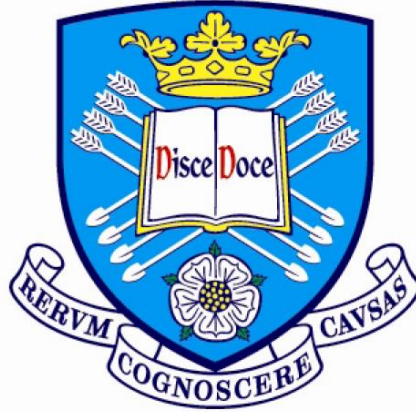


The University of Sheffield



**Sensorless Torque and Rotor Position Control for
Interior Permanent Magnet Machines at Standstill
and Low Speed**

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ABSTRACT

This thesis is focused on sensorless torque control of interior permanent magnet synchronous machines at standstill and low speed. It can be divided into two closely related sub-topics, i.e. sensorless rotor position estimation and torque control at standstill and low speed.

For sensorless rotor position estimation, a fast initial rotor position estimation method is firstly proposed based on high frequency rotating voltage injection. Compared to conventional methods, the proposed method can detect the initial rotor position faster by comparing the three phase high frequency current amplitudes as a result of high frequency rotating voltage injection. Secondly, a novel incremental inductance measuring method is proposed based on both high frequency rotating and pulsating voltage injections. Different from conventional methods which use either high frequency rotating or pulsating voltage injection for measuring the incremental inductances, the proposed method is immune to the delay effects and does not require a complicated closed-loop position observer. Thirdly, the cross-coupling effect can result in position estimation error in high frequency signal injection based sensorless control and deteriorates the torque control performance. A novel method is proposed for offline measuring the cross-coupling effect induced position error by injecting high frequency pulsating voltages in the dq -axes. The measured errors are then used for online compensating the cross-coupling effect. Compared to conventional methods using the offline measured relationship between incremental inductances for cross-coupling effect compensation, the proposed method is simple to implement and has better dynamic performance.

For improving the torque control performance, a simultaneous sensorless rotor position and torque estimation method is proposed for standstill and low speed operation. With the proposed method, the rotor position is estimated by injecting high frequency square wave voltages to the estimated dq -axes, which are simultaneously used for identifying the incremental inductances. The identified incremental inductances are subsequently used to calculate the electromagnetic torque by approximating the dq -axis flux linkages with curve-fitting technique. Besides, a whole speed range sensorless operation method is proposed using a single closed-loop position observer. Compared to conventional methods, unity-length position error signal can be obtained from the saliency model based on high frequency rotating voltage injection without using a scaling factor.

Various simulations and experiments have been carried out to verify the effectiveness of the developed methods.

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ABSTRACT

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LIST OF SYMBOLS

I_{dh}^r, I_{qh}^r	Amplitudes of the dq -axis HF currents	A
$I_{dh}^{\hat{r}}, I_{qh}^{\hat{r}}$	Amplitudes of the estimated dq -axis HF currents	A
I_n	Amplitudes of negative sequence high frequency current	A
I_p	Amplitudes of positive sequence high frequency current	A
$Amp_i_{abch}^s$	Amplitudes of three phase high frequency currents	A
i_{posi}^s	HF current induced by u_{posi}^s	A
i_{nega}^s	HF current induced by u_{nega}^s	A
$i_{dq}^{\hat{r}}$	Stator currents in estimated synchronous rotating frame	A
$i_{\alpha\beta}^s$	Stator currents in stationary two phase frame	A
i_{dq}^r	Stator currents in synchronous rotating frame	A
S_h	Power spectrum density	dB
L_d, L_q and L_{dq}	Apparent dq -axis and mutual inductances	H
L_{dh}, L_{qh} and L_{dqh}	Incremental dq -axis and mutual inductances	H
L_{Δ}	$L_{\Delta} = 0.5 * (L_{qh} - L_{dh})$	H
L_{Σ}	$L_{\Sigma} = 0.5 * (L_{dh} + L_{qh})$	H
T_e	Electromagnetic torque	Nm
T_L	Load torque	Nm
T_{cross}	Torque component contributed by cross-coupling effect	Nm
θ_{middle}	Angle at the middle of a sector	Rad
θ_{bound}	Angle at the boundary of a sector	Rad
θ_{VCT}	Continuous position output by vector tracking observer	Rad
$\hat{\theta}_r$	Estimated rotor position	Rad

θ_r	Rotor position	Rad
$\tilde{\theta}_r$	$\theta_r - \hat{\theta}_r$	Rad
ω_e	Electrical angular speed	Rad/s
$\hat{\omega}_e$	Electrical estimated angular speed	Rad/s
ω_h	Angular frequency of the injected signal	Rad/s
u_{nega}^s	Backward rotating HF voltage	V
$V_{discrete}$	Discrete voltage vector	V
E_{ex}	Extended back EMF	V
$E_{\alpha\beta}^{ex}$	Extended back EMF in stationary reference frame	V
$E_{\alpha\beta}^a$	Back EMF in stationary reference frame for active flux	V
u_{posi}^s	Forward rotating HF voltage	V
$u_{\alpha\beta}^s$	Stator voltages in stationary reference frame	V
$u_{dq}^{\hat{r}}$	Stator voltages in estimated synchronous rotating frame	V
ψ_a	Active flux	Wb
Y_{Δ}	$Y_{\Delta} = 0.5(1/L_q - 1/L_d)$	1/H
Y_{Σ}	$Y_{\Sigma} = 0.5(1/L_d + 1/L_q)$	1/H
p	Differential operator	
P	Number of pole pairs	
$S(\theta_r)$	$S(\theta_r) = \begin{bmatrix} \cos(2\theta_r) & \sin(2\theta_r) \\ \sin(2\theta_r) & -\cos(2\theta_r) \end{bmatrix}$	
γ	$\gamma = (L_{qh} - L_{dh})/(L_{qh} + L_{dh})$	
λ	$\lambda = L_{dqh}/L_{qh}$	

LIST OF ABBREVIATIONS

AC	Alternate current
AM	Amplitude modulation
ADC	Analog to digital converter
BPF	Band pass filter
BLAC	Brushless AC
BLDC	Brushless DC
DC	Direct current
DTC	Direct torque control
DFT	Discrete Fourier transform
FEA	Finite element analysis
EMF	Electromagnetic force
FOC	Field oriented control
HDM	Heterodyning demodulation
HF	High frequency
HPF	High pass filter
INFORM	Indirect flux detection by online reactance measurement
IPMSM	Interior permanent magnet machine
LPF	Low pass filter
LUT	Lookup table
MTPA	Maximum torque per ampere
MTPV	Maximum torque per voltage
NSC	Negative sequence current
PM	Permanent magnet

PLL	Phase locked loop
PSC	Positive sequence current
PI	Proportional integral
PWM	Pulse width modulation
SNR	Signal to noise ratio
SVM	Space vector modulation
SRF	Stationary reference frame
SDM	Synchronous demodulation
SRFF	Synchronous rotating reference frame
SFF	Synchronous shaft filter
SPMSM	Surface permanent magnet synchronous machine
SSRF	Synchronous stationary reference frame
ZVCD	Zero vector current derivative

CHAPTER 1

GENERAL INTRODUCTION

1.1 Introduction

Permanent magnet synchronous machines (PMSMs) have been widely used in both domestic and industrial applications due to excellent features of high efficiency, high torque density and high power density [JAH86], [ZHU07]. The commonly used control methods are direct torque control (DTC) and field oriented control (FOC). For DTC, the torque is directly controlled by regulating the stator flux linkages and the torque angle [TAK86], while FOC is based on decoupling control of the dq -axis currents [BLA72]. For the control of PMSMs with either DTC or FOC, the rotor position is essential. In FOC, the rotor position is needed to transform the currents, voltages and flux linkages in the stationary reference frame to the synchronous rotating frame, while in DTC, the rotor position is required to select the optimum voltage vectors [BUJ04]. Generally, the rotor position can be obtained by hardware sensors, e.g., Hall-effect sensors, encoders, and resolvers. These sensors would increase cost and decrease the reliability of the system due to additional cables and circuits. Consequently, elimination of the position sensors in PMSM drives has been the focus of intensive research for more than two decades [WAN20].

The rotor position can be estimated from the measured terminal currents or voltages which contain the position information. These methods are commonly referred to as sensorless methods. According to the models that are employed for position estimation, these sensorless methods can be classified into saliency model based methods [JAN95] and fundamental model based methods [HEJ11]. The saliency model based methods make use of the saliency variation of inductance or resistance with rotor position and are suitable for standstill and low speed applications [RAC10], while the fundamental model based methods utilize back electromagnetic force (EMF) or flux linkage signals and have better performance at medium and high speeds. The main focus of this thesis is the sensorless control of PMSM at standstill and low speed based on the saliency model.

An important premise for saliency model based sensorless methods is the existence of saliency variation with rotor position, i.e. resistance saliency or inductance saliency. The resistance saliency is usually caused by the different eddy current losses on dq -axes [YAN12a],

while the inductance saliency is caused by asymmetric structure on dq -axes or intentionally magnetic saturation [BIA07]. The eddy currents caused by the injected HF currents can cause Joule losses in the magnetic materials. The Joule loss can be modelled by resistances in the equivalent HF circuit. If the dq -axis equivalent HF resistances are different, such saliency variation of the equivalent HF resistances can be used for sensorless position estimation. The surface-mounted permanent magnet machine (SPMSM) has normally negligible saliency ratio, but saliency variation can also be created by magnetic saturation [LEE20]. Interior permanent magnet synchronous machine (IPMSM) has high saliency ratio due to salient pole structure and is the main focus of this thesis.

Saliency model based sensorless method is usually implemented by exploiting the saliency variation through test signal injection. The injected signal can be continuous high frequency (HF) voltage, current, discrete voltage, or the PWM waveforms [RAU07]. The rotor position can be extracted from the responses caused by the above injection signals with different demodulation methods [JAN95], [YOO11]. The sensorless performance can be affected by magnetic saturation, cross-coupling effect [LI09], multi-saliencies [REI08], load transient [BRI00], and delay effects [XU17a]. These factors are either due to the inherent properties of the magnetic materials in the machine or caused by the signal processing process for extracting the rotor position. In following sections, these factors will be introduced in detail.

As mentioned before, torque control can be achieved through FOC or DTC. No matter which method is adopted, feedback of the electromagnetic torque is essential for monitoring or control purpose [MAR20]. Hardware based torque sensors, such as strain gauges, torsional displacement can be installed to measure the shaft torque, which are expensive and require extra room and cables. Instead, the torque is usually estimated using the product of flux linkage and current. The flux linkage observer is affected by resistance variation, magnetic saturation. Various parameter identification schemes can be used to improve the estimation accuracy, such as those based on the fundamental model [LIU11], [ICH06] or those based on HF signal injection techniques [MAR20], [XU14]. In position sensorless control, the torque estimation becomes more challenging as the position estimation error can further deteriorate the estimation accuracy. Besides, to improve the disturbance rejection ability, load torque estimation based on the mechanical model is also desirable [WAN19a]. For many applications, whole speed range sensorless operation is required. Hence, the switch between saliency model based sensorless methods and fundamental model based sensorless methods should be considered.

In the following section, various position sensorless methods and the switch between different sensorless methods will be presented first. Then, different torque control and torque estimation methods will be introduced. Finally, the scope and major contributions of the research will be outlined.

1.2 Rotor position Sensorless Control Methods

There has been a large variety of researches on the sensorless control from standstill to high speed over the past decades as shown in Fig. 1.1(a). In general, position sensorless control methods can be classified as shown in Fig. 1.1(b).

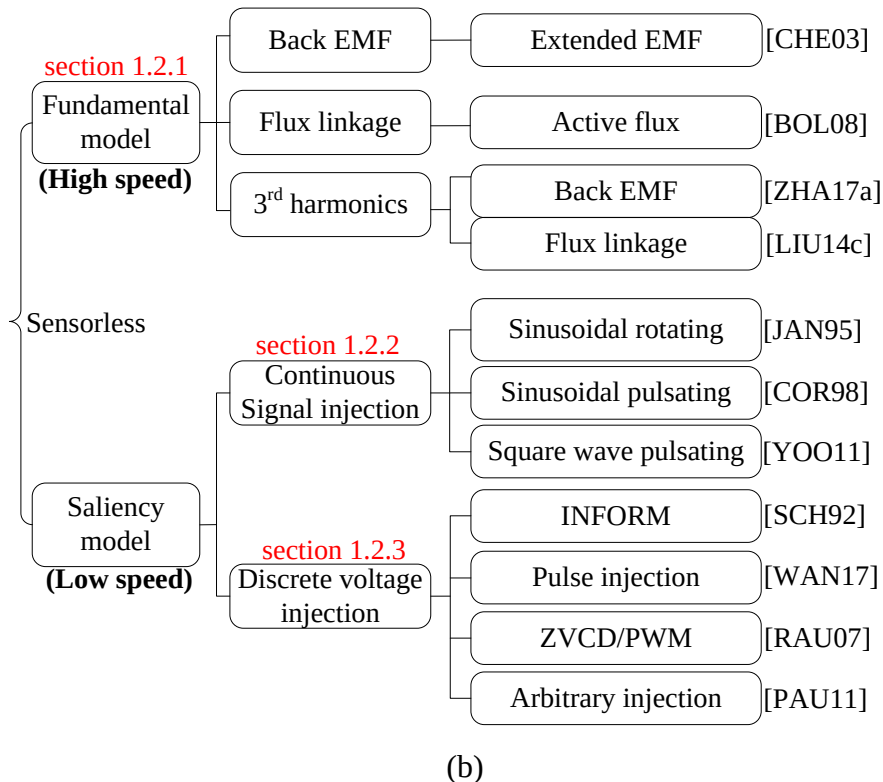
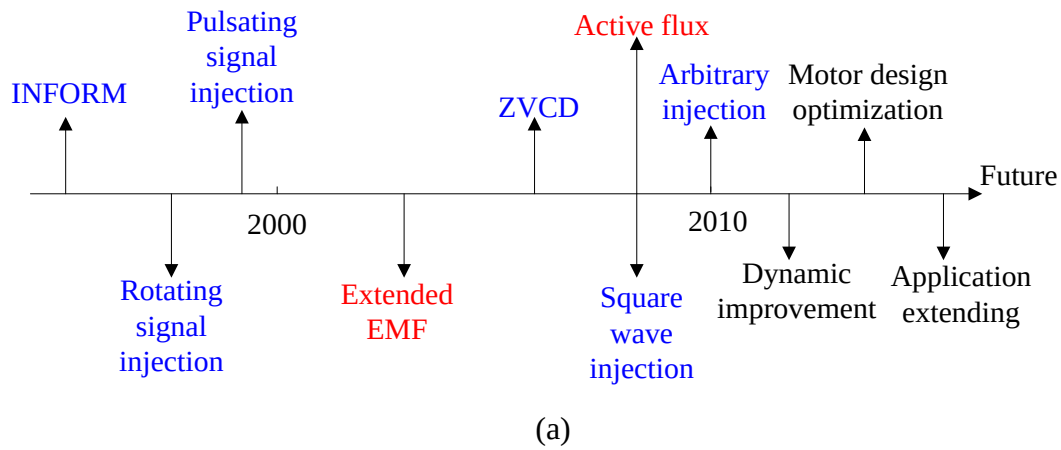


Fig. 1.1. Sensorless control methods. (a) Development history. (b) Classifications.

The fundamental model based methods exploit the back EMF or flux linkages induced by permanent magnet (PM), while the saliency based methods detect the variation of inductance or resistance caused by geometry structure or saturation. Methods that are based on fundamental frequency model can be further divided into the back-EMF-based schemes [MAT96], the flux-linkage-based schemes [HU98] and the 3rd harmonics based methods [LIU14a]. These methods have satisfactory performance at medium and high speeds, but may fail at standstill and low speed due to low signal to noise ratio of the back-EMF signals. The saliency generally appears as spatial inductance variation caused by machine geometry and magnetic saturation, or spatial resistance variation induced by eddy current.

Spatial saliency can be detected by additional excitation signal injection or just making use of the PWM excitations. The injected signals can be continuous waveforms, such as sinusoidal rotating, sinusoidal or square wave pulsating signals. Discrete signal injection based sensorless methods inject voltage pulses or make use of the current variations during PWM voltage vectors. The indirect flux detection by on-line reactance measurement (INFORM) method measures the current responses induced by the vectors which are applied in different positions [SCH96]. The pulse injection method injects two voltage pulses in opposite directions and extracts the current variations for position estimation [WAN17a]. The zero voltage vector current derivative (ZVCD) method calculates the rotor position by measuring the current variations during zero voltage vectors. The arbitrary injection method estimates the position from the differences between the measured currents and the calculated model currents.

In actual applications, the saliency model based methods and fundamental model based methods are usually combined to provide whole speed range sensorless operation [YAN17], [XIA20], [YOU18].

1.2.1 Fundamental model based sensorless methods

The back EMF and flux linkages induced by permanent magnet can be used to observe the rotor position as shown in Fig. 1.2.

