambda_s(k+1)$ is derived as

$$\lambda_s(k+1) = \lambda_s(k) + T_s(\mathbf{v}_s(k) - R_s \mathbf{i}_s(k)) \quad (6.4)$$

By referring to (6.2), it is seen that $\mathbf{i}_s(k+1)$ is also required for the prediction of $T_e(k+1)$. Theoretically, stator current is predicted based on (6.5),

$$\frac{d\mathbf{i}_s}{dt} = -\frac{R_s}{L_q} \mathbf{i}_s + \frac{\omega_{re}}{L_q} \mathbf{j}(\lambda_s - L_q \mathbf{i}_s) + \frac{1}{L_q} \mathbf{v}_s \quad (6.5)$$

where $\mathbf{j}$ represents $+90^\circ$ phase shift, L_q denotes the stator inductance in q -axis and ω_{re} is rotor electrical speed. By using the Euler forward method, $\mathbf{i}_s(k+1)$ is derived as:

$$\begin{aligned} \begin{bmatrix} i_{s\alpha}(k+1) \\ i_{s\beta}(k+1) \end{bmatrix} &= \begin{bmatrix} 1 - \frac{R_s}{L_q} T_s & -\omega_{re} T_s \\ \omega_{re} T_s & 1 - \frac{R_s}{L_q} T_s \end{bmatrix} \begin{bmatrix} i_{s\alpha}(k) \\ i_{s\beta}(k) \end{bmatrix} \\ &+ \begin{bmatrix} 0 & \frac{\omega_{re}}{L_q} T_s \\ -\frac{\omega_{re}}{L_q} T_s & 0 \end{bmatrix} \begin{bmatrix} \lambda_{s\alpha}(k) \\ \lambda_{s\beta}(k) \end{bmatrix} + \frac{T_s}{L_q} \begin{bmatrix} v_{s\alpha}(k) \\ v_{s\beta}(k) \end{bmatrix} \end{aligned} \quad (6.6)$$

Subsequently, $T_e(k+1)$ is predicted as

$$T_e(k+1) = \frac{3}{2} P(\lambda_s(k+1) \times \mathbf{i}_s(k+1)) \quad (6.7)$$

To compensate for one cycle digital delay, the torque and flux have to be further

predicted for one more step ahead by shifting forward one sampling cycle:

$$T_e(k+2) = \frac{3}{2}P(\lambda_s(k+2) \times \mathbf{i}_s(k+2)) \quad (6.8)$$

In (6.8) the predictions of λ_s^{k+2} and $\mathbf{i}_s^{k+2}$ are obtained as:

$$\lambda_s(k+2) = \lambda_s(k+1) + T_s(\mathbf{v}_s(k+1) - R_s \mathbf{i}_s(k+1)) \quad (6.9)$$

$$\begin{aligned} \begin{bmatrix} i_{s\alpha}(k+2) \\ i_{s\beta}(k+2) \end{bmatrix} &= \begin{bmatrix} 1 - \frac{R_s}{L_q} T_s & -\omega_{re} T_s \\ \omega_{re} T_s & 1 - \frac{R_s}{L_q} T_s \end{bmatrix} \begin{bmatrix} i_{s\alpha}(k+1) \\ i_{s\beta}(k+1) \end{bmatrix} \\ &+ \begin{bmatrix} 0 & \frac{\omega_{re}}{L_q} T_s \\ -\frac{\omega_{re}}{L_q} T_s & 0 \end{bmatrix} \begin{bmatrix} \lambda_{s\alpha}(k+1) \\ \lambda_{s\beta}(k+1) \end{bmatrix} \\ &+ \frac{T_s}{L_q} \begin{bmatrix} v_{s\alpha}(k+1) \\ v_{s\beta}(k+1) \end{bmatrix} \end{aligned} \quad (6.10)$$

2) Iterative cost function minimization

To obtain optimal torque and flux control, the value of cost function (6.11) has to be evaluated with all the possible VVs $\mathbf{v}_s(k+1)$. In (6.11), notation k_f represents the weighting factor of stator flux error. For three-phase, two-level inverter, the number of possible VVs is 7. It is noted that the parameter k_f needs to be tuned to achieve the best dynamic response. A good starting point to select k_f is (6.12):

$$J = |\mathbf{T}_e^{ref} - \mathbf{T}_e^{k+2}| + k_f \left| |\lambda_s^{ref}| - |\lambda_s^{k+2}| \right| \quad (6.11)$$

$$k_f = \frac{T_{rated}}{\lambda_{rated}} \quad (6.12)$$

where T_{rated} and λ_{rated} are the rated torque and stator flux, respectively.

6.3 Proposed Duty Ratio Controller and Improvement of Flux Regulation

6.3.1 PI controller based duty ratio adjustment

As mentioned above, the application of only one VV in an entire sampling cycle inevitably generates high torque ripples. To further diminish the torque ripples, two voltage vectors, including one active VV and one zero VV, are used in the proposed method. This means VV $\mathbf{v}_s(k+1)$ becomes the combination of two VVs $\mathbf{v}_{s1}(k+1)$ and $\mathbf{v}_{s2}(k+1)$. Define the time duration of active VV $\mathbf{v}_{s1}(k+1)$ as t_1 . Then, the time duration of zero VV $\mathbf{v}_{s2}(k+1) = \mathbf{0}$ is $(T_s - t_1)$. In this way, $T_e(k+2)$ can be re-derived from (6.8), (6.9), and (6.10) as

$$T_e(k+2) = \frac{3}{2}P \left\{ (M_1 N_2 - M_2 N_1) + \left[v_{s\alpha}(k+1) \left(N_2 - \frac{M_2}{L_q} \right) + v_{s\beta}(k+1) \left(-N_1 + \frac{M_1}{L_q} \right) \right] t_1 \right\} \quad (6.13)$$

where

$$\begin{cases} K_1 = 1 - \frac{R_s}{L_q} T_s \\ K_2 = \omega_{re} T_s \end{cases}, \begin{cases} M_1 = \lambda_{s\alpha}(k+1) - T_s R_s i_{s\alpha}(k+1) \\ M_2 = \lambda_{s\beta}(k+1) - T_s R_s i_{s\beta}(k+1) \end{cases}$$

$$\begin{cases} N_1 = K_1 i_{s\alpha}(k+1) - K_2 i_{s\beta}(k+1) + \frac{K_2 \lambda_{s\beta}(k+1)}{L_q} \\ N_2 = K_2 i_{s\alpha}(k+1) + K_1 i_{s\beta}(k+1) - \frac{K_2 \lambda_{s\alpha}(k+1)}{L_q} \end{cases}$$

