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TORQUE RIPPLE MINIMIZATION OF SWITCHED RELUCTANCE MOTORS
USING SPEED SIGNAL BASED PHASE CURRENT PROFILING

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ABSTRACT

The potential applications of an SRM as industrial drives are mainly inhibited due to the high torque ripple. This thesis presents a new control algorithm for torque ripple minimization of Switched Reluctance Machines (SRM). SRMs have significant ripple in the total output torque production due to the unique physical structure and torque production mechanism. Each phase of an SRM is excited separately and sequentially. During phase commutation the torque production responsibility does not transit from one phase to another phase instantly which causes torque ripple. The amount of torque ripple is significantly high if no corrective measures are taken to smooth it.

The high torque ripple of SRM causes vibration and acoustic noise in the motor drive system. The contemporary approaches to minimize the torque ripple of SRM are based on improved machine design or controller design or combination of both.

In this thesis the focus is to minimize torque ripple through new controller design. It is a challenge to reduce the torque ripple through control techniques, as the characteristic of the machine is quite nonlinear. Estimating or measuring the torque is either complex or costly. Since the torque ripple can manifest itself on the speed information, the speed controller can achieve torque ripple reduction to some extent, but the bandwidth of the controller is limited by the system inertia.

In this work the ripple on the speed information is used in a closed-loop control, to minimize the torque ripple during phase commutation. The speed signal is obtained

through a speed sensor or an estimator, which is less complicated and more cost-effective than using a torque sensor. In spite of being filtered by the inertia, the ripple information can be extracted from the speed signal using signal processing.

This thesis shows that a properly extracted signal has an acceptable correlation in terms of shape with the torque ripple. The acquired ripple information is used for shaping the SRM phase current during commutation and the results show that the proper shaping of current during the commutation portion could minimize the torque ripple significantly.

DEDICATION

I dedicate my work to my family.

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CHAPTER I

INTRODUCTION

1.1 Switched Reluctance Motors (SRM)

The concept of a Switched Reluctance Motor (SRM) was provided early in 1838. It was then used to propel a railway locomotive on the Glasgow-Edinburgh railway [1]. As only mechanical switches were used, full utilization of the SRM capabilities was not possible at that time. SRMs again came to the attention of researchers and industries with the advent of fast-acting power semiconductors in 1970. The basic design and working principles of the machine were established during the 1970s and 1980s [1] by professor Lawerenson's group.

An SRM has excellent potential for variable speed drive applications, because of some features such as high torque density, wide speed range of operation capability, inherent fault tolerability, absence of rotor magnets and low construction cost. The absence of rotor magnets makes an SRM a good competitor of permanent magnet motors (PM).

The advent of modern power electronics and microcontroller capabilities has made the SRM a possible candidate in different applications over PM machines [1-3]. An

SRM can be sized based on application requirements such as big traction motors to small system applications such as washing machine drives.

1.2 Advantages of SRMs

The unique stator and rotor construction of an SRM has some exceptional advantages compared to other types of motor. This type of construction is simple and cheap. Rotor construction is just combination of simple sheets of laminated steel and does not have winding or permanent magnets, brushes or commutator settings. All SRMs are driven by a drive circuit known as a converter. Having a stator coil between switches eliminates the possibility of shoot-through faults. The SRM converter also does not require bi-directional current, which reduces the number of power electronic switches in the converter.

The torque-speed characteristics curves of the SRMs can be tailored more easily than the induction and PM machines during the motor design stage by changing motor geometric parameters. SRMs have high initial starting torque without excessive in-rush current because of high self-inductance of the stator winding. The open circuit voltage and short circuit current are small during faults in an SRM. Due to the absence of permanent magnet and rotor windings, the rotor stays cooler during the operation and it has relatively lower inertia. As the stator phases are independent of each other, the drive can still run the motor in case of the failure of one or more phases. The fluxes in the SRM can be weakened easily, enabling wide speed range of operation.

1.3 Disadvantages of SRM

The most common disadvantages of the SRM are the high torque ripple and the acoustic noise. The construction and excitation pattern of the SRM are the causes of high torque ripple in this motor. Due to high torque ripple, the capacitor equipment on the dc bus is large to filter the ripple in the dc bus current. Also because of the circumferential shape, the stator vibrates as radial magnetic force is induced in it and this vibration creates acoustic noise in the SRM.

Air-gap magnetic fields in the SRM are produced only by the stator excitation compared to PM motors which have fields from the magnet into the air-gap too. Thus the KVA ratings of the SRM converter are high compared to the converter of PM motors.

1.4 Research Objectives and Motivation

Extensive research has been done for developing advanced control mechanisms to improve the performance of the SRM. One of the main objectives of this research is to minimize the torque ripple so that the performance and application areas of the motor can be broadened.

There are mainly two approaches to reduce the torque ripple. One approach is by improving the magnetic design of the machine and the other one is by using sophisticated electronic control algorithms. In the magnetic design approaches, by changing the stator and rotor pole structures, the amount of torque ripple can be reduced, but not eliminated completely. The design approach also affects the specific motor outputs.

There are different control methods for torque ripple minimization. Fig. 1.1 shows the overall block diagram of the SRM speed controller which includes torque ripple minimization mechanisms. One of the methods to implement torque ripple reduction is Torque-Sharing-Function (TSF) techniques. This method controls the rate of change of torque during commutation period according to a predefined TSF. Different TSFs can be used to reduce the amount of ripple. Torque is related to phase current and rotor position.

Figure 1.1 General block diagram of an SRM control.

Excitation angles to switch the phase on-off can be adjusted to minimize the torque ripple. Overlapping of the phases can be increased to gain additional control freedom. A high speed digital signal processor with a good amount of memory space is required to implement these control algorithms. Developing a torque ripple reduction method with less computation time and smaller data storage is desired.

A new method of ripple minimization has been proposed in the thesis to reduce the complexity of the controllers and parameter dependencies. In the proposed method, the estimated speed signal is used for torque ripple estimation and minimization. In the conventional control system, the bandwidth of the controller is not high enough to compensate for the torque ripple especially at higher inertia values. An additional feedback loop is proposed in this research to overcome the bandwidth limitation to compensate for the torque ripple.

Torque sensors are very costly and it is not always possible to integrate them with the system. On the other hand, a position sensor is an inseparable part of an SRM control since each phase excitation is done based on rotor position information. The new method here uses the speed signal which can be estimated from the position signal.

1.5 Thesis Organization

Chapters I and II of this thesis present a brief history, the basic configurations and the main working principles of SRMs. Different SRM converter configurations, their advantages and disadvantages, are also represented in detail. Research objectives and motivation for this thesis is also presented in Chapter II. Chapter III presents an extensive literature review of existing techniques for the torque ripple minimization in SRMs and provides a brief discussion about the proposed method. Chapter IV presents a detailed analysis of the proposed methods and analysis of the methods from a control point of view. Simulation results are presented in Chapter V. Chapter VI shows experimental validation of the proposed method. Conclusion and future work are presented in Chapter VII.

CHAPTER II

PRINCIPLES OF SRM DRIVE: OPERATION AND CONTROL

2.1 Introduction

The basic structure of SRMs, principles of operation, arithmetic equations governing the dynamics, typical converter topologies, and basic control principles are covered in this chapter.

2.2 SRM Machine Structure

The SRM is a doubly-salient machine, i.e. both the stator poles and rotor poles have saliency in the structure. The SRM structure can be different based on the number of phases and the number of rotor and stator poles. The choice of phase and pole numbers depends on the applications of that particular SRM. The selection of the number of rotor poles of an SRM has been expressed by Eqn. 2.1 [4],

$$N_r = N_s \pm km. \quad (2.1)$$

Here N_s is the number of stator poles and the value of the integer k can be chosen such that the modulus of k in terms of the number of phases should not be equal to zero.

Higher numbers of stator and rotor pole combinations reduces torque ripple. Also a good starting torque can be provided by increasing the number of poles.

In the SRM structure, only stator poles have concentrated windings. Diametrically opposite poles are connected in series or in parallel or in both series-parallel combinations to form the phase of an SRM. In a series connection, currents in the coils are the same but the voltage across each coil is lower than the total DC bus voltage. In fact, the voltage across each coil will be the total supply voltage divided by the number of coils. In a parallel connection, the voltages across each coil are the same but the total supply current in each coil will be the total current divided by the number of coils.

For an SRM drive the fundamental frequency in Hz is,

$$f = \frac{N}{60} N_r. \quad (2.2)$$

where N is the motor speed in rev/m and N_r is the number of rotor poles. The step angle is an important parameter for SRM design. It determines the control frequency per rotor revolution. The rotor positions at which each phase needed to be excited is determined by the step angle. The “step angle” or “stroke” of an SRM in rad is given by,

$$\varepsilon = \frac{2\pi}{N_p N_{rep} N_r}. \quad (2.3)$$

Stator pole arc (β_s) and rotor pole arc (β_r) are also very important parameters for SRM design. Torque is produced when the rotor pole and stator pole start overlapping

each other. To produce continuous torque, the pole arc must be greater than the step angle.

To produce the largest variation of phase inductance with respect to rotor position, the inter-polar arc of the rotor must be greater than the stator pole arc. The necessary condition is given by

$$\frac{2\pi}{N_r} - \beta_r > \beta_s. \quad (2.4)$$

2.3 Principle of Operation

The excited phase of the stator produces flux in the air-gap and attracts the closest rotor pole to minimize the reluctance of the magnetic path and also sets a stable equilibrium position. That position is also known as the ‘aligned’ position for the excited stator and rotor pole. The phase inductance is also at a maximum value for that particular ‘aligned’ position. As the rotor moves, due to saliency in the structure, the air gap changes between the stator and rotor pole. Due to changes in the air gap, the phase inductance decreases gradually from its maximum value at the ‘aligned’ position to its minimum value at the ‘unaligned’ position. At the unaligned position, the rotor pole is exactly halfway between two poles of the same phase. Unsaturated aligned and unaligned inductances are the two key reference positions for SRM control. The waveform of the ideal motoring current is a series of pulses and these pulses are synchronized with the positive or rising slope of the inductance. If the inductance profile is considered as linear, then the ideal torque waveform should be the same in shape as the ideal motoring current.