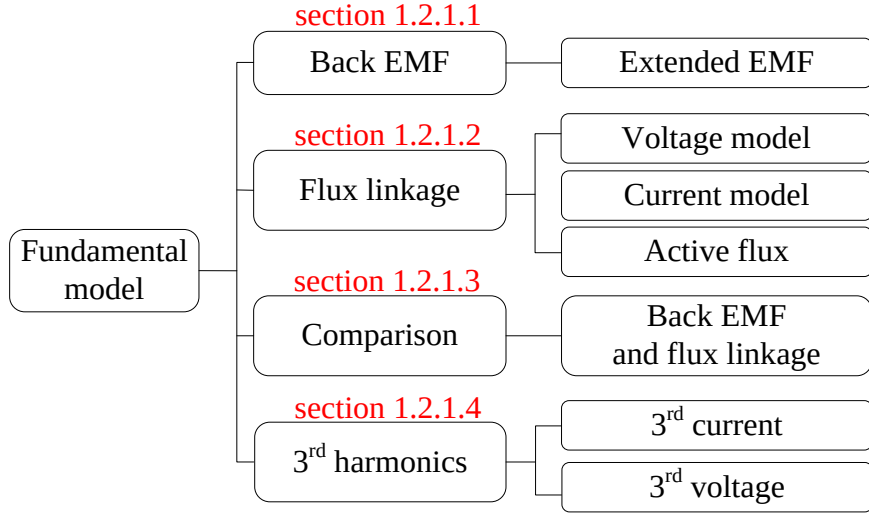


Fig. 1.2. Fundamental model based sensorless control methods.

1.2.1.1 Back EMF

(a) Machines with non-salient pole

The voltage equations of PMSM with non-salient pole are given by

$$\begin{bmatrix} u_d \\ u_q \end{bmatrix} = \begin{bmatrix} R_s + L_s p & -\omega_e L_s \\ \omega_e L_s & R_s + L_s p \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \omega_e \begin{bmatrix} 0 \\ \psi_{pm} \end{bmatrix} \quad (1.1)$$

where R_s is the phase resistance, L_s is the phase inductance, ω_e is the rotor electrical angular speed, p is the differential operator, ψ_{pm} is the permanent magnet (PM) flux-linkage, u_d , u_q are the dq -axis voltages, i_d , i_q are the dq -axis currents, respectively. The cross-coupling inductance is not considered in this model.

For sensorless rotor position estimation based on this model, as the actual rotor position is not known, the voltages and currents are expressed in the estimated synchronous rotating reference frame $\hat{d}\hat{q}$ as shown in Fig. 1.3.

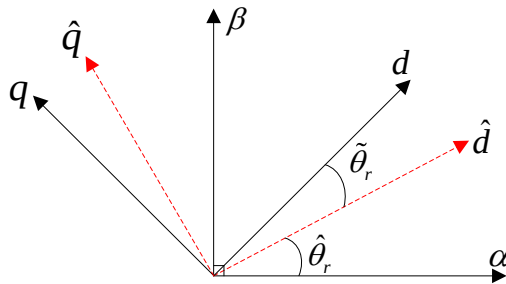


Fig. 1.3. Relationship between the actual dq and the estimated $\hat{d}\hat{q}$ frame.

$\hat{\theta}_r$ is the position of the estimated d -axis and $\tilde{\theta}_r$ is the error between the actual and estimated rotor positions. Voltage equations in the estimated frame are obtained from (1.1) by coordinate transformation as

$$\begin{bmatrix} u_d^{\hat{r}} \\ u_q^{\hat{r}} \end{bmatrix} = \begin{bmatrix} R_s + L_s p & -\omega_e L_s \\ \omega_e L_s & R_s + L_s p \end{bmatrix} \begin{bmatrix} i_d^{\hat{r}} \\ i_q^{\hat{r}} \end{bmatrix} + \begin{bmatrix} -\omega_e \psi_{pm} \sin(\tilde{\theta}_r) \\ \omega_e \psi_{pm} \cos(\tilde{\theta}_r) \end{bmatrix}, \tilde{\theta}_r = \theta_r - \hat{\theta}_r \quad (1.2)$$

where “ $\hat{\cdot}$ ” means the estimated reference frame. As can be seen from (1.2), the term $-\omega_e \psi_{pm} \sin(\tilde{\theta}_r)$ is the back EMF on the estimated d -axis and is proportional to the position error when $\tilde{\theta}_r$ is very small [MAT96]. Based on this phenomenon, the rotor position can be estimated by controlling $\omega_e \psi_{pm} \sin(\tilde{\theta}_r) = 0$ with the phase-locked-loop (PLL) observer [YOO11] shown in Fig. 1.4.

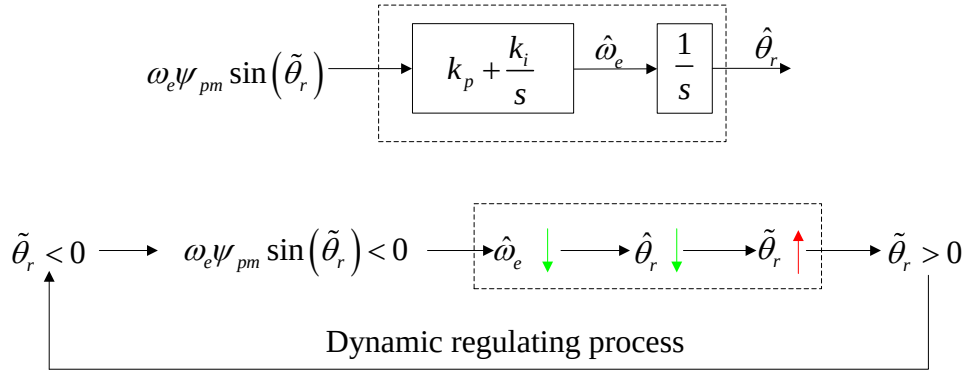


Fig. 1.4. PLL observer for estimating rotor position and its dynamic regulating process.

With this type of observer, the estimated position $\hat{\theta}_r$ is dynamically adjusted until $\omega_e \psi_{pm} \sin(\tilde{\theta}_r)$ is controlled to zero. The ripples in the estimated position due to sampling noise can also be suppressed due to the low pass nature of the observer.

(b) Machines with salient pole

The above sensorless position estimation method is only applicable to machines with non-salient pole. For machines with salient pole, the voltage equations in the dq -axis frame are given by

$$\begin{bmatrix} u_d \\ u_q \end{bmatrix} = \begin{bmatrix} R_s + L_d p & -\omega_e L_q \\ \omega_e L_d & R_s + L_q p \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \omega_e \begin{bmatrix} 0 \\ \psi_{pm} \end{bmatrix} \quad (1.3)$$

where L_d, L_q are the dq -axis apparent inductances. Due to asymmetry of the dq -axis inductances, the back EMF cannot be directly extracted in the estimated dq -axis frame. To solve this problem, the extended back EMF concept was proposed for machines with salient pole [MOR02] [CHE03] as

$$\begin{bmatrix} u_d^{\hat{r}} \\ u_q^{\hat{r}} \end{bmatrix} = \begin{bmatrix} R_s + L_d p & -\omega_e L_q \\ \omega_e L_q & R_s + L_d p \end{bmatrix} \begin{bmatrix} i_d^{\hat{r}} \\ i_q^{\hat{r}} \end{bmatrix} + E_{ex} \begin{bmatrix} -\sin(\tilde{\theta}_r) \\ \cos(\tilde{\theta}_r) \end{bmatrix} \quad (1.4)$$

$$E_{ex} = \omega_e \psi_{pm} + \omega_e (L_d - L_q) i_d + (L_q - L_d) p i_q \quad (1.5)$$

where E_{ex} is amplitude of the extended back EMF. By controlling $E_{ex} \sin(\tilde{\theta}_r) = 0$ with the same PLL observer in Fig. 1.4, the rotor position can also be obtained.

In actual applications, accuracy of the estimation is affected by many other nonlinear factors, such as inverter nonlinearity, back EMF harmonics and parameter variations. The voltages used for calculating the back EMF are usually different from the actual measured terminal voltages due to distortions caused by the inverter nonlinearities, including dead time, parasitic capacitance and conduction voltage drop [MUN99] [LIU14b]. This can be mitigated by applying various model based compensation strategies [ZHA04] or terminal voltage sampling technique [CHE20].

Due to spatial harmonics of the flux linkages, the back EMF contains $(6k \pm 1)$ harmonics [JUN10], [PAR14], in which the 5th and 7th harmonics are dominant. It is difficult to suppress the back EMF harmonic distortions without sacrificing the dynamic performance. In [ZHA16a], adaptive vector filters are used to compensate the back EMF harmonics. In [LIN16], adaptive notch filters are used to minimize the influence of asymmetric back EMF. As shown in (1.4), calculation of the back EMF requires stator resistance and inductances. These parameters vary with temperature, fundamental currents and manufacturing tolerance. The sliding mode observer for position estimation was employed accounting for the parameter variations and shows good performance [CHI09] [BER14]. Besides, the disturbance observer has been used with good performance in disturbance rejection and dynamic operation [PAR14] [XIA16]. The problems and solutions for back EMF based sensorless control methods are summarized in Fig. 1.5.

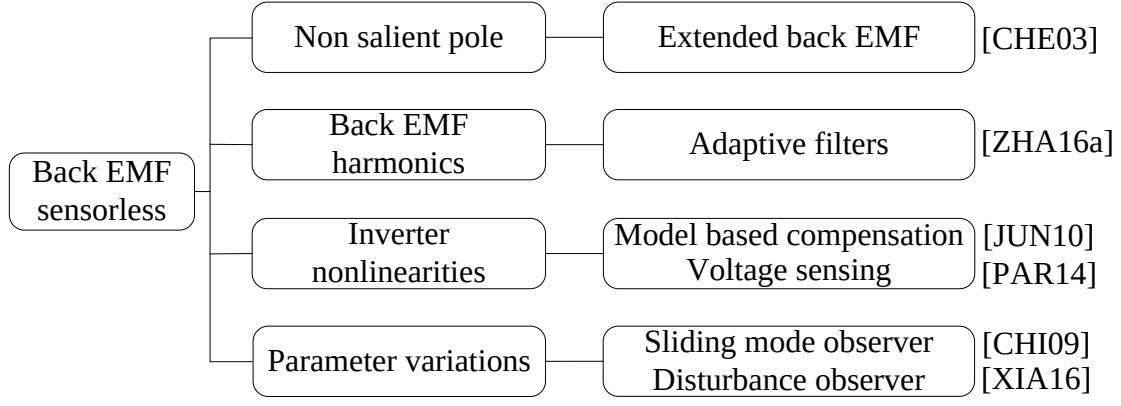


Fig. 1.5. Summary of back EMF based sensorless methods.

1.2.1.2 Flux linkages

The flux linkages induced by the permanent magnet can also be used for estimating the rotor position [SHE02], [BOL08].

(a) Machines with non-salient pole

The flux-linkages for machines with non-salient pole in the stationary reference frame are expressed as

$$\begin{bmatrix} \psi_\alpha \\ \psi_\beta \end{bmatrix} = \begin{bmatrix} L_s & 0 \\ 0 & L_s \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \psi_{pm} \begin{bmatrix} \cos(\theta_r) \\ \sin(\theta_r) \end{bmatrix} \quad (1.6)$$

where ψ_α, ψ_β and ψ_{pm} are the stator flux linkages and permanent flux linkage, θ_r is the rotor position, i_α, i_β are the two phase stationary stator currents, L_s is the phase inductances. The rotor position can be obtained as

$$\theta_r = \tan^{-1} \left(\frac{\psi_\beta - L_s i_\beta}{\psi_\alpha - L_s i_\alpha} \right) \quad (1.7)$$

The stator flux linkages can be calculated by integrating the back EMF voltages as

$$\psi_\alpha = \int (u_\alpha - i_\alpha R_s) dt \quad (1.8-a)$$

$$\psi_\beta = \int (u_\beta - i_\beta R_s) dt \quad (1.8-b)$$

where R_s is the phase resistance. Accuracy of the integration operation is affected by drift and voltage distortions [HU98]. If there are offsets in the voltages, resistance or measured currents, the calculated flux linkages would contain dc offset and become saturated due to integration.

$$\begin{bmatrix} u_\alpha \\ u_\beta \end{bmatrix} = (R_s + L_q p) \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + p \begin{bmatrix} \psi_\alpha \cos(\theta_r) \\ \psi_\alpha \sin(\theta_r) \end{bmatrix} = (R_s + L_q p) \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \begin{bmatrix} E_\alpha^a \\ E_\beta^a \end{bmatrix} \quad (1.12)$$

where $E_\alpha^a = p[\psi_\alpha \cos(\theta_r)]$, $E_\beta^a = p[\psi_\alpha \sin(\theta_r)]$ are the back EMF induced by the active flux linkage. The rotor position can be obtained from the voltages and currents in the two phase stationary frame as

$$\theta_r = \tan^{-1} \left(\frac{\psi_\beta - L_q i_\beta}{\psi_\alpha - L_q i_\alpha} \right) = \tan^{-1} \left(\frac{-(u_\alpha - R_s i_\alpha - L_q p i_\alpha)}{u_\beta - R_s i_\beta - L_q p i_\beta} \right) = \tan^{-1} \left(\frac{-E_\alpha^a}{E_\beta^a} \right) \quad (1.13)$$

As mentioned before, the voltage integrations for calculating the stator flux linkages suffer from the dc offset and parameter variations, especially at standstill and low speed. Alternatively, the flux linkages can be calculated with the current model as

$$\psi_d = \psi_{pm} + L_d i_d \quad (1.14-a)$$

$$\psi_q = L_q i_q \quad (1.14-b)$$

The current model based flux linkage observer requires permanent flux linkage ψ_{pm} , dq -axis inductances L_d, L_q which vary with i_d, i_q . Besides, a rotor position sensor (or sensorless algorithm) is essential to obtain the dq -axis positions. The current model is not dependent on the speed and thus is of better performance at low speed compared to the voltage model.

(c) Hybrid flux linkage observer

To improve the performance of flux linkage observer, a hybrid flux linkage observer combining the voltage model and current model, which is similar to the closed-loop hybrid [JAN94] flux observer for an induction machine is used as shown in Fig. 1.7(a).

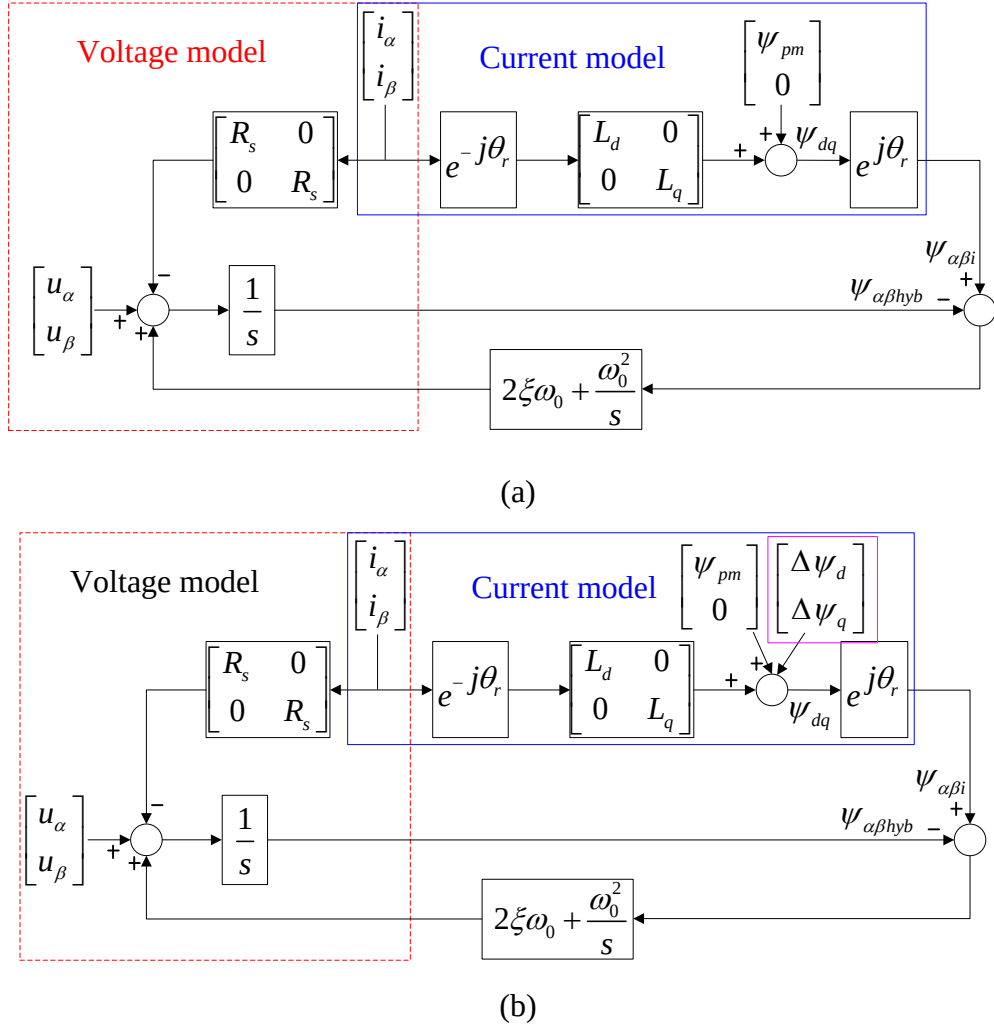


Fig. 1.7. Hybrid flux observer. (a) Existing flux observer. (b) Modified flux observer accounting for the dq-axis flux variations.

where $\psi_{\alpha\beta hyb}$ are the hybrid observed flux linkages. As can be seen that the hybrid flux can be expressed from $\psi_{\alpha\beta i}$ obtained from current model and $\psi_{\alpha\beta v}$ from voltage model as

$$\psi_{\alpha\beta hyb} = \frac{s^2}{s^2 + 2\xi\omega_0 s + \omega_0^2} \psi_{\alpha\beta v} + \frac{2\xi\omega_0 s + \omega_0^2}{s^2 + 2\xi\omega_0 s + \omega_0^2} \psi_{\alpha\beta i} \quad (1.15)$$

The feedback compensator damping ratio ξ and crossover frequency ω_0 determine the time domain characteristics of the observer. At standstill and low speed, the current model is dominant, while at high speed the voltage model is dominant. As the current model is more sensitive to parameter variations, the estimation accuracy may be deteriorated below the crossover frequency due to flux linkage variations.