On the basis of (6.13), if the dead-beat based torque control principle $T_e(k+2) = T_e^*$ is applied [86], the resultant time duration of active VV can be derived as

$$t_1 = \frac{\frac{2T_e^*}{3P} - M_1 N_2 + M_2 N_1}{v_{s\alpha}(k+1) \left(N_2 - \frac{M_2}{L_q} \right) + v_{s\beta}(k+1) \left(-N_1 + \frac{M_1}{L_q} \right)} \quad (6.14)$$

However, the direct application of (6.14) can lead to flux-deregulation as reported in

[86]. This thesis proposes a PI controller based duty ratio determination mechanism as shown in Figure 6.2. It is seen from this figure that the duty cycle d is dynamically determined by two PI controllers. One PI controller regulates the torque error and the other regulates the flux error. The larger output of the two PI controllers will be selected as the final duty cycle output. In the worst case, when the duty cycle becomes unity, it turns into a conventional MPTC with $(k+2)$ step prediction. The time duration of active VV is given by $t_1 = d * T_s$.

Figure 6.2 Proposed duty cycle control method.

6.3.2 Pseudo Algorithm of the Proposed Method

The proposed control algorithm can be summarized as follows:

Step 1. Measure the stator current $\mathbf{i}_s(k)$ and rotor speed $\omega_{re}(k)$, estimate the stator flux $\lambda_s(k)$ using (6.3), and set $\mathbf{v}_s(k)$ as the optimal voltage vector applied in the previous sampling cycle.

Step 2. Predict the stator flux $\lambda_s(k+1)$ and stator current $\mathbf{i}_s(k+1)$ using (6.4) and (6.6), respectively.

Step 3. Use torque PI regulator and flux PI regulator to find a duty ratio d and t_1 .

Step 4. Predict the torque $T_e(k+2)$ and stator flux $\lambda_s(k+2)$ using (6.8) to (6.9). Then, evaluate the value of cost function J_i using (6.11) for each VV $\mathbf{v}_s^i(k+1)$.

Step 5. Find the minimum cost function among J_i . Then, determine the corresponding optimal VV $v_s(k + 1)$ with its time duration. Return to Step 1.

6.3.3 Potential flux de-regulation of dead-beat based control

Despite its simplicity, the deadbeat-based torque regulation in (6.14) produces the potentially optimal time duration by considering only the minimization of torque error. It ignores the flux regulation. Even though flux error is still taken into account in cost function (6.11), the selected optimal VV, which originally satisfies both torque and flux control as in the conventional MPTC, is already adjusted by deadbeat-based torque control. By giving preference to torque control, flux de-regulation might be caused since the limited control freedom of changing only the VV may not be sufficient for flux regulation under some circumstances. In other words, the application of the deadbeat principle equivalently increases the weighting of torque control in the optimization. Moreover, the increased weighting is varying dynamically with respect to the changing of operating conditions, such as load and speed. Consequently, it is difficult to ensure proper flux regulation under all operating conditions.

6.4 Simulation Results

To verify the proposed method, simulation studies are performed in Matlab/Simulink (refer to Appendix B for the simulation model) and the results are presented in this section. To compare the performance with that of [86] the same parameters of PMSM are used and are given in Table 6.1. The sampling frequency of both conventional MPTC and the proposed MPTC is set at 5kHz.

The first simulation evaluates the steady-state performance of the proposed method against the conventional FCS-MPTC method. The torque, stator flux, and current waveform harmonics are compared. The results of classic FS-MPTC are shown in Figure 6.3(a) and Figure 6.4(a), and the results of proposed MPTC are shown in Figure 6.3(b) and Figure 6.4(b), respectively.

Table 6.1 PMSM parameter for simulation study

Rated torque	4 Nm
Number of pole pairs	2
Stator resistance	5.5 Ω
Permanent magnet flux linkage	0.297 Wb
d-axis inductance	0.0868 H
q-axis inductance	0.1697 H
DC link voltage	200 V

Comparing Figure 6.3 (a) and (b), it is observed that the proposed method significantly reduces the torque ripples and stator current harmonics and at the same time obtains a smoother flux response compared to the conventional MPTC. There's an approximately 40% torque ripple reduction by using the proposed method. What's more, the flux ripples are used by near 67% compared with the conventional method. The effect of torque and flux ripple reduction could also be reflected in the stator current. The total harmonic distortion (THD) analysis on the stator current produced by two different methods are plotted in Figure 6.4. By comparing Figure 6.4 (a) and (b), it is seen that the current THD of classic MPTC is estimated as 6.09% while the proposed MPTC only at around 2.62%, leading to a 43% reduction on the THD. Based on this, the THD of stator current is estimated as 6.09% and 2.62% for conventional MPTC and the proposed method, respectively. Hence, the THD is attenuated by roughly 43% using the proposed method. From the above discussions, it can be concluded that the proposed duty cycle based MPTC method can significantly improve the steady-state performance of PMSM drives.

The second simulation compares the transient response of the proposed method against classical MPTC. The transient speed and torque response of classical MPTC and proposed MPTC are presented in Figure 6.5 (a) and (b), respectively. By comparing the speed response, both obtain a fast step speed response of approximately 0.08s for a step speed change from 300rpm to 600rpm.

Figure 6.3 Steady-state response of conventional MPTC(a) and proposed MPTC (b) at 300rpm with 3.2Nm load.

Figure 6.4 Current harmonics of conventional MPTC(a) and proposed MPTC (b) at 300rpm with 3.2Nm load.