A stroke in SRM is known as the cycle of torque production associated with one current pulse.

The number of stator and rotor poles is unequal in an SRM and this ensures not all the rotor poles get aligned with the stator poles. If such condition occurs, the generated torque would be zero. Instantaneous torque is produced in an SRM due to attraction between stator and rotor poles. The phases are excited one after another to create rotation in the rotor. That rotor rotation produces continuous torque. To control the stator phase excitation rotor position is required at each instant of time. Based on the rotor position information a controller can be designed which would control the turn on angle, the turn off angle, and the winding current of each excited stator winding. The SRM can be rotated in either direction by changing the sequence of phase excitation in the controller. The torque production principle and operation of an SRM is similar to that of a stepper motor. The two most important features of an SRM from the control point of view are torque per ampere maximization and torque ripple minimization. For torque maximization the phase inductance must have the largest possible variation with rotor position. In that case the torque ripple also increases. The higher number of rotor and stator poles will help minimize the torque ripple if the described method of the per-phase-excitation pattern is followed.

2.4 Dynamics of the SRM

The general equation of stator current in one phase of the SRM can be described as,

$$V_{ph} = iR + \frac{d\lambda}{dt} \quad (2.5)$$

where V_{ph} is the DC bus voltage, i is the instantaneous phase current, R is the winding resistance and λ is the flux linkage in the stator. To maximize the utilization of the magnetic circuit, the SRM is always driven in such a way that there is always high magnetic saturation.

In SRMs, the flux-linkage λ is a function of rotor position and stator current represented as,

$$\lambda = \lambda(i, \theta). \quad (2.6)$$

Due to saturation in the magnetic circuit, λ is a nonlinear function of stator phase current and rotor position.

The general equation of an SRM can now be written as,

$$V_{ph} = iR + \frac{\partial \lambda}{\partial i} \frac{di}{dt} + \frac{\partial \lambda}{\partial \theta} \frac{d\theta}{dt}. \quad (2.7)$$

If magnetic linearity ($\lambda = L(\theta)i$) is assumed then V_{ph} can be represented as,

$$\begin{aligned}
V_{ph} &= iR + \frac{d(L(\theta)i)}{dt} \\
&= iR + L(\theta) \frac{di}{dt} + i \frac{dL(\theta)}{d\theta} \omega
\end{aligned} \tag{2.8}$$

where, $\omega = \frac{d\theta}{dt}$ is the rotor angular speed, θ is the rotor angular position and $L(\theta)$ is the instantaneous phase inductance. The third term in the equation is considered as the “back-EMF” term. It plays the same role as the back-EMF in other motors. The back-EMF voltage is dependent on the stator phase current, instantaneous rate of change of phase inductance and speed. The back-EMF is zero if the phase current is zero or the phase inductance is constant relative to the rotor position.

2.5 Energy Conversion

The power balance relationship can be used to evaluate the energy conversion process of the SRM. From Eq. (2.5), multiplying both sides by i , the instantaneous power can be expressed as,

$$\begin{aligned}
P &= V_{ph}i = i^2R + (L(\theta)i \frac{di}{dt} + i^2 \frac{dL(\theta)}{d\theta} \omega) \\
&= i^2R + \left(L(\theta)i \frac{di}{dt} + \frac{1}{2} i^2 \frac{dL(\theta)}{d\theta} \omega \right) + \frac{1}{2} i^2 \frac{dL(\theta)}{d\theta} \omega \\
&= i^2R + \frac{d}{dt} \left(\frac{1}{2} Li^2 \right) + \frac{1}{2} i^2 \frac{dL}{d\theta} \omega.
\end{aligned} \tag{2.9}$$

The first term represents the stator winding loss, the second term represents the rate of change of magnetic stored energy, and the third term represents the mechanical output

power. To represent the electromechanical energy conversion properly, a nonlinear analysis of the saturated magnetic circuit is required. The analysis is done based on magnetization curves and the magnetization curve is a plot of flux linkage λ versus current i , with the rotor position as a varying parameter. The energy conversion process in an SRM is explained by magnetic energy and co-energy from flux linkage vs current characteristics. Figure 2.1 shows the magnetic energy W' and the co-energy W for aligned position.

Figure 2.1 Energy and co-energy diagram in SRM excitation cycle.

Considering constant current and neglecting the stator resistance, the total electrical energy input in an SRM is equal to the sum of stored magnetic energy W and co-energy W' . The co-energy, W' is converted to mechanical output and magnetic stored energy W can be retrieved as electrical energy. The definition of energy ratio[4] for SRM is,

$$ER = \frac{W'}{W+W'}. \quad (2.10)$$

Here W' is the energy converted into mechanical work and W is the energy returned to the source through regeneration with the converter.

According to eqn. (2.10) for linear SRM at most 50% of the total energy can be converted to mechanical energy. If an SRM is operated under saturation, the co-energy portion increases, which also increases the SRM efficiency.

2.6 Torque Production Mechanism

In the SRM, the torque is produced using the reluctance principle of electric machines. The salient structure of the rotor creates variation in the reluctance due to air-gap changes and tends to move to the minimum reluctance position. Mathematically these can be expressed as,

$$T_{ph}(\theta, i) = \frac{\partial W'(\theta, i)}{\partial \theta}, i=\text{constant} \quad (2.11)$$

where co-energy W' is,

$$W' = \int_0^i \lambda(\theta, i) di \quad (2.12)$$

and where $\lambda = Li$ and L is the phase inductance. Thus the co-energy for linear magnetic characteristics can be expressed as,

$$W' = \frac{1}{2} Li^2. \quad (2.13)$$

The instantaneous phase torque can be expressed as,

$$T_{phase} = \frac{1}{2} i^2 \frac{dL}{d\theta}. \quad (2.14)$$

The total instantaneous torque of a machine is given by the sum of the phase torques,

$$T_{instantaneous}(\theta, i) = \sum_{ph 1} T_{ph}(\theta, i) \quad (2.15)$$

An idealized inductance profile and torque profile for one electrical cycle considering ideal constant current excitation is shown in Fig. 2.2. As seen from the Eqn. 2.10, motoring torque is produced during rising inductance slope and generating torque is produced during falling inductance region.

Figure 2.2 Ideal inductance, current and torque profile.

The rising inductance region is between the unaligned position of the rotor and stator to the aligned position and the falling inductance region is from the aligned position to the unaligned position.

For motoring operation, phase currents need to be present during the rising inductance region and for generation operation the current should be present during the falling inductance region. For optimum drive performance the positioning of phase currents with respect to rotor position should be appropriate.

2.7 Converter Topologies

The SRM phase currents can be unipolar as torque production does not depend on the direction of current flow. The stator phases of an SRM are also electrically isolated from one another. Based on these two unique features of SRMs, many different power circuit configurations can be developed. Motor construction and the number of phases also affects the type of converter to be designed. Converter design also can be dependent on particular applications of the SRM [4].

The most commonly used converter for an SRM is the bridge converter because of its flexibility and versatility in four quadrant operation. The typical bridge converter for a three phase operation is shown Fig. 2.3.

Figure 2.3 Classic bridge converter [4] for SRM drive.

As shown in Fig. 2.4, during the magnetization period, both the switches are turned on to connect the DC sources to the phase winding and transfer energy from the power supply to the motor. After the phase current is established in phase one, the switch is controlled to be turned on or off to regulate the phase current. During commutation, or just after the aligned position, both switches of a phase are turned off and demagnetization is done through the two freewheeling diodes.

Figure 2.4 (a) Magnetization, (b) Free-wheeling and (c) Demagnetization.

With this type of converter, the windings of each phase can be controlled independently. If phase overlap is desired in the design of the control system of SRM, this type of

converter is highly important. This converter configuration is also suitable for high power drives. But the disadvantage is that it requires two switches and two diodes per phase.

Different types of converters are being used to reduce number of switches and increase drive efficiencies. A split capacitor converter is shown in Fig. 2.5. It has one switch per phase but the DC supply is also being divided by using two capacitors. In this type of converter, when a phase is energized, either the lower or the upper capacitor DC voltage appears across the phase with respect to the midpoint of the two capacitors. So for torque production, only one-half of the DC supply voltage can be used. This converter can be used only with even number of phases as to maintain balanced power in the supply capacitors.

Figure 2.5 Split-capacitor converter [4] for SRM drive.

To minimize switching losses by reducing the number of switches a new type of winding technology is introduced called bifilar winding. This type of winding increases

the cost, by adding an extra winding which has made the topology complex and unpopular.

For low speed operation and if PWM current control is applied for the entire operation, another type of converter topology is suggested by Miller [1]. As shown in Fig. 2.6, the proposed converter has one switch common to all the phases. The main limitation for this type of converter circuit is that at high speed, after turning-off one phase, it cannot be demagnetized quickly because the DC source can be connected in reverse order across the demagnetizing phase winding. This is due to having one common switch in the configuration which intermittently keeps turning on and off when PWM scheme is being used.

Figure 2.6 Miller converter [4].

2.8 Outer Loop Speed or Torque Controller

The outer loop controller can be either speed or torque controller. Both of these controller types are discussed in the following subsections.

2.8.1 Torque Controller

For producing motoring torque in an SRM the motor phase current is synchronized with the rotor position. By using a torque controller, the reference phase currents are generated and a current regulator is used to regulate the reference currents into the phase winding to produce necessary torque. The current regulation of the phase is shown in Fig. 2.7, which produces the required output shaft torque. Either the upper, or lower, or both the switches are alternatively turned on and off to maintain the required reference current. This type of controller is known as hysteresis current controller.

Figure 2.7 Hysteresis current controller to follow a current profile.

Figure 2.8 Outer speed control loop of SRM.

2.8.2 Speed Controller

The outer speed loop commands either the phase currents or the torque to regulate the SRM speed. The speed control loop determines the speed from the encoder using sensorless estimation as feedback to generate the necessary reference current or torque. A complete system block diagram is shown in Fig. 2.8.

2.9 Control Approaches

The performance of an SRM highly depends on the appropriate positioning of the phase excitation current based on rotor position. The turn-on time, dwell angle and the magnitude of the phase current determine the torque and motor efficiency. The operating speed of a motor determines the type of the control to be employed.