In [YOO09], a modified flux observer is proposed by calculating the flux variations with the output voltages of the current regulators as shown in Fig. 1.7(b) where the flux errors $\Delta\psi_d, \Delta\psi_q$ can be obtained from the current PI output v_{dPI}, v_{qPI} as

$$\begin{bmatrix} \Delta\psi_d \\ \Delta\psi_q \end{bmatrix} = \begin{bmatrix} \psi_d - \hat{\psi}_d \\ \psi_q - \hat{\psi}_q \end{bmatrix} = \frac{1}{\omega_e} \begin{bmatrix} 0 & 1 \\ -1 & 0 \end{bmatrix} \left(\begin{bmatrix} v_{dPI} \\ v_{qPI} \end{bmatrix} - \begin{bmatrix} R_s & 0 \\ 0 & R_s \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} \right) \quad (1.16)$$

where $\hat{\psi}_d, \hat{\psi}_q$ are the estimated dq -axis flux linkages. (1.16) can be derived from the steady state voltage equations, neglecting the current derivatives as

$$\begin{bmatrix} R_s & 0 \\ 0 & R_s \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 & -\omega_e \\ \omega_e & 0 \end{bmatrix} \begin{bmatrix} \psi_d \\ \psi_q \end{bmatrix} = \begin{bmatrix} v_{dPI} \\ v_{qPI} \end{bmatrix} + \begin{bmatrix} 0 & -\omega_e \\ \omega_e & 0 \end{bmatrix} \begin{bmatrix} \hat{\psi}_d \\ \hat{\psi}_q \end{bmatrix} \quad (1.17)$$

$-\omega_e \hat{\psi}_d, \omega_e \hat{\psi}_q$ are the feedforward decoupling voltages that are added to the PI outputs.

Similar to the back EMF, harmonics may also exist in the flux linkages and result in ripples in the estimated results. The compensation methods for back EMF model are also applicable for flux linkage model as shown in Fig. 1.8.

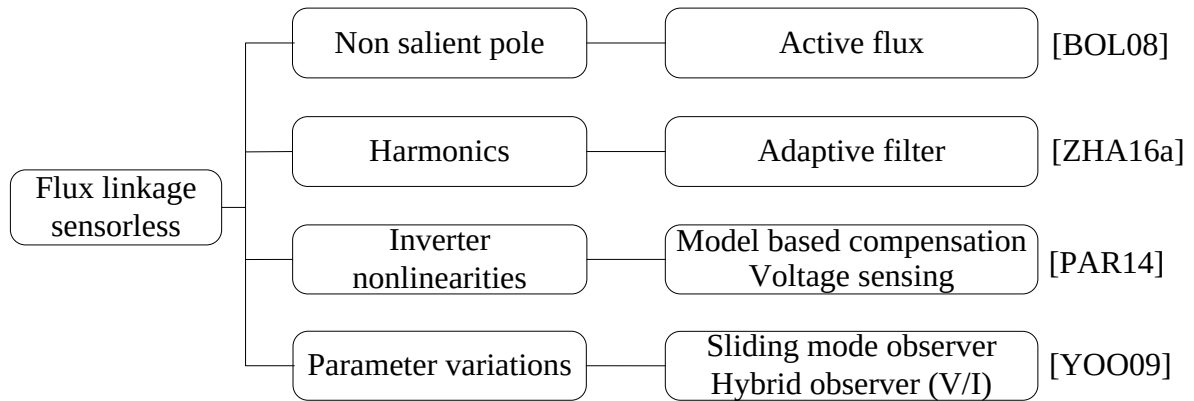


Fig. 1.8 Summary of flux linkage based sensorless methods.

1.2.1.3 Comparison between back EMF and flux linkage methods

Back EMF and flux linkage methods for position estimation are all based on the fundamental electrical machine model. They can both be expressed in the two phase stationary reference frame, i.e. $\alpha\beta$ frame, and the two phase synchronous rotating reference frame, i.e. dq frame. Comparisons between the back EMF model and flux linkage model are made as follows.

(a) Comparison between back EMF and flux linkage methods

In previous section, the back EMF and flux linkage based sensorless methods are introduced. It is interesting and meaningful to compare these two methods as shown in Table 1.1.

Table 1.1 Comparison between back EMF and flux linkage based sensorless methods

	Back EMF	Flux linkage
General equations	$e_x = u_x - R_s i_x - L_y \frac{di_x}{dt}$	$\psi_x = \int (u_x - R_s i_x) dt - L_y i_x$
Calculation	Differential	Integral
Parameters required	Resistance, inductance	Resistance, inductance
Robust to noise	Low	High
Dynamic	High	Low

where x,y refer to the corresponding reference frames, i.e., α , β , d , q . The flux linkage based method calculation is more robust to noise compared to the back EMF based method as the integration behaves as a low pass filter to the noise, while differential operation may introduce noise for the back EMF based method. However, the integration for the flux linkage based method also introduces delay and results in decreased dynamic performance.

(b) Comparison between extended back EMF and active flux

The extended back EMF [CHE03] and active flux [BOL08] are similar in that they transform a machine with salient pole to a machine with non-salient pole as presented before. This largely simplifies the position estimation for IPMSM and enables a unified sensorless theory in medium and high speeds for ac machines. However, there are also differences between them. The active flux concept refers to the torque-producing flux in the electromagnetic torque formulas of ac machine [BOL08], while the extended back EMF concept does not have clear physical meaning. Detailed comparison is shown in Table 1.2.

Table 1.2 Comparison between extended back EMF and active flux

	Extended back EMF(ex)	Active flux(a)
Definition	$E_{ex} = \omega_e \psi_f + \omega_e (L_d - L_q) i_d + (L_q - L_d) p i_q$	$\vec{\psi}_a = \vec{\psi}_s - L_q \vec{I}_s = \vec{\psi}_f + (L_d - L_q) \vec{i}_q = \psi_a e^{j\theta_r}$
Dq-frame voltages	$\vec{u}_{dq}^{ex} = (R_s + L_d p) \vec{i}_{dq} + j\omega_e L_q \vec{i}_{dq} + \vec{E}_{dq}^{ex}$	$\vec{u}_{dq}^a = (R_s + L_q p) \vec{i}_{dq} + j\omega_e L_q \vec{i}_{dq} + \vec{E}_{dq}^a$
Dq-frame back EMF	$\vec{E}_{dq}^{ex} = jE_{ex}$ (d-axis back EMF is zero)	$\vec{E}_{dq}^a = p\vec{\psi}_a = (L_d - L_q) p i_d + j\omega_e \vec{\psi}_a $ (d-axis back EMF is nonzero)
$\alpha\beta$-frame voltages	$\vec{u}_{\alpha\beta}^{ex} = (R_s + L_d p) \vec{i}_{\alpha\beta} + j\omega_e (L_q - L_d) \vec{i}_{\alpha\beta} + \vec{E}_{\alpha\beta}^{ex}$	$\vec{u}_{\alpha\beta}^a = (R_s + L_q p) \vec{i}_{\alpha\beta} + \vec{E}_{\alpha\beta}^a$
$\alpha\beta$-frame back EMF	$\vec{E}_{\alpha\beta}^{ex} = E_{ex} [-\sin(\theta_r) + j \cos(\theta_r)]$ (aligned with q-axis/ no additional back EMF vector)	$\vec{E}_{\alpha\beta}^a = \omega_e \vec{\psi}_a [-\sin(\theta_r) + j \cos(\theta_r)] - (L_q - L_d) e^{j\theta_r} p i_d$ (not aligned with q-axis additional back EMF vector $-(L_q - L_d) e^{j\theta_r} p i_d$)

The main difference between the extended back EMF and active flux is that the extended back EMF $\vec{E}_{dq}^{ex}$ is aligned with the q-axis as shown in Fig. 1.9(a), while the active flux model based back EMF $\vec{E}_{dq}^a$ has an additional back EMF component $(L_d - L_q) p i_d$ on the d-axis as shown in Fig. 1.9(b).

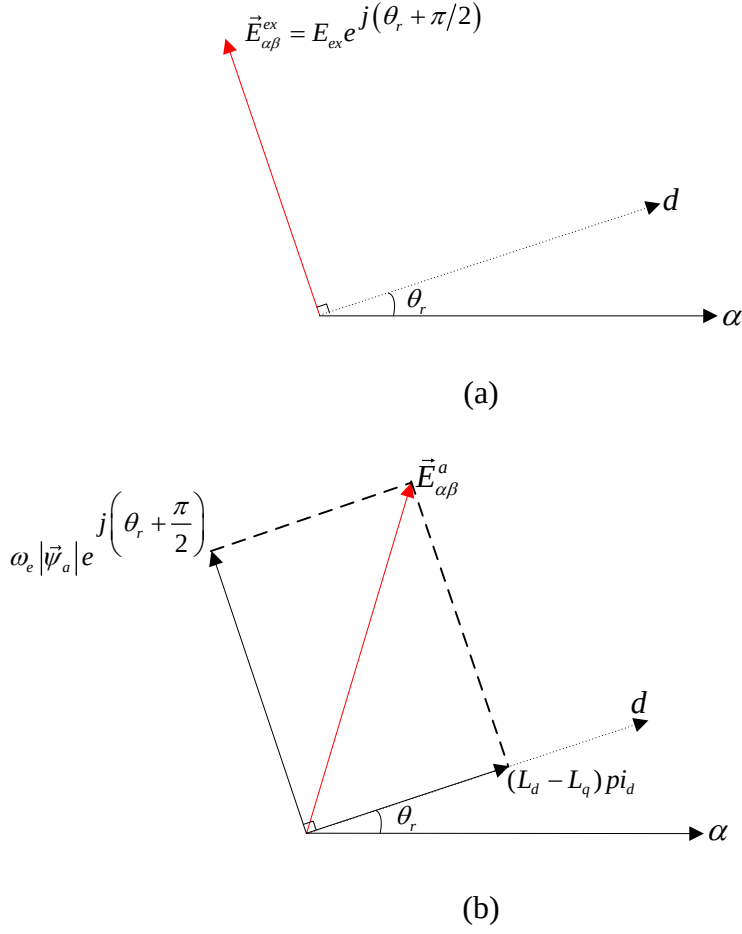


Fig. 1.9 Vector diagram for extended back EMF and active flux in $\alpha\beta$ -frame. (a) Extended back EMF. (b) Active flux.

In [XIA20], the extended back EMF is used for unified sensorless control over the whole speed range. The presence of pi_q in E_{ex} extends the possibility of position estimation to the low speed region with HF q -axis excitation. Similarly, the active flux concept is used for low speed sensorless control in [PES19] due to the existence of pi_d in the back EMF, which guarantees sufficient amplitude of the active flux signal.

1.2.1.4 Third harmonics

Conventionally, the back EMF harmonics are disturbances to the fundamental model based sensorless methods. However, as the harmonic component also contains rotor information, it can be utilized for sensorless control if the magnitude is high enough. The 3rd harmonic back EMF as shown in Fig. 1.10 is usually used in sensorless control in both brushless DC (BLDC) [SHE06] and brushless AC (BLAC) machine drives [LIU14a].

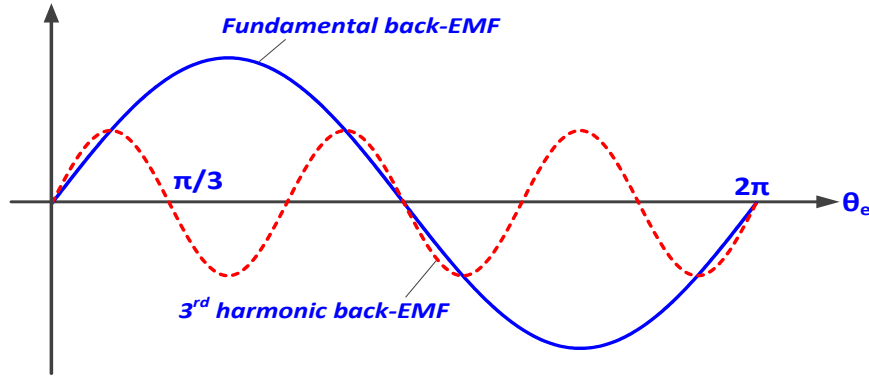


Fig. 1.10. Third harmonic back EMF waveform.

The 3rd harmonic back EMF can be measured by detecting the voltage difference between the neutral point of the stator windings and the central point of the Y-connected balanced resistance network which is in parallel with the three phase terminals [SHE06]. The detected discrete zero crossings can be used for high speed BLDC control [SHE04] or for continuous position output [LIU14a]. As the frequency of the PWM harmonics is much higher than the fundamental frequency, the noise caused by common mode PWM harmonics can be filtered by low pass filters or a band pass filters. Besides, as the full range of the voltage sensors and A/D converters can be utilized for sampling the third harmonics, it has high signal to noise ratio. The 3rd harmonic back EMF based method is less sensitive to pulse width modulation noise and measurements of winding resistance. Although the 3rd harmonic back EMF based sensorless method can achieve good results in some applications, it is only applicable for machines with the 3rd harmonics and requires additional wires and cost for accessing the neutral point.

Instead of directly measuring the 3rd harmonic back EMF using voltage sensor for position detection, the 3rd harmonic current (also called zero sequence current) induced by the 3rd harmonic back EMF can also be utilized. A premise for utilizing the zero sequence current is that zero sequence current can flow in the windings. For a general three phase machine without

neutral point connection, zero sequence current cannot circulate. Whereas in a machine with three phase separate windings (open winding configuration) [ZHA16b] [ZHA17], the zero sequence current can be made use of.

As zero sequence current can circulate in the three phase open windings, the zero sequence current can be sampled and used to reconstruct the 3rd harmonic back EMF waveform as shown in Fig. 1.11(a). Alternatively, the 3rd harmonic back EMF can also be seen as equal to the applied zero sequence voltage if the zero sequence current is regulated to zero by open-winding controllers as demonstrated in Fig. 1.11(b). The two approaches detect the 3rd harmonic back EMF both by utilizing the zero sequence current, but are implemented in different ways.

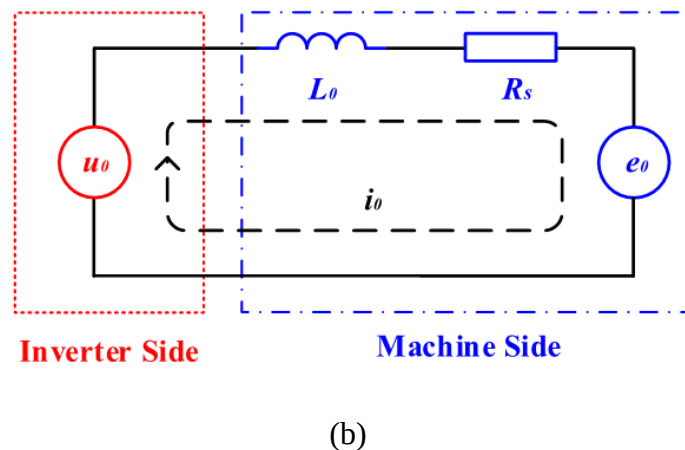
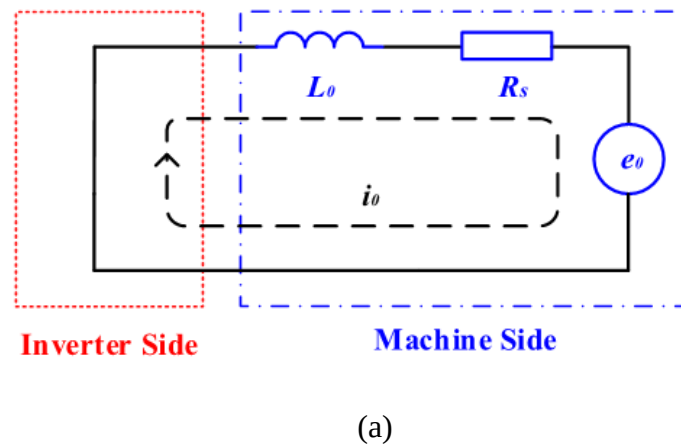


Fig. 1.11 Two ways of utilizing 3rd harmonic for position estimation in machines with open winding configuration. (a) Zero sequence current circulating. (b) Zero sequence current not circulating.

1.2.2 Saliency model based sensorless methods with continuous signal injection

Instead of using fundamental model for position estimation, the saliency variation caused by structure and magnetic saturation can also be used for position estimation. The saliency information can be detected by injecting various test signals, i.e. continuous high frequency signal, discrete voltage signal as shown in Fig. 1.12.