Thus, the transient speed response of the proposed method is equivalent to the classic method. Nonetheless, if look at the torque ripple as shown in subplots 2 of Figure 6.5 (a) and (b), the proposed method is superior in torque ripple and flux ripple reduction at low load condition and still 20% smaller torque and flux ripple when it reaches torque limit region.

Figure 6.5 Performance of step speed test: (a) speed, torque, and flux responses of PMSM controlled by classical MPTC, (b) speed, torque and flux responses of PMSM controlled by the proposed method

Figure 6.6 Performance of torque reversal test: (a) speed, torque, and flux responses of PMSM controlled by classical MPTC, (b) speed, torque and flux responses of PMSM controlled by the proposed method.

In addition, the performance of the proposed method is also examined in the scenario of dynamic torque reversal when the motor is held standstill. A full load of 4 Nm is applied to test the full load performance at zero speed. The results are given in Figure 6.6. By comparing Figure 6.6 (a) and (b), it is visible that the dynamic torque response is nearly the same under two different control approaches. However, the proposed method generates smaller torque ripples and flux ripples at full load.

Figure 6.7 Illustration of flux de-regulation problem of dead-beat based duty cycle control in (a) and excellent flux regulation of PI-based duty cycle control in (b).
reference in red and actual in blue.

The third simulation evaluates the flux-deregulation problem seen in the dead-beat based approach in [86]. In this simulation, a step speed increase from 0 to 300rpm is implemented with 0.5Nm load; then at 0.5s, a step load of 3.2Nm is applied. At 0.6s, the motor speed step from 300rpm to 600rpm, carrying the 3.2Nm load. The problem of flux de-regulation is illustrated in Figure 6.7(a). It is observed from subplot 3 of Figure 6.7 (a) that the stator flux goes out of control frequently under step load and step speed changes. This is mainly caused by the neglecting of flux regulation in the deadbeat-based duty ratio determination. In comparison, by using the proposed PI-

based duty cycle determination method, the same simulation is carried out and a well-regulated stator flux is obtained together with excellent torque and speed responses as shown in Figure 6.7 (b). Therefore, it can be concluded that the proposed method is capable of reducing torque ripples without sacrificing flux regulation performance.

6.5 Conclusion

This chapter proposes an improved model predictive torque control (MPTC) approach by employing PI regulated duty ratio calculation. Compared to the conventional MPTC, the proposed approach contributes to significantly reduce the torque ripples and stator current harmonics for PMSM drives. Additionally, it maintains the advantages of fast response and less rotor position dependency while avoiding the undesirable flux-deregulation phenomenon

CHAPTER 7 Conclusion and Future Works

7.1 Conclusion

A. General Summary

This thesis focuses on the performance improvement control for direct torque controlled PMSM drives, with flywheel as the potential application. The main research area including torque ripple reduction, speed range extension, robustness enhancement through sensor-less direct torque control and control methods verification on the dual airgap AFPM flywheel motor.

In the first place, the torque ripple reduction of DTC is investigated. Emerging techniques to address the torque ripple problem can be classified into three groups: (1) Model Predictive Torque Control (MPTC) which select an optimal voltage vector(VV) and apply it on the next full control cycle (one VV); (2) Duty cycle based MPTC which finds an optimal active voltage vector and zero voltage vector (two VVs) for the next control cycle and (3), Space Vector Modulation (SVM) MPTC which essentially uses two active vectors and one zero vectors for the next control cycle (three VVs). In Chapter 6, this thesis contributes an improved duty cycle based MPTC which finds the optimal duty cycle by a PI controller. It obtains excellent torque and flux regulation with fixed switching frequency. In addition, SVM based DTFC methods and Deadbeat SVM based MPTC methods are also explored by this thesis in Chapter 4 and Chapter 5 respectively. Especially, in Chapter 5, the predictive deadbeat control is proposed in this thesis. Compared to the existing work, this method is implemented in the stationary reference frame and has a simplified nature. It produces very smooth torque and flux responses using SVM. Moreover, the usage of prediction techniques compensates for the one cycle digital delay and thus, is desirable for the digital control system.

In the second place, fast and stable speed control is generally preferred in a wide speed range for high-speed PMSM drives, especially important for the flywheel storage system. Field Weakening (FW) techniques are generally used to extend the

speed range to allow extra energy storage. Most of the existing FW methods for DTFC based drives have constrained constant power speed range (CPSR) due to the limited usage of DC-link voltage. Moreover, many of these FW approaches are heavily dependent on machine parameters. This can lead to degraded performance of drives or even control instability under some circumstances. To extend the CPSR and simultaneously enhance the parameter robustness of FW operation, this thesis proposes a new robust FW algorithm for DTFC based SPMSM drives in Chapter 4. The proposed algorithm achieves extended CPSR and reduced parameter dependency.

It's worth mentioning that most of the direct torque methods are all less dependent on the accuracy of machine parameters and rotor position, which are inherently more robust than the FOC based methods. However, the stator flux observer plays a vital role in the accurate estimation of flux, torque, angle, and speed. Various sensor-less observers have been proposed in the literature to save the cost of such sensors and to improve the drives' reliability. This thesis studied three stator flux observers and evaluated its performance and compensator gain setting method in Chapter 3. Then the flux observer-based control is experimentally tested out in Chapter 5. It could potentially be used for all the DTC methods discussed in this thesis.

Lastly, the dual airgap AFPM flywheel machine is studied. The machine model is being introduced in Chapter 2. The parameters are being used for simulation study across Chapter 3 to Chapter 5, and it could also be applied to Chapter 6. In addition, the flywheel prototype has been experimentally tested with classic MPTC and dead-beat SVM based MPTC in Chapter 5.

B. Comparison of Methods from Chapter 4 to Chapter 6

Now that there are various different direct torque control methods presented from Chapter 4 to Chapter 6, namely, DTFC-SVM in Chapter 4, DB-SVM PTC in Chapter 5, Duty Cycle PTC in Chapter 6, and the classic MPTC being used as benchmark method in Chapter 5 and Chapter 6, it would be an interesting idea to compare these methods in terms of torque ripple reduction, speed range extension, robustness enhancement, computation simplicity, to show merits and demerits of each method. Although Table 2.2 gives a simple comparison on the torque ripple and switching

frequency, it's too general. In this section, since the simulation models for each of the methods are available as in Appendix B, the AFPM flywheel motor model is being used for simulation with each of the control methods, to give a more concrete comparison. The comparisons are conducted as below.

(1). Torque ripple reduction

Since the torque ripple is affected by speed control parameters and the load condition, the comparison is done by running the AFPM model in torque control mode, and applying the 1.0 Nm torque for both torque reference and load torque. This shall make the motor running in standstill condition, eliminating the impact of speed control. In addition, the sampling, control and switching frequency is set at 10kHz and simulation step length is 0.5us. The DC voltage is set at 150V. The same AFPM parameters as shown in Table 3.1 is used for all the simulations. Then the torque ripple is quantified by using standard deviation. The results of the torque ripple plot are shown in Figure 7.1:

Figure 7.1 Torque Regulation Ripple for Four Compared Methods under 1Nm Load and Reference

The data from the above simulation plot are then analyzed to compare the steady-

state torque ripple. Because the four methods have slightly different dynamic response time, data after 0.01s are used for comparison to make the comparison result more representative. The mean and standard deviation for each method is calculated in Table 7.1. Note that the benchmark method for this comparison is the classic MPTC. The number “1” means the torque ripple obtained by using this method, while a number lower than this means lower ripple, and vice versa. For example, 0.7 means the torque ripple is 70% of the that obtained classic MPTC methods.

From this set of data, it is seen that, overall, all three methods achieved good torque tracking performance, as the mean value is very close to the target torque reference 1 Nm. In terms of steady-state torque ripple, the DTFC method achieves the lowest torque ripple while the classic MPTC is the highest. DB-SVM PTC is slightly lower than DTFC and Duty-cycle MPTC ranks the third.

Table 7.1 Mean and Standard Deviation of Four Compared Methods

Control Method	Classic MPTC	DTFC; (Chapter 4)	DB-SVM MPTC (Chapter 5)	Duty MPTC (Chapter 6)
Mean (Nm)	0.997539	0.999939	1.004791	1.000474
Standard Deviation (Nm)	0.061431	0.002311	0.006953	0.015039
Compare with MPTC:	1	0.037624	0.113181	0.244812

(2). Speed range extension

The benchmark for speed range extension is based on no-load base speed which is 2170rpm for 150Vdc voltage. The conventional FW algorithm using (4.5) and (4.6) as introduced in Chapter 4. From the data in Chapter 4, the max speed for using the conventional field weakening method can achieve 4700rpm while the proposed field weakening method can achieve 5500rpm. Given that the base speed is 2170rpm, the proposed method extends the base speed by 2.53 times and the conventional method extends the base speed by 2.16 times.

In fact, the speed extension range depends on max available bus voltage, the load,

maximum phase current, and machine parameters. Different parameters used for simulation/experiment will result in a different extension range. Thus, a theoretic analysis and calculation of the speed extension range is necessary and is given here.

The stator flux vector λ_s equation given in (7.2) is depicted in the dq frame as shown in Figure 7.2. Note that the current vector i_s is located at the origin and the $L_s i_s$ is parallel to i_s . The circle is given by $L_s i_{s\_max}$. When the motor operates in MTPA mode, the $L_s i_s$ term will rise along the DA line until the maximum current is reached at point A. The vector length OA is the maximum stator flux amplitude $|\lambda_{s\_max}|$ given by (7.2). The minimum stator flux length $|\lambda_{s\_min}|$ happens at point B given by (7.3). During field weakening, the stator flux vector may be located at any point in the sector ABD depending on the actual load condition and current angle.

$$\lambda_s = \lambda_r + L_s i_s \quad (7.1)$$

Figure 7.2 Field Weakening Operating Region

$$|\lambda_{s\_max}| = \sqrt{\lambda_{pm}^2 + (L_s I_{max})^2} \quad (7.2)$$

$$|\lambda_{s\_min}| = \lambda_{pm} - L_s I_{max} \quad (7.3)$$

In steady state, the phase voltage is given by:

$$v_s(j\omega) = R_s i_s(j\omega) + j\omega \lambda_s(j\omega) \quad (7.4)$$

Then when the λ_s reaches the $|\lambda_{s\_max}|$, the resulting maximum voltage is illustrated as shown in Figure 7.3 (a). The max speed in this condition can be solved from (7.5):

$$V_{max}^2 = (R_s I_{max})^2 + (\omega \lambda_{s\_max})^2 - 2 R_s I_{max} \omega \lambda_{s\_max} \cos(\pi - \delta) \quad (7.5)$$

Figure 7.3 Voltage at maximum and minimum stator flux point

In this thesis study, equation (7.6) is used which gives a smaller speed:

$$\omega_{re} = \frac{V_{max} - R_s I_{max}}{|\lambda_{s\_max}|} \quad (7.6)$$

Similarly, when the stator flux is reduced to the minimum, the phase voltage is shown in the triangle as depicted in Figure 7.3 (b), and the speed given by:

$$\omega_{re} = \frac{\sqrt{V_{max}^2 - (R_s I_{max})^2}}{|\lambda_{s\_min}|} \quad (7.7)$$

In the real case, the maximum speed will be smaller than (7.7) as the available torque becomes zero when the stator flux reaches point B.

In Chapter 4, without the field-weakening, the maximum phase voltage is limited by the SVPWM in the linear modulation region. Thus, the maximum available phase voltage is:

$$V_{max1} = V_{dc}/\sqrt{3} \quad (7.8)$$

The conventional FW also uses the $V_{dc}/\sqrt{3}$ as max voltage. The proposed FW is able to push the maximum voltage to (7.9):

$$V_{max2} = 2V_{dc}/\pi \quad (7.9)$$

Without considering other factors, the available voltage using (7.9) will be 10.26% higher than using (7.8):

$$V_{max2}/V_{max1} = 2\sqrt{3}/\pi \quad (7.10)$$

The base speed without field-weakening can be derived by substituting (7.2) and (7.8) into (7.6):

$$\omega_{re\_base} = \frac{V_{dc}/\sqrt{3} - R_s I_{max}}{\sqrt{\lambda_{pm}^2 + (L_s I_{max})^2}} \quad (7.11)$$

Due to the fact that the field weakening is meaningful only when the load is less than the full load, and for the flywheel application, the steady-state load should be very small, the no-load base speed is preferred. This can be found by simply set I_{max} to zero in (7.11):

$$\omega_{re\_base\_noload} = \frac{V_{dc}/\sqrt{3}}{\lambda_{pm}} \quad (7.12)$$

The maximum speed for conventional field-weakening can be found by substituting (7.2) and (7.8) into (7.7):

$$\omega_{re\_max\_convFW} = \frac{\sqrt{(V_{dc}/\sqrt{3})^2 - (R_s I_{max})^2}}{\lambda_{pm} - L_s I_{max}} \quad (7.13)$$

The maximum speed for proposed field-weakening can be derived by substituting (7.2) and (7.9) into (7.7):

$$\omega_{re\_max\_propFW} = \frac{\sqrt{(2V_{dc}/\pi)^2 - (R_s I_{max})^2}}{\lambda_{pm} - L_s I_{max}} \quad (7.14)$$

A numerical case study with different parameters is then summarized in Table 7.2 below. It is seen from the table that a single parameter change makes a different speed extension ratio. The ratio will become more different from different load conditions.

Table 7.2 Speed Range Extension Numerical Case Study (No Load)

Parameters	Case 1	Case 2	Case 3	Case 4
Vdc (V)	150	150	75	150
I _{max} (A _{pk})	5	2.5	5	5
R _s (Ohm)	4.65	4.65	4.65	4.65
L _s (mH)	26.8	26.8	26.8	13.4
λ _{pm} (V·s/rad)	0.1905	0.1905	0.1905	0.1905
ω _{re_base_noload} (rad/s)	454.5	454.5	227.3	454.5
ω _{re_max_convFW} (rad/s)	1120.8	607.0	354.7	512.9
ω _{re_max_propFW} (rad/s)	1278.1	679.0	433.4	584.8
ω _{re_max_convFW} /ω _{re_base_noload}	2.47	1.34	1.56	1.13
ω _{re_max_propFW} /ω _{re_base_noload}	2.81	1.49	1.91	1.29
ω _{re_max_propFW} /ω _{re_max_convFW}	1.14	1.12	1.22	1.14

In summary, the proposed field-weakening method in Chapter 4 excels the conventional field-weakening method for the following two reasons:

- 1). Proposed method pushes the voltage to the maximum of $2V_{dc}/\pi$ which is 10.26% higher than the maximum $V_{dc}/\sqrt{3}$ by conventional method. This is achieved through the flux adjustment PI controller to achieve highest voltage.
- 2). Proposed method pushes the current to the maximum current limit to ensure the flux can be reduced as low as possible, so that the maximum achievable speed can be higher. This is achieved through the torque adjust PI. The conventional field-weakening purely based on machine parameters and thus minimum stator flux cannot guaranteed.

From the above summary, it is also seen that the principle applies to the control methods in other chapters as well.

(3). Robustness enhancement

For robustness enhancement, all the four methods studied in this thesis uses a stator flux observer and removed the encoder as position/speed feedback. The amount of robustness enhancement is the same for all four methods in terms of position feedback sensor dependency.

In terms of the field-weakening capability with respect to the machine parameters, the proposed field-weakening is able to tolerate 20% stator resistance and 20% inductance detune, as demonstrated in Chapter 4, Figure 4.9 to Figure 4.12.

(4). Computation simplicity

In terms of computation simplicity, the amount of calculations required for the “TqFluxControlAndSwitching” block as in Appendix B.3 to B.6 is given in Table 7.3. The data is extracting from the code generation report by generating the model into generic C code using “ert.tlc” on Intel X86 CPU. The PWM generation part was excluded as it is handled by CPU hardware.

Table 7.3 Computation resource report for the four compared methods

Control Method	Classic MPTC	DTFC; (Chapter 4)	DB-SVM MPTC (Chapter 5)	Duty MPTC (Chapter 6)
Lines of Code	228	181	236	271
Accumulated Stack Size (bytes)	422	213	254	405
Complexity:	29	27	31	36

A screenshot of the generated step function report for the MPTC model is given in Figure 7.4 as an example.

View function metrics in a call tree format or table format. Accumulated stack numbers include the estimated stack size of the function plus the maximum of the accumulated stack size of the subroutines that the function calls.

View: Call Tree | [Table](#)

Function Name	Accumulated Stack Size (bytes)	Self Stack Size (bytes)	Lines of Code	Lines	Complexity
[+] TqFluxControlAndSwitching_step1	422	300	228	387	29
[+] TqFluxControlAndSwitching_initialize	24	0	10	34	1
TqFluxControlAndSwitching_step0	3	3	22	57	3
rtIsInf	0	0	1	4	2

Figure 7.4 Code Generation Report Screenshot for MPTC

7.2 Future Works

Despite the progress made so far in the thesis, due to time constraints, there're potentially several other topics that are worth studying. Future works are planned in the following areas:

A. Sensor-ed and Sensor-less Performance Difference Study

This thesis only explores one stator flux observer and the performance difference from the control using encoder feedback is worth quantitatively studied. In addition, some other interesting stator flux observers, for example, the one introduced by [87], could be comparatively evaluated. The same characteristic function analysis method

could be applied to this observer.

B. Axial direction displacement control

So far this thesis focus on the rotating direction torque and flux control for the dual air-gap AFPM flywheel motor. The control of the axial direction force was only being addressed by [34] with classical PI control. Special sensor is required to measure the z-axis displacement (for example, Micro-Epsilon Co. DT110-T). All other control methods have not been addressed yet. It would be interesting to combine the z-axis control with the torque and flux control, or at least, to quantify to how much extent the two have coupled.

C. Flywheel in the energy storage system

This thesis has been focus on the control of the flywheel motor itself. The control of the flywheel as part of a larger system needs to be studied. For example, when the flywheel charges and discharges rapidly, the dc bus voltage need to be regulated. As pointed out by [10], the operating point will change under rapid discharge control, this leads to the degraded performance with the PI controller.

Appendix A Derivation of (5.7)

The deadbeat based DTFC target to let $k+2$ step prediction meets the reference value:

$$T_e^* = T_e(k+2) \quad (\text{A.1})$$

$$|\lambda_s^*| = \sqrt{\lambda_{s\alpha}^2(k+2) + \lambda_{s\beta}^2(k+2)} \quad (\text{A.2})$$

From A.2, we have

$$\lambda_{s\alpha}^2(k+2) + \lambda_{s\beta}^2(k+2) - |\lambda_s^*|^2 = 0 \quad (\text{A.3})$$

To predict $\lambda_{s\alpha\beta}(k+2)$ and $T_e(k+2)$, $i_{s\alpha\beta}(k+2)$, $\lambda_{s\alpha\beta}(k+1)$ and $i_{s\alpha\beta}(k+1)$ are required:

$$\begin{bmatrix} \lambda_{s\alpha}(k+1) \\ \lambda_{s\beta}(k+1) \end{bmatrix} = \begin{bmatrix} \lambda_{s\alpha}(k) \\ \lambda_{s\beta}(k) \end{bmatrix} + T_s \left(\begin{bmatrix} v_{s\alpha}(k) \\ v_{s\beta}(k) \end{bmatrix} - R_s \begin{bmatrix} i_{s\alpha}(k) \\ i_{s\beta}(k) \end{bmatrix} \right) \quad (\text{A.4})$$

$$\begin{aligned} \begin{bmatrix} i_{s\alpha}(k+1) \\ i_{s\beta}(k+1) \end{bmatrix} &= \left(1 - \frac{T_s R_s}{L_q}\right) \begin{bmatrix} i_{s\alpha}(k) \\ i_{s\beta}(k) \end{bmatrix} \\ &+ \frac{T_s w_{re}}{L_q} \begin{bmatrix} -(\lambda_{s\beta}(k) - L_q i_{s\beta}(k)) \\ (\lambda_{s\alpha}(k) - L_q i_{s\alpha}(k)) \end{bmatrix} + \frac{T_s}{L_q} \begin{bmatrix} v_{s\alpha}(k) \\ v_{s\beta}(k) \end{bmatrix} \end{aligned} \quad (\text{A.5})$$

(A.4) and (A.5) can be calculated from the k -step known variable. To find $i_{s\alpha\beta}(k+2)$, A prediction from (A.4) and (A.5) and the unknown $v_{s\alpha\beta}(k+1)$ is conducted as shown in (A.6):

$$\begin{aligned}
 \begin{bmatrix} i_{s\alpha}(k+2) \\ i_{s\beta}(k+2) \end{bmatrix} &= \left(1 - \frac{T_s R_s}{L_q}\right) \begin{bmatrix} i_{s\alpha}(k+1) \\ i_{s\beta}(k+1) \end{bmatrix} \\
 &+ \frac{T_s w_{re}}{L_q} \begin{bmatrix} -(\lambda_{s\beta}(k+1) - L_q i_{s\beta}(k+1)) \\ (\lambda_{s\alpha}(k+1) - L_q i_{s\alpha}(k+1)) \end{bmatrix} \\
 &+ \frac{T_s}{L_q} \begin{bmatrix} v_{s\alpha}(k+1) \\ v_{s\beta}(k+1) \end{bmatrix}
 \end{aligned} \tag{A.6}$$

As the $\lambda_{s\alpha\beta}(k+2)$ prediction is given by (A.7)

$$\begin{aligned}
 \begin{bmatrix} \lambda_{s\alpha}(k+2) \\ \lambda_{s\beta}(k+2) \end{bmatrix} &= \begin{bmatrix} \lambda_{s\alpha}(k+1) \\ \lambda_{s\beta}(k+1) \end{bmatrix} \\
 &+ T_s \left(\begin{bmatrix} v_{s\alpha}(k+1) \\ v_{s\beta}(k+1) \end{bmatrix} - R_s \begin{bmatrix} i_{s\alpha}(k+1) \\ i_{s\beta}(k+1) \end{bmatrix} \right)
 \end{aligned} \tag{A.7}$$

Now, bring (A.7) back to (A.3):

$$\begin{aligned}
 &[\lambda_{s\alpha}(k+1) + T_s v_{s\alpha}(k+1) - T_s R_s i_{s\alpha}(k+1)]^2 \\
 &+ [\lambda_{s\beta}(k+1) + T_s v_{s\beta}(k+1) \\
 &- T_s R_s i_{s\beta}(k+1)]^2 - |\lambda_s^*|^2 = 0
 \end{aligned} \tag{A.8}$$

Define the following two auxiliary variables:

$$M_1 = \lambda_{s\alpha}(k+1) - T_s R_s i_{s\alpha}(k+1)$$

$$M_2 = \lambda_{s\beta}(k+1) - T_s R_s i_{s\beta}(k+1)$$

Then (A.8) can be simplified as:

$$\begin{aligned} & [M1+T_s v_{s\alpha}(k+1)]^2 + [M2+T_s v_{s\beta}(k+1)]^2 - |\lambda_s^*|^2 \\ & = 0 \end{aligned} \quad (\text{A.9})$$

On the other hand, the k+2 torque prediction is given by:

$$\begin{aligned} T_e^* = T_e(k+2) = & \frac{3}{2}P \left(\lambda_{s\alpha}(k+2)i_{s\beta}(k+2) \right. \\ & \left. - \lambda_{s\beta}(k+2)i_{s\alpha}(k+2) \right) \end{aligned} \quad (\text{A.10})$$

Substitute (A.10) with (A.6) and (A.7), and re-arrange, we have:

$$\begin{aligned} & [M1+T_s v_{s\alpha}(k+1)] \\ & * \left[\left(1 - \frac{T_s R_s}{L_q} \right) i_{s\beta}(k+1) \right. \\ & \quad \left. - \frac{T_s \omega_{re}}{L_q} (\lambda_{s\alpha}(k+1) - L_q i_{s\alpha}(k+1)) \right. \\ & \quad \left. + \frac{T_s}{L_q} v_{s\beta}(k+1) \right] - [M2+T_s v_{s\beta}(k+1)] \\ & * \left[\left(1 - \frac{T_s R_s}{L_q} \right) i_{s\alpha}(k+1) \right. \\ & \quad \left. + \frac{T_s \omega_{re}}{L_q} (\lambda_{s\beta}(k+1) - L_q i_{s\beta}(k+1)) \right. \\ & \quad \left. + \frac{T_s}{L_q} v_{s\alpha}(k+1) \right] = \frac{2T^*}{3P} \end{aligned} \quad (\text{A.11})$$

We then define:

$$N_1 = \left(1 - \frac{T_s R_s}{L_q} \right) i_{s\alpha}(k+1) + \frac{(\omega_{re} T_s)}{1} i_{s\beta}(k+1) - \frac{\omega_{re} T_s}{L_q} \lambda_{s\beta}(k+1)$$

$$N_2 = \left(1 - \frac{T_s R_s}{L_q}\right) i_{s\beta}(k+1) + \frac{T_s w_{re}}{L_q} (\lambda_{s\alpha}(k+1) - L_q i_{s\alpha}(k+1))$$

The torque equation in (A.11) can be simplified as:

$$\begin{aligned} & [M1 + T_s v_{s\alpha}(k+1)] \left[N2 + \frac{T_s}{L_q} v_{s\beta}(k+1) \right] \\ & - [M2 + T_s v_{s\beta}(k+1)] \left[N1 \right. \\ & \left. + \frac{T_s}{L_q} v_{s\alpha}(k+1) \right] = \frac{2T^*}{3P} \end{aligned} \quad (A.12)$$

Let $V_\alpha = T_s v_{s\alpha}(k+1)$ and $V_\beta = T_s v_{s\beta}(k+1)$, (A.12) turns into the form of (5.7):

$$V_\beta = \frac{\frac{2T^*}{3P} + M2N1 - M1N2}{\frac{M1}{L_q} - N1} + \frac{\frac{M2}{L_q} - N2}{\frac{M1}{L_q} - N1} V_\alpha \quad (A.13)$$

Appendix B Simulation Models

B.1 Top Level Overall Model shared across Chapter 3 to Chapter 6

B.2 Speed Control and Field Weakening Model

This part of the model sets the speed reference, torque reference, and flux reference. The conventional field weakening part in which servers adjust the torque and flux reference is also included here. This model is also shared from Chapter 3 to Chapter 6.

B.3 DTFC Model for Chapter 3 and Chapter 4

This is the “TqFluxControlAndSwitching” block for DTFC in Chapter 3 and 4.

B.4 MPTC Model for Chapter 5 and Chapter 6

This is the “TqFluxControlAndSwitching” block for MPTC which is used as a benchmark control method in Chapter 5 and 6.

B.5 DB-SVM-PTC Model for Chapter 5

This is the “TqFluxControlAndSwitching” block for the Deadbeat-SVM PTC method in Chapter 5.

B.7 Stator Flux, Torque, Speed and Angle Observers for Chapter 3 to Chapter 6

This is the “StatorFlux_Tq_Angle_Speed_Observer” block for Chapter 3 to Chapter 6. The Observer 3 is selected in the variant subsystem

Observer 1 Model

Observer 2 Model

Observer 3 Model and the code:

```
function [Iap_est,Ibt_est,Flux_Ap,Flux_Bt,Flux_Act_Ap,Flux_Act_Bt,Theta] =  
Observer_dq(Valpha,Vbeta,Ialpha,Ibeta,ModelParam)  
  
global Rs,  
global flux_r,  
global Ld,  
global Lq,  
global Ts,  
%value declaired as persistent cannot be output. so _k1 means previous  
%value.  
persistent Flux_Ap_k1,  
persistent Flux_Bt_k1,  
persistent IntQ_Tempk1,  
persistent IntD_Tempk1,  
persistent IntQ_Satk1,  
persistent IntQ_UnSatk1,  
persistent IntD_Satk1,  
persistent IntD_UnSatk1,  
  
persistent Theta_est,  
  
if isempty(Flux_Ap_k1)  
    Flux_Ap_k1=(0);  
    Flux_Bt_k1=(-flux_r);
```

```
IntQ_Tempk1=(0);
IntD_Tempk1=(0);

IntQ_Satk1=(0);
IntQ_UnSatk1=(0);

IntD_Satk1=(0);
IntD_UnSatk1=(0);

Theta_est=(0);
end

Kp_Apest=ModelParam(1);%kp_obs
Ki_Apest=ModelParam(2); %ki_obs
Kp_Btest=ModelParam(1);%kp_obs
Ki_Btest=ModelParam(2);%kp_obs

Int_Max=3*Rs; %%2*Rs
Int_Min=-3*Rs; %%-2*Rs
CosEst=cos(Theta_est);
SinEst=sin(Theta_est);
%%Group 1
Flux_d_est=Flux_Ap_k1*CosEst + Flux_Bt_k1*SineEst; %%Correct Flux_d_est
Flux_q_est=-Flux_Ap_k1*SineEst + Flux_Bt_k1*CosEst; %%Correct Flux_q_est
```

```
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
```

```
Id_est=(Flux_d_est - flux_r)/Ld;
Iq_est=Flux_q_est/Lq;
%%Group 1
Iap_est=Id_est*cosEst - Iq_est*sinEst; %%Correct Iap_est
Ibt_est=Id_est*sinEst + Iq_est*cosEst; %%Correct Ibt_est
%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%%
```

```
Iap_est_err = (Ialpha - Iap_est);
Ibt_est_err = (Ibeta - Ibt_est);
PropQ_Temp=Kp_Apest*Iap_est_err;
PropD_Temp=Kp_Btest*Ibt_est_err;
IntQ_Temp=IntQ_Tempk1 + (Ki_Apest*Iap_est_err + (IntQ_Satk1 - IntQ_UnSatk1))*Ts;
%%Vap_comp_k1=IntQ_Satk1;
IntD_Temp=IntD_Tempk1 + (Ki_Btest*Ibt_est_err + (IntD_Satk1 - IntD_UnSatk1))*Ts;
%%Vbt_comp_k1=IntD_Satk1;
if(IntQ_Temp>Int_Max)
    IntQ_Temp=Int_Max;
end
if(IntQ_Temp<Int_Min)
    IntQ_Temp=Int_Min;
end
if(IntD_Temp>Int_Max)
    IntD_Temp=Int_Max;
end
```

```
if(IntD_Temp<Int_Min)
    IntD_Temp=Int_Min;
end
IntQ_UnSatk=PropQ_Temp + IntQ_Temp;
IntD_UnSatk=PropD_Temp + IntD_Temp;
if(IntQ_UnSatk>Int_Max)
    IntQ_Satk=Int_Max; %%Vap_comp=IntQ_Satk;
else
    IntQ_Satk=IntQ_UnSatk;
end
if(IntQ_UnSatk<Int_Min)
    IntQ_Satk=Int_Min; %%Vap_comp=IntQ_Satk;
else
    IntQ_Satk=IntQ_UnSatk;
end
if(IntD_UnSatk>Int_Max)
    IntD_Satk=Int_Max; %%Vbt_comp=IntD_Satk;
else
    IntD_Satk=IntD_UnSatk;
end
if(IntD_UnSatk<Int_Min)
    IntD_Satk=Int_Min; %%Vbt_comp=IntD_Satk;
else
    IntD_Satk=IntD_UnSatk;
end
CompAp_k=Valpha - Ialpha*Rs + IntQ_Satk*1;
CompBt_k=Vbeta - Ibeta*Rs + IntD_Satk*1;
```

```
Flux_Ap= Flux_Ap_k1 + (CompAp_k)*Ts;
Flux_Bt= Flux_Bt_k1 + (CompBt_k)*Ts;
%%Active Flux Estimation
Flux_Act_Ap=Flux_Ap - Lq*Ialpha; %Debug
Flux_Act_Bt=Flux_Bt - Lq*Ibeta; %Debug
Theta=atan2(Flux_Act_Bt,Flux_Act_Ap);

%update k1 value

Flux_Ap_k1=Flux_Ap;
Flux_Bt_k1=Flux_Bt;
IntQ_Tempk1=IntQ_Temp;
IntD_Tempk1=IntD_Temp;

IntQ_Satk1=IntQ_Satk;
IntQ_UnSatk1=IntQ_UnSatk;

IntD_Satk1=IntD_Satk;
IntD_UnSatk1=IntD_UnSatk;

Theta_est=Theta;
end
```

B.8 Advanced FieldWeakening

This is the “AdvancedFW” block mainly studied in Chapter 4. The output ports are feedback into the “SpeedRegulator” block for enhanced field weakening

Appendix C Clarke and Park Transforms

The transform of motor vectors (here we use current to illustrate) from a-b-c frame to $\alpha-\beta$ reference frame is called Clarke Transform given by:

$$\begin{bmatrix} i_\alpha \\ i_\beta \\ i_0 \end{bmatrix} = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix}$$

where i_0 is always 0 in a balanced three-phase system. The coefficient $2/3$ is to make energy conservation. Conversely, the inverse Clarke transformation is defined as

$$\begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} = \begin{bmatrix} 1 & 0 & 1 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} & 1 \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} & 1 \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \\ i_0 \end{bmatrix}$$

If expressed in non-matrix form, and drop the zero sequence component, the Clarke transform can be expressed as

$$\begin{cases} i_\alpha = i_a \\ i_\beta = (2i_b + i_a) / \sqrt{3} \end{cases}$$

And the inverse transform is:

$$\begin{cases} i_a = i_\alpha \\ i_b = -\frac{1}{2}i_\alpha + \frac{\sqrt{3}}{2}i_\beta \\ i_c = -\frac{1}{2}i_\alpha - \frac{\sqrt{3}}{2}i_\beta \end{cases}$$

Park Transform is used to convert a vector from the static $\alpha-\beta$ reference frame to the rotating $d-q$ reference frame, given by

$$\begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} \cos \theta_e & \sin \theta_e \\ -\sin \theta_e & \cos \theta_e \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix}$$

Note that the angle θ_e is defined as the angle between the d-axis and α axis. The inverse Park transform is simply a change of sign at the sine components:

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \begin{bmatrix} \cos \theta_e & -\sin \theta_e \\ \sin \theta_e & \cos \theta_e \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix}$$

Author's Publications

Journal Papers Submitted for Possible Publication

1. H. Yang, X. Zhang, Y. Wang, and G. Foo, "A Duty Cycle based Finite-set Model Predictive Torque Control for Permanent Magnet Synchronous Motor Drives"
2. H. Yang, X. Zhang, Y. Wang "A Simplified Deadbeat Predictive Torque Control Strategy for PMSM Drives using SVM".

Conference Papers

3. H. Yang, X. Zhang, Y. Wang, and G. Foo, "A Duty Cycle based Finite-set Model Predictive Torque Control for Permanent Magnet Synchronous Motor Drives", (Accepted by and Presented on Southern Power Electronics Conference (SPEC) 2018, Singapore)

Publications During Ph.D. Period but not Covered by Thesis

4. H. Yang, Y. Wang, and F. Lim, "Design of an energy sharing dual-escalator system based on the dual active bridge converter," in *Electric Machines & Drives Conference (IEMDC), 2013 IEEE International*, 2013, pp. 1424-1429: IEEE.
5. H. Yang, P. Wu, and Y. Wang, "Design Issues for the active clamp forward converter in micro-inverter applications," in *Power Electronics for Distributed Generation Systems (PEDG), 2015 IEEE 6th International Symposium on*, 2015, pp. 1-4: IEEE.
6. L. Teng, H. Yang, and Y. Wang, "Model reference tracking control of linear motor with dead-zone via switched systems subjected to time-varying delay," In 42nd Annual Conference of the IEEE Industrial Electronics Society, Florence, Italy, 2016.

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