2.9.1 Parameters to Control

SRM control parameters are turn-on angle, turn-off angle and the phase current. The dwell angle is the difference between turn-off angle and turn-on angle. In current

controlled drives, current is regulated directly. In voltage controlled drives, current is regulated indirectly by changing the phase voltage. At low speed, due to negligible back-emf, the current rises instantly after turn-on. The current can be regulated by controlling the supply voltage or regulating the current level. The performance of the SRM depends on the type of control used to control the motor. With increasing speed, the back-emf increases and it opposes the supplied dc voltage. Phase advancing and phase turn-off are adjusted to produce motoring torque properly. Lack of proper adjustment of turn-on and turn-off will produce either zero or negative phase torque.

At high speed, an SRM operates in single pulse mode and the control is done by advancing the turn-on angle and also adjusting the dwell angle. At high speed when the current magnitude is high and the rotor is in appropriate position the back-emf exceeds the applied bus voltage. This is the reason behind the decrement of current even if a positive voltage is applied during the positive inductance slope. The turn-on angle and dwell angle value is adjusted according to the speed value using proper controller. At turn-off both the switches are turned-off so that a negative voltage is applied across the phase to demagnetize the phase quickly. When the rotor passes the aligned position the back-emf voltage polarity is reversed and instead of opposing the current it might increase the value. In high-speed operation, the back-emf is greater than applied DC bus voltage. This causes the conduction window to shrink to apply any kind of control. The only control parameters are the dwell angle and the advance angle. Dwell angle is used to regulate the torque and advance angle can be adjusted to increase the efficiency.

2.9.2 Calculation of an Advance Angle

In the ideal case, the turn-on angle is advanced in such a way that the desired value of the reference current is reached just at the onset of the pole overlap where the positive inductance slope starts. In the unaligned position the inductance value is small and remains almost constant. So in the unaligned position, back-EMF can be neglected. Neglecting phase resistance, Eqn. 2.5 can be written as,

$$\begin{aligned} V_{ph} &= L(\theta) \frac{di}{dt} \\ &= L(\theta) \frac{\Delta i}{\Delta \theta} \omega. \end{aligned} \quad (2.16)$$

The advance angle θ_{adv} can be calculated based on unaligned inductance L_u , ω , V_{dc} and the reference current i^* .

$$\theta_{adv} = L_u \omega \frac{i^*}{V_{dc}}. \quad (2.17)$$

2.9.3 Voltage Controlled Drive

When precise torque control is not an issue, the low performance drive with PWM voltage control and adjustable duty cycle would be sufficient to control the SRM. A block diagram of a voltage controlled drive is presented in the Fig. 2.9.

Figure 2.9 Voltage controlled drive [4].

An angle controller takes rotor speed information as an input and gives necessary turn-on and turn-off angles. The commutator block provides the necessary gating signals based on control inputs and rotor position information. The PWM generation is the inner loop, and for precise speed control, an outer speed feedback loop is added to the system. An additional current controller in the system will further adjust the duty cycle.

2.10 Advance Control Strategies

Based on the applications of the SRM, various performance indexes such as maximized torque per ampere, maximum efficiency or minimum torque ripple can be optimized. Suppose an SRM is to be used in a traction application which needs high efficiency and wide speed range operation. There are some applications where torque ripple is a critical issue, such as electric power steering in car. However torque per ampere maximization and ripple minimization is not possible at the same time. The drive efficiency is related to torque per ampere maximization. For high efficiency, theta-on and theta-off needs to be determined accurately. Precise angle determination needs an

accurate SRM model and online parameter identification. For torque ripple minimization accurate SRM modeling is also important in order to control the phase currents during commutation. In high performance drives to minimize torque ripple, a commutator and torque controller work together to generate the necessary gating signals. The torque controller can be implemented through a model or table used on SRM characteristics.

2.11 Conclusion

This chapter briefly explains various SRM aspects from a mathematical point of view, in-terms of various drives configuration that can be used with the SRM and also from various applications point of view. The SRM advantages and limitations are described in brief here. The motivation for the thesis is also to develop an approach to make the operability of an SRM suitable for industrial use.

CHAPTER III

TORQUE RIPPLE MINIMIZATION OF SRMS

3.1 Introduction

The torque production mechanism in an SRM is different from any other conventional machine. The mechanical output torque can be produced by applying independent pulse currents for each phase in synchronization with the rotor position. Torque ripple occurs during commutation when stator current commutates from one phase to another. The electromagnetic torque can be altered by either changing the phase current excitation or the inductance profile. Alteration in the current profile requires advance control techniques, and alteration in the inductance profile requires advance machine design techniques. In magnetic design based ripple minimization the number of phases, the number of rotor poles, and the stator and rotor poles structure can be controlled. Modifications of these design parameters affect the motor performance. There are many control based torque ripple reduction approaches. The excitation angles can be varied to reduce the torque ripple during commutation. The phase currents can also be controlled during the commutation interval. This chapter review origination torque ripple in an SRM, different torque control methods to minimize ripple and limitations of these methods.

3.2 Origination of Torque Ripple in SRM

Due to that unique doubly salient structure of an SRM torque pulsation is inherent in the machine operation [5]. For torque production, an SRM uses variable reluctance principle through air-gap variation. The inductance of the stator winding is a function of both current and rotor position, and the relationship is nonlinear due to saturation effects, which makes the torque production nonlinear. The definition of torque-angle-current characteristics and the magnetization of the phase are nonlinear. These nonlinear characteristic curves along with the magnetization pattern actually specify the amount of torque ripple during commutation. The torque verses position curve for a specific value of current in two consecutive phases is shown in Fig. 3.1. At θ_c , the commutation occurs

Figure 3.1 Torque-current-angle characteristics of SRM and torque dip.

and due to the nonlinear torque production by each phase, the shared torque from each phase is not equal to the total torque and a ‘torque dip’ occurs.

The torque ripple is defined as the difference between the maximum and minimum instantaneous torque compared to the average torque as described in Eqn. 3.1. The

instantaneous maximum and minimum value of torque is measured during steady state operation of the machine.

$$\text{Torque ripple} = \frac{T_{inst(maximum)} - T_{inst(minimum)}}{T_{average}} \times 100\%. \quad (3.1)$$

3.3 Design Based Torque Ripple Minimization Approaches

For the generation of reluctance-based motoring torque, a unique structure is used for SRM machines. In an SRM, the stator has windings, and when the windings are excited, it creates flux in the system. The flux passes through the rotor which is made of laminated steel. The geometric structure of the stator and the rotor affects the flux linkage characteristics of the SRM. The electromagnetic flux linkage phenomenon is highly nonlinear in characteristics for which the torque production process in SRM is also nonlinear. This nonlinearity causes uneven torque to be produced during commutation by both the incoming and outgoing phases, and this uneven sharing of torque creates high torque ripple during commutation.

With appropriate design, partial torque ripple minimization can be achieved. The torque-current-position ($T-i-\theta$) characteristics curves per phase of the SRM can provide some indications of torque ripples to be occurred in the system. Torque dips occur in certain ranges of rotor position when both the outgoing and incoming phases are excited, and based on the amount of current in both phases, the torque is created. Certain structural parameters affect the $T-i-\theta$ characteristics curves, i.e. the ripple percentage of the machines. The design parameters are stator-rotor pole overlapping, air-gap, stator

pole arc (β_s) and rotor pole arc (β_r), material used for rotor construction, number of poles and number of phases [4]. In [2, 6] it has been shown that the torque ripple is highly sensitive to stator and rotor pole arcs. The inductance overlap ratio, K_L can be expressed as the ratio of inductance overlap of two adjacent phases to the angle over which inductance is changing [7].

$$K_L = 1 - \frac{\text{Stroke Angle}}{\min(\beta_s, \beta_r)}. \quad (3.2)$$

In Eqn. 3.2 the '*Stroke Angle*' is defined as,

$$\text{Stroke Angle} = \frac{2\pi}{N_{ph}N_{rep}N_r}. \quad (3.3)$$

The overlap region is dependent of '*stroke angle*'. In Eqn. 3.3 N_{ph} , N_{rep} and N_r respectively represent the number of phases, repetition number, and number of rotor poles. By changing the value of N_{ph} , N_{rep} and N_r the value of '*stroke angle*' can be changed. If the value of '*stroke angle*' is reduced, the amount of torque overlap between phases is increased. The overlap ratio, K_L provides a good measurement of torque overlap for the commutating phases. It is observed that as K_L increases the torque dip decreases and the average torque increases [4]. In [8] the effect of stator and rotor pole arcs in motor performance are presented. It was also shown that with sufficient stator and rotor pole overlapping the torque production during commutation can be increased. Widening of stator pole arc suppress torque ripple and increase the torque value [6, 9]. But a wider stator pole will reduce winding area and will increase the copper loss. By increasing the number of strokes per mechanical rotation, torque ripple can be minimized. The number of strokes per mechanical revolution is defined as,

Figure 3.2 Stator and rotor pole arc of an SRM

$$Strokes_{(per\ mech.rev.)} = N_{ph}N_rN_{rep}. \quad (3.4)$$

From Eqn. 3.4 it is observed that strokes per mechanical revolution can be increased by increasing the number of rotor poles. But a higher value of N_r would decrease the saliency ratio [1, 10]. A lower saliency ratio would cause higher volt-amps, more copper loss as the switching frequency increases, and lower torque output. The number of phases can also be increased for a higher number of strokes. For an additional number of phase, additional stages of switches added to the control circuit which increases controller complexity and costs. Also a higher number of phases will increase the fringing effect in the magnetic circuit of the machine. The repetition number can also be increased to lower the ‘stroke angle’. According to [5] if N_{rep} is doubled then double torque will be produced due to the increase in the angular rate of change of flux linkage. But higher N_{rep} will make the geometric design complicated.

Different geometric design parameters of an SRM have a significant effect on the minimization of the torque ripple. But modification of these parameters beyond certain ranges sometimes causes performance degradation for the machine. So a change of either one or more design parameters requires optimization to achieve the proper value for certain applications or operating conditions. Using only the design approach minimize torque ripple is not possible completely. So controller based approaches are also used to achieve better results. Optimization also includes several control parameters along with design parameters [11, 12]. As stator and rotor poles have a direct impact on the inductance profile which can be used for ripple reduction, these are the most used approaches for design based ripple minimization [13-18]. In [13], optimal machine design was obtained for a given current waveform. A gradient based method is used to optimize the stator tooth shapes. The design is implemented using finite element analysis software and for an actual machine such precise stator shaping is not possible. A new type of stator pole tip geometry has been investigated in [14]. The new geometry was designed to have more flux-linkage, thus obtaining more magnetic saturation compared to conventional design at the same current level. The new modified stator pole geometry results in more fringing flux from the stator pole. This reduce effective air gap between stator and rotor. The total reluctance of the motor is thus reduced due to the lower air gap. More flux will pass through the rotor i.e. more flux-linkage would take place which will increase the saturation of the motor. In the proposed design, a higher torque will be produced from unaligned to a certain angle when stator and rotor poles are overlapping. Another paper [15] studies the impact of stator pole shape modification on electromagnetic torque. In [15] the tapered shape of the stator pole is compared with

project shape. The results are verified for the 12/8 pole machine using finite element analysis. From the results, it was observed that the projected pole shape shows less torque ripple compared to the tapered pole shape. The projected pole arc at the unaligned position provides wider torque-current-position characteristics curves, which causes reduced torque ripple during overlapping. In [16], two different stator shapes are analyzed for ripple reduction performance. One shape is the taper type structure of the stator. The taper type stator pole structure decreases the reluctance of the stator pole sections which would reduce the overall reluctance of the machine. Another structure is observed using a stator face with non-uniform air gap. The air gap profile between the stator and the rotor poles is an important parameter for changing the torque profile through inductance. A stator pole face with non-uniform air gap design is made such that the air gap gets narrower as the rotor pole overlaps with the stator pole. The inductance profile becomes flatter near the aligned position. This helps to reduce the torque ripple.

Studies have also been done based on only rotor shape design for torque ripple minimization. New types of rotor shapes are proposed in [17, 18]. In [17] the rotor pole shape with the asymmetry lamination layer angle is proposed. The fringing flux creates torque ripple just before the overlapping of the stator and the rotor pole. In [18] a notched tooth type rotor design is proposed in the forward rotating direction to reduce fringing flux. This new rotor design approach produces linear inductance profile as fringing flux during overlapping position of stator and rotor pole, get reduced. [19] used a genetic optimization algorithm on the results of a magneto-static field analysis at constant current. The rotor is subdivided into smaller arcs with specific radius and angular length for each arc. These are the decision variables for the optimization algorithm. By

optimization, substantial reduction in ripple is achieved. By considering dynamic conditions of torque and current, the performance of the algorithm can be improved and more ripple reduction is possible. The algorithm is computationally very time consuming. Also material nonlinearity is not considered in optimization calculations.

Research has been done to modify both the stator and rotor pole geometry and to reach an optimized value for a reduced torque ripple in an SRM. Optimizations of the values are done mostly due to minimize the performance degradation of the machine because of geometrical structure change. In [20] a neural network based approach for optimal design considering the stator and rotor geometric parameters is presented. The outcome of this optimization procedure is the improvement of average torque with minimum ripple. Predictions of torque ripple characteristics are provided for different motor configurations varying the stator and rotor pole arcs. By using an industrial motor, a motor model was developed and field solutions are used to validate the model. Then, using field solutions, various motor geometries are obtained. A generalized regression neural network is used for interpolation of the numerical data. The model developed using finite element analysis is not properly matched with the experimental one, especially in the motoring torque region. Also the neural network based computations are complex and time consuming. In [21], a particle Swarm optimization is used for pole arc optimization for maximum average torque and minimum torque ripple. A comparison of convergence characteristics of particle Swarm optimization and neural network based on genetic algorithm for optimization of SRM geometry is provided. According to that analysis particle Swarm optimization method converges much faster than genetic algorithm based neural network optimization. In terms of robustness (minimum, average

and standard deviations) of particle swarm is much better than genetic algorithm. A numerical multi-objective optimization method for rotor shape optimization is shown in [22]. The objectives are high starting torque, higher average torque and lower ripple in torque. However this new design uses asymmetrical poles and variable air-gap. Initially the rotor design was for high starting torque. Later the structure was optimized for other machine performance parameters, one of which is torque ripple. In [23], a new method for calculating the average torque has been presented. Later the average torque is used as a parameter to compare different machine design models. The method has been verified for an 8/6 SRM. Different pole arcs for stator and rotor are analyzed with finite element based machine models. After rigorous calculations, an optimum stator and rotor pole arc is obtained.

By extensive literature review it is observed that in the design based ripple minimization method, finite element analysis (FEA) based model has been used at large scale. The FEA based models sometimes suffer from the inaccuracies due to some assumptions during calculations. Moreover the calculation time is very high for such analysis.

3.4 Controller Based Ripple Minimization Approach

Controller based torque ripple minimization approaches are most popular for SRMs [45-50]. Control of torque can be direct or indirect for ripple minimization [24]. For the direct control method, instantaneous torque is controlled, and for the indirect method, flux linkage or phase currents are controlled by shaping the torque. Phase current

profiling can be done either online or offline [24]. All the control based approaches of the SRM require torque (T) characteristics of the machine with respect to current (i) and rotor position (θ). This is popularly known as the T - i - θ characteristics of SRM. In most of the control based approaches, the T - i - θ characteristic is stored in a tabular format. But the static T - i - θ characteristics based control models have some limitations. The static T - i - θ table does not consider the losses that occurred during dynamic operation. Mutual coupling effects are also neglected. So the table cannot provide the precise value of current. An analytical expression can also be used to represent the T - i - θ characteristics. Using an analytical expression, it is difficult to describe the highly nonlinear torque distributions of an SRM. High computation time and memory storage are also critical issues for any model-based algorithms.

The methods of torque control are mainly based on torque sharing and current profiling [25-28]. Other methods are instantaneous [29-31] torque control, fuzzy logic, and neural network based [32 -35] control etc. Torque sharing function (TSF) and current profiling based approaches are studied in detail at [36]. An ideal TSF is defined to provide torque sharing between individual phases to minimize commutation ripple. The TSF can be sinusoidal, cubic or exponential functions [37-40]. In [36] the author presented a method which translates the reference torque to a reference current waveform based on analytical expression. An optimization criterion is being applied to the TSF.

In the instantaneous torque control, direct torque production is controlled to minimize torque ripple [37-38]. The direct instantaneous torque control structure proposed in [29] is simple, but during implementation the complex switching rules in the commutation region is the main limitation of these methods. In [31] an iterative learning

method is proposed which takes care of controller disadvantages due to model independencies. It also has the limitation of a complex switching structure.

In SRM controls, accuracy of the model is crucial as the machine model and the parameters are highly nonlinear. For a complete model based control system each motor has to be characterized. Even two machines built using the same process might need to be characterized individually. To develop control methods which are independent of any model research has been done on fuzzy logic and neural network based control [32-35]. These type of methods have no dependencies on the model and are not much effected by parametric variation of the system. Fuzzy logic based control proposed in [32,33] produces good results. These controllers have online adaptation features. But the computation time taken by such complex algorithms is a fact to consider for some applications where accurate responses are required for certain application requirements. Also the current reference produced by these methods sometime produce high initial current. The neural network based methods are also being considered for torque ripple minimization [34, 35]. The computation complexities and high initial currents are still a problem of these methods as with the fuzzy logic based ones.

There are several other methods to minimize torque ripple. The technique in [38] adopted a linear torque sharing during commutation and reduced peak current requirement of each phase. In [39] a sinusoidal TSF and fixed frequency PWM has been proposed for current regulation but this approach has been limited for low current application. In [40] another TSF has been proposed which is based on optimization of turn off angles and overlap regions. An online method has been proposed in [41] which

shows good results and not dependent on a machine model. But special types of machine design is necessary and the algorithm is computationally intensive.

Many other methods have also been tried on ripple minimization of an SRM. In [42] a bio-inspired model for ripple minimization has been presented. A model is developed here based on a simple model of mammal's limbic system and emotional process. A field oriented control based model, similar as the permanent magnet synchronous machine, was tried to be adopted in the control of an SRM in [43]. SRM inductance was modeled in a dq reference frame which was used with a complex switching strategy to establish smooth control. The method accuracy is dependent on smooth generation of the inductance profile which is the main drawback.

3.5 Conclusion

In order to overcome these limitations of model based approaches, a new way of phase current shaping is proposed here which is suitable for online implementations, contains less complex calculations and requires less memory storage. The suggested method does not use the motor $T-i-\theta$ characteristics to implement torque-current conversion.

CHAPTER IV

TORQUE RIPPLE MINIMIZATION OF SWITCHED RELUCTANCE MOTORS USING SPEED SIGNAL PROCESSING

4.1 Introduction

In this chapter a new controller based torque ripple minimization method has been proposed, verified and implemented. The proposed method is based on current profiling by using signal processed speed signal. Any ripple occurrence in the torque signal is reflected in the speed through the electromechanical dynamics of the drive system. The speed signal contains transient signal, DC level shift and noises. By using the filtering technique on speed signal, required information about torque ripple are collected. The torque ripple information from the processed speed signal has been used to minimize torque ripple using feedback controller. The method has been developed in this study are highly suitable for low to medium speed applications. Most of the controller based ripple minimization methods discussed in Chapter 3 is dependent of the SRM $T-i-\theta$ characteristics model. For the model dependent methods, the computation time are high and good size of memory storage are required in the embedded controller. The model based control system suffers from the inaccuracy problems, as the proper model of an SRM cannot be developed because of high degree of nonlinearity. One of the main

objectives of this work here is to develop a controller based torque ripple reduction technique for SRMs which is independent of any model based $T-i-\theta$ characteristics and flux-linkage information. A model independent torque ripple reduction technique can be applicable to any SRM with minimum modification. Also the computational complexity of such a method will be much less compared to other methods as nonlinear characteristics are avoided to generate torque ripple cancellation information. As a result the model inaccuracy effects will be much less for the proposed system.

In the closed loop speed control system, the controller output would respond to reduce any ripple in speed as well as the ripple in torque. But the performance is limited to the bandwidth of the speed controller. So the outer loop speed controller cannot generate the required reference signal to cancel out the torque ripple. For outer loop speed control the mostly used and industrially accepted controller is proportional-integral-derivate (PID) controller. The PID based controller bandwidth is affected by the settings of the gains used to tune the system. For a feedback based system not any gains can be used, otherwise the system would be unstable. So for low bandwidth PI based speed controllers, additional compensation is added in the feedback system based on extracted torque ripple information from the speed signal. The compensation has improved the total system bandwidth by providing proper reference to minimize the torque ripple.

This chapter presents general outer loop speed control of an SRM with and without the proposed compensation. A simplified model of the SRM speed control system has been developed and analyzed to show the effect of inertia, gains and system parameters on the performance of the controllers.

4.2 Torque and speed relation from electromechanical equation

A new algorithm for the torque ripple minimization of an SRM has been proposed here. In the proposed method torque ripple information has been extracted after applying signal processing on the speed signal. From the electromechanical relation of the machine (Eqn. 4.1), it can be observed that any changes in the instantaneous torque are reflected in the motor rotational speed as acceleration. The electromechanical torque balance equation can be represented as,

$$\frac{d\omega}{dt} = \frac{1}{J}(T_e - T_L). \quad (4.1)$$

here T_L , J , and T_e represent load torque, motor inertia, and electromagnetic torque generation respectively. At steady state, any changes in T_e is equivalent to the derivative of speed signal, $\frac{d\omega}{dt}$, on the other hand load variation would also result in fluctuation in the motor speed. Since the frequency of the torque ripple produced by the phase commutation is directly linked to the speed and motor design, the torque ripple information can be extracted by proper signal processing. We found that using band pass filter centered at the pole passing frequency of the motor operation and taking first order derivate of the speed signal, we can extract a wave-shape which co-relates very well with the torque ripple.

Figure 4.1 Block diagram of speed signal processing.

The center frequency of the filter is then adaptively updated based on the speed of the motor. Figure 4.1 shows block diagram of speed signal processing steps and Fig. 4.2 shows the electromagnetic torque and processed wave-shape of the speed signal obtained from the simulation of a (12/8) SRM operating at 500 rpm. It is evident from the Fig. 4.2 that both the torque and processed speed signal have excellent similarity between them. At 500 rpm the pole passing frequency is 200 Hz [44], which is seen clearly from Fast Fourier Transform (FFT) of the torque and processed speed signal as shown in Fig. 4.3.

Figure 4.2 Electromagnetic torque output and wave shape of processed speed signal.

Figure 4.3 FFT of electromagnetic torque output and processed speed signal.

4.3 Proposed Speed Signal Based Torque Ripple Minimization System

The block diagram of the proposed control system is presented in Fig. 4.4. The outer loop PI based speed controller provides the reference current for the hysteresis current regulator. In the conventional speed control system of an SRM, the outer speed control loop generates reference torque values for a certain speed command. The bandwidth of the speed controller is relatively kept low then the additional current is generated by the proposed controller to compensate for the torque ripple. Output of the speed controller and the propose torque ripple compensation blocks are added together and commanded as a total reference current to a current regulator block.

Figure 4.4 Block diagram of the proposed system.

In the proposed algorithm of ripple minimization no a-priori knowledge of torque-current-position characteristics of the machine is required. In the inside loop of torque controller for ripple minimization current profiling during commutation method is applied. The required current profile is generated from the processed speed signal of the system. The speed signal is processed through a band-pass filter and a first order derivate to remove DC components from the signal. The first order derivate is used to achieve more dynamic behavior after the band-pass filtering. After filtering and derivative operation, the processed signal is multiplied with negative gain and added to the current

command from the PI speed controller. The compensation gain is adjusted to certain value so that minimum torque ripple is obtained and system stability is maintained. The hysteresis current controller has been used to regulate the current. A conventional commutation scheduler and asymmetric H-bridge converter has been used for voltage application in the SRM. A current reference is generated and added to the existing reference current to reduce torque ripple. The presented method does not require any machine model to calculate current reference for desired torque generation. It has less complex calculations, so implementation is easier and faster compared to other SRM control systems.

4.4 Proposed System Model in s-domain for Analysis

A block diagram of the proposed system in s-domain has been presented in Fig. 4.5. In the block diagram the current controller and SRM model together are represented by a gain, which is obtained from $T-i-\theta$ characteristics. In the s-domain speed command is taken as an input to the system. The error between commanded speed and actual speed is taken as an input to the system. The error between commanded speed and actual speed is the input for PI speed controller outer control loop. The torque ripple is represented as a disturbance, mechanical dynamics are represented as an inertia and viscous damping. The torque ripple compensation unit is modelled as the cascaded connection of band-pass filter, differentiator and a gain.

The parameter in the block diagram are represented as,

$\omega_{command}$ = Commanded speed in radian per second

ω_{actual} = Actual speed in radian per second

Figure 4.5 Proposed system block diagram in s-domain

K_p = Proportional gain of PI speed controller

K_i = Integral gain of PI speed controller

K_s = SRM gain

K_c = Compensation gain

W_h = High pass filter cut-off frequency in radian per second

W_l = Low pass filter cut-off frequency in radian per second

J = Motor inertia in kg-m^2

B = Motor damping factor

Torque ripple = Torque ripple input to the system as disturbance

4.5 Impact of Inertia on the Speed Signal

The inertia of the system plays an important role in the proposed control method. The processing of speed ripple information is one of the key factor of the proposed compensation method. If the inertia is high, the load mechanics act as a low pass filter for

the torque and not much ripple would appear in the output speed. This has been represented mathematically in Eqn. 4.2 as,

$$\frac{d\omega}{dt} = \frac{1}{J}(T_e - T_L). \quad (4.2)$$

From Eqn. 4.2 it is evident that if inertia (J) is high, the $\frac{d\omega}{dt}$ value will be very low. The effect of inertia is studied by updating the system in Fig. 4.5 and eliminating the outer speed control dynamics. T_e represents the average torque value and T_{e_Ripple} is considered as an input as shown in Fig. 4.6.

Figure 4.6 s-domain model of the proposed system for the torque and ripple as input.

The closed loop transfer function, $H(s)$ of the simplified system can be expressed as,

$$H(s) = \frac{s^2 + (W_h + W_l)s + W_l W_h}{Js^3 + [J(W_h + W_l) + B + K_s K_c W_l]s^2 + [JW_l W_h + B(W_h + W_l)]s + BW_h W_l} \quad (4.3)$$

$$= \frac{\omega_{actual}(s)}{T_{ripple}(s)}$$

A combination of a fixed frequency sinusoidal wave and DC signal has been used to represent the average torque and torque ripple input to the system. The output of the system has been observed for high and low inertia value. The values are used for simulating the system is presented in the Table 4.1.

Table 4.1 Simulation parameters to observe inertia effect on the proposed system.

Variable	Values	Unit
J (Inertia)	0.0005(low) and 0.01(high)	kg-m ²
K_s (SRM represented as a linear gain model of current to torque)	0.286	Nm/A
K_c (Compensation gain)	0.04	
W_h (High-pass filter cut-off frequency)	$2\pi \times 100$	rad/sec
W_l (Low-pass filter cut-off frequency)	$2\pi \times 800$	rad/sec
B (Viscous Friction Coefficient)	0.1	Nm sec

The output signal of the system for low inertia and high inertia has been shown in Fig. 4.7 (a) and 4.7 (b), respectively.

Figure 4.7 (a) Output signal for low inertia value.

Figure 4.7 (b) Output signal for high inertia value.

As evident from Fig. 4.7 (a) and (b) that the inertia value is affecting the amount of ripple magnitude in the output signal. For higher inertia, it is observed that signal

ripple magnitude is decreased. A good amount of ripple magnitude information in the output signal would certainly be helpful for the extraction process of the signal. As in the practical scenario speed signal will be estimated using the rotor position signal. A high resolution position encoder is required to get more samples of rotor position value. A low resolution position encoder might also be good for running the motor. But the speed estimation from low resolution position encoder would not contain the data about the torque ripple during the commutation period. So because of the lack of data about complete signal the proper signal might not be reconstructed after signal processing. Also the noise affects the position encoder signal. If magnitude is high for speed ripple then it will not be corrupted much by noise magnitude. Based on how much is the motor inertia, the filter parameters on the proposed system requires adjustment. In the simulation a first order based filter system has shown good results because no noise was considered in the speed signal. But in the experiments we have to use higher order and more sophisticated digital filters in processing speed signal.

4.6 Impact of Inertia on the System Response

The simplified model of Fig. 4.6 has been analyzed again for the system response for three different inertia values. For the system behavior analysis in terms of inertia, pole-zero plot function of MATLAB has been used. The system close loop transfer function in Eqn.(4.3), has highest third order in the denominator and second order numerator. So the system has highest three poles and two zero's at most. The values chosen for the system simulation are presented in Table 4.2.

Table 4.2 Simulation parameters to observe the effect of inertia on the system response.

Variable	Values	Unit
J (Inertia)	0.0001(low) , 0.001 (medium) and 0.01(high)	kg-m ²
K_s (SRM is represented as a linear gain model of current to torque)	0.286	Nm/A
K_c (Compensation gain)	0.04	
W_h (High-pass filter cut-off frequency)	$2\pi \times 100$	rad/sec
W_l (Low-pass filter cut-off frequency)	$2\pi \times 800$	rad/sec
B (Viscous Friction Coefficient)	0.1	Nmsec

Figure 4.8 Pole-zero map of $H(s)$.

The system pole-zero plot is shown in Fig. 4.8. From the Fig. 4.8 it is observed that one pole and two zero's are on the real axis and two conjugate poles are on the upper

and lower part of the pole-zero plot. For the analysis of the system response at different inertia values, the movement of these conjugate poles in the plane is observed. The inertia values are selected as high, medium and low numerical values. So for high inertia value any torque ripple input will not cause much ripple in the output speed signal because the system response for high inertia becomes critically damped or under damped. On the

Figure 4.9 (a) Pole-zero plot of $H(s)$ for low inertia.

Figure 4.9 (b) Pole-zero plot of $H(s)$ for medium inertia.

Figure 4.9 (c) Pole-zero plot of $H(s)$ for high inertia.

other hand for low inertia, the damping ratio is also low. Any change in the instantaneous torque shows up as a significant amount of ripple magnitude on the speed. The three pole-zero plots of the system for high, medium and low inertia values are shown in Fig. 4.9(a), (b) and (c) respectively. From the three plots it is observed that for different inertia values, the two complex conjugate poles have different locations on the plane. As the inertia increases, the conjugate poles get closer to the real axis, which suggests that system response time gets smaller.

As the inertia increases the poles move away from the imaginary axis which suggests that the system damping coefficient increases. From this analysis, we can conclude that inertia has an impact on the oscillation of the speed signal which would determine the capability of the proposed controller in reducing torque ripple.

4.7 System Compensation with Feedback

In the previous section the effect of inertia on the system response has been described. If the system response is not satisfactory for a certain value of inertia, by using the feedback loop compensation gain, the system's significant conjugate pole position

can be modified to some extent to achieve desired system response. In the feedback loop K_c , W_h and W_l can be changed to tune the system response. Three values of k_c where $k_{c3} > k_{c2} > k_{c1}$ are used to see their effects on the system response and presented in Fig. 4.10. As k_c increases system gets less damped which makes it possible to extract more information from the speed signal.

Figure 4.10: Pole-zero plots for the simplified system for same inertia but increasing values of k_c where $k_{c3} > k_{c2} > k_{c1}$.

4.8 Band width of Proportional Integral (PI) as Speed Controller of SRM

Figure 4.11 A simplified model of the conventional speed control system of an SRM.

Having additional compensator or the torque ripple increases the overall bandwidth of the speed controller. To show the improvements on the bandwidth of the speed controller, let's consider only the operation of the outer speed controller in the proposed system. Figure 4.12 shows the simplified model of the outer speed controller only, where $G(s)$ in the system plant, $H(s)$ in the controller. The whole system can be represented as a negative feedback system with a torque ripple as sinusoidal input. The close loop transfer function of the system can be represented as,

$$\begin{aligned} e(s) &= T_{ripple}(s) + T(s) \\ &= T_{ripple}(s) - H(s)\omega_{actual}(s) \end{aligned}$$

From the block diagram we can see,

$$\begin{aligned} \omega_{actual}(s) &= G(s)e(s) \\ &= G(s)(T_{ripple}(s) - H(s)\omega_{actual}(s)) \\ \frac{\omega_{actual}(s)}{T_{ripple}(s)} &= \frac{G(s)}{1+G(s)H(s)}. \end{aligned} \tag{4.4}$$

The unity feedback transfer function $G(s)H(s)$ for the conventional speed control system in Fig. 4.11 can be presented as,

$$G(s)H(s) = \frac{K_s K_p s + K_i}{J s^2 + B s} . \quad (4.5)$$

Table 4.3 Simulation parameters to observe the effect of inertia on the controller BW.

Variable	Values	Unit
J (Inertia)	0.0005(low) and 0.05(high)	kg-m ²
K_s (SRM is represented as a linear gain model of current to torque)	0.286	Nm/A
K_p (PI controller proportional gain)	0.165	-
K_i (PI controller integral gain)	0.9	-
B (Viscous Friction Coefficient)	0.1	Nm sec

The bode plot for the system at Fig. 4.11 and for the transfer function of Eqn. 4.6, has been presented in Fig. 4.12 (a) and (b). The plot at Fig. 4.12 (a) is for low value inertia system and the plot at Fig. 4.14 (b) is for high inertia without any compensation. At pole passing frequency open loop gain of the system is -22.9 dB for low inertia and -62.7dB for high inertia.

Figure: 4.12(a) Bode plot for low inertia value without compensation.

Figure: 4.12(b) Bode plot for high inertia value without compensation.

Figure 4.13 Model of the torque ripple input based proposed system.

Let us now consider the speed control system of an SRM with the proposed compensation technique in the feedback loop. The s-domain model of the proposed method is shown in Fig. 4.13. The system open loop transfer function $G(s)H(s)$ is derived and presented in Eqn.(4.7).

$$G(s)H(s) = \frac{K_s(K_c W_l + K_p)s^3 + [K_s K_i + K_s K_p(W_h + W_l)]s^2 + K_s[K_i(W_h + W_l) + K_p W_h W_l]s + K_i K_s W_h W_l}{Js^4 + [J(W_h + W_l) + B]s^3 + [B(W_h + W_l) + W_h W_l J]s^2 + BW_h W_l s} \quad (4.7)$$

The system shown in Fig. 4.6 and the system in Fig. 4.13 has difference in the input signal. The system in Fig. 4.6 has actual speed signal as input and no PI controller transfer function was considered in that system. So the system closed loop transfer function in Eqn. 4.3 has two conjugate poles in the system as shown in pole-zero plot in Fig. 4.8. The system in Fig. 4.15 has a PI controller in the feedback loop which has a input of negative actual speed signal and proposed system in feedback with the actual speed command. The transfer function in Eqn. 4.7 has been used to generate the bode plots and it is an open loop transfer function of the proposed system. The system has

three zeros in the system which are evident in Fig. 4.16(a) and (b). The numerical values to generate the bode plots of the system in Eqn. 4.7 are in table 4.4.

Table 4.4 Simulation parameters to observe the effect of inertia on the controller BW with compensation.

Variable	Values	Unit
J (Inertia)	0.0001(low) , 0.001 (medium) and 0.01(high)	kg-m ²
K_s (SRM is represented as a linear gain model of current to torque)	0.286	Nm/A
K_c (Compensation gain)	0.04	-
K_p (PI controller proportional gain)	0.165	-
K_i (PI controller integral gain)	0.9	-
W_h (High-pass filter cut-off frequency)	$2\pi \times 1000$	rad/sec
W_l (Low-pass filter cut-off frequency)	$2\pi \times 1400$	rad/sec
B (Viscous Friction Coefficient)	0.1	Nm sec

Bode plots of the proposed compensation method with the above numerical values in the speed control system of a SRM has been provided in Fig. 4.14 (a) and (b).

Figure 4.14(a) Bode plot for low inertia value with compensation.

Figure 4.14(b) Bode plot for high inertia value with compensation.

The plot at Fig. 4.14 (a) is for low inertia value with proposed compensation and the plot at Fig. 4.14 (b) is for high inertia value with the proposed compensation method. These bode plots of the proposed compensation method has been compared with the bode plots of the conventional system. The magnitude plots for all the cases have been compared at a certain frequency. The magnitude improvement is evident for the proposed system.

The bode plots are scaled at frequencies between 0.1 to 10000 rad/s. The magnitudes of the conventional system and proposed system with compensation are compared at 1300 rad/s (close to pole passing frequency at 500 RPM). At that frequency the magnitude of conventional system from Fig. 4.12(a) and (b) at low and high inertia are respectively

-22.9 dB and -62.7 dB. As expected at higher inertia attenuation is more than the lower inertia. But for the conventional system for both the low and high inertia the magnitude gain is low enough to recover any signal at that particular frequency.

In the proposed system the bode plots at Fig. 4.14(a) and (b) are presented for similar low and high inertia values of the conventional system. For the proposed system the magnitude gains are +6.81dB and -33.1dB for the low and high inertia at 1300 rad/s. If the values are compared with the conventional system gain it is an improvement over the conventional system. As magnitude gains are improved the required signal at certain desired frequency can be extracted.

4.9 Conclusion

It is apparent from the analyses of this chapter that using proposed method system performance can be modified so that we can extract excellent torque ripple

information from the speed ripple information and these information can be used to cancel torque ripple in an SRM. The effect of system inertia also been analyzed as high system inertia causes much of the ripple information to be filtered and not much ripple is appeared in the speed information.

CHAPTER V

SIMULATION RESULTS

5.1 Introduction

The proposed torque ripple reduction method for SRM has been applied to two different types of machines with different characteristics and simulation results have been presented to show the improvement in torque ripple in both cases.

5.2 Switched Reluctance Motor Model

A good magnetic flux linkage model of any SRM machine is required to design and simulate a controller for an SRM. The main challenge of acquiring SRM properties is to how much accurately the inherent magnetic non-linearity of the machine is being modelled. Proper determination of motor characteristics is important for designing and achieving accurate results as the actual application of the SRM to be used, in simulation for any new motor control algorithm. Analytical or lookup table based model can be used to simulate the dynamic behavior of the SRM.

The two most popular analytical models in SRM research are the Spong model [29] and the Arthur Radun model [11]. In Spong model a simplified nonlinear model of

an SRM is being presented which is based on curve fitting technique. The torque relation function in the Spong model requires numerical solution to get the value of the required current for a desired torque [12]. But using Spong model any real machine characteristics is difficult to match due to lack of associated geometric parameters. Another popular analytical model of an SRM is Arthur Radun's geometrical model [11]. This model considers parameters, based on material magnetic property and geometry of the machine. This model does not use actual machine data or results of finite element analysis. Iron saturation considered in the model and the model is applicable for the rotor positions when the stator and rotor poles overlap. But the $T-i-\theta$ characteristics of this model does not represent proper machine model at unaligned position and aligned rotor positions.

The lookup table base model contains flux linkage $\lambda(i, \theta)$ and static torque $T(i, \theta)$ data table for different current level ' i ' and rotor position ' θ ' values. Based on the input current and rotor positions, necessary flux linkage and torque values are interpolated from the lookup table.

5.3. Finite Elements Based Analytical Models of SRM

Simulations results of the proposed control method to minimize torque ripple of SRM are implemented for two different machines. Both the motor models are lookup table based model and the data tables are obtained from finite element analysis of the motor.

One of the SRM analytical model for simulation with the proposed method is a 24/16 poles SRM (SRM-1) machine. The geometric parameters of the machine is presented in the Table 5.1.

Table 5.1 SRM-1 Machine parameters.

Parameter Description	Parameter Value
Configuration:	24/16 pole
Number of phase, N_{ph}	3
Number of repetition, N_{rep}	4
Number of parallel path, N_{par}	4
Number of series path, N_{ser}	2
Number of turns per pole, N_p	23
Geometry:	
Material: M19 Steel	
Stack length, L_{stk}	54.8 mm
Radius to rotor yoke, R_{ry}	80.3 mm
Radius to rotor pole tips, R_g	100 mm
Radius to stator yoke, R_{sy}	127.9 mm
Machine outer radius, R_{out}	150 mm
Air-gap length, g	0.5 mm
Stator pole width, $\theta_{ps}=\beta_s*N_{rep}$	32^0
Rotor pole width, $\theta_{pr}=\beta_r*N_{rep}$	32^0

The flux linkage and $T-i-\theta$ characteristic of the SRM obtained from finite element analysis have been presented in Fig. 5.1 and Fig. 5.2.

Figure 5.1 $T-i-\theta$ characteristics of SRM-1.

Figure 5.2 Flux linkage characteristics of SRM-1.

The other SRM (SRM-2) characteristic that is used for simulation are presented in Table 5.2.

Table 5.2 SRM-2 Machine parameters.

No. of Phases	3
No. of Stator Poles	12
No. of Rotor Poles	8
Power	300 W
Phase Resistance	2.2 Ω
Minimum Inductance	6.553 mH
Maximum Inductance	28.25 mH
Stack Length	1.889"
Air Gap	0.015"
Radius to air gap at rotor	3.2724"
Radius to outside rotor yoke	2.4"
No. of Turns	75
Rotor Pole width	15°

The 300 W SRM (SRM-2) has applied to drive washing machine based load. This can also be a good candidate for any servo type systems after eliminating the torque ripple. The SRM has the following characteristics from the finite element based model analysis.

Figure 5.3 Flux linkage vs current characteristics of SRM-2.

Figure 5.4 $T-i-\theta$ characteristics of SRM-2.

5.4 Simulation Results of SRM-1

The proposed control algorithm has been simulated using SIMULINK software. The proposed algorithm is applied for an SRM-1 designed for traction application at 2800 rpm and at 100 N-m load torque. First the SRM-1 is tested for conventional PI based speed control. The conventional control block diagram is presented in Fig. 2.8.

Figure 5.5 and 5.6 show the associated torque ripple and speed ripple operating at 2800 rpm. For conventional controller the measured torque ripple is 30.68% and speed ripple peak to peak difference is 1.6 rpm. PI speed controller output reference with conventional controller and new reference with proposed controller are shown in Fig. 5.7 and Fig. 5.8.

Reference current generated from PI controller and actual current using hysteresis current control is presented in Fig. 5.9 and Fig. 5.10. As observed the closed loop speed controller tries to generate a current shape to minimize speed ripple if a speed controller is used in the outer loop. But due to bandwidth limitation the proper shape cannot be generated to minimize torque ripple.

Fig. 5.11 and 5.12 represent the simulated torque and speed of the SRM-1. The improvement in torque ripple is visible compared to Fig 5.5 and speed improvement compared to Fig. 5.6. The torque ripple is 4.4% and peak to peak speed ripple value is 0.2 rpm.

Figure: 5.5 Torque of SRM-1 at 2800 rpm with conventional controller.

Figure: 5.6 Speed ripple of SRM-1 at 2800 rpm with conventional controller.

Figure: 5.7 Current reference at 2800 rpm with conventional controller for SRM-1.

Figure: 5.8 Reference current of SRM-1 at 2800 rpm with proposed controller.

Figure 5.9 Reference phase current with conventional PI speed controller (SRM-1).

Figure 5.10 Actual phase current with conventional PI speed controller (SRM-1).

Figure 5.11 Torque of SRM-1 2800 rpm with the proposed controller.

Figure 5.12 Speed ripple of SRM-1 at 2800 rpm with the proposed controller.

Figure 5.13 Reference phase current with the proposed controller (SRM-1).

Figure 5.14 Actual current of one phase with proposed controller (SRM-1).

5.5 Simulation Results of SRM-2

Simulations have been performed for 300W SRM (SRM-2). The torque and speed waveforms with conventional speed control are presented in Fig. 5.15 and 5.16. The torque ripple percentage with conventional control is 138.25% and speed ripple is around 2.4 rpm. The reference phase current and actual phase current is shown in Fig. 5.17 and Fig. 5.18. The reference does not have much variation due to bandwidth limitation of the PI controller.

The SRM-2 was simulated at 500 rpm and with 1 Nm load torque. The same model is used to verify the proposed method at the same speed and load torque. The results are shown in Fig. 5.21 and 5.22. The torque ripple percentage with proposed method is 42.12% and speed ripple is 0.3 rpm. From the plots of reference and actual currents of one phase for the proposed method in Fig. 5.19 and in Fig. 5.20, it is evident that current profile for the proposed system has required shape to minimize torque and speed ripple.

As seen from the simulation the proposed method is much simpler to implement and does not contain complex equations of torque current relationships to minimize torque ripple. A proportional controller with proper gain is enough to generate necessary shape in feedback loop to minimize ripple. So the proposed method is highly effective for real time application because of low computation time.

Figure 5.15 Torque at 500 rpm with conventional controller (SRM-2).

Figure 5.16 Speed at 500 rpm with conventional controller (SRM-2).

Figure 5.17 Filtered Speed at 500 rpm with conventional controller (SRM-2).

Figure 5.18 Derivative of the filtered speed at 500 rpm with conventional controller (SRM-2).

Figure 5.19 Reference current of one phase with proposed controller (SRM-2).

Figure 5.20 Actual current of one phase with proposed controller (SRM-2).

Figure: 5.21 Torque at 500 rpm with proposed controller (SRM-2).

Figure: 5.22 Speed at 500 rpm with proposed controller (SRM-2).

CHAPTER VI

EXPERIMENTAL VERIFICATION

6.1 Introduction

The proposed control methods for torque ripple minimization are verified in Chapter V using analytical model of the SRM in the MATLAB/Simulink environment. The experimental verification of the proposed method is presented in this Chapter using 300 W SRM drive. The SRM used for experiments has been commercially applied as drive motor for washing machine. Some modifications has been done on the converter and motor to make it suitable for experimental setup. For controlling the motor operation a dSPACE based control system is used. To interface the controller and the converter board of the SRM an interface board has been developed. The sections of this Chapter have detailed explanations of the experimental process.

6.2 Hardware Setup

For experimental verification of the proposed method, hardware setup has been developed using SRM motor, converter, interface board, controller and dynamometer. As the motor and converter used in commercial application there was no available parameter

and characteristics data of the motor. Also the converter circuit board has no available technical documents. The machine dimensional parameters have been measured and then the measured values are used for finite element analysis to obtain the motor characteristics. dSPACE software is used to implement the control algorithm for the SRM. For dSPACE control the actuators are the gate signals for the converter switches. Analog sensors provides rotor position and phase current measurements to the controller. Interface board has been developed to condition the analog measurements from the sensors, digital signal from the position encoder and output PWM signals from the dSPACE. An analog to digital converter is used to get the digital equivalent measurement of analog phase current values. The current sensor circuitry includes a scaling portion and an antialiasing circuit which also act as buffer for the controller input too. The whole setup is driven by regulated DC supply designed for laboratory applications. The load requirement of the SRM is implemented using a dynamometer. For the output signal monitoring of the system digital oscilloscope is used and for signal acquisition current and voltage probes are used. dSPACE also has the signal display capabilities where it can show the acquired signal in real time. A block diagram of the hardware setup is presented in Fig. 6.1 which shows the interfacing of the individual systems.

Figure 6.1 Experimental hardware setup block diagram.

6.3 Determination of SRM parameters for modelling

For experimental verification of the proposed method a 300 W SRM from Maytag Neptune washing machine was used. The machine has no characteristics data available from the manufacturer. For the simulation purposes a model is developed by measuring the physical dimensions of the motor.

Figure 6.2: a) Stator of the experimental SRM.

Figure 6.2: b) Rotor of the experimental SRM.

Figure 6.2: c) Stator and rotor of the experimental SRM.

Figure 6.2: d) Assembled experimental SRM.

Figure 6.2 Experimental SRM structure of stator and rotor.

The number of turns in the stator winding is determined by using a passive method as without completely unwinding the stator winding it cannot be determined. First the motor was supplied with a fixed load, current and torque average was obtained using a power analyzer. The power analyzer provides average torque value after filtering the noisy torque signal obtained using the Magtrol dynamometer controller box. The geometric parameters are measured using appropriate measuring devices. All the measured values are put into a geometry based model of SRM[11]. The model generates the static $T-i-\theta$ characteristics of the machine. Initially the number of turns was estimated from the measurement of a diameter of a winding and total width of the bunch. The results from model is then compared with the experimental result. The number of turns is adjusted accordingly to match with the experimental measurements. For more accuracy finite element analysis software Flux2D is used to generate the $T-i-\theta$ characteristics of the motor being measured. A table of measured geometric parameters is presented in Table 6.1.

Table 6.1 Geometric parameters of 300 W 12/8 experimental SRM.

No. of Phases	3
No. of Stator Poles	12
No. of Rotor Poles	8
Power	300 W
Phase Resistance	2.2 Ω
Minimum Inductance	6.553 mH
Maximum Inductance	28.25 mH

Stack Length	1.889"
Air Gap	0.015"
Radius to air gap at rotor	3.2724"
Radius to inside stator yoke	3.8665"
No. of Turns	75
Rotor Pole width	15°

6.4 SRM Converter

The SRM comes with a custom built for industrial application converter as shown in Fig. 6.3. The converter is classic asymmetric H-bridge configuration based converter for three phase operation. The power rating of the converter is 300 W to match the power rating of the motor to be driven. The converter board has rectifier section as the whole board was operated from a single phase AC supply. The rectifier section of the board was bypassed as the output was a fixed DC supply without any external control and the output DC is not a purely regulated one. For more control on the supply and to get regulated DC supply a lab based DC supply was used. Initially the converter had it's own controller IC which was removed to provide control signal from the dSPACE controller. The board has a gate drive circuitry based on IR2101 chip which drives both the upper and lower switches of individual each phases. The MOSFET model IRF644 has a peak voltage rating of 250 V and peak current rating of 14 A. Additional diodes were used for freewheeling and demagnetization of SRM.

Figure 6.3 SRM converter used for experimental setup.

6.5 Gate Drive Interface Circuit

To provide switching signal to the MOSFETS through the gate drives from the dSPACE controller an interface board was built using proto board. The interface board is a vital part of the hardware as it provides protection, and conditioning for the signals. It also acts as a buffer and provides isolated ground for digital controller and power converter. The diagram of the gate driver is shown in Fig. 6.4.

Figure 6.4 Gate Drive Interface.

The gate signal commands come from the dSPACE controller and passes through a buffer IC SN74LVC541A. Opto-couplers are used between the controller and gate drive circuit to provide galvanic isolation. Separate DC supply was used for buffer and opto-coupler side. The interface board was designed to interface both 5V and 3.3V volt logic levels using SN74LVC541A as shown in Fig. 6.5.

Figure 6.5 Gate drive interface board.

6.6 Encoder Interface Board

Rotor position information is vital for the control of SRM machine. Precise speed information is required for the implementation of the proposed ripple minimization method. The rotary optical encoder model AE0631024R00.75YGY from Adison Electric fits into the motor has a resolution of 1024 pulse per revolution. The encoder required 5V supply and pull up resistor to the output channels to drive TTL load. As the encoder outputs goes to the input of the controller the MC14504BCP buffer is used for protection and conditioning. The circuit diagram of the encoder interface is shown in Fig. 6.6.

Figure 6.6 Encoder interface circuit.

6.7 Phase Current Sensor Circuit

The current sensor model used LEM55 is used for current measurements. Analog conditioning circuit is designed and implemented to interface the output of the current sensor and provide input to the dSPACE controller as shown in Fig. 6.7.

Figure 6.7 Current conditioning circuits.

The conditioning circuit has voltage follower as buffer, scaling, low-pass filtering and level shifter. The low pass filter is for filtering noise and also acts as an anti-aliasing filter. The cut-off frequency of the low pass filter is given as:

$$f_c = \frac{1}{2\pi RC} \quad (6.1)$$

with the element values used in the circuit the calculated cut-off frequency is $f_c = 48.23$ kHz.

6.8 dSPACE controller

The controller used in experimental validation of the proposed method is dSPACE. In dSPACE, control blocks can be developed in Matlab/Simulink then converted into equivalent C-code for operation. The signal can also be captured and saved into log file for further analysis. The hardware unit can communicate with the host

computer through Ethernet port. The system is equipped with 900 Mhz IBM PPC, 16 bit resolution of ADC and 0.0003 Hz to 150 KHz PWM generation feature.

6.9 Dynamometer and SRM Coupling

After successful testing of the hardware, controller and motor at no load conditions it was connected to a Magtrol eddy current hysteresis dynamometer. The dynamometer was controlled using the Magtrol DSP6000 controller unit for the application of load torque. Several power supplies have been used to power the different components of the setup. Data acquisition and real time control of the system was performed using the DSPACE Control Desk 3.7.4. Oscilloscopes and current probes were also used for observing the current wave shapes. The experiments were carried out at DC bus voltage of 120 V and the switching was limited to 30 kHz. The entire control system had a control loop has a 33 μ s of interrupt rate. A PI based speed controller was implemented for speed control with the gains being selected through a trial and error method until a satisfactory performance was achieved. The focus of this thesis is not speed control hence this portion didn't get much attention. The overall system performance would also improve if a better PI controller or any other speed controller is implemented. Fig. 6.9 shows a picture of the entire experimental platform used for testing.

Figure 6.8 Experimental setup.

6.10 Experimental Results

In this section experimental results of the proposed method are presented. Initially conventional control of the SRM has been applied. The block diagram of the conventional control has been shown in Fig. 6.10. In the conventional system outer loop speed controller has an actual speed calculation block from rotor position.

Figure 6.9 Conventional control method for SRM.

The motor speed is calculated using a frequency calculation blocks. From the position encoder, pulse stream is generated by the channel 'A' output. The time period of the

pulse varies with the speed of the system. So if the pulse width can be measured speed can be measured. The measured output is then scaled to match the rpm of the motor. To generate the reference current for the regulator a PI speed controller is used in the outer loop. The reference current value is generated according to the commanded speed. A hysteresis current controller has been used to regulate the phase currents.

6.10.1 Speed Signal Processing

The processing of speed signal is important for the proposed method to shape the phase currents in minimizing the torque ripple. A noisy and low resolution position signal will not provide enough information about the ripple in the electromechanical torque of the machine. For a good position signal acquisition a high resolution encoder (1024 PPR) is used in the system. From the one channel (Channel A) output pulse variation of the system speed has been measured. The measured speed is then filtered using infinite impulse response (IIR) digital filter and first order derivative is used to get the torque ripple information. With conventional controller the motor was ran at a fixed speed (500 rpm) and different loading conditions. The torque ripple should increase as we increase the load torque so as the speed ripple. These changes are evident from the recovered speed signals after signal processing.

Figure 6.10 (a) Speed estimation at $T_L = 0$ Nm.

Figure 6.10 (b) Filtered speed estimation at $T_L = 0$ Nm.

Figure 6.10 (c) Derivative of the filtered speed signal at $T_L = 0$ Nm.

Figure 6.10 (d) FFT Spectrum of the filtered speed signal at $T_L = 0$ Nm.

Figure 6.11(a) Speed signal at $T_L = 0.5$ Nm.

Figure 6.11 (b) Filtered speed signal at $T_L = 0.5$ Nm.

Figure 6.11 (c) Derivative of the filtered speed signal at $T_L = 0.5$ Nm.

Figure 6.11 (d) FFT spectrum of the filtered speed signal at $T_L = 0.5$ Nm.

Figure 6.12 Filtered speed signal at different loading conditions.

Figure 6.10, 6.11 and 6.12 show plots of speed signal processing with different loading condition. Figure 10(a) and 11(a) show acquired speed signals using position encoder. The signal has been acquired with a high resolution position encoder so the speed signal contains enough ripple information to be extracted using signal processing.

Figure 6.10(b) and 6.11(b) show the plot of filtered signal using IIR band-pass filter designed using MATLAB filter design tool. The selection of band-pass filter cut-off frequencies is important. As the SRM was running as 500 rpm the lower cut-off frequency is chosen as 80 Hz because at 500 rpm according to pole-passing frequency calculation the torque ripple should occur at 200 Hz cycle. Upper limit selection depends on noise presented the signal, stability of the system and how much information about frequencies is required to represent the torque ripple accurately. Because the more accurate information we get from the filtered signal the more accurate reference current shaping can be done. At loaded condition amount of ripple that can be extracted from the signal processing increase three times. Filtered speed signal is not actually the proportion of torque as seen from the electromechanical relation of motor and load. In order to get similar signal about torque ripple derivative of the filtered speed signal has been taken. The derivative output for no-load and loaded condition has been presented in Fig. 6.10 (c) and 6.11(c). This signal is used in feedback loop for reference current shaping for ripple minimization of the proposed system. If we compare the signals of filtered speed and derivate of filtered speed we see that these are different. The derivative output provides more dynamics about torque ripple signals. The FFT of the filtered speed signal for no-load and at loaded condition are presented at Fig. 6.10(d) and 6.11(d). As the system runs at 500 rpm speed we see a high component at 200 Hz in the spectrum plot of both the no-

load filtered speed signal and filtered speed signal at 0.5 Nm load. The derivatives of filtered speed signal at different loads are presented together in Fig. 6.12.

6.10.2 Implementation of the proposed method

For validation of the proposed method the system was ran at certain load and speed. The ripple information was used in the closed loop control to shape the current references. As we see from the plot at Fig. 6.11(c), the derivative of the filtered speed signal has a similar shape with the torque ripple. As we are trying to minimize commutation torque ripple we use the shape appeared at the pole passing frequency. A selection window is used to select the portion of the signal and that portion was added to the reference current. The system was started with a conventional control system and then load was applied to the system. When the whole system reached to steady state the compensation system then started. The results during steady state of the proposed method are presented in Fig. 6.13 and 6.14. These data were captured using dSPACE data capturing facility. The raw data was imported and processed using MATLAB to see the results in plots. As we see from the Fig. 6.13 and 6.14, ripple magnitude reduces after application of the proposed compensation method. With conventional system, the ripple magnitude is 1.8 unit and with proposed method the ripple magnitude is 0.5 unit. The torque ripple reduces to 72.2 % of the original one. The oscilloscope capture of the actual phase currents using current probes are shown in Fig. 6.15 and 6.16.

Figure 6.13 Actual phase current and filtered speed signal with conventional control.

Figure 6.14 Actual phase current and filtered speed signal with proposed control.

Figure 6.15 Actual phase current with conventional control.

Figure 6.16 Actual phase current with the proposed control.

6.11 Conclusion

The proposed torque ripple control method has been implemented experimentally successfully. The torque ripple is actively controlled and reduced significantly. No prior motor characteristics are required to implement the control. The proposed method can also be used instead of costly torque sensor to sense torque ripple shape from speed signal of the SRM. The speed signal can be estimated from high resolution position encoder signal which does not require additional sensors in the drive system. With all the benefits, the proposed method has excellent potential for industrial based applications.

CHAPTER VII

CONCLUSION AND FUTURE WORK

7.1 Conclusions

In this thesis a new control based torque ripple minimization method has been proposed. The proposed method uses a signal that has been acquired in real time to create the phase current profile without the requirement of pre-calculated $T-i-\theta$ characteristics. The new method has been verified for two different types of SRM machine model in simulations. For both of the machine the proposed method shows good performance in torque ripple reduction. Application of both hardware and software based filters are common to remove system noise and disturbances. The proposed system contains two PI based controller and proper tuning of all the gain parameters are required to achieve satisfactory system performance. Also proper estimation of speed signal is important as the shaping information for phase current reference comes from speed signal.

7.2 Future Work

It has been shown that torque ripple minimization of SRM is possible using real time acquisition of speed. But high performance operation the proposed system needs more tuning of gains and system parameters. The proposed ripple reduction method can be applied to medium performance drive system where ripple sensitivity is not very high. Food processor, washing machines etc. can be a good example of such system. By this method good amount of ripple can be reduced without adding any extra sensor in the system. This will reduce per unit cost of the system. The method can also be used as a low cost redundant system beside actual torque sensor based drive system. In order to make the system suitable for high performance drive application the following future developments can be followed.

1. Improve system stability and reduce system processing time by optimization of all the system parameters.
2. Harmonic content of the filtered speed signal can be linked to the torque ripple production.
3. Explore methods so that using low resolution position sensor can be used.

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