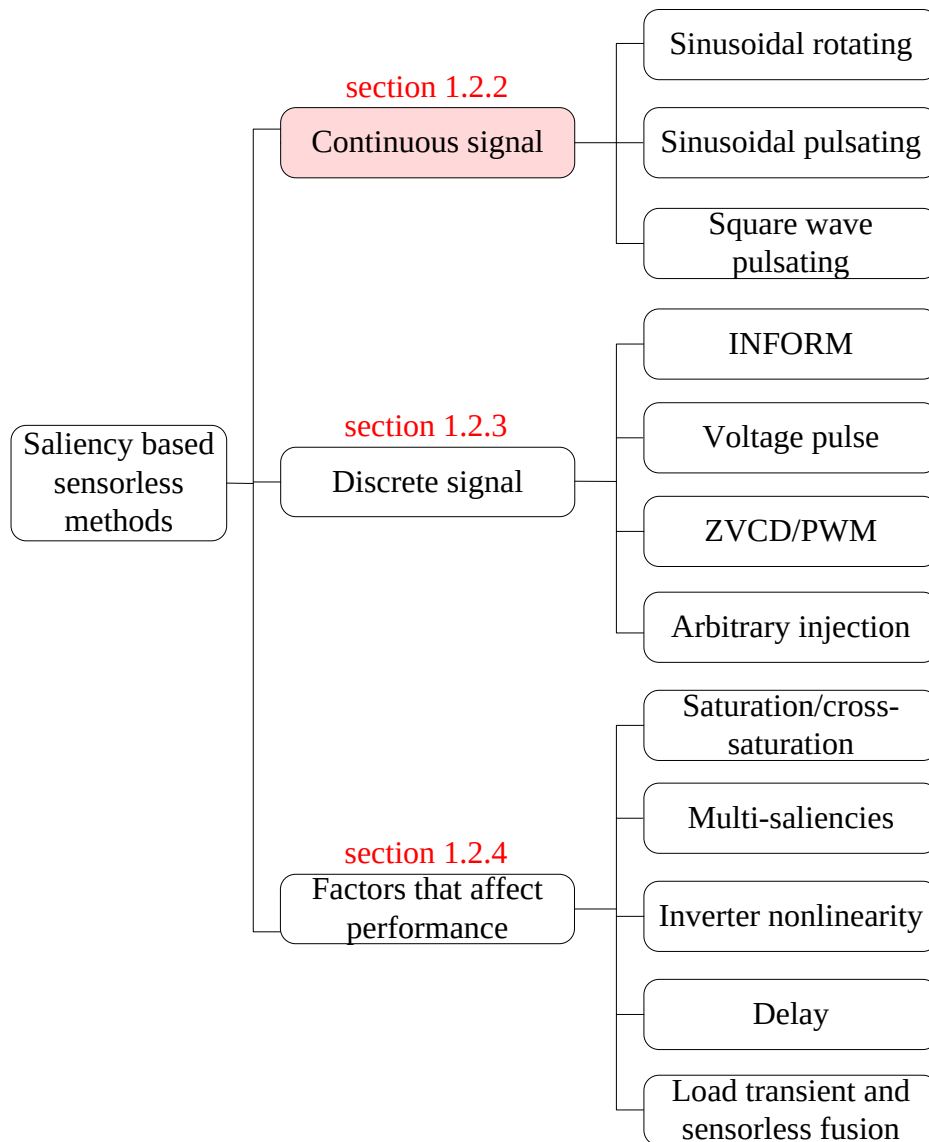
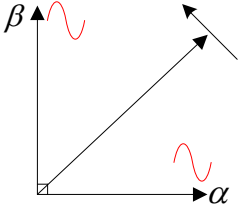
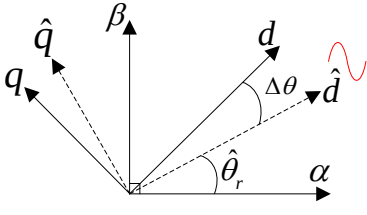
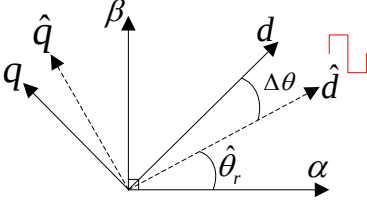
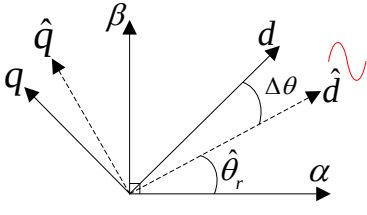
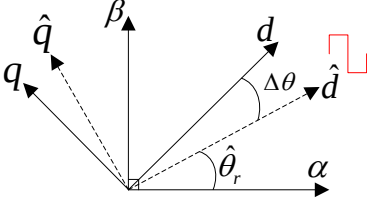


Fig. 1.12. Category of saliency tracking based sensorless methods.

In this section, the continuous signal injection based sensorless control methods will be introduced. The widely used signals are HF sinusoidal rotating signal [JAN95], sinusoidal pulsating signal [HA01] and pulsating square wave voltage signal [YOO11] as shown in Table 1.3. The rotating signal is injected to the stationary reference frame, while the pulsating signal is injected to the estimated d -axis.

Table 1.3 Conventional HF continuous signal injection based sensorless control methods

Injection signal		Injection frame	Demodulation signal
Rotating	Sinusoidal		HF current
	HF voltage		HF voltage
Pulsating	HF current		Zero sequence voltage
	[SEI15]		[XU16]
	Sinusoidal		HF current
	HF voltage		HF voltage
Pulsating	HF voltage		Zero sequence voltage
	HF current		Flux linkage [GUG06]
Pulsating	HF square wave voltage		HF square wave current
	[YOO11]		Zero sequence voltage
Others		Novel injection frames Novel injection signals Novel demodulation strategies	

As the synthesized injected HF voltage is a rotating space vector or a pulsating space vector in the vector diagram, it is therefore called “rotating injection” or “pulsating injection”. The waveform of the pulsating signal can be sinusoidal or square wave. The injected signal can be high frequency voltage or current. The response signals used to demodulate the rotor position, can be high frequency current, high frequency voltage, high frequency flux linkage [GUG06] or zero sequence voltage [XU16]. Voltage injection is easy to implement by adding the HF voltage command to the reference, while injecting current requires closed-loop control of the HF currents [BAS11]. The resulting HF currents or voltages are extracted for position estimation if HF voltages or currents are injected. The zero sequence HF voltages which contain position information can also be utilized for position estimation [ALM17], [XU16].

The techniques mentioned above measure responses of the injected signals and extract the rotor position with various demodulation methods. Hence, the saliency tracking based sensorless method can be seen as system with three stages as shown in Fig. 1.13. They are injection signal, injection frame, and demodulation method. The first part specifies what

types of signals are injected. The second part defines where the signals are injected and the last part provides solutions of how to demodulate the rotor position from the responses.

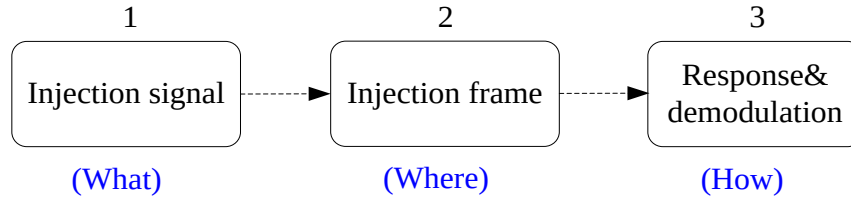


Fig. 1.13. General structure of HF continuous signal injection based sensorless control methods.

Except the three basic stages, there are many other factors that affect the estimation performance, such as magnetic saturation, inverter nonlinearities and delay. They are also very important for improving the sensorless performances, i.e. accuracy, convergence. These issues will be introduced in detail later. In the next section, the basic principles for sensorless control will be presented first.

1.2.2.1 HF sinusoidal rotating signal injection based sensorless control

The HF voltage equations of IPMSM in the synchronous rotating reference frame (SRRF) at standstill and low speed are given by

$$\begin{bmatrix} u_{dh}^r \\ u_{qh}^r \end{bmatrix} = \begin{bmatrix} L_{dh} & 0 \\ 0 & L_{qh} \end{bmatrix} P \begin{bmatrix} i_{dh}^r \\ i_{qh}^r \end{bmatrix} \quad (1.18)$$

where $u_{dh}^r, u_{qh}^r, i_{dh}^r, i_{qh}^r$ are the HF voltages and currents in the SRRF, L_{dh}, L_{qh} are the dq -axis incremental inductances. Here, it is noted that the incremental inductance is different from the apparent inductances used in fundamental model based sensorless control. The apparent inductance at a specific working point is defined as the ratio between the flux linkage increment from zero current to the specific working current point and the current, while the incremental inductance is the slop of the tangent line at a specific working current point on the flux linkage-current characteristic. In this model, the resistance is neglected due to low magnitude compared to the HF reactance. At standstill and low speed, the back EMF is also not considered due to low amplitude.

(a) Sensorless position estimation using negative sequence current

For analysing the rotating injection in the stationary reference frame, voltage equations in the stationary synchronous reference frame (SSRF) are as

$$\begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} = \begin{bmatrix} L_{\Sigma} - L_{\Delta} \cos(2\theta_r) & -L_{\Delta} \sin(2\theta_r) \\ -L_{\Delta} \sin(2\theta_r) & L_{\Sigma} + L_{\Delta} \cos(2\theta_r) \end{bmatrix} P \begin{bmatrix} i_{\alpha h}^s \\ i_{\beta h}^s \end{bmatrix} \quad (1.19)$$

By injecting HF voltages as (1.20), which are superimposed on the fundamental voltages, the HF currents in the SSRF are calculated from (1.19), (1.20) as (1.21).

$$\begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} = u_h \begin{bmatrix} \cos(\theta_h) \\ \sin(\theta_h) \end{bmatrix}, \theta_h = \omega_h t \quad (1.20)$$

$$i_{\alpha h}^s + ji_{\beta h}^s = I_p e^{j(\theta_h - \pi/2)} + I_n e^{-j(\theta_h - 2\theta_r - \pi/2)} \quad (1.21)$$

$$I_p = \frac{(L_{qh} + L_{dh})u_h}{2\omega_h L_{dh} L_{qh}}, I_n = \frac{(L_{qh} - L_{dh})u_h}{2\omega_h L_{dh} L_{qh}} \quad (1.22)$$

where u_h, θ_h, ω_h are the amplitude, angle and frequency of the injected signal. The responses are composed of positive sequence current (PSC) and negative sequence current (NSC). Only NSC contains rotor position information. To extract position from the NSC, the demodulation strategies as shown in Fig. 1.14 are usually used.

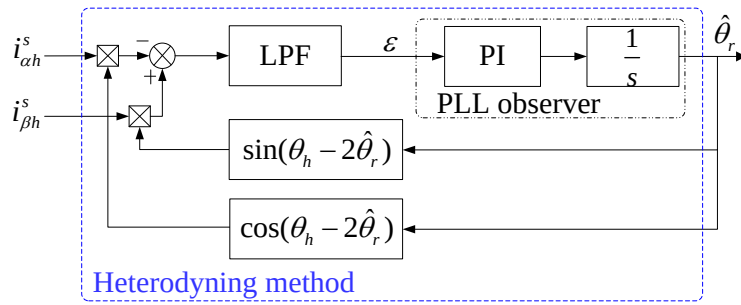
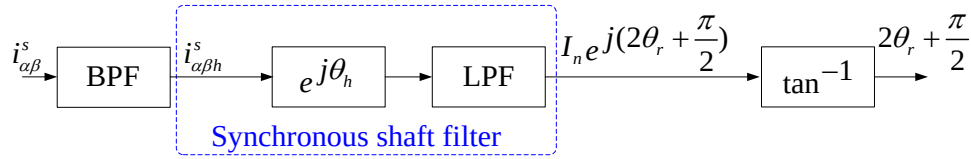


Fig. 1.14. Methods for demodulating rotor position based on HF rotating injection. (a) Synchronous shaft filter. (b) Heterodyning method.

The BPF is used to separate the fundamental currents and the HF currents. The synchronous shaft filter (SSF) moves positive sequence current to HF domain and filters it out, leaving NSC. However, direct inverse tangent calculation may introduce much noise. Thus, the heterodyning

method as shown in Fig. 1.14(b) is used to get smooth position output. The LPF is used to remove the HF component after the heterodyning operation. The rotor position can be obtained by controlling $\varepsilon=0$ with a PLL observer, i.e.

$$\varepsilon = I_n \sin \left[2(\theta_r - \hat{\theta}_r) \right] \quad (1.23)$$

When the position error $\tilde{\theta}_r = \theta_r - \hat{\theta}_r$ is small, ε is approximated as $2I_n(\theta_r - \hat{\theta}_r)$. Then, the transfer function of the position observer in Fig. 1.14(b) is given by

$$\begin{aligned} \frac{\hat{\theta}_r}{\theta_r} &= \frac{2I_n k_p s + 2I_n k_i}{\tau s^3 + s^2 + 2I_n k_p s + 2I_n k_i} \\ &\approx \frac{2I_n k_p s + 2I_n k_i}{s^2 + 2I_n k_p s + 2I_n k_i} \text{ (neglecting LPF)} \end{aligned} \quad (1.24)$$

The coefficients τ, k_p, k_i are the time constants of the LPF, proportional and integral parameters of the PI. The criteria for reliable convergence is that the error signal ε should have positive slope near the point $\tilde{\theta}_r = 0$ [YAN17]. The transfer function diagram of the position observer is shown in Fig. 1.15(a). The error signal ε with position error $\tilde{\theta}_r$ is plotted in Fig. 1.15(b) with the highlighted stable regions.

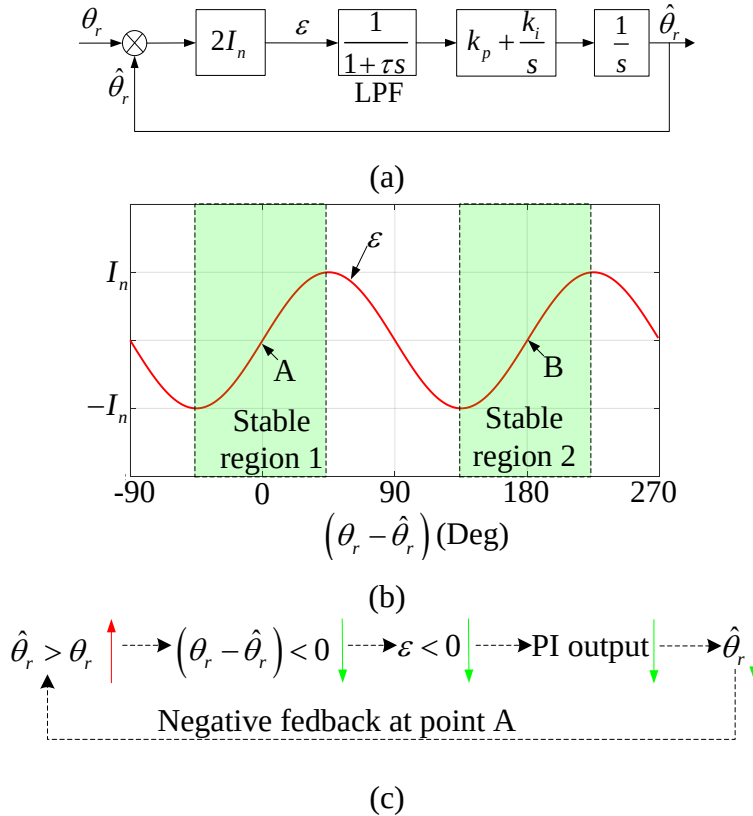


Fig. 1.15. Closed-loop control block diagram of the heterodyning method. (a) Transfer block. (b) Stable region for convergence. (c) Dynamic process for convergence.

It is observed that A and B are stable convergence points due to positive slopes of the error signal. For instance, dynamic process of convergence at point A is shown in Fig. 1.15(c). In one electrical period, there are two convergence points. The initial polarity can be detected by making use of different magnetic saturation levels at the north pole and south pole, and secondary saliencies, which will be introduced later. The dynamic performance of the position estimation performance is dependent on I_n which varies with the working currents. Thus, sufficient saliency level (or I_n) should be guaranteed for reliable convergence.

(b) Sensorless position estimation using zero sequence voltage

Except making use of the HF current response for position estimation, the zero sequence voltage can be measured by the circuit in Fig. 1.16 and utilized for position estimation. It can be seen that access to the neutral point is required.

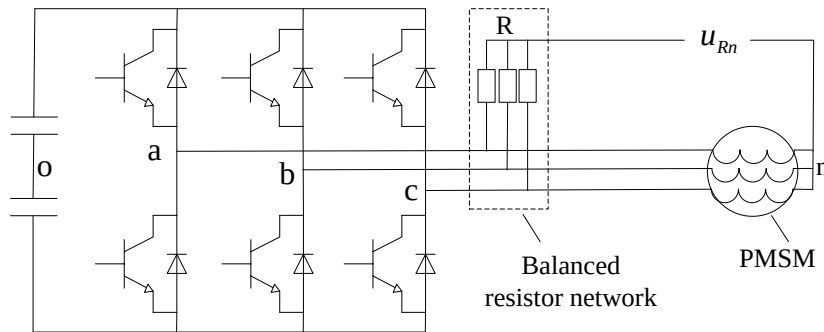


Fig. 1.16. Circuit for extracting the zero sequence HF voltage u_{Rn} for sensorless position estimation.

The three phase voltage equations neglecting resistances and back EMF can be given by

$$\begin{bmatrix} u_{an} \\ u_{bn} \\ u_{cn} \end{bmatrix} = \begin{bmatrix} L_{aa} & M_{ab} & M_{ac} \\ M_{ba} & L_{bb} & M_{bc} \\ M_{ca} & M_{cb} & L_{cc} \end{bmatrix} \begin{bmatrix} di_a/dt \\ di_b/dt \\ di_c/dt \end{bmatrix} \quad (1.25)$$

where $L_{aa}, L_{bb}, L_{cc}, M_{ab}(M_{ba}), M_{ac}(M_{ca}), M_{bc}(M_{cb})$ are the incremental self and mutual inductances respectively as

$$\begin{aligned} L_{aa} &= L_1 - L_2 \cos(2\theta_r) \\ L_{bb} &= L_1 - L_2 \cos(2\theta_r + 2\pi/3) \\ L_{cc} &= L_1 - L_2 \cos(2\theta_r - 2\pi/3) \\ M_{ab} &= M_{ba} = M_1 - M_2 \cos(2\theta_r - 2\pi/3) \\ M_{bc} &= M_{cb} = M_1 - M_2 \cos(2\theta_r) \\ M_{ca} &= M_{ac} = M_1 - M_2 \cos(2\theta_r + 2\pi/3) \end{aligned} \quad (1.26)$$

where L_1, L_2, M_1, M_2 are the amplitudes of dc and second harmonics of HF self and mutual inductances. The three phase balanced voltages u_{ao}, u_{bo}, u_{co} are given by

$$\begin{bmatrix} u_{ao} \\ u_{bo} \\ u_{co} \end{bmatrix} = \begin{bmatrix} u_{an} + u_{nR} + u_{Ro} \\ u_{bn} + u_{nR} + u_{Ro} \\ u_{cn} + u_{nR} + u_{Ro} \end{bmatrix} \quad (1.27)$$

For HF rotating voltage injection to the stationary frame [XU16], the injected three phase voltages are expressed in (1.28). The zero sequence voltage u_{Rn} can be obtained from (1.25)-(1.27) as (1.29).

$$\begin{aligned} u_{ao} &= u_h \cos(\omega_h t) \\ u_{bo} &= u_h \cos(\omega_h t - 2\pi/3) \\ u_{co} &= u_h \cos(\omega_h t + 2\pi/3) \end{aligned} \quad (1.28)$$

$$u_{Rn} = V_{o1} \cos(\omega_h t + 2\theta_r) - V_{o2} \cos(\omega_h t - 4\theta_r) \quad (1.29)$$

$$V_{o1} = \frac{u_h (L_0 - M_0)(L_2 - M_2)}{2L_0^2 + 2M_0^2 - 4L_0M_0 - 2L_2M_2 - L_2^2/2}, \quad (1.30-a)$$

$$V_{o2} = \frac{u_h (L_2 - 2M_2)(L_2 - M_2)}{2L_0^2 + 2M_0^2 - 4L_0M_0 - 2L_2M_2 - L_2^2/2} \quad (1.30-b)$$

The zero sequence voltage contains rotor position. Due to higher amplitude of V_{o1} compared to V_{o2} , the first term in (1.29) is used for position estimation by controlling $\varepsilon=0$, i.e.

$$\varepsilon = LPF \left[2u_{Rn} \sin(\omega_h t + 2\hat{\theta}_r) \right] = V_{o1} \sin(2\theta_r - 2\hat{\theta}_r) \quad (1.31)$$

As can be seen in (1.30-a), V_{o1} is not dependent on the injection frequency, while the amplitude of negative sequence current in (1.22) is inversely proportional to the injection frequency. Thus, compared to the method utilising negative sequence current for position estimation, higher injection frequency can be injected for the zero sequence voltage method. For zero sequence voltage based method, it allows the full range of the sensors and A/D converters to be used for sampling the zero sequence voltage and improves the signal to noise ratio. However, for conventional HF currents based method, the current sensors and A/D converters are scaled to measure the maximum fundamental currents magnitude. This means the full range of the sensors and A/D converters cannot be fully utilized for measuring the HF currents. Therefore, the zero sequence voltage method is less sensitive to noise compared to the HF current based method.

1.2.2.2 HF sinusoidal/square wave pulsating signal injection

For HF pulsating signal injection, the signal is usually injected to the estimated d -axis as shown in Fig. 1.17. Injection in q -axis is not preferred as HF current on q -axis will produce HF permanent magnet torque ripple. The voltage equations in the estimated synchronous rotating reference frame (ESRRF) can be expressed as

$$\begin{bmatrix} u_{dh}^{\hat{r}} \\ u_{qh}^{\hat{r}} \end{bmatrix} = R^{-1}(\theta_r - \hat{\theta}_r) \begin{bmatrix} L_{dh} & 0 \\ 0 & L_{qh} \end{bmatrix} R(\theta_r - \hat{\theta}_r) p \begin{bmatrix} i_{dh}^{\hat{r}} \\ i_{qh}^{\hat{r}} \end{bmatrix}, \quad (1.32)$$

$$R(x) = \begin{bmatrix} \cos(x) & \sin(x) \\ -\sin(x) & \cos(x) \end{bmatrix}$$

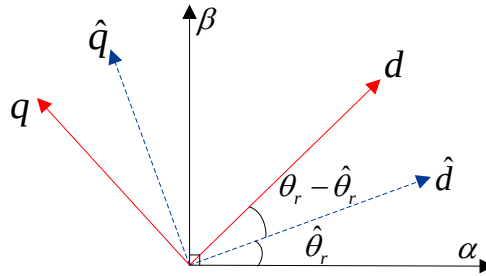


Fig. 1.17. Pulsating signal injection frame.

where $u_{dh}^{\hat{r}}, u_{qh}^{\hat{r}}, i_{dh}^{\hat{r}}, i_{qh}^{\hat{r}}$ are the HF voltages and currents in the ESRRF, $\tilde{\theta}_r = \theta_r - \hat{\theta}_r$ is the angle between the actual d -axis and the estimated d -axis, the superscript “ $\hat{r}$ ” means the estimated reference frame. The injected signal is usually sinusoidal wave or square wave as follows.

(a) Sinusoidal pulsating voltage waveform injection

By injecting pulsating HF voltage as (1.33) to the stator winding, the HF currents on the ESRRF can be solved as (1.34).

$$\begin{bmatrix} u_{dh}^{\hat{r}} \\ u_{qh}^{\hat{r}} \end{bmatrix} = \begin{bmatrix} u_h \cos(\omega_h t) \\ 0 \end{bmatrix} \quad (1.33)$$

$$\begin{bmatrix} i_{dh}^{\hat{r}} \\ i_{qh}^{\hat{r}} \end{bmatrix} = \begin{bmatrix} I_1 + I_2 \cos[2(\theta_r - \hat{\theta}_r)] \\ I_2 \sin[2(\theta_r - \hat{\theta}_r)] \end{bmatrix} \sin(\theta_h) \quad (1.34)$$

$$I_1 = \frac{u_h (L_{qh} + L_{dh})}{2\omega_h L_{dh} L_{qh}}, I_2 = \frac{u_h (L_{qh} - L_{dh})}{2\omega_h L_{dh} L_{qh}} \quad (1.35)$$

As can be seen, the HF currents due to pulsating voltage injection are amplitude-modulation waveforms. The q -axis HF current amplitude $I_{qh}^{\hat{r}}$ can be extracted by synchronous demodulation for position estimation as shown in Fig. 1.18. The rotor position is obtained by controlling $I_2 \sin(2(\theta_r - \hat{\theta}_r)) = 0$ with the PLL observer.

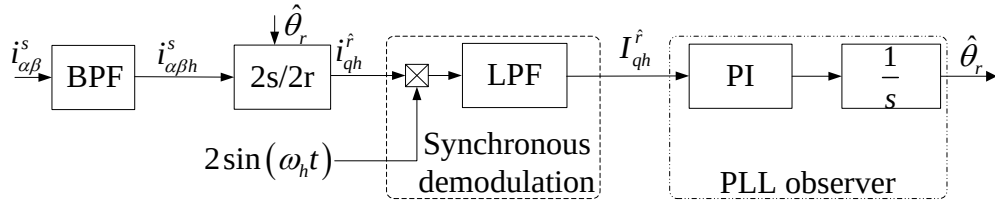


Fig. 1.18. Pulsating voltage injection based position demodulation method.

Similar to that of HF rotating voltage injection, the zero sequence voltage for pulsating voltage injection can also be extracted with the same measuring circuit in Fig. 1.16 for position estimation as

$$u_{Rn} = V_o \cos(\omega_h t) \cos(2\theta_r + \hat{\theta}_r) \quad (1.36)$$

$$V_o = \frac{u_h (L_0 - M_0)(L_2 - M_2)}{2L_0^2 + 2M_0^2 - 4L_0M_0 - 2L_2M_2 - L_2^2/2} \quad (1.37)$$

$$\varepsilon = LPF(u_{Rn} * 4\cos(\omega_h t - 3\hat{\theta}_r)) = V_o (\sin(2\tilde{\theta}_r) + \sin(2\theta_r + 4\hat{\theta}_r)) \quad (1.38)$$

As can be seen, by controlling the error signal $\varepsilon=0$, there will exist significant six harmonic in the estimated result. To reduce the ripple, a modified pulsating injection is proposed as shown in Fig. 1.19 in which the pulsating voltage is injected to the q -axis of an estimated frame which rotates anticlockwise at twice of the estimated angular speed.

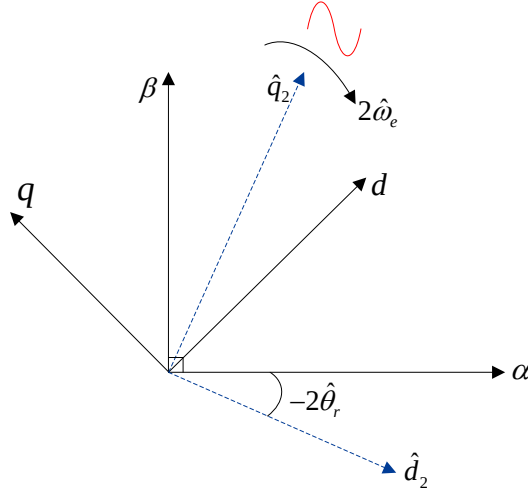


Fig. 1.19. Modified injection frame for zero sequence voltage method.

With this novel injection method, the zero sequence voltage due to pulsating voltage injection can be expressed as (1.39). By controlling $\varepsilon=0$ in (1.41) with PLL observer, the position can be estimated and the six harmonic error is largely reduced as V_{o1} is much bigger than V_{o2} .

$$u_{Rn} = V_{o1} \cos(\omega_h t) \sin(2\tilde{\theta}_r) - V_{o2} \cos(\omega_h t) \sin(4\theta_r + 2\tilde{\theta}_r) \quad (1.39)$$

$$V_{o1} = \frac{u_h (L_0 - M_0)(L_2 - M_2)}{2L_0^2 + 2M_0^2 - 4L_0M_0 - 2L_2M_2 - L_2^2/2} \quad (1.40)$$

$$V_{o2} = \frac{0.5u_h (L_2 - 2M_2)(L_2 - M_2)}{2L_0^2 + 2M_0^2 - 4L_0M_0 - 2L_2M_2 - L_2^2/2}$$

$$\varepsilon = LPF(u_{Rn} \cos(\omega_h t)) \approx 2V_{o1} \sin(2\tilde{\theta}_r) (V_{o1} \gg V_{o2}) \quad (1.41)$$

As can be seen, both the sinusoidal rotating and pulsating voltage injection methods can be used for sensorless position control and there is not a substantial difference between the two techniques [BIA07]. They have some characteristics in common as

- ① The signal processing is complicated with BPF and LPF.
- ② The injection frequency cannot be too high due to limited sampling point during one injection cycle. Generally, for switching frequency 10kHz, the injection sinusoidal frequency is usually below 2kHz.

(b) Pulsating square waveform injection

Pulsating square wave injection [YOO11] was proposed as shown in Fig. 1.20. Instead of injecting sinusoidal voltage, pulsating square wave voltage as (1.42) is injected to the estimated d-axis.

$$u_{dh}^{\hat{r}} = \begin{bmatrix} V_h, 2kT_s \leq t < (2k+1)T_s \\ -V_h, (2k+1)T_s \leq t < 2(k+1)T_s \end{bmatrix}, k = 0, 1, 2, \dots \quad (1.42)$$

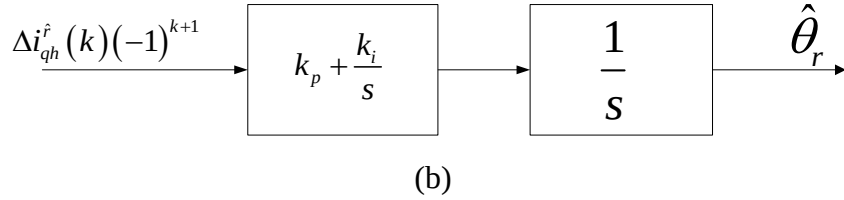
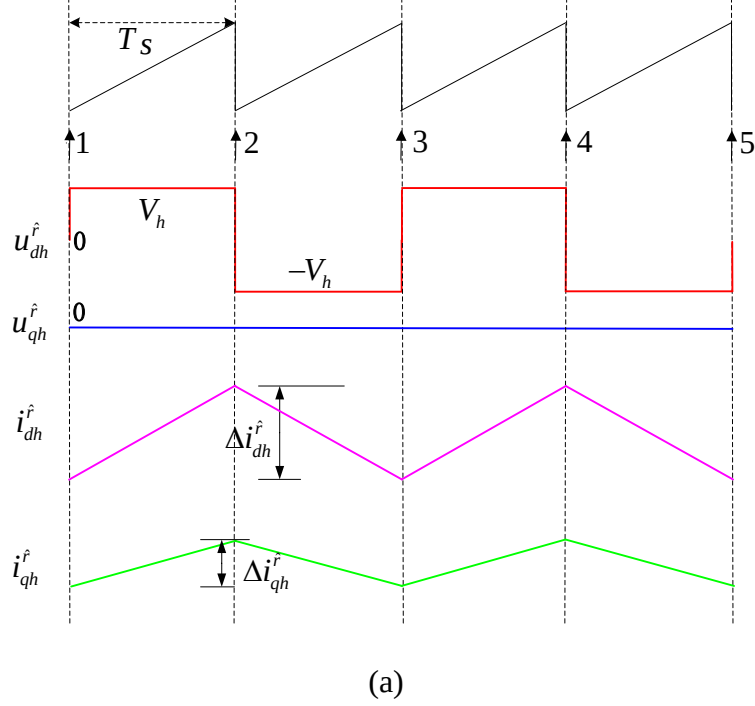


Fig. 1.20. Pulsating square wave voltage injection. (a) Injection sequence. (b) Position demodulation.

where T_s is the sampling period, V_h is the amplitude of the injection square wave voltage. By substituting the injected voltages into the voltage equations (1.32) in the estimated reference frame, the dq -axis currents can be solved as

$$\begin{bmatrix} i_{dh}^{\hat{r}} \\ i_{qh}^{\hat{r}} \end{bmatrix} = \begin{bmatrix} \chi_1 + \chi_2 \cos[2(\theta_r - \hat{\theta}_r)] \\ \chi_2 \sin[2(\theta_r - \hat{\theta}_r)] \end{bmatrix} \int \begin{bmatrix} u_{dh}^{\hat{r}} \\ u_{qh}^{\hat{r}} \end{bmatrix} dt \quad (1.43)$$

$$\chi_1 = \frac{L_{qh} + L_{dh}}{2L_{dh}L_{qh}}, \chi_2 = \frac{L_{qh} - L_{dh}}{2L_{dh}L_{qh}} \quad (1.44)$$

The current variations between adjacent sampling points can be solved from (1.43) as

$$\begin{bmatrix} \Delta \hat{i}_{dh}^r(k) \\ \Delta \hat{i}_{qh}^r(k) \end{bmatrix} = \begin{bmatrix} I_a + I_b \cos[2(\theta_r - \hat{\theta}_r)] \\ I_b \sin[2(\theta_r - \hat{\theta}_r)] \end{bmatrix} (-1)^{k-1}, k=1,2,3... \quad (1.45)$$

$$I_a = \frac{(L_{qh} + L_{dh})V_h T_s}{2L_{dh}L_{qh}}, I_b = \frac{(L_{qh} - L_{dh})V_h T_s}{2L_{dh}L_{qh}} \quad (1.46)$$

By controlling $\varepsilon = \Delta \hat{i}_{qh}^r(k)(-1)^{k+1}$ with the PLL observer, the rotor position can be estimated. The minimum sampling period can be half of the switching frequency if sampling at both the bottom and peak of the PWM carrier waveform can be performed. In this case, the injection frequency is the same as the switching frequency and the conventional BPF and LPF can be eliminated. If the switching frequency exceeds the upper limit of hearing, the HF noise problem is also mitigated. However, at the same time, with high injection frequency (small T_s), the value of I_b is also small which reduces the SNR of the error signal. Similar to pulsating sinusoidal voltage injection, the HF zero sequence voltage can also be used for sensorless position estimation. For improving the estimation accuracy, the method in Fig. 1.19 by injecting the anticlockwise rotating frame can also be used.

According to previous presentation on continuous signal injection based sensorless control, Table 1.4 shows the comprehensive comparison.

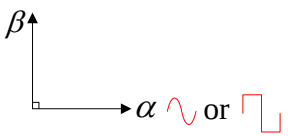
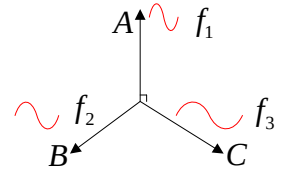
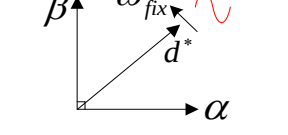
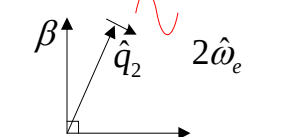
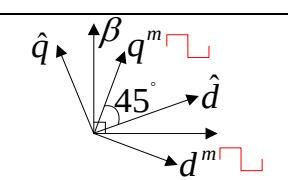
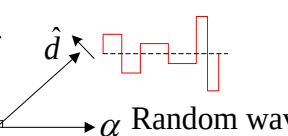
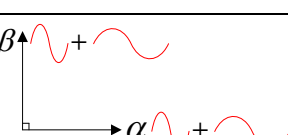
Table 1.4 Comparison between continuous signal injection based sensorless control methods

	Injection frequency	Demodulation methods
Sinusoidal rotating	Medium	① Negative sequence current ② Zero sequence voltage (Insensitive to injection frequency and noise)
Sinusoidal pulsating	Medium	① Q-axis HF current amplitude ② Zero sequence voltage (Insensitive to injection frequency and noise)
Square wave pulsating	High	① Q-axis HF current variation ② Zero sequence voltage (Insensitive to injection frequency and noise)

1.2.2.3 Other continuous signal injection methods

In this section, some other novel continuous injection methods will be introduced as summarized in Table 1.5.

Table 1.5 Summary of other novel continuous signal injection based sensorless methods

	Novel injection frame	Novel injection signals	Novel demodulation
References	Injection strategy	Novelties	Benefits
[LIU14c] [LIU14d]		Pulsating injection in the stationary frame	Stability↑
[TAN17]		Pulsating injection in the ABC frame ($f_1 \neq f_2 \neq f_3$)	Stability↑ Compensate delay
[LUO16]		Pulsating injection to estimated fixed rotating frame	Stability↑ No observer Compensate delay
[XU16]		Pulsating injection to anticlockwise rotating frame	Harmonics↓ Injection frequency↑
[LI19]		Pulsating injection to the measured reference frame	Compensate cross-coupling effect
[WAN17b]		Pseudo-random injection	Reduced noise↓
[CHE14] [SHU21]		Multi-frequency injection	Decoupling multi-saliencies/reduced noise↓
[BOL11]	For pulsating injection	DFT technique	Delay immune
[LI20]	For pulsating injection	Moving average filter	No BPF filters
[SHU21]	For rotating injection	HF current amplitudes	No observer

Their novelties are on the injection frame (where), injection signal type (what) and demodulation strategy (how) which have been described as the three key components in the HF signal injection based sensorless control in Fig. 1.13. Some benefits can be gained with these methods as follows. No matter how novel these methods are, their basic principles are still the saliency model of the machine.

(a) Novel injection frames

In [LIU14c], [LIU14d], the pulsating signals are injected in the stationary frame, while conventionally the pulsating signal is injected to the estimated frame. As the injection is performed in the stationary reference frame, the stability is improved.

In [TAN17], pulsating sinusoidal signals of different frequencies are injected to the three phases which bring the benefits of increased stability and delay effect compensation.

In [LUO16], pulsating voltage is injected to the d-axis of the estimated frame which is rotating at fixed frequency. Good steady and dynamic performances can be achieved.

In [XU16], for making use of zero sequence voltage for position demodulation, pulsating signal is injected to q -axis of an estimated frame which rotates anticlockwise at two times of the estimated rotor speed.

In [LI19], pulsating square wave is injected to the measured reference frame for the purposed of compensating the cross-coupling effect.

(b) Novel injection signals

Instead of changing the injection frame, some novel methods modify the injection signals.

In [WAN17b], pseudo-random square voltage wave is injected to the estimated d -axis. As the pseudo-random signal has a wider frequency spectrum, the sharp noise is largely reduced.

In [CHE14], signals of multiple frequencies are injection to the stationary reference frame. The PSCs and NSCs for different frequencies are extracted to decouple the multi-saliencies.

In [SHU21], rotating signals with multiple frequencies are injected to the stationary frame for reducing the noise level due to reduced peaks at the injection frequency.

(c) Novel demodulation strategies

Some novel methods use different signal processing techniques for position demodulation.

In [BOL11], the discrete Fourier transform (DFT) technique is introduced to extract the amplitude of the HF carrier current, by which the uncertain delay effects can be mitigated.

In [LI20], the BPFs and LPFs which are conventionally used in pulsating injection based sensorless method are neglected by employing moving average filter.

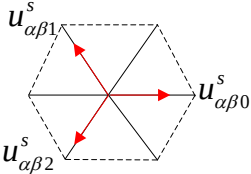
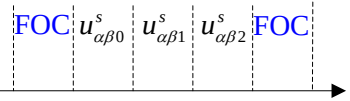
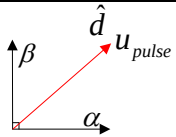
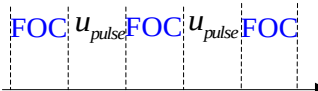
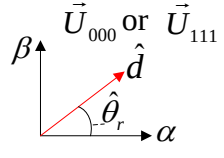
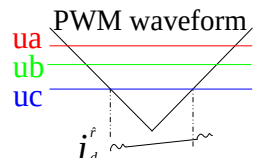
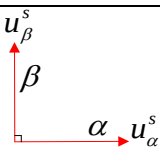
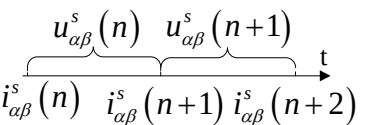
In [SHU21], instead of using negative sequence current, the amplitudes of three phase HF currents are extracted for fast initial position estimation. The conventional PLL observer is not used.

The continuous signal injection methods mentioned above have largely enriched the ways for sensorless position estimation. In the following section, saliency tracking based position estimation methods by injecting discrete signals will be presented.

1.2.3 Saliency model based sensorless methods with discrete signal injection

In the previous section, the continuous signal injection sensorless methods are characterized by the use of BPFs and LPFs for extracting the HF currents. The methods in this section do not rely on digital filters and merely make use of the current variations during discrete voltage vectors as shown in Table 1.6. They usually inject discrete signals to the machine or just make use of the PWM waveforms in FOC. Hence, they are called discrete signal injection based sensorless methods.

Table 1.6 Summary of the discrete signal injection based sensorless methods

Method	Injection frame	PWM sequence	Characteristics
1.2.3.1 INFORM			Current variations 3 injection vectors
1.2.3.2 Pulse injection			Current variations 1 injection vector
1.2.3.3 ZVCD/PWM			Current variations 1 injection vector
1.2.3.4 Arbitrary injection			Current variations Arbitrary vector

1.2.3.1 INFORM method

The indirect flux detection by online reactance measurement (INFORM) method is based on real time inductance measurement by voltage pulse injection [SCH96]. It is based on the HF model as

$$\begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} = \begin{bmatrix} L_{\Sigma} - L_{\Delta} \cos(2\theta_r) & -L_{\Delta} \sin(2\theta_r) \\ -L_{\Delta} \sin(2\theta_r) & L_{\Sigma} + L_{\Delta} \cos(2\theta_r) \end{bmatrix} p \begin{bmatrix} i_{\alpha h}^s \\ i_{\beta h}^s \end{bmatrix} \quad (1.47)$$

$$u_{\alpha\beta k}^s = U_h e^{j \frac{2\pi}{3} k}, k = 0, 1, 2 \quad (1.48)$$

When the voltage vectors $u_{\alpha\beta k}^s$ shown in (1.48) are injected to the machine as shown in Fig. 1.21.

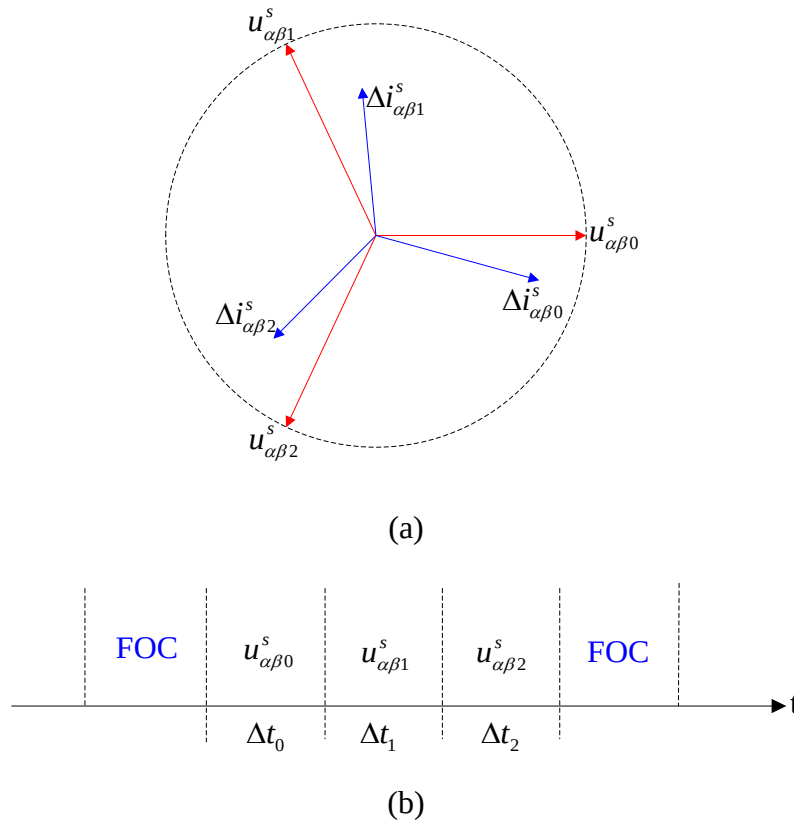


Fig. 1.21. INFORM demonstration. (a) Injection frame. (b) Injection sequence.

The current variations due to different voltage vector injections can be expressed as

$$\frac{\Delta i_{\alpha\beta k}^s}{\Delta t_k} = \frac{U_h}{L_{\Sigma}^2 - L_{\Delta}^2} [L_{\Sigma} + L_{\Delta} e^{j \left(2\theta_r - \frac{2\pi}{3} k \right)}], k = 0, 1, 2 \quad (1.49)$$

where Δt_k is the duration of the corresponding voltage vector. The rotor position can be obtained by adding the current derivatives in the three injections cycles as

$$\sum_{k=0,1,2} \frac{\Delta i_{\alpha\beta k}^s}{\Delta t_k} e^{j\frac{2\pi}{3}} = \frac{3L_\Delta U_h}{L_\Sigma^2 - L_\Delta^2} e^{j2\theta_r} \quad (1.50)$$

In actual applications, injection in one direction is usually performed with two opposite voltage pulses of the same magnitude to reduce the current distortions [SCH96]. The INFORM process may increase the implementation complexity and current sampling requirements due to very short injection time [XIE16]. The current distortions will also contribute to the torque ripples, noises and vibrations. To lower current ripples and harmonics, the INFORM can also be integrated in the PWM pattern [HOF17].

1.2.3.2 Voltage pulse injection method

In the INFORM method, at least three injection voltage vectors are injected along phase axes. The method using at least two separate injection vectors is also reported in [FRE10], utilizing the current variation behaviour under different injection voltage vectors. In this section, the voltage pulse injection method can estimate the rotor position with a single injection voltage vector. The current variation under the single testing voltage vector is directly related to the position error as shown in Fig. 1.22.

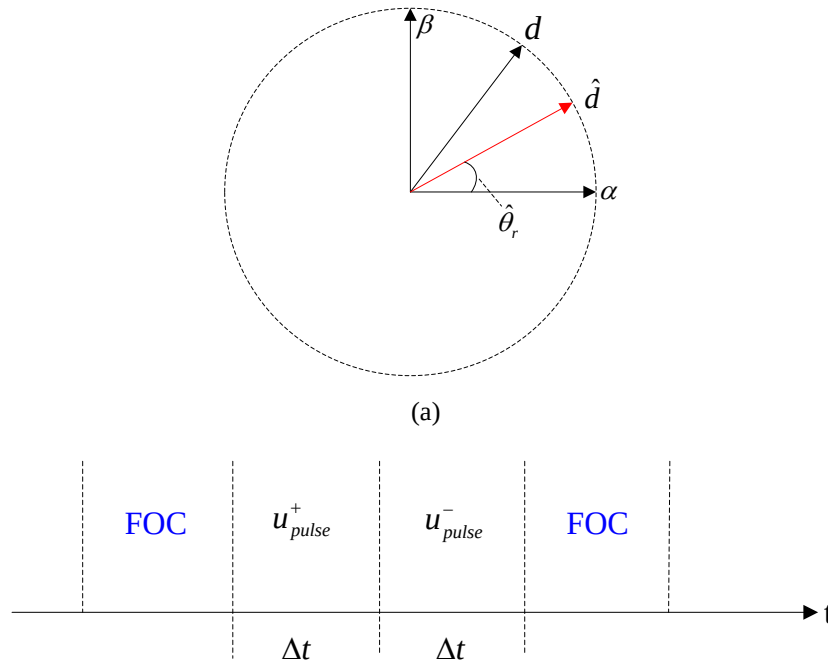


Fig. 1.22. Pulse injection method. (a) Injection frame. (b) Injection sequence.

When a voltage vector $U_h e^{j\hat{\theta}_r}$ is injected to the estimated d -axis, the resultant current variation in the estimated frame are calculated as

$$\frac{\Delta \hat{i}_{dq}^r}{\Delta t} = \frac{\Delta i_{\alpha\beta}^s}{\Delta t} e^{-j\hat{\theta}_r} = \frac{U_h}{L_\Sigma^2 - L_\Delta^2} \left[L_\Sigma + L_\Delta e^{j(2\theta_r - 2\hat{\theta}_r)} \right] \quad (1.51)$$

The imaginary part of the current variation can be expressed as

$$\text{Im}\left(\frac{\Delta \hat{i}_{dq}^r}{\Delta t}\right) = \frac{U_h L_\Delta}{L_\Sigma^2 - L_\Delta^2} \sin(2\theta_r - 2\hat{\theta}_r) \quad (1.52)$$

The rotor position can be obtained by controlling $\text{Im}\left(\frac{\Delta \hat{i}_{dq}^r}{\Delta t}\right) = 0$ with a PLL observer. BPFs and LPFs are not used in the position demodulation. To reduce the current distortion and compensating the inverter nonlinearities, a negative voltage pulse can be injected following the positive pulse. The sharp noise caused by pulse current can be reduced by adopting a low frequency pulse injection strategy [WAN17a].

1.2.3.3 ZVCD/PWM method

In the zero voltage vector current derivative (ZVCD) method, the current derivative on the estimated d -axis during zero voltage vector is measured for position estimation [RAU08]. Similar to the INFORM and pulse injection methods, digital filters are not required. If the zero voltage vector in a PWM period is not long enough for reliable current sampling, additional zero voltage vector can be inserted [XIE16].

Zero voltage vector in the pulse width modulation technique is shown by $\vec{U}_{000}, \vec{U}_{111}$ in Fig. 1.23(a). The vector $\vec{U}_{000}$ means all the lower power devices are switched on, while $\vec{U}_{111}$ means all the upper power devices are switched on. In actual applications, $\vec{U}_{000}$ is usually used for ZVCD as phase current sampling can be performed at the low side with low cost sampling hardware. A typical PWM sequence waveform is shown in Fig. 1.23(b).

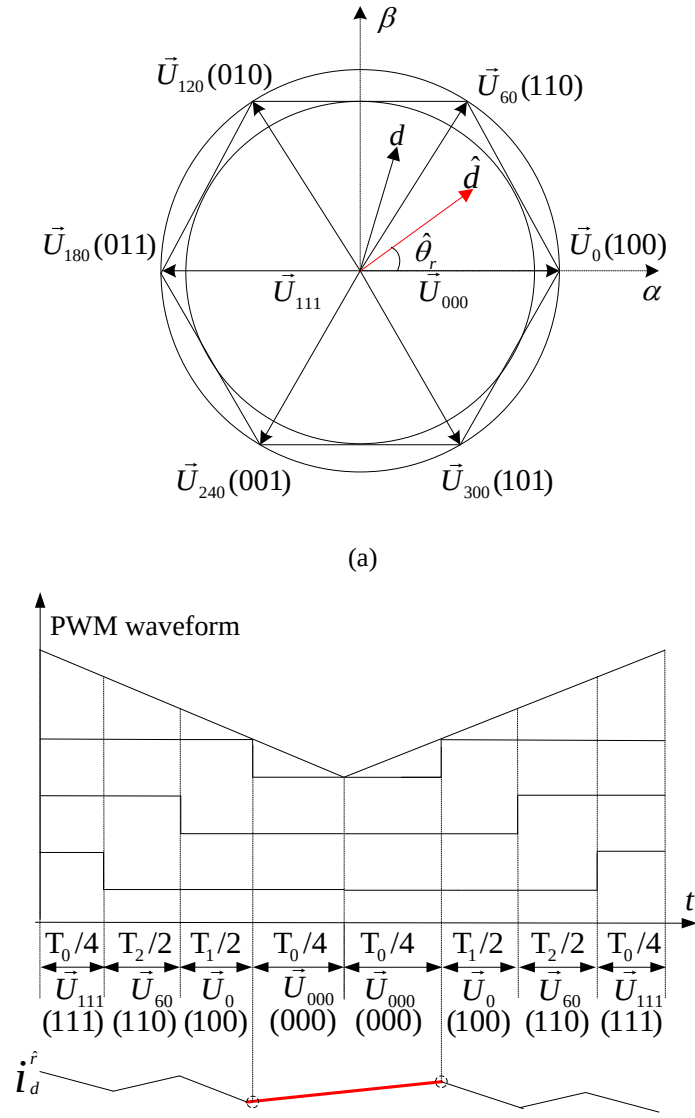


Fig. 1.23. ZVCD method. (a) Injection frame. (b) Injection sequence.

The general voltage equation for IPMSM are given by

$$p \begin{bmatrix} i_d \\ i_q \end{bmatrix} = \begin{bmatrix} -\frac{R_s}{L_d} & \frac{L_q}{L_d} \omega_e \\ -\frac{L_d}{L_q} \omega_e & -\frac{R_s}{L_q} \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} \frac{u_d}{L_d} \\ \frac{u_q - \omega_e \psi_f}{L_q} \end{bmatrix} \quad (1.53)$$

In the zero voltage vector interval, the dq -axis voltages are all zero. By transforming the voltage equations to the estimated reference frame, the voltages are derived as

$$p \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} = R(\tilde{\theta}_r) \begin{bmatrix} -\frac{R_s}{L_d} & \frac{L_q}{L_d} \omega_e \\ -\frac{L_d}{L_q} \omega_e & -\frac{R_s}{L_q} \end{bmatrix} R^{-1}(\tilde{\theta}_r) \begin{bmatrix} \hat{i}_d \\ \hat{i}_q \end{bmatrix} + R(\tilde{\theta}_r) \begin{bmatrix} 0 \\ -\frac{\omega_e \psi_f}{L_q} \end{bmatrix} + \tilde{\omega}_e \begin{bmatrix} \hat{i}_q \\ -\hat{i}_d \end{bmatrix} \quad (1.54)$$

where $\tilde{\theta}_r = \theta_r - \hat{\theta}_r$ is the difference between the actual position and the estimated d -axis. By setting the d -axis reference current to zero in the FOC, $\hat{i}_d$ is regarded as zero. The speed difference $\tilde{\omega}_e$ is very small and does not dominate in the equation. Then, the rotor position can be obtained by using the d -axis current derivative as

$$p\hat{i}_d \approx \left[\left(\frac{R_s}{L_d} - \frac{R_s}{L_q} \right) \hat{i}_q - \frac{\psi_f}{L_q} \omega_e \right] \tilde{\theta}_r \quad (1.55)$$

As can be seen, the current derivative measured on the estimated d -axis is proportional to the position error. By controlling $p\hat{i}_d=0$, the rotor position can be estimated.

To spare enough time for current samplings at the start and the end of the zero voltage vector, the maximum FOC output voltage should be limited. This zero voltage vector can be extended by inserting a separate zero voltage vector between adjacent FOC control cycles [XIE16] or applying the zero voltage vector extension scheme [WAN18]. To reduce the error caused by sampling noise, distortion and nonlinearity, a method by sampling the currents for several consecutive times in each start and end of zero voltage vector is utilized [WAN18].

It is also noted that except for making use of the current derivative during zero voltage vector, the current derivatives during active voltage vectors of the PWM waveform can also be used for position estimation [GAO07].

1.2.3.4 Arbitrary injection method

The arbitrary injection method is a novel sensorless position estimation method which also relies on the current derivative [PAU11]. The voltage equations in the stationary frame is expressed as

$$\begin{bmatrix} u_\alpha^s \\ u_\beta^s \end{bmatrix} = R_s \begin{bmatrix} i_\alpha^s \\ i_\beta^s \end{bmatrix} + \begin{bmatrix} L_\Sigma - L_\Delta \cos(2\theta_r) & -L_\Delta \sin(2\theta_r) \\ -L_\Delta \sin(2\theta_r) & L_\Sigma + L_\Delta \cos(2\theta_r) \end{bmatrix} p \begin{bmatrix} i_\alpha^s \\ i_\beta^s \end{bmatrix} + \omega_e \psi_{pm} \begin{bmatrix} -\sin(\theta_r) \\ \cos(\theta_r) \end{bmatrix} \quad (1.56)$$

The current derivatives in the stationary reference frame are derived as

$$p \begin{bmatrix} i_{\alpha}^s \\ i_{\beta}^s \end{bmatrix} = (Y_{\Sigma} I + Y_{\Delta} S(\theta_r)) \left(\begin{bmatrix} u_{\alpha}^s \\ u_{\beta}^s \end{bmatrix} - \overbrace{R_s \begin{bmatrix} i_{\alpha}^s \\ i_{\beta}^s \end{bmatrix}}^{\vec{u}_R^s} - \overbrace{\omega_e \psi_{pm} \begin{bmatrix} -\sin(\theta_r) \\ \cos(\theta_r) \end{bmatrix}}^{\vec{u}_{EMF}^s} \right) \quad (1.57)$$

where $Y_{\Sigma} = \frac{1}{2} \left(\frac{1}{L_d} + \frac{1}{L_q} \right)$, $Y_{\Delta} = \frac{1}{2} \left(\frac{1}{L_q} - \frac{1}{L_d} \right)$, $S(\theta_r) = \begin{bmatrix} \cos(2\theta_r) & \sin(2\theta_r) \\ \sin(2\theta_r) & -\cos(2\theta_r) \end{bmatrix}$.

The current variation during one sampling period is expressed as

$$\Delta i_{\alpha\beta}^s(n) = (Y_{\Sigma} I + Y_{\Delta} S(\theta_r)) (u_{\alpha\beta}^s(n) - \vec{u}_R^s(n) - \vec{u}_{EMF}^s(n)) \Delta t \quad (1.58)$$

The difference between two adjacent current variations is then expressed as

$$\Delta i_{\alpha\beta}^s(n+1) - \Delta i_{\alpha\beta}^s(n) = (Y_{\Sigma} I + Y_{\Delta} S(\theta_r))^* \left(\underbrace{u_{\alpha\beta}^s(n+1) - u_{\alpha\beta}^s(n)}_{\Delta \vec{u}_{\alpha\beta}} - \underbrace{(\vec{u}_R^s(n+1) - \vec{u}_R^s(n))}_{\Delta \vec{u}_R \approx 0} - \underbrace{(\vec{u}_{EMF}^s(n+1) - \vec{u}_{EMF}^s(n))}_{\Delta \vec{u}_{EMF} \approx 0} \right) \Delta t \quad (1.59)$$

The resistive voltage drop $\vec{u}_R^s$ and back EMF voltage drop $\vec{u}_{EMF}^s$ can be seen as constant during two consecutive PWM cycles [LAN10]. Thus, $\Delta \vec{u}_R^s$ and $\Delta \vec{u}_{EMF}^s$ are nearly zero. The rotor position can be calculated by (1.61). General signal processing flow for obtaining the rotor position is shown in Fig. 1.24.

$$Y_{\Delta} S(\theta_r) \Delta \vec{u}_{\alpha\beta} \Delta t = \underbrace{[\Delta i_{\alpha\beta}^s(n+1) - \Delta i_{\alpha\beta}^s(n)] - Y_{\Sigma} \Delta \vec{u}_{\alpha\beta} \Delta t}_{\Delta \vec{e}_{\alpha\beta}} \quad (1.60)$$

$$\theta_r = \frac{1}{2} \tan^{-1} \left(\frac{\Delta u_{\alpha} \Delta e_{\beta} + \Delta u_{\beta} \Delta e_{\alpha}}{\Delta u_{\alpha} \Delta e_{\alpha} - \Delta u_{\beta} \Delta e_{\beta}} \right) \quad (1.61)$$

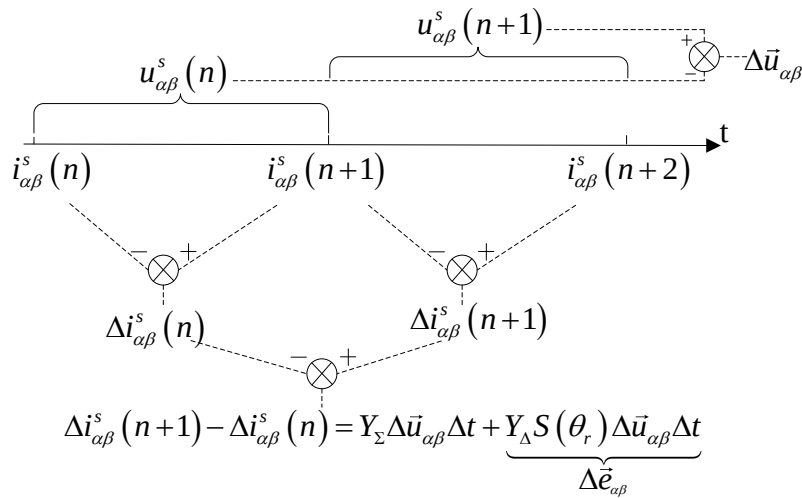


Fig. 1.24. Arbitrary injection method demonstration.

This method does not require any predetermined injection waveform and has low dependency on speed, resistance. However, there are still two drawbacks as

- ① The measurement of current variations is sensitive to noise, sampling delay and distortion.
- ② The value of Y_{Σ} is required. To obtain the value of Y_{Σ} , additional signal injection is usually employed. In [BAR15], a square wave spatially rotating 120° in the stationary reference frame every switching cycle is injected.

The above discrete signal excitation based sensorless methods are all independent of digital filters. However, they are all sensitive to sampling noise and distortion due to the measurement of current derivatives. General comparison of saliency model based sensorless methods including continuous signal injection and discrete signal injection methods is shown in Table 1.7.

Table 1.7 General comparison of saliency model based sensorless methods

General types	Injection methods	Bandwidth	Sampling sensitivity	Parameter sensitivity	Noise emission
Continuous signal	Sinusoidal rotating	Medium	<i>Low</i>	No	High
	Sinusoidal pulsating	Medium	<i>Low</i>	No	Medium
	Pulsating square wave	<i>High</i>	High	No	Medium
Discrete Signal	INFORM	Medium	High	No	Medium
	Pulse injection	Medium	High	No	Medium
	ZVCD/PWM	<i>High</i>	High	No	<i>Low</i>
	Inherent PWM	<i>High</i>	High	<i>Yes</i>	<i>Low</i>

1.2.4 Common issues for saliency model based sensorless methods

Ideally, with the above saliency model based sensorless methods, the position can be estimated without error. However, in actual applications, there are usually steady state errors and harmonics in the estimated results. In heavy load applications, the estimation may even diverge [BIA07]. In this section, the issues that affect the performance of saliency model based sensorless methods are presented as shown in Fig. 1.25. Saturation, cross-coupling effect and multi-saliencies are attributed to properties of the magnetic materials [REI08], while the delay

and load transient are caused by signal processing. The estimation is also affected by inverter nonlinearities, i.e. forward voltage drop, switching delay and parasitic capacitances which can be classified as hardware issues.

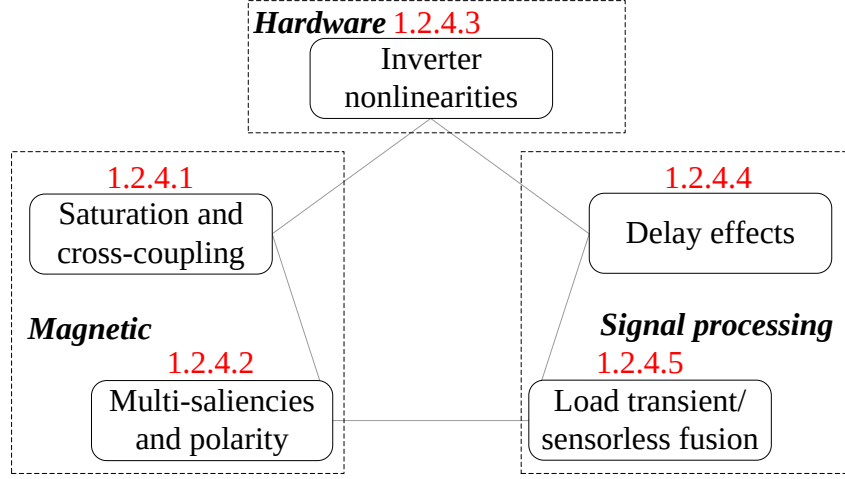


Fig. 1.25. Issues that affect performance of saliency model based sensorless methods.

1.2.4.1 Saturation and cross-coupling effect

The saliency model considering magnetic saturation and cross-coupling effect is given by

$$\begin{bmatrix} u_{dh}^r \\ u_{qh}^r \end{bmatrix} = \begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{qh} \end{bmatrix} p \begin{bmatrix} i_{dh}^r \\ i_{qh}^r \end{bmatrix} \quad (1.62)$$

where the mutual incremental inductances L_{dqh}, L_{qdh} describe the cross-coupling effect between d -axis and q -axis. Their values are equal due to reciprocity theorem [MIN17]. The impacts of saturation and cross-coupling effect on the sensorless control of IPMSM are as follows.

(a) Saturation

An important premise of saliency based position estimation is the existence of saliency variation, i.e. $L_{qh} \neq L_{dh}$. However, with the increase of the load current, the magnetic material is saturated which results in decreased inductance. The saliency variation may disappear due to reduced difference between L_{qh} and L_{dh} . For example, in HF rotating voltage injection based sensorless control, the error signal $\frac{(L_{qh}-L_{dh})}{2\omega_h L_{dh} L_{qh}} \sin(2\tilde{\theta}_r)$ may be zero when L_{dh} equals L_{qh} . This will result in zero input of the position observer and failure of the estimation.

For small saliency ($L_{dh} \approx L_{qh}$), higher injection voltage can be adopted for increasing the amplitude of the error signal. Specifically, for SPMSM with normally negligible saliency ratio,

saliency variation can be created by intentional magnetic saturation [KAN19] or modifying the current trajectory on which optimal saliency level can be achieved [LEE20].

(b) Cross-coupling effect

The cross-coupling effect can result in offset in the estimated position. The sinusoidal rotating voltage injection based sensorless method is analysed for example. When the cross-coupling effect, i.e. L_{dqh}, L_{qdh} is considered, the HF currents induced by HF rotating voltage injection are given by

$$i_{\alpha h}^s + j i_{\beta h}^s = I_p e^{j(\theta_h - \pi/2)} + I_n e^{-j(\theta_h - 2(\theta_r - \Delta\theta_m) - \pi/2)} \quad (1.63)$$

$$\Delta\theta_m = 0.5 \tan^{-1} \left(\frac{2L_{dqh}}{L_{qh} - L_{dh}} \right) \quad (1.64)$$

As can be seen, there is an offset angle $\Delta\theta_m$ caused by the cross-coupling incremental inductances. By using the heterodyning method in section 1.2.1, the position error signal that is sent to the position observer is

$$\varepsilon = I_n \sin(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \quad (1.65)$$

where $\hat{\theta}_r$ is the estimated rotor position, which can be obtained by controlling $\varepsilon = 0$ with the PLL observer, i.e.

$$\hat{\theta}_r = \theta_r - \Delta\theta_m \quad (1.66)$$

As can be seen, there is a position estimation error $\Delta\theta_m$ in HF rotating voltage injection based sensorless control caused by the cross-coupling effect. It has been verified that the error is inherent to the machine and not dependent on specific injection signals [GUG06]. If such error is not compensated, it will result in significant performance degradation especially at heavy load conditions. However, the existence of cross-coupling effect also helps to extend the feasible region for sensorless control [BIA07a]. For high performance sensorless control, the cross-coupling effect induced error should be compensated. Several compensation methods are presented in Fig. 1.26.

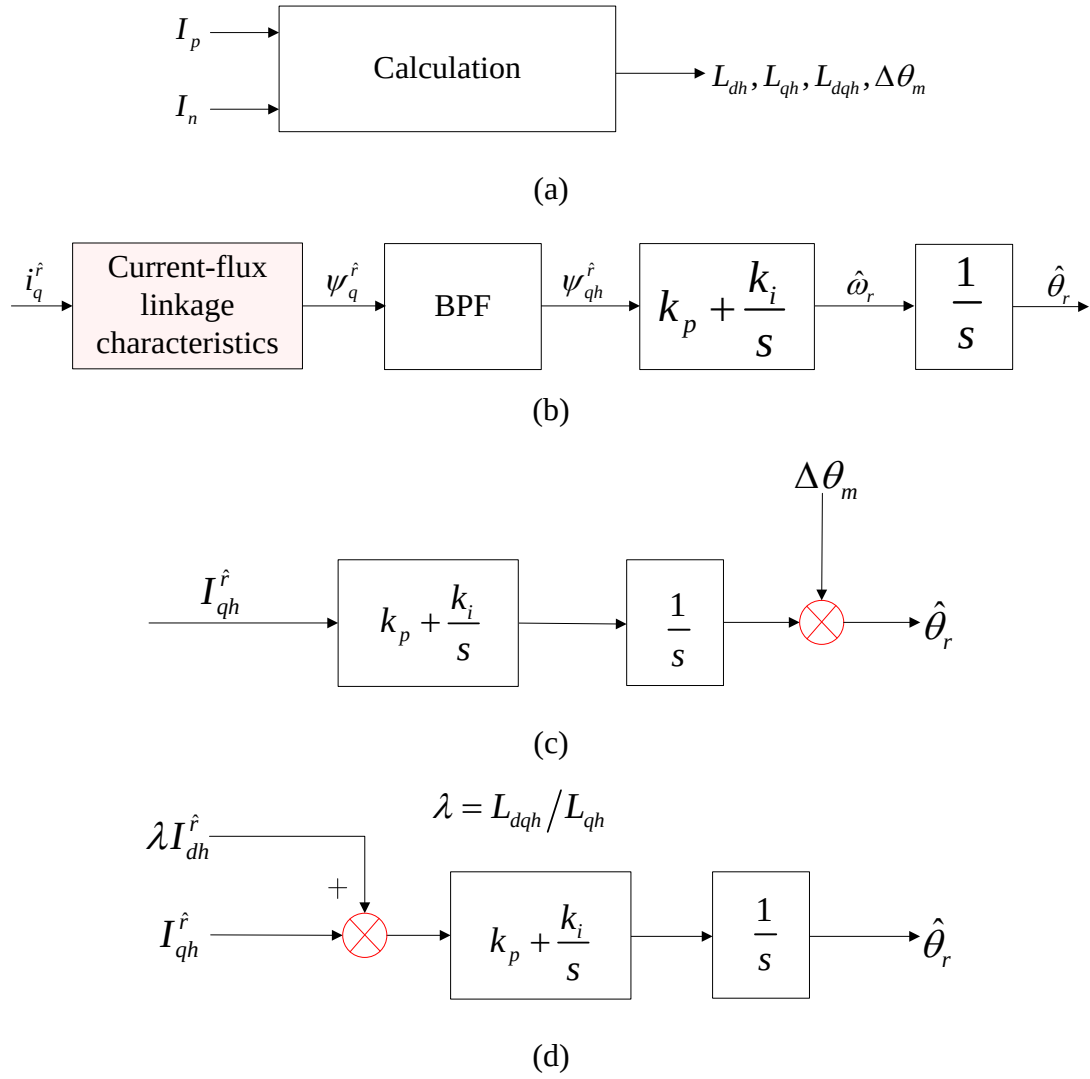


Fig. 1.26. Category of cross-coupling effect compensation methods. (a) Calculation approximation [REI08], [MUR10]. (b) Flux modulation based on offline measured flux linkage-current characteristics [GUG06]. (c) Offline measured errors (angle compensation) [FRE10]. (d) Offline measured ratio between incremental inductances (current compensation) [LI09].

As shown in Fig. 1.26, the compensation methods can be divided as online and offline methods. The online methods obtain the cross-coupling information from the online measured HF currents or voltages without any pre-measured-knowledge of the machine characteristics, while the offline methods use the offline measured machine data to online compensate the cross-coupling effect. The flux demodulation method extracts the HF flux linkage based on the offline measured current-flux linkage characteristics which contain the cross-coupling effect information [GUG06]. The error caused by the cross-coupling effect can be offline measured and used to online compensate the estimated position. The ratios between incremental

inductances can also be utilized to compensate the cross-coupling by modifying the input current signal of the position observer [LI09].

The online method is based on the assumption of fixed d-axis incremental inductance. This assumption is only feasible for some specific machines as the d -axis incremental inductance usually also varies with dq -axis currents due to magnetic saturation and cross-coupling effect. The offline methods modify the input or output of the position observer to make it converge at $\Delta\theta = 0$.

In highly saturation conditions, as I_n in the observer input signal ε , i.e. $I_n \sin(2\Delta\theta - 2\Delta\theta_m)$, is very small, there is a possibility that the trace $\varepsilon - \Delta\theta$ will not have a convergence point during $\Delta\theta \in [-180^\circ, 180^\circ]$. In such situation, the angle compensation and current compensation can be combined to guarantee reliable convergence at $\Delta\theta = 0$ [MAN20].

1.2.4.2 Secondary-saliencies and polarity

In previous HF model, the incremental inductances in the dq -frame are seen as only dependent on the dq -axis currents. However, they are also varying with rotor position which is caused by spatial inductance harmonics (also called secondary saliencies). The HF model in the stationary frame considering spatial harmonics can be expressed as

$$L_{\alpha\beta} = \begin{bmatrix} L_\Sigma + \sum_{h=1}^{\infty} L_{\Delta h} \cos(2h\theta_r) & -\sum_{h=1}^{\infty} L_{\Delta h} \sin(2h\theta_r) \\ -\sum_{h=1}^{\infty} L_{\Delta h} \sin(2h\theta_r) & L_\Sigma - \sum_{h=1}^{\infty} L_{\Delta h} \cos(2h\theta_r) \end{bmatrix} \quad (1.67)$$

where h is the harmonic order ($h=1$ represents the primary saliency), $L_{\Delta h}$ is the amplitude of the spatial inductance. For rotating voltage injection [REI08], the HF currents will contain current harmonics caused by the secondary saliencies as

$$i_{\alpha\beta h}^s = I_p e^{j(\theta_h - \pi/2)} + \sum_{h=1}^{\infty} I_{nh} e^{-j(\theta_h - 2h\theta_r - \pi/2)}, I_{nh} = \frac{u_h L_{\Delta h}}{2\omega_h (L_\Sigma^2 - L_{\Delta h}^2)} \quad (1.68)$$

For pulsating voltage injection, the estimated q -axis HF current is as

$$\hat{i}_{qh}^r = \left[I_n \sin(2\theta_r - 2\hat{\theta}_r) + \sum_{h=1}^{\infty} I_{nh} \sin(2\theta_r - 2h\hat{\theta}_r) \right] \sin(\omega_h t) \quad (1.69)$$

The HF current components with $h>1$ are called the secondary saliency currents. It is difficult to separate the primary currents with the secondary saliency currents as their frequencies are too close. Hence, there will be ripples in the estimated position. In [REI08], a neural network

is introduced to decouple secondary saliencies. In [CHE16], the multi-saliencies are offline measured with pulse voltage injection and used for online compensation. In [CHE14], with multi-signal injection method, multi-saliencies can be removed with only the primary saliency left. In many applications, due to relatively low magnitudes, multi-saliencies can be neglected in machines with high primary saliency. Although, secondary saliencies can result in estimation ripples, they can also be utilized for polarity detection as described below.

Polarity detection is an important topic in saliency tracking based sensorless control. In section 1.2.21, it has been revealed that there is an ambiguity of π in the estimated position as convergence can be reached at both $\Delta\theta=0$ and $\Delta\theta=\pi$. This is due to that the saliency undergoes two cycles in one electrical period. To distinguish the polarity, three kinds of methods can be used as shown in Fig. 1.27.

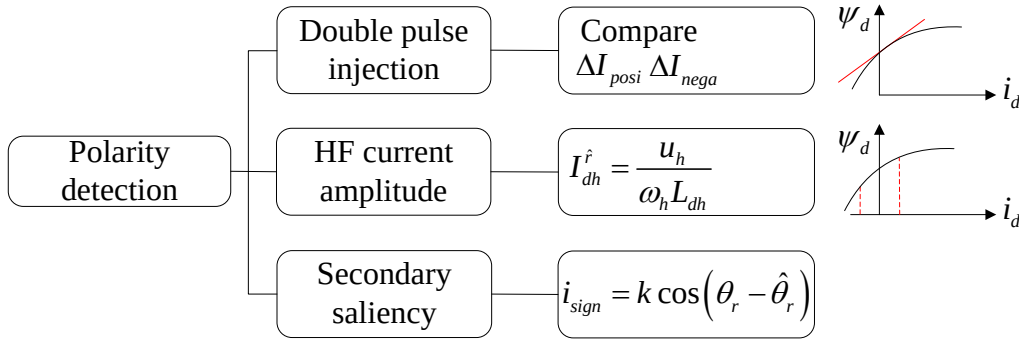


Fig. 1.27. Polarity detection methods.

The double pulses injection method is performed with two opposite voltage pulses injected to the estimated d -axis. The current variation caused by voltage injection on north-pole is higher than that on south-pole due to different saturation levels [HOL08]. In [GON11], polarity is detected by the d -axis HF current amplitude variation under different d -axis fundamental currents, while in [JEO05], [XU17b] [KIM04], the secondary saliency is used to indicate the polarity. General comparison between these methods are shown in Table 1.8.

Table 1.8 Comparison between different polarity detection methods

	Double pulses	Amplitude variations	Secondary saliencies
Concept complexity	<i>Simple</i>	<i>Simple</i>	<i>Complex</i>
S/N ratio	<i>High</i>	<i>High</i>	<i>Low</i>
Robustness	<i>High</i>	<i>High</i>	<i>Low</i>
Response	<i>Slow</i>	<i>Slow</i>	<i>Quick</i>
Computation burden	<i>Low</i>	<i>Low</i>	<i>Medium</i>
Integration with sensorless algorithm	<i>No</i>	<i>Yes</i>	<i>Yes</i>

1.2.4.3 Inverter nonlinearity

The saliency tracking based sensorless methods are usually carried out with a general purpose three phase inverter. The voltage output is affected by nonlinear inverter characteristics which include dead time, turn on/off times, parasitic capacitance and voltage drop [GUE05]. The inverter nonlinearities can distort the HF voltage and result in current harmonics which may induce ripples in the estimated position. Hence, the impacts of the inverter nonlinearities should be analysed and compensated.

The conventional saturation model indicates that the voltage distortion caused by inverter nonlinearities is determined by the current level and a curve describing the relationship between current and voltage errors can be plotted by experiment [PAR12]. This model is capable of analysing the voltage loss for fundamental current waveforms, but cannot be directly used for analysing the HF voltage loss. For modelling the impact of inverter nonlinearities on HF voltage injection, the HF equivalent resistance model was introduced in [GON11]. This model employed small signal linearization of the current-voltage loss trace to simplify the analysis. The HF voltage loss is modelled as the voltage drop on the equivalent HF resistance which is modulated by the fundamental phase current.

The voltage error is described as

$$\Delta u_{err} = f(i_{xf}) + f'(i_{xf})i_{xh} = \text{sign}(i_{xf}) \cdot \Delta U + R_{xh} \cdot i_{xh} \quad (1.70)$$

where i_{xf}, i_{xh} denote the fundamental current and the HF current of phase x, ΔU is the saturation voltage error, and R_{xh} is the HF equivalent resistance as shown in Fig. 1.28.

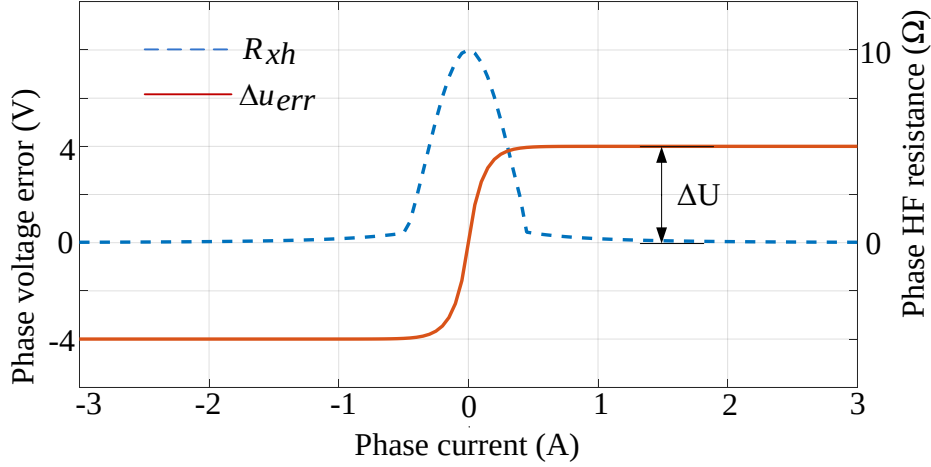


Fig. 1.28. HF equivalent resistance model.

With the HF phase resistances, the equivalent dq-axis HF resistances can be model as

$$R_{dh} = \frac{2}{3} \left[R_{ah} \cos^2(\theta_r) + R_{bh} \cos^2\left(\theta_r - \frac{2\pi}{3}\right) + R_{ch} \cos^2\left(\theta_r + \frac{2\pi}{3}\right) \right] \quad (1.71)$$

$$R_{qh} = \frac{1}{3} \left[R_{ah} \sin(2\theta_r) + R_{bh} \sin\left(\theta_r - \frac{4\pi}{3}\right) + R_{ch} \sin\left(\theta_r + \frac{4\pi}{3}\right) \right] \quad (1.72)$$

where R_{ah} , R_{bh} , R_{ch} are the equivalent phase HF resistances. The HF phase resistance can be curve-fitted with analytical expressions as

$$R_{xh} = \frac{2\Delta U * k * e^{-ki_{xf}}}{\left(1 + e^{-ki_{xf}}\right)^2}, x = a, b, c \quad (1.73)$$

where k is the parameter only determined by the dead time and switching devices, ΔU is the saturation value of the voltage error. For bus voltage 150V and dead time $2\mu s$ in the test rig, $k=0.5$ and $\Delta U=2.6V$ are identified. Such model can be used to compensate the error caused by inverter nonlinearities. However, accurate inverter parameters or offline measurements are required for setting up this model [PAR12]. In actual applications, the effect of compensation is sensitive to the sampling noise and distortions. In [GON11], the positive sequence current is used to compensate the ripple caused by inverter nonlinearities without knowing inverter parameters. In [CHO08], the pulsating signal injection axis is switched for minimal zero current clamping effects.

1.2.4.4 Delay effects

In the position demodulation process, delay effects in current sampling and voltage synthesis should be considered. There are two kinds of delays in HF voltage injection based sensorless control, i.e. the voltage synthesis delay and the sampling, filtering delay.

(a) Voltage synthesis by PWM waveform

Firstly, the voltage synthesis delay in current is mainly caused by the way of PWM voltage is generated [BAE03] as shown in Fig. 1.29. At t_0 , the voltage reference is transformed to the three phase PWM register compare values. But at t_0 , the PWM registers are only calculated and stored in the so called shadow registers, but not take effect. At t_1 , the comparison values in the shadow registers are transferred to the actual compare registers and the PWM waveforms will be generated during the period $[t_1, t_2]$. The voltage is approximately to be generated in the middle of the PWM period $[t_1, t_2]$. Therefore, the delay between the voltage reference and the actual output voltage is about $1.5T_{PWM}$.

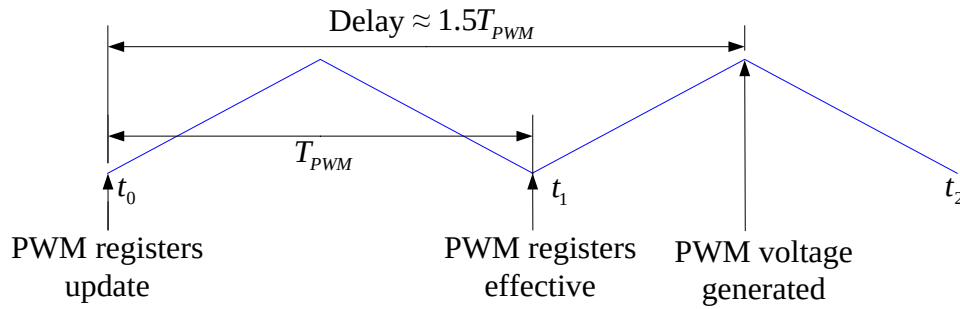


Fig. 1.29. Delay caused by PWM voltage synthesis.

(b) Delay caused by sampling and filtering

For HF voltage injection based sensorless control, the HF currents used for position estimation are obtained by sampling circuit and digital filters which will introduce delay to the actual currents. The accurate delay is difficult to determine as it is related to sampling hardware and filter parameters.

The HF currents considering delay for rotating voltage injection is expressed as

$$i_{\alpha\beta h}^s = I_p e^{j\left[\left(\theta_h + \varphi_{delay}\right) - \frac{\pi}{2}\right]} + I_n e^{-j\left[\left(\theta_h + \varphi_{delay}\right) - 2\theta_r - \frac{\pi}{2}\right]} \quad (1.74)$$

where φ_{delay} is the phase delay caused by voltage synthesis and sampling, filtering. With the position estimation methods in section 1.2.2.1, the delay effects will result in $\varphi_{delay}/2$ error in the estimation result for rotating voltage injection based sensorless control as

$$\theta_r - \hat{\theta}_r = \frac{1}{2} \varphi_{delay} \quad (1.75)$$

In [XU17a], a PI regulator is used to suppress the position errors by tracking the phase shift caused by the delays. For pulsating voltage injection, the delay effect will result in HF current amplitude decrease as

$$\begin{aligned} I_{qh}^{\hat{}} &= LPF \left[2I_{qh} \sin(2\tilde{\theta}_r) \sin(\omega_h t + \varphi_{delay}) * \sin(\omega_h t) \right] \\ &= I_{qh} \cos(\varphi_{delay}) \sin(2\tilde{\theta}_r) \end{aligned} \quad (1.76)$$

As can be seen, the amplitude of the position error signal is reduced. This will impact the bandwidth of the position observer, but not result in error in the estimated position. In [BOL11], the discrete Fourier transform (DFT) technique is introduced to calculate the HF current amplitude which is not affected by the delay effects.

1.2.4.5 Load transient and fusion of sensorless methods

For continuous HF signal injection based sensorless method, the load transient should also be considered for improving the dynamic performance. At steady state, there are no interferences between the HF and fundamental currents as the band pass filters can separate them as there frequencies are far apart. However, under load transient, the step load currents contain widespread frequency spectrum and are coincided with the frequency of the HF currents. Therefore, there will be interference between the load current and the HF currents, which will result in position estimation error.

There are typically three kinds of methods for eliminating the adