A modified hysteresis-based DTC strategy for synchronous reluctance machines in high speed range

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**A MODIFIED HYSTERESIS-BASED
DTC STRATEGY
FOR
SYNCHRONOUS RELUCTANCE
MACHINES IN HIGH SPEED RANGE**

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Summary

Recently, synchronous reluctance machines (SynRMs) have gained popularity in certain variable-speed applications. In terms of torque and efficiency these machines demonstrate similar properties to induction machines. However, they have lower rotor losses than induction machines. Because of its ability to regulate flux levels, “spinning” core losses can be eliminated. This makes them preferable to permanent-magnet machines in scenarios where machines rotate without load torque. Furthermore, they are able to be applied to extremely high speeds. A flywheel energy storage system is a typical application which utilizes above-mentioned advantages.

During high speed range, torque capability is of major concern. For classical hysteresis-based DTC, torque capability is reduced due to the fact that stator flux is weakened. In other words, the constant torque range is limited. In this thesis, a modified hysteresis-based direct torque control is proposed to extend the constant torque region. Consequently, torque capability is enhanced in high speed range and hence speed response time is lessened. The proposed method is based on the classical hysteresis-based DTC structure. Like the classical hysteresis-based DTC, the proposed method is implemented based on the measured currents and machine parameters.

The modified hysteresis-based DTC is established by combining the classical hysteresis-based DTC with a flux error status modifying block. The block features a simple structure and an easily-executed algorithm, which renders it possible to fulfill even in a logic circuit. The flux error status block adjusts the flux error status produced by the flux hysteresis controller. After this, the new flux error status is fed into the switching table to produce the optimum voltage vector. It is worth mentioning that the modified flux error status is determined by flux position in the stationary reference frame $\alpha\text{-}\beta$, which is fixed in space. Also, the moment to activate this block is a major factor to consider. Since the objective of this incorporated block is to increase torque capability in dynamic conditions, this flux error status modifying block should be deactivated under steady states. In this thesis, the torque error is analyzed continuously to decide whether the incorporated block should be triggered or not. A

simulation is carried out by SIMULINK/MATLAB to confirm the suggested theory. A further comparison is made between the classical and the proposed hysteresis-based DTC methods. However, the modified DTC has relatively large torque ripples. Therefore, the torque hysteresis controller is replaced with a PI controller to address intrinsic disadvantages of hysteresis-based DTC.

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List of Symbols

v_a, v_b, v_c	stator phase voltages, V
v_d, v_q	d and q axis stator voltages, V
v_α, v_β	α and β axis stator voltages, V
v_x, v_y	x, y axis voltage, V
v_s	stator voltage, V
$V_1, V_2, \dots, V_8$	stator voltage vectors
V_{smax}	maximum stator voltage, V
i_a, i_b, i_c	stator phase currents, A
i_d, i_q	d and q axis stator currents, A
i_α, i_β	α and β axis stator currents, A
i_x, i_y	x, y axis stator currents, A
i_s	stator current, A
I_{smax}	maximum stator current, A
$\varphi_\alpha, \varphi_\beta$	α and β axis stator flux, Wb
φ_d, φ_q	d and q axis stator flux, Wb
φ_x, φ_y	x and y axis stator flux, Wb

φ	actual stator flux vector, Wb
φ_s^*	reference stator flux vector, Wb
φ_r	rotor flux, Wb
φ_{rx}	x axis rotor flux, Wb
φ_{ry}	y axis rotor flux, Wb
$\Delta\varphi_b$	bandwidth of the stator flux controller, Wb
T_e^*	reference torque, Nm
$\Delta\tau_b$	bandwidth of the torque controller, Nm
T	actual electromagnetic torque, Nm
T_r	reactive torque, Nm
T_{e_vlim}	torque limit imposed by the voltage limit, Nm
T_{e_clim}	torque limit imposed by the current limit, Nm
P	number of pole pairs
R_s	stator resistance, Ω
δ	torque angle, rad
θ_s	stator flux position, rad
$\theta_{s,0}$	initial stator flux position, rad
θ_r	rotor position, rad

$\theta_{r,0}$	initial rotor flux position, rad
γ	the angle between stator current and d axis
L_d, L_q	d, q axis inductance, H
S_a, S_b, S_c	states of phase a, b, c switches
ω_s	electrical speed of the stator flux, rad/s
ω_{re}	electrical speed of the rotor flux, rad/s
ω_r	mechanical rotor speed, rpm
τ	output of the torque controller
ψ^+	output of the stator flux controller
ψ^-	output of the modified stator flux error status block
I, II,..., VI	region number
k	k^{th} switching time interval
CNT_φ	counter for stator flux error
CNT_T	counter for torque error
N_{ϕ_lim}	counter limit for stator flux error
N_T_lim	counter limit for torque error
i	i^{th} voltage vector
C _{upper}	Upper triangular carrier

C_{lower}	Lower triangular carrier
T_{COM}	Output of PI controller
T_{CU}, T_{CL}	Results of comparison of T_{COM} and two carriers

List of Abbreviations

SynRM	synchronous reluctance machine
DTC	direct torque control
AC	alternating current
THD	total harmonic distortion
MTPA	maximum torque per ampere
MTPF	maximum torque per flux
SVM	space vector modulation
DC	direct current
IPM	Interior permanent magnet machine
PMASynRM	Permanent magnet assisted synchronous reluctance machine
CSF	Constant switching frequency

Chapter 1 Introduction

1.1 Background

Over the past few years, synchronous reluctance machines have been competitive alternatives to induction machine in certain variable-speed applications. Similar to induction machines, these machines exhibit torque and efficiency and properties [1]. However they have lower rotor losses [2-5]. “Spinning” core losses can be eliminated due to its ability to regulate flux levels. This makes them preferable to permanent-magnet machines in scenarios where machines rotate without torque. Furthermore, they are applied to extremely high speeds. A flywheel energy storage system is a suitable application which utilizes above-mentioned advantages.

Direct torque control (DTC) becomes intensively attractive because of its simplicity, fast response and parameter robustness. It is applicable to various types of AC machines, including induction and synchronous machines. It is known that DTC achieves fast torque response without need for rotating coordinate transform, pulse-width-modulation and accurate rotor position. A substantial amount of papers have investigated the performance of DTC in AC machines [6-9]. Both advantages and disadvantages are analyzed. Because of SynRM advantages mentioned above, the DTC applied to SynRM is proposed by many researchers [10] [11]. However, both [10] and [11] introduce SVM producing the maximum possible voltage vector reference and simultaneously regulating stator flux and torque. This imposes a burden of computation of reference voltage for the control strategy. As a result, classical hysteresis-based DTC control strategy is adopted owing to its simplicity.

Another issue for DTC controlled SynRM is regulating flux reference during different speeds. This leads to different control strategies for machines. On one hand, for electric drives below base speed, voltage and current limits are not reached. Optimal system efficiency can be achieved by maximum torque per

ampere (MTPA) when copper loss is dominant power loss. A few examples of efficiency –optimization solution are proposed for the DTC strategy. In [12] S. Bolognani illustrates an online procedure of the automatic search for the MTPA operating point for synchronous reluctance machines. But a random pattern is injected into the stator flux reference signal to retrieve information of the MTPA point from the sampled current magnitude, imposing a huge burden of computation. R. E. Betz carries out an analysis of ideal SynRMs and also gives detailed consideration to saturation and iron-loss aspects [13]. For simplicity, saturation and iron-loss are ignored in this thesis and hence the flux reference in low-speed operating region is constant. On the other hand, at constant power region, the stator flux reference is selected from reference torque such that the minimum current produces the maximum torque (MTPF) when voltage and current reach their respective limits [14]. J. Faiz derives a condition of MTPF for interior permanent magnet machines, and with minor revision, it is directly applicable to SynRM. Since torque is closely associated with flux reference, torque capability should be derived under voltage and current limits.

During high speed range, the main focus is torque capability. In general, SVM is usually incorporated to fully utilize the inverter voltage into flux weakening mode, resulting in a higher torque capability in the flux weakening region. In normal case, the mode of flux weakening is decided by the voltage reference [15]. In other words, if the reference exceeds voltage vector limits enclosed by the hexagonal boundary, the inverter would work in flux weakening mode [16]-[18]. In [16], [17] and [18], many researchers propose a switching strategy approximating voltage reference with vector that lies on the hexagonal boundary. The only difference between them is the definition of approximate voltage vectors. Instead of switching two active voltage vectors in a sample period in aforementioned papers, A. Tripathi in [19] switch only one voltage vector, transforming stator flux locus from a circle to a hexagon shape. This is because the voltage momentarily stays at tips of hexagon. Due to the fact that the optimum one of the two active voltage vectors is switched, this method further increases torque capability. However, it can not be

applied to hysteresis-based DTC because there is no reference voltage available. And SVM imposes a burden of computation. Moreover, it is not directly applicable to SynRM.

As a consequence, a modified flux weakening strategy for hysteresis-based DTC controlled SynRM is proposed in this thesis. The simulation result is included to validate the proposed method. Also an improvement is added to address the intrinsic disadvantage of hysteresis-based DTC.

1.2 Motivation

As mentioned above, synchronous reluctance machines are usually applied to occasions where high speed range is required. As such, it is desirable to accelerate torque response and thus reduce speed response time. One solution is to generate a voltage reference that moves along hexagonal boundary. But it is not applied to the hysteresis-based structure. Another solution available for hysteresis structure is introduced in [20]. However, the presented method still has a huge burden of computation due to its instantaneously varying rotor flux position. Also Auzani Bin Jidin offers a better solution [21]. He suggests an approximation of the rotor flux position to the stator flux position. The flux weakening scheme produces a six-step voltage in torque acceleration, which reduces response time to a large extent. At the same time it intensively reduces complexity of switching an optimized voltage. Nevertheless, this method appropriate for an induction machine is not directly applicable to a synchronous machine.

In consequence, this thesis is focused on proposing a modified hysteresis-based DTC strategy applicable to a synchronous reluctance machine, aimed at reducing speed response time and meanwhile making system stable and robust. Moreover, an improvement is included to deal with intrinsic disadvantage of hysteresis-based DTC.

1.3 Scope of Work

The outline of this thesis is as follows.

In the first chapter, literature review is done to explain the motivation and objective of this thesis.

In Chapter 2, mathematical model of SynRM drive and reference frame and their transformation commonly used in a SynRM are investigated respectively in two sections.

After discussion about mathematical model of SynRM, Chapter 3 is focused on classical hysteresis DTC control, which the modified hysteresis DTC method is based on. The first section discusses the voltage source inverter and synchronous reluctance machine. The inverter states are represented by eight possible voltage vectors which can be categorized into two types: active and zero voltage vectors. In the second section, hysteresis controllers and switching table are introduced. The outputs of the hysteresis controllers are fed into the switching table, which is used to generate the appropriate voltage vector. The third section contributes to the analysis of how the amplitude and direction of the stator flux are controlled by selecting the appropriate stator voltage vector. In the fourth section, control techniques aimed for higher efficiency are introduced under different speed range. Also necessary derivation of torque capability imposed by voltage and current limits is demonstrated.

Chapter 4 describes a flux error status modifying block that increases torque capability in high speed range. Simulation is conducted to prove effectiveness of modified DTC method and comparison is drawn with the classical method.

Chapter 5 presents CSF-based DTC principle, which is aimed at reducing torque ripples faced with modified DTC in Chapter 4. The simple structure is retained while speed response time is faster compared with the classical method.

The last chapter constitutes conclusion and the future work to improve the performance of the proposed methods.

Chapter 2 Reference Frame and SynRM Mathematical Model

The main focus of this thesis is to determine an effective and proper control strategy for synchronous reluctance machines. A brief introduction to SynRM mathematical model along with the concept of reference frame and its transformation used in high-performance SynRM drive is presented in this chapter.

2.1 Reference Frame and Coordinates Transformation

Although synchronous reluctance machines might have different rotor construction structures, their mathematical models are the same. For better illustration of behavior of synchronous reluctance machine, different reference frames are discussed at first as shown in Figure 2.1.

For this thesis, four reference frames are incorporated and classified into two types based on whether they are rotating or not.

- Rotating coordinates: stator flux reference frame x - y ; rotor flux reference frame d - q ;
- Fixed coordinates: stator reference frame abc ; stationary reference frame α - β ;

It is noted that x - y and d - q coordinates rotate synchronously with stator and rotor flux respectively. And stator and rotor flux are fixed on x and d axis individually. Particularly at steady states, x -axis has a constant leading displacement δ with d -axis, which is described as torque angle. As for fixed coordinates, a -axis is aligned with α in space. Also axis a , b , and c have a 120 degrees displacement with each other.

Figure 2.1 Different reference frames

In machine drives, coordinates transformation is required for variables to satisfy different needs of a synchronous reluctance machine's parameter determination. Here two coordinates transformation are introduced.

2.1.1 Park's Transformation

Park's transformation, also called direct-quadrature-zero transformation, is a mathematical transformation that projects variables of three phase fixed stator reference frame to a rotating reference frame. In this way, the three phase sinusoidal variables are reduced into two DC variables. This simplifies calculations for the control of inverters and analysis of three phase synchronous machines.

Knowing the rotor position θ_r , measured by an absolute encoder, the three line variables (stator voltages, stator currents) can be transformed into rotor flux reference frame d-q, given by the equation as follow:

$$\begin{bmatrix} x_d \\ x_q \\ x_0 \end{bmatrix} = \frac{2}{3} \begin{bmatrix} \cos \theta_r & \cos(\theta_r - 2\pi/3) & \cos(\theta_r + 2\pi/3) \\ -\sin \theta_r & -\sin(\theta_r - 2\pi/3) & -\sin(\theta_r + 2\pi/3) \\ 1/2 & 1/2 & 1/2 \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \quad (1)$$

where x_d , x_q , x_0 are voltage, current, flux linkage in dq axis respectively, and the same as for x_a , x_b , x_c .

In a particular case where x_a , x_b , x_c are balanced, x_0 is zero.

Conversely, the inverse park's transformation is:

$$\begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} = \begin{bmatrix} \cos(\theta_r) & -\sin(\theta_r) & 1 \\ \cos(\theta_r - 2\pi/3) & -\sin(\theta_r - 2\pi/3) & 1 \\ \cos(\theta_r + 2\pi/3) & -\sin(\theta_r + 2\pi/3) & 1 \end{bmatrix} \begin{bmatrix} x_d \\ x_q \\ x_0 \end{bmatrix} \quad (2)$$

Since the main variables to control in field oriented control (FOC) are i_d and i_q , park's transformation is indispensable for FOC. But it is not essential for direct torque control, because direct torque control can be implemented both in rotor reference frame d-q and stationary reference frame α - β . Because of abovementioned requirements, accurate rotor position is demanded for field oriented control. Otherwise the control system would be unstable because of subsequent estimated parameters. This is one of main advantages of DTC in comparison to FOC. Moreover, owing to the fact that accurate rotor position is not needed for DTC, sensorless control becomes possible. Since sensorless control is not the key point in this thesis, details of sensorless control would not be explained in detail.

2.1.2 Clarke Transformation

Clarke transformation, also known as alpha-beta transformation, is a mathematical transformation employed to simplify the control of inverters by projecting three phase variables to stationary reference frame α - β . One application is the generation of reference signal used for space vector modulation. Another one is to calculate torque reference and stator flux without the need for actual rotor position.

The Clarke transformation is given by

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

Reversely, the inverter Clarke transformation is:

$$\begin{bmatrix} x_\alpha \\ x_\beta \\ x_\gamma \end{bmatrix} = \begin{bmatrix} 1 & 0 & 1 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} & 1 \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} & 1 \end{bmatrix} \begin{bmatrix} x_a \\ x_b \\ x_c \end{bmatrix} \quad (4)$$

Similarly, when $x_a + x_b + x_c = 0$, x_γ equals zero

2.2 Mathematical Model of SynRM

Synchronous reluctance machines have emerged as early as induction machines. One of the main benefits of synchronous reluctance machines is lower rotor losses while maintaining high torque capability as induction machines. Furthermore, the stator of a synchronous reluctance machine is built from a cylinder in identical manner as induction machines and only the rotor features salient poles. In other words, the stator of both can be constructed in the same assemble line, which would definitely reduce manufacturing cost. Also the properly distributed three phase windings produce sinusoidal waves at the air gap.

Synchronous reluctance machines can be divided into two types with regard to power supplies line start SynRM and inverter fed machines, as depicted in Figure 2.2. In retrospect, studies of line-start synchronous reluctance machines are quite aged [22] [23]. They are designed for directly powered from

utility. Thus, the instantaneous current can lead to the deterioration of machine performance as well as a huge impact on the power grids.

Nowadays, a more common type of synchronous reluctance machine is inverter fed. A SynRM with a cage can operate in open loop but their performance is rather poor with that of the cageless type because the starting cage windings lead to torque ripples and even instability during operation. Later, owing to the rapid development of power electronic devices, converters and control strategies, extensive efforts are devoted to the design of such machines and optimization of their configuration.

Figure 2.2 Classification of synchronous reluctance machines

Main attention is attached to the improvement of rotor structures of cageless synchronous reluctance machines And SynRM can be classified into four categories based on their rotor configurations: rotor

punching laminated SynRM, discrete-rotor-pole, axially laminated rotor SynRM and permanent magnet assisted SynRM.

Figure 2.3 Rotor punching laminated SynRM

Figure 2.4 Discrete-rotor-pole SynRM

Figure 2.5 Axially-laminated-rotor SynRM

A SynRM employs a laminated punching to create salient effects as shown in Figure 2.3[28]. The saliency ratio of this type of rotor is only two or three because of the need for rotor bridges. Advancement is made in [29] as shown in Figure 2.4. Owing to the discrete poles mounted on a non magnetic shaft, the salient ratio could be capable of four to five. Also a comparison of frame with equivalent induction machines is drawn in this thesis. A larger salient ratio is achieved in axially laminated rotor SynRMs as shown in Figure 2.5[30]. Paths of minimum reluctance in the direction of the laminations and the maximum reluctance in the path normal to the laminations can be achieved when the magnetic laminations are bent. For axially laminated rotor SynRMs, saliency ratio could reach seven to eight. The main advantage of high saliency ratio is to essentially eliminate relatively low power factor and efficiency compared with induction machines with identical stators. In addition, a wide constant power region is attained. These days, interest in another type of SynRM has been booming [31] [32] [33]. When a proper quantity of permanent magnets is added, the power factor and torque capability are hugely increased as shown in Figure 2.6. It possesses similar properties to an interior permanent magnet machine (IPM) so that it is called permanent magnet assisted synchronous reluctance machine (PMASynRM). But, what makes it competitive to traditional IPM is that the amount of permanent magnets utilized and its

counterpart flux linkages are smaller. Moreover, the size of inverter is diminished when power factor is improved [34].

Figure 2.6 Permanent magnet assisted SynRM

As can be seen in Figure 2.1, there exist various types of reference frame. As a result, the expression of a synchronous reluctance machine can be established in different reference frames.

With the following assumptions, the behavior of a synchronous reluctance machine is shown as Figure 2.7 below.

- Magnetic saturation and core losses are omitted

Figure 2.7 Synchronous reluctance machines in rotor flux reference frame

In the rotor flux reference frame, equations of SynRMs' behavior are shown below (5)-(9),

$$\begin{aligned} v_d &= L_d \cdot \frac{di_d}{dt} + i_d \cdot R_s - \omega_{re} L_q i_q \\ v_q &= L_q \cdot \frac{di_q}{dt} + i_q \cdot R_s + \omega_{re} L_d i_d \end{aligned} \quad (5)$$

$$\begin{aligned} \varphi_d &= L_d i_d \\ \varphi_q &= L_q i_q \end{aligned} \quad (6)$$

$$\varphi_s = \sqrt{\varphi_d^2 + \varphi_q^2} = \sqrt{(L_d i_d)^2 + (L_q i_q)^2} \quad (7)$$

$$T_e = \frac{3}{2} P (\varphi_d i_q - \varphi_q i_d) = \frac{3}{2} P (L_d - L_q) i_d i_q \quad (8)$$

$$\delta = \tan^{-1} \frac{\varphi_d}{\varphi_q} \quad (9)$$

where the subscript d, q represents d -axis and q -axis components of variables, ω_{re} is electrical speed of rotor flux, L_d and L_q are the d and q axes stator inductance. φ_s, T_e are stator flux and electromagnetic torque respectively. δ is defined as torque angle between the stator flux φ_s and rotor flux φ_r .

The voltage of a SynRM in the rotor flux d - q reference frame is shown in (5). The components and amplitude of stator flux linkage are given in (6) and (7) respectively. The electromagnetic torque is given in (8). The angle between stator flux linkage and rotor flux linkage is given in (9).

For stationary reference frame α - β , the electromagnetic torque and voltage can be expressed in another way (10)-(12).

$$v_\alpha = i_\alpha \cdot R_s + \frac{d\varphi_\alpha}{dt} \quad (10)$$

$$v_\beta = i_\beta \cdot R_s + \frac{d\varphi_\beta}{dt} \quad (11)$$

$$T_e = \frac{3}{2} P (\varphi_\alpha i_\beta - \varphi_\beta i_\alpha) \quad (12)$$

where the subscript α, β represent α -axis and β -axis components of variables.

These equations are very useful especially for sensorless control. It is because they do not introduce park transformation. However, they will not be used in this thesis.

To better illustrate maximum torque capability, the torque expression in the stator flux x - y reference frame should be developed. The components of voltage, stator flux and rotor flux of SYNRM in x - y reference frame are given in (13)-(16).

$$\begin{aligned} v_x &= R_s \cdot i_x + \frac{d\varphi_x}{dt} \\ v_y &= R_s \cdot i_y + \omega_s \varphi_s \end{aligned} \quad (13)$$

$$\begin{aligned} \varphi_x &= \varphi_s = L_q i_x + \varphi_{rx} \\ \varphi_y &= 0 = L_q i_y + \varphi_{ry} \end{aligned} \quad (14)$$

$$\begin{aligned} \varphi_{rx} &= \varphi_r \cos \delta \\ \varphi_{ry} &= -\varphi_r \sin \delta \end{aligned} \quad (15)$$

$$\varphi_r = (L_d - L_q) \cdot i_d \quad (16)$$

where the subscript x, y represent α -axis and β -axis components of variables.

In the x-y stator flux reference frame, stator flux linkage is aligned with x-axis. Consequently, the torque expression can be derived by the product of φ_s and i_y as shown in (17). Substituting (16) into (15), i_y can be expressed with respect to δ as shown in (18). According to (17) and (18), T_e can be obtained in terms of δ as displayed in (19). According to (9), $\sin \delta$ is replaced by i_d, i_q, L_d, L_q and (20) is obtained. A further simplification of (19) based on (20) is given in (21). By replacing i_d, i_q with (22) in (21), equation (23) is attained. Equation (21) and (23) are convenient to derive maximum torque capabilities under both low-speed and flux-weakening regions. The details of derivation will be discussed in Chapter 3.

$$T_e = \frac{3}{2} P \varphi_s i_y \quad (17)$$

$$i_y = \frac{(L_d - L_q) \cdot i_d \sin \delta}{L_q} \quad (18)$$

$$T_e = \frac{3P}{2} \frac{(L_d - L_q)}{L_q} \varphi_s i_d \sin \delta = \frac{3P \varphi_s \varphi_r \sin \delta}{2L_q} \quad (19)$$

$$\sin \delta = \frac{L_q i_q}{\sqrt{(L_d i_d)^2 + (L_q i_q)^2}} \quad (20)$$

$$T_e = \frac{3P}{2} \frac{(L_d - L_q)}{\sqrt{(L_d i_d)^2 + (L_q i_q)^2}} \varphi_s i_d i_q = \frac{3P}{2} (L_d - L_q) i_d i_q \quad (21)$$

$$i_d = i_s \cos \gamma, \quad i_q = i_s \sin \gamma \quad (22)$$

$$T_e = \frac{3P(L_d - L_q)}{2} i_s^2 \sin \gamma \cos \gamma = \frac{3P(L_d - L_q)}{4} i_s^2 \sin 2\gamma \quad (23)$$

2.3 Conclusion

This chapter constitutes introduction of various types of SynRMs, reference frames and their transformation. Mathematical models of a SynRM in diverse reference frame are built for subsequent analysis of DTC strategy in the following chapters.

Chapter 3 Classical Direct Torque Control

Direct torque control, initially proposed by I. Takahashi [6] and by M. Depenbrock [35] has gained its popularity because of its simplicity, good performance and robustness. Unlike FOC, DTC principle is not based on reproducing the electromagnetic behavior of a dc motor but is aimed at a complete exploitation of the flux and torque-producing capabilities of a SynRM fed by a voltage source inverter (VSI). DTC features no current regulators and PWM signal generators, where the switching state selection is dependent on a switching table. In other words, DTC reduces the dependence on parameters of synchronous reluctance machines and improves the dynamic of flux and torque response.

The DTC drive constitutes five parts: torque and flux comparators, switching table, voltage source inverters, torque and flux estimators, and SynRM. The diagram is given in the following Figure 3.1.

Figure 3.1 Basic modules for classical DTC

The reference torque T_e^* is produced by a PI controller whose input is the error between the speed reference and the actual speed. And the reference stator flux ϕ_s^* is generated by MTPA and constrained by

voltage limits. The actual torque and flux are estimated by equation (7) and (8). Therefore, the torque and flux estimator will not be discussed in the following subchapter. Then the torque and flux errors are fed into the torque and flux hysteresis controllers and hence the output of hysteresis controllers are Ψ^+ and τ , which is used to select the opportune voltage vectors. At last the chosen voltage vector is fed into a SynRM.

3.1 Voltage Source Inverter and SynRM

For a three-phase machine with sinusoidally balanced distribution as shown in Figure 3.2, the expression for the stator voltage v_s is given in equation (24) when the phase a winding correlates with the α axis.

$$v_s = \frac{2}{3}(v_a + v_b e^{j\frac{2}{3}\pi} + v_c e^{j\frac{4}{3}\pi}) \quad (24)$$

Figure 3.2 An inverter fed SynRM

where v_a , v_b , and v_c are the instantaneous values of line-to-neutral voltages. The phase voltages v_a , v_b , and v_c are dictated by the states of bi-directional switches S_a , S_b and S_c when a SynRM is fed by an inverter. v_a can be either V_{dc} if S_a is one or 0 if S_a equals 0. And so it is for S_b and S_c as shown Figure 3.3.

Figure 3.3 Inverter switch states

Therefore there are eight possible voltage vectors: $V_1(100)$, $V_2(110)$,... $V_8(111)$, among which $V_1(100)$, $V_2(110)$,... $V_6(101)$, are active voltage vectors and $V_7(000)$, and $V_8(111)$ are zero voltage vectors that lie on the center of the hexagon. Each active voltage vector is displaced by 60 degrees as demonstrated in Figure 3.4.

Figure 3.4 Eight possible voltage vectors

And these eight voltage vectors can be indicated in (25).

$$v_s(S_a, S_b, S_c) = \frac{2}{3} V_{dc} (S_a + S_b e^{j\frac{2}{3}\pi} + S_c e^{j\frac{4}{3}\pi}) \tag{25}$$

It is easily observed that the length of voltage vector is $\frac{2}{3} V_{dc}$ and table of space vectors and switching states is obtained as follows.

Table 3.1 Space Vectors and Switching States

Voltage Vector	Switching States(Three Phases)	Vector Definition
----------------	--------------------------------	-------------------

V_1	100	$\vec{V}_1 = \frac{2}{3}V_{dc}$
V_2	110	$\vec{V}_2 = \frac{2}{3}V_{dc}e^{j\frac{\pi}{3}}$
V_3	010	$\vec{V}_3 = \frac{2}{3}V_{dc}e^{j\frac{2\pi}{3}}$
V_4	011	$\vec{V}_4 = \frac{2}{3}V_{dc}e^{j\pi}$
V_5	001	$\vec{V}_5 = \frac{2}{3}V_{dc}e^{j\frac{4\pi}{3}}$
V_6	101	$\vec{V}_6 = \frac{2}{3}V_{dc}e^{j\frac{5\pi}{3}}$
V_7	000	$\vec{V}_7 = 0$
V_8	111	$\vec{V}_8 = 0$

3.2 Hysteresis Controllers and Switching Table

To choose the voltage vector for torque and stator flux control, two hysteresis controllers are utilized as shown in Figure 3.5. The hysteresis band designed for stator flux is asymmetrical to the vertical axis since the amplitude of stator flux can not be negative. In this case, the actual stator flux would be $\Delta\phi_{BW}$ larger than the reference at most or smaller than the reference. As for stator flux controller, the hysteresis band for torque controller is symmetrical since torque can be either positive or negative. Under this circumstance, the maximum error between the actual and the reference is $\pm\Delta\tau_{BW}/2$. $\Delta\phi_{BW}$, $\Delta\tau_{BW}$ are the bandwidth of the flux and torque hysteresis controllers respectively.

Figure 3.5 Hysteresis controllers for torque and stator flux

where $\Delta\tau = T_e^* - T$, and $\Delta\varphi = \varphi_s^* - \varphi$. In other words, the torque and flux error equals the subtraction of the actual from the reference. τ and ψ^+ are the outputs of the flux and torque hysteresis controllers respectively.

When $\Delta\tau > \Delta\tau_{BW}/2$, $\tau = 1$ and the torque has to be increased. When $\Delta\tau < -\Delta\tau_{BW}/2$, $\tau = 0$ and the torque has to be decreased. If $-\Delta\tau_{BW}/2 \leq \Delta\tau \leq \Delta\tau_{BW}/2$, τ remains the previous value. It still holds true for ψ except that $\Delta\tau_{BW}/2$ is replaced by $\Delta\varphi_{BW}$ and 0 substitutes $-\Delta\tau_{BW}/2$.

Table 3.2. Switching Logic of Classical Two-level Hysteresis DTC

<i>Flux error status</i>	<i>Torque Error status</i> / <i>Sector</i>	I	II	III	IV	V	VI
$\psi^+ = 1$	$\tau = 1$	V_2	V_3	V_4	V_5	V_6	V_1
$\psi^+ = 1$	$\tau = 0$	V_6	V_1	V_2	V_3	V_4	V_5
$\psi^+ = 0$	$\tau = 1$	V_3	V_4	V_5	V_6	V_1	V_2
$\psi^+ = 0$	$\tau = 0$	V_5	V_6	V_1	V_2	V_3	V_4

Based on the information of flux error status, torque error status, sector, the six-vector switching table is given in Table 3.2. It is observed that stator flux rotates in the counterclockwise direction when $\tau = 1$ and in the clockwise direction when $\tau = 0$. As for amplitude of stator flux, it is enlarged when $\psi^+ = 1$, and

reduced when $\psi^+ = 0$. Consequently, the torque is controlled by controlling the rotational direction of the stator flux, provided that its amplitude is kept constant.

Since choice of the voltage vector is dependent on the position of the stator flux, it is worthwhile to investigate the determination of the sector number of stator flux, which is implemented in the stationary reference frame α - β . As seen from Table 3.3, the criteria to decide sector number are as follows.

Table 3.3 Determination of sector number

Sector	Criteria for sector number
I	If $\varphi_\alpha > 0$ and $-\varphi_s/2 < \varphi_\beta < \varphi_s/2$
II	If $\varphi_\alpha > 0$ and $\varphi_s/2 < \varphi_\beta < \varphi_s$
III	If $\varphi_\alpha < 0$ and $\varphi_s/2 < \varphi_\beta < \varphi_s$
IV	If $\varphi_\alpha < 0$ and $-\varphi_s/2 < \varphi_\beta < \varphi_s/2$
V	If $\varphi_\alpha < 0$ and $-\varphi_s < \varphi_\beta < -\varphi_s/2$
VI	If $\varphi_\alpha > 0$ and $-\varphi_s < \varphi_\beta < -\varphi_s/2$

For instance, when $\varphi_\alpha > 0$ and $-\varphi_s/2 < \varphi_\beta < \varphi_s/2$, the stator flux is located in the sector I. Consequently, the possible voltage vectors might be V2, V3, V5, V6. A further selection of the voltage vector requires the output of the hysteresis controllers. If $\psi^+ = 1$ and $\tau = 1$, V2 is chosen then.

3.3 Control of Amplitude and Rotation Direction of the Stator Flux

Figure 3.6 Stator voltage vectors

The stator flux can be stipulated as:

$$\varphi_s = \int (v_s - R_s i_s) dt \quad (26)$$

During the switching interval, the voltage applied to a SynRM is constant and equation (26) can be rearranged as (27) in the discrete time domain.

$$\varphi_s(k) = v_s \Delta t - R_s i_s \Delta t + \varphi_s(k-1) \quad (27)$$

If the stator resistance is ignored, equation (27) can be rewritten as in equation (28). It implies that the tip of the stator flux linkage, namely, φ_s , will move in the direction of the applied voltage vector as depicted in Figure 3.6.

$$\varphi_s(k) = v_s \Delta t + \varphi_s(k-1) \quad (28)$$

To select the voltage vector controlling the amplitude of the stator flux, the voltage vector region is divided into six regions: I—VI. The distance of the tip of the applied voltage vector will move is dominated by duration of a time interval of which the stator voltage is applied. In each region, two adjacent voltage vectors are switched alternatively to keep the switching frequency minimum. Furthermore, alternatively switching two adjacent voltage vectors will help maintaining the amplitude of stator flux within an error band because one of the two adjacent is responsible for increasing the amplitude of stator flux and the other accounts for decreasing. For example, voltage vector V_2 cause the amplitude of stator flux to increase while voltage vector V_3 results in amplitude reduction provided that the stator flux φ_s is located at region I and rotates in the counterclockwise direction. In this case, the bandwidth of amplitude of stator flux error as shown in Figure 3.7, governs the performance of the control strategy.

If the stator flux revolves reversely, the opposite voltage vector should be chosen. For instance, when φ_s is in region I, voltage vectors V_2 and V_3 are supposed to be replaced by V_5 , V_6 respectively.

Figure 3.7 The control of the amplitude of stator flux linkage

As for zero voltage vectors, although they are not used in this thesis, it is still necessary to discuss their effects on the position of stator flux. When zero voltage vector is applied, the stator flux stays at the original location.

Figure 3.8 stator flux trajectory in MATLAB

In Figure 3.8, the speed of a SynRM is increased from zero to 400 rpm and the initial stator flux is zero. During the acceleration of a SynRM, the stator flux is increased from zero to a constant value and kept constant at its rated value. Since the hysteresis band of the flux hysteresis controller is quite small, the trajectory of the stator flux is circular in reality.

3.4 Control Techniques

The basic principle of direct torque control for a SynRM is to control two main variables, namely the electromagnetic torque and stator flux. It features simple algorithm, easy digital implementation and robust operation. A block diagram in detail is given in Figure 3.9

Figure 3.9 The block diagram of classical hysteresis-based DTC controlled SynRM

As given in Figure 3.9, the rotor position is detected with an absolute encoder and the electrical rotor speed is calculated by a derivative block. Through the torque and flux observer, it is easy to obtain the actual torque, stator flux as well as its angle in the d-q reference frame. The actual torque T and stator flux φ are attained from equation (1), (6) and (8). As for the torque reference T_e^* , it is resulted from a PI controller and constrained by a torque limiter, which takes voltage and current limits into consideration. The stator flux reference φ_s^* is the minimum value computed from the flux weakening and MTPA blocks. The reference torque and stator flux will be investigated in the following section.

In order to achieve high efficiency during high speed range, various control techniques are presented before, such as $i_d = 0$ control, maximum torque per ampere control, unity power factor control, field weakening control. For $i_d = 0$ control, the dc-link voltage is not fully utilized and hence the efficiency of the drive is rather low [36]. On the other hand, the torque per ampere is quite low with unity power factor control. Consequently, in this thesis, only MTPA and field weakening control are considered [37].

3.4.1 Maximum Torque per Ampere (MTPA) Trajectory

Maximum torque per ampere control is initially proposed by R.E Betz et al to achieve highly efficient operation of SynRM below base speed [13]. To obtain fast transient speed response and high torque, the

stator current phase angle γ should be controlled to derive maximum torque. The relationship between stator current and its phase angle is attained by setting the derivative of (23) with respect to γ to zero.

Figure 3.10 Effect of current phase angle γ

Obviously, electromagnetic torque is maximum when γ equals $\pi/4$ or $3\pi/4$ as shown in Figure 3.10. This leads to equivalence of absolute value of i_d and i_q . Because the two main variables to control in DTC are torque and stator flux, it is imperative to transform the trajectory from angle-torque plane to torque-flux plane. Under this condition, the relationship between φ_s and T_e is derived as shown in equation (29).

$$|T_e| = \frac{3P(L_d - L_q)}{2(L_d^2 + L_q^2)} \varphi_s^2 \quad (29)$$

Therefore, the reference stator flux amplitude for MTPA is given in equation (30), where T_e represents the reference torque calculated from PI speed controller.

$$\varphi_s^* = \sqrt{\frac{2}{3P} \frac{(L_d^2 + L_q^2)}{(L_d - L_q)}} |T_e| \quad (30)$$

The trajectory satisfying MTPA characteristic represented by equation (30) is drawn on flux-torque plane in Figure 3.11, where torque is horizontal axis, and torque is vertical axis. According to analysis Figure 3.11, the MTPA trajectory is independent from machine speed and merely dictated by machine parameters.

Figure 3.11 MTPA in φ_s - T_e plane

In normal cases, the amplitude of stator flux in constant torque region is kept constant to attain high torque capability.

3.4.2 Field Weakening Control

When the speed increases, the back emf, namely $\omega\varphi$, increases simultaneously and hence the stator voltage has to increase to maintain the torque or current control. But if the speed is increased further, the

back emf would touch the voltage limit imposed by both inverters and SynRMs. Moreover, the torque or current controller would saturate and eventually torque or current control would be lost. This is clearly illustrated in [38] by Zhe Zhang. By simulation, when a small speed reference is fed, the actual speed could follow the reference well. While for large speed reference, the speed controller fails. This implies that the back emf is too large that dc link can not provide enough voltage to maintain torque and flux control. Thus, stator flux is weakened for continuous control of speed. Obviously, the torque capability is largely constrained by voltage and current limits.

A. Voltage Limit

Normally, the voltage limit is expressed as follows:

$$\sqrt{v_d^2 + v_q^2} \leq V_{smax} \quad (31)$$

In the steady state, the first term in equation (5) can be omitted. Substituting (5) into (31), the voltage limit can be described as (32).

$$\sqrt{(R_s i_d - \omega_{re} L_q i_q)^2 + (R_s i_q + \omega_{re} L_d i_d)^2} \leq V_{smax} \quad (32)$$

Equation (32) can be further simplified as (33) in high speed range since the resistor term is negligible compared with the speed term. It can be concluded from (33) that the stator flux during high speed range is inversely proportional to the electrical rotor speed when the voltage limits imposed by the inverter and SynRM is constant.

$$\varphi_s^* \leq \frac{V_{smax}}{|\omega_{re}|} \quad (33)$$

B. Current Limit

Normally, current limitation equation is given in (34)

$$\sqrt{i_d^2 + i_q^2} \leq I_{smax} \quad (34)$$

The shape of the current limit is a circle centered on the origin in the i_d - i_q plane. Also the I_{smax} is constrained by both the inverter and SynRM.

C. The Resultant Torque Capability Imposed by Voltage and Current Limits

When the rotor speed is above the base speed, stator current and voltage limits are reached. In this case flux weakening (FW) should be employed and MTPA control should be modified. Consequently, by adding resultant torque limits imposed by voltage and current limits, operation at high speed range is realized.

Current limitation exerts an impact on the electromagnetic torque shown in (35). Due to the power equivalence of a synchronous reluctance machine, the relationship is given in (35), where v_s , i_s and ω_m represents stator voltage, current and mechanical rotor speed respectively. And $\omega_{re} = P\omega_m$.

$$\left(\frac{3}{2} v_s i_s\right)^2 = (T_e \omega_m)^2 + (T_r \omega_m)^2 \quad (35)$$

On one hand the left term represents apparent power in the whole system. On the other hand, the first term at right side is called active power and the latter one is defined as the reactive power which reflects power stored in energy storage elements. It is noteworthy that T_r depicted as reactive torque is estimated as in (36).

$$T_r = \frac{3P}{2} (\varphi_\alpha i_\alpha + \varphi_\beta i_\beta) \quad (36)$$

At steady states in high speed range, the maximum value of stator voltage is established as mentioned in (33). Substituting it into (35) and taking (36) into account, (37) is determined.

$$T_{e_clim} = \pm \sqrt{\left| \left(\frac{3}{2} P I_{smax} \varphi_s^* \right)^2 - (T_r)^2 \right|} \quad (37)$$

Next voltage limitation on the torque limit is examined below. When the rotor speed is further increased, the maximum torque capability with respect to stator flux linkage is employed. The aforementioned equation (8) can be transformed into (38) taking (6) and (9) into account.

$$T_e = \frac{3}{2} P \frac{(L_d - L_q)}{L_d L_q} \varphi_d \varphi_q = \frac{3}{4} P \frac{(L_d - L_q)}{L_d L_q} \varphi_s^{*2} \sin 2\delta \quad (38)$$

Obviously, when torque angle δ is $\pi/4$ or $3\pi/4$ i.e. $|\varphi_d| = |\varphi_q|$, maximum torque under voltage limitation is obtained as shown in (39).

$$T_{e_vlim} = \pm \frac{3}{4} P \frac{(L_d - L_q)}{L_d L_q} \varphi_s^{*2} \quad (39)$$

3.5 Conclusion

In this chapter, the classical hysteresis-based DTC is comprehensively examined. At first, voltage source inverter and SynRM are explored. The eight switching states are transferred into eight corresponding voltage vectors. In section 2, the principle of two hysteresis controllers are analyzed and the switching table to choose voltage vector is studied. In the third section, it can be observed that how the applied

voltage vectors shape the tip of the stator flux. In the fourth section, the control techniques used in this thesis are discussed to derive the reference torque and stator flux.

Chapter 4 A modified hysteresis-based DTC strategy for synchronous reluctance machine

4.1 Principle of the Modified Hysteresis-based DTC Strategy for SynRM

Figure 4.1 The block diagram of hysteresis-based DTC controlled SynRM with modified flux weakening

As shown in Figure 4.1, the modified hysteresis-based DTC is employed for SynRMs. Unlike the classical hysteresis-based DTC, a flux error status modification block is included. The output of the flux controller ψ^+ is converted to a new flux error status ψ^- , resultantly making a new selection of the applied voltage vector.

If a large load torque is applied, a large torque error is produced. Consequently, the torque controller produces a signal status that causes the actual torque to increase. That is to say, no zero voltage vectors are selected under this circumstance as mentioned in the last chapter. This method bears some resemblance to DTC-SVM in which the reference voltage moves along the hexagonal boundary. Therefore, no zero voltage vectors are applied. A fast torque response is attained, owing to the fact that no voltage zero voltage vectors are chosen.

Nevertheless, all abovementioned methods do not provide the fastest torque response. It is because of two possible active voltage vectors switched alternatively. Equation (40) is introduced which was derived in [20] to tell how to select the optimum voltage vector.

$$\frac{dT_e}{dt} = C\varphi_r \left[\frac{2}{3}V_{dc} \sin(\theta_i - \theta_{r,0}) - \omega\varphi_{s,0} \cos(\theta_{s,0} - \theta_{r,0}) \right] \quad (40)$$

where C is a constant dependent on the machine parameters and $\frac{2}{3}V_{dc}$ is the magnitude of the applied voltage. θ_k , $\theta_{r,0}$ and $\theta_{s,0}$ are the angles of the stator voltage vector, initial rotor flux and stator flux respectively. φ_r , and $\varphi_{s,0}$ are the amplitude of the rotor flux and initial stator flux. The second term in (40) is independent from the applied voltage vector. As such, the maximum rate of change of the torque as described in (40) can be simplified as (41)

$$\left[\frac{dT_e}{dt} \right]_{max} = \max_i \{ \sin(\theta_i - \theta_{r,0}) \} \quad (41)$$

The rotor flux angle can be considered as equivalent to the stator flux angle as what is done in [39]. In this case, equation (41) mentions that the fastest torque response can be obtained when the voltage angle is 90 degrees with regard to the stator flux angle. As such, a voltage vector that has the largest tangential component would be selected. Assume that the stator flux moves in the counterclockwise direction and its amplitude is proportional to the product of the applied voltage vector and a switching interval. Let us consider a special example in Figure 4.2. The stator flux enters the sector I at t_1 and leaves precisely at half of this sector. Within this period, the voltage vector V_2 and V_3 are switched alternatively. Consequently, the stator flux is increased and decreased alternatively as shown in Figure 4.2. It can be seen from the torque trajectory that the variation of the torque differs when different voltage vectors are applied. When voltage vector V_2 is chosen, higher torque rate is achieved in comparison to the voltage vector V_3 . Because of the selection of the other voltage vector, fastest torque response can not be attained

Figure 4.2 The output of the torque and stator flux

To select the optimum voltage vector, the flux error status is changed before it is given to the switching table. The voltage vector sector is further separated into two subsectors: subsector 1 (unshaded area) and subsector 2 (shaded area), as portrayed in Figure 4.3. And the criteria for subsector number are given in Table 4.1. Provided that the modified flux error status block is activated and stator flux, rotating in the counterclockwise direction, is situated at subsector 1, the output of the flux hysteresis controller, namely ψ^+ would be modified to a new flux error status ψ^- equal to 0, before it is fed into the switching table.

As the stator flux is located at subsector 2, the block will generate a signal equal to 1. On this condition, instead of switching two voltage vectors in a subsector, only the one that has the largest tangential voltage component would be selected.

Figure 4.3 The digital outputs of the modified flux weakening block corresponding to the optimized voltage vector in each subsector

Table 4.1 Determination of subsector number

Sector number	Subsector number	Criteria for subsector number
I	1	If $\varphi_\alpha > 0$ and $-\varphi_s/2 < \varphi_\beta < 0$
	2	If $\varphi_\alpha > 0$ and $0 < \varphi_\beta < \varphi_s/2$
II	1	If $\varphi_\alpha > 0$ and $\varphi_s/2 < \varphi_\beta < \sqrt{3}\varphi_s/2$
	2	If $\varphi_\alpha > 0$ and $\sqrt{3}\varphi_s/2 < \varphi_\beta < \varphi_s$
III	1	If $\varphi_\alpha < 0$ and $\sqrt{3}\varphi_s/2 < \varphi_\beta < \varphi_s$

	2	If $\varphi_\alpha < 0$ and $\varphi_s/2 < \varphi_\beta < \sqrt{3}\varphi_s/2$
IV	1	If $\varphi_\alpha < 0$ and $0 < \varphi_\beta < \varphi_s/2$
	2	If $\varphi_\alpha < 0$ and $-\varphi_s/2 < \varphi_\beta < 0$
V	1	If $\varphi_\alpha < 0$ and $-\sqrt{3}\varphi_s/2 < \varphi_\beta < -\varphi_s/2$
	2	If $\varphi_\alpha < 0$ and $-\sqrt{3}\varphi_s/2 < \varphi_\beta < -\varphi_s/2$
VI	1	If $\varphi_\alpha > 0$ and $-\varphi_s < \varphi_\beta < -\sqrt{3}\varphi_s/2$
	2	If $\varphi_\alpha > 0$ and $-\sqrt{3}\varphi_s/2 < \varphi_\beta < -\varphi_s/2$

Another benefit of application of one voltage vector in a subsector is fully utilizing dc link voltage and hence improving the torque capability during high speed range. Normally, the shape of the reference voltage limits is a circle enclosed by a hexagonal boundary as portrayed in Figure 4.4. But obviously dc link voltage is not fully used because there exists a blank area between the hexagonal boundary and the circular voltage reference. In consequence, to further improve the utilization of the dc link voltage, the circular reference voltage is pushed outward a bit by adding the flux error status block to select the optimum voltage vector.

From the aforementioned analysis, one voltage vector is constantly applied to an inverter in one subsector. For instance, when the stator flux enters subsector 2 of sector VI, V2 is selected which would make the flux to increase. On the condition that the stator flux is in the subsector 1 of sector I, V2 is still picked but it causes the flux to decrease this time. In other words, instead of switching two adjacent voltage vectors in a subsector, only one voltage vector is chosen, which might make the flux out of

control, or even render the whole control system collapse. As such, the moment to trigger and cease the flux error status modification block is worth a detailed discussion.

Figure 4.4 Voltage reference and limits

The moment to activate the modified flux error status block is worth paying attention to. The aim for adding the block is to increase torque capability and thus reduce speed response time. Thus, when reference torque T_e^* is suddenly changed and the torque comparator output 1 for N_{T_lim} consecutive times, the modified flux weakening is activated. However, the continuous selection of one voltage vector would definitely deteriorate the performance of stator flux, since the amplitude of stator flux is always increased. This would undoubtedly make the whole system collapse. To avoid the breakdown, restrictions are imposed. If the stator flux is out of bands for N_{ϕ_lim} consecutive times, the modified flux weakening would be terminated. N_{T_lim} is the counter number that estimated torque crosses the hysteresis boundary and N_{ϕ_lim} represents the number of times that estimated flux is out-of bounds. It is worth mentioning that the choice of N_{ϕ_lim} is independent from N_{T_lim} . But the selection of N_{ϕ_lim} has an impact on the performance of the proposed method. If N_{ϕ_lim} is chosen too large, the flux and torque loop would lose

control, because the failing control of flux would result in an erroneous calculation of the estimated torque, even a collapse for the whole system. If N_{φ_lim} is selected as a very small number, the performance of the proposed method would largely deteriorate. It is because synchronous reluctance machine would frequently return from the proposed method to the classical hysteresis DTC. Since the magnitude of the torque T_e is related to the square of the stator flux φ_s^{*2} according to equation (38), more stringent rules should be imposed on the counter number for the torque N_{T_lim} to ensure that the system is stable. Consequently, a smaller number is chosen for N_{T_lim} as compared with N_{φ_lim} . In this thesis, the parameters N_{T_lim} and N_{φ_lim} are tuned on a trial and error basis to give a satisfactory output response. N_{T_lim} is chosen as 4 and N_{φ_lim} is selected as a slightly larger number 40. Actually, the counter number for the torque N_{T_lim} reached its maximum value when the counter number for the stator flux N_{φ_lim} reached 40. For better illustration of the moment to activate the proposed method, flowchart of modification of flux status is depicted below in Figure 4.5.

Figure 4.5 Flow chart of the proposed method of modification of flux status

4.2 Simulation Results

To verify the feasibility of the proposed modified flux weakening strategy, a hysteresis-based DTC controlled SynRM drive system is established in MATLAB/SIMULINK. Parameters of the SynRM drive system is specified in Table 4.2.

Table 4.2 Parameter of the SynRM drive system

Number of pole pairs	P	2
Stator resistance	R_s	5.15Ω
d-axis inductance	L_d	0.323H
q-axis inductance	L_q	0.0595H
dc-link voltage	V_{dc}	140V
Rated phase-phase voltage	V_{rated}	380V
Rated phase current	I_{max}	3A
Inertia	J	33.5e-3kg m^2

To prove faster torque response, a step torque reference is applied from -3.5 Nm to 3.5 Nm. As shown in Figure 4.6, for the classical method, it takes 0.0043 seconds to reach the reference torque, while in contrast the proposed method spends 0.0012 seconds less time.

Figure 4.6 Torque response time

Figure 4.7 Constant torque and power region

Another simulation is done to explain why higher torque capability is achieved with the proposed method. As drawn in Figure 4.7, the blue line and red line represent torque response over speed with modified

method and classical method respectively. When the speed is above the cut-off point, the SynRM enters the constant power region. Before that it is called constant torque region. For the classical method, the cut-off point is 585 rpm while the boundary is 650rpm for the proposed. Therefore the constant torque region is extended. During the constant power region, the torque for the proposed method is above that of the classical. So the torque capability is enlarged during high speed range. Figure 4.8 confirms the result as well.

Figure 4.8 Comparison of the performance of the torque capability for modified and classical hysteresis-based DTC

In Figure 4.8, simulation is done to explain how higher torque capability is achieved in the flux weakening region. In this case, flux counter is removed and thus flux status is continuously forced to be altered once torque error counter reaches its limit. Like Figure 4.7, the blue line and red line represent torque response with modified method and classical method respectively. A step speed reference is abruptly applied at 0.8s. And it is quite interesting to observe that the constant torque region is terminated at approximately 1s but the duration of the constant torque region for the proposed is 0.074s longer, a 9

percent of enhancement. During constant power region, the actual torque for the proposed is still higher than that of the classical.

A complete comparison is drawn between DTC without and with modified flux weakening when load torque is 0 and 1 Nm respectively in Figure 4.9 and 4.10. Initially, the synchronous reluctance machine works at steady-state speed equal to 400rpm. At 0.8133 second, speed reference changes from 400 rpm to 1340 rpm. As indicated in Figure 4.9 and 4.10, at first, the estimated torque and stator flux instantaneously increase to their maximum values, the relationship of which is determined by MTPA. When the machine speed further grows up, flux is reduced to avoid saturation. And in the meanwhile, torque capability is lessened affected by voltage and current limits. In other words, the SynRM enters the flux weakening region.

(a)

Figure 4.9 Comparison of performance between (a) DTC and (b) modified DTC when torque is 0 by simulation

Figure 4.10 Comparison of performance between (a) DTC and (b) modified DTC when torque is 1 Nm by simulation

Under no load condition, the speed response time with modified flux weakening strategy is 1.5942s while it is 1.7482s for classical method. Mathematically, $(1.7482-1.5942) / 1.7482 = 8.81\%$. Thus, 8.81 percent of improvement is achieved.

When the load torque equals 1 Nm, the effectiveness of modified flux weakening strategy becomes more pronounced. With modified flux weakening strategy, the speed response time is reduced from 4.1687s to 3.3624s. Thus, $(4.1687-3.3624) / 3.3624 = 19.34\%$. 19.34 percent of advancement is obtained.

The faster speed response is plausible from Figure 4.9 and 4.10. It is because the higher torque capability is obtained during high speed range. However, the stator flux is compromised to achieve faster speed response. It can be easily observed that with modified flux weakening strategy, stator flux has larger ripples than that of the classical method.

Figure 4.11 Trajectory of the stator flux

Another interesting result is the trajectory of the stator flux shown in Figure 4.11. In comparison with Figure 3.8, when the speed of a SynRM exceeds its base speed, the amplitude of the stator flux is reduced. In addition, the trajectory of the stator flux is not a pure circle because of the application of the proposed method. When the static state is achieved, the stator flux tends to be a value near zero.

4.3 Conclusion

A simple and dynamic modified flux weakening strategy for hysteresis-based DTC controlled synchronous reluctance motors is discussed in this chapter. The proposed method achieves faster torque response and higher torque capability during high speed range while maintaining the whole system stable

by adding flux error counter. Another benefit of the proposed strategy is its simplicity because of the elimination of SVM block, normally adopted for modified flux weakening operation. Simulation is carried out by MATLAB/SIMULINK to verify the effectiveness of the proposed method.

Chapter 5 Modified DTC with CSF Strategy

5.1 Introduction

The main objective of the proposed method in Chapter 4 is to boost torque capability during high speed range and thus reduce speed response time. However, there exists two main disadvantages in the modified hysteresis-based DTC, namely variable inverter switching frequency and high torque ripples. Due to the variation of the time taken for the actual torque to touch the upper and lower hysteresis bands under different operating conditions, the switching frequency in a hysteresis-based DTC is variable. The torque ripple is closely linked with torque hysteresis band. If the value of torque hysteresis band is too large, the torque ripple would be large. On the other hand, if the band is set too small, the faster transit of torque requires smaller sampling period and hence more accurate processor. Alternatively, sampling period can be reduced to obtain smaller torque ripples. But this also depends on a faster digital processor. Several methods have been proposed to deal with these problems. One of the interesting methods is to introduce dithering technique [40]. Utilizing this method, the torque and flux errors are superimposed with high-frequency triangular waveforms whose amplitudes equal their corresponding hysteresis bands. In this case, the hysteresis torque comparators are forced to follow that of triangular waveforms. Nevertheless, this condition holds true only when the slope of the triangular frequency is close to the positive and negative slopes of the torque and flux error, because the gradient of torque error is determined by the operating conditions.

Switching is performed at regular intervals for the other one of the approaches. Two or more voltage vectors are selected instead of applying a single voltage vector during one sampling period. Consequently, by choosing the appropriate voltage vectors are combined to synthesize the desired stator voltage. In [41]-[55], the switching periods are divided into two states. The difference among them is the way where duty cycles are calculated. For example, in [41]-[44], it employs predictive control. For [45]-[48], it applies an

equation to obtain minimum average torque ripple. Fuzzy logic is introduced in [49]-[52]. Another approach is to use space vector modulation [53] [54]. A further improvement is to combine SVM with fuzzy logic [55]. But for abovementioned methods, the structure of hysteresis-based DTC can not be retained and complicates the control architecture in machine control.

5.2 CSF-based DTC and Simulation Results

A completely different method is to replace the torque hysteresis controller with constant switching frequency (CSF) controller as depicted in Figure 5.1

Figure 5.1 Constant frequency torque controller.

The torque error firstly goes through a PI controller and hence generates T_{COM} . This signal is then compared with two triangular carriers. One is larger than zero and the other is less than zero. And these two carriers are 180 degrees out of phase to horizontal axis. T_{CU} , T_{CL} are the results of comparison between T_{com} with two carriers, which are subsequently fed into the torque status logic to yield τ .

The Torque status logic in Figure 5.1 can be described as follows

$$\tau = \begin{cases} 1, & \text{when } T_{CU} \geq C_{upper} \\ \tau(k-1), & \text{when } C_{lower} \leq T_{CU} \leq C_{upper} \\ 0, & \text{when } T_{CL} \leq C_{lower} \end{cases} \quad (42)$$

where $\tau(k-1)$ is the previous torque error status.

In this case, the simple structure of hysteresis-based DTC is maintained as shown in Figure 5.2. It features simple implementation and other merits of hysteresis-based DTC, while torque ripple is substantially undermined.

Figure 5.2 Block diagram of hysteresis-based DTC with constant switching frequency controller.

The goal is to fulfill constant switching frequency and thus reduce torque ripples. Therefore the frequency and peak value of triangular waveforms are specified at constants. The parameters of hysteresis-based DTC and CSF-based DTC are specified in Table 5.1. The gain values of K_p and K_i are restricted to ensure that the absolute slope value of the output of PI controller does not exceed the absolute slope of triangular carrier. Due to highly non-linear properties of the system, PI parameters are chosen by trial and error. Thus the critical frequency ω_c equals $K_i / K_p = 12826/10 = 1.2826$ kHz. It is worth mentioning that the PI controller has a desaturation device to facilitate the stepping out of the saturation region when the output of the PI controller is larger than 2 or less than -2. That is why the magnitude of the triangular carrier was

chosen as 4. The sampling period is 25 μ s, i.e., the frequency is 40 kHz. Considering that triangular carrier is generated by a FGPA in real applications, its frequency is set lower than the sampling frequency and its actual value is by trials of fine tuning in simulation. As such, the triangular frequency is about 3.03 kHz.

Table 5.1 Parameters of hysteresis-based DTC and CSF-based DTC drives

Hysteresis-based DTC		CSF-based DTC	
Flux hysteresis band	$1/2^{13}\text{Wb}$	Flux hysteresis band	$1/2^{13}\text{Wb}$
Rated Flux	0.62 Wb	Rated flux	0.62 Wb
Torque hysteresis band	$1/2^4\text{Nm}$	Torque rated	3.5Nm
	3.5Nm	Proportional gain, K_p	10
Torque rated		Integral gain, K_i	12826
		Carrier frequency	3.03kHz
		Peak to peak of carrier	4

Simulation had been done to verify CSF-based DTC with modified flux weakening when load torque equals 0 and 1 Nm respectively. As seen in Figure 5.3 (a) and (b), the synchronous reluctance machine can achieve 1340rpm under CSF-based DTC strategy, when load torque equals 0 and 1 Nm.

(a)

(b)

Figure 5.3 Simulation of performance of CSF-based DTC with modified flux weakening when torque is 0 (a) and 1 Nm (b)

(a)

(b)

Figure 5.4 Comparison by simulation of torque ripples between modified DTC and CSF-based DTC when load torque equals (a) 0 and (b) 1 Nm respectively

Furthermore, two aspects are analyzed to indicate the strength of CSF-based method with modified flux weakening. One is torque ripple and the other is torque response time.

As depicted in Figure 5.4 (a) and (b), the torque ripple is reduced in comparison with the modified method, especially during high speed range. The blue line represents electromagnetic torque of modified DTC as mentioned in Chapter 4 while the red line denotes that of CSF-based DTC. It is easily observed that the torque ripple has been tremendously decreased by CSF based DTC.

On the other hand, the speed response time is discussed in Figure 5.5. The speed response time for 0 load torque is 1.678s, a 3.86% improvement compared with classical method. For 1 Nm load torque, the response time is 3.926s, a 6.18 percent of improvement.

(a)

(b)

Figure 5.5 Comparison by simulation of speed response time between classical method and CSF-based DTC when load torque equals (a) 0 and (b) 1 Nm

5.3 Conclusion

With CSF-based DTC, torque ripples, which are relatively large in modified DTC, have been addressed. At the same time, the speed response is still faster than the classical method. In summary, the CSF-based DTC not only has simple structure of DTC, but also features small torque ripples and constant switching frequency.

Chapter 6 Conclusion and Recommendations

6.1 Conclusion

In this thesis, two modified methods, namely modified DTC and CSF-based DTC, are proposed in comparison with classical method.

In Chapter 4, a modified hysteresis-based DTC strategy for synchronous reluctance machine is firstly proposed. The drive system is a combination of the classical DTC strategy and a flux error status modification block. In other words, the drive is composed of two hysteresis controllers, switching table, voltage source inverter, and synchronous reluctance machine. The proposed method has demonstrated better performance in comparison with the classical one with respect to the torque response and torque capability during high speed range.

In order to enlarge the torque capability and reduce torque response time, the optimum voltage reference is chosen instead of switching two voltage vectors alternatively in a subsector. Since the generation of the voltage vector is dependent on the output of the hysteresis controllers, a flux error status modification block is added before the output of the flux hysteresis controller is fed into the switching table. The hysteresis controller is a system not only dependent on its current value but also its previous states. In a word, the proposed method fulfills its goal with simple implementation.

Simulation is done to confirm the proposed method. The torque capability is enlarged because the constant power region is extended and thus the faster speed response is achieved. The faster torque response is attained because the voltage vector that hinders the acceleration of the electromagnetic torque is removed when the stator flux is tracked. When the stator flux is out of the hysteresis band for five consecutive times, the stator flux is considered to be out of control; otherwise the stator flux is thought to be tracked. Another point worth mentioning is that the merit of the proposed method is even more

pronounced when the load torque is larger. This is because the actual torque is still kept the same high level as no load condition and the percentage is higher after subtracted by load torque.

Although the proposed method in Chapter 4 hugely improves the speed response time, the torque ripple is relatively large. Consequently, in Chapter 5, the torque hysteresis controller is replaced with a constant switching frequency controller, which smoothes the torque ripples. Moreover, the response time is still faster in comparison with the classical method. In a nutshell, CSF-based DTC not only has simple structure, but also features reduced torque ripples and constant switching.

6.2 Recommendations

The main objective in this thesis is to boost torque capability during high speed range and thus reduce speed response time. As the focus of this thesis is to propose a modified method of hysteresis DTC, synchronous reluctance machine is simplified without taking magnetic saturation and core losses, which requires an improvement of the existing method. Moreover, although large torque ripples have been tackled in Chapter 5, the speed response time becomes relatively large compared with modified method in Chapter 4. This requests us to propose an even more advanced method to improve the performance of CSF-based DTC.

Author's Publications

Xiang Shuguang, Zhang Xinan, Gilbert Foo Hock Beng, "A modified hysteresis-based DTC strategy for synchronous reluctance motors in high speed range," IEEE Power Electronics and Drive Systems, June 2015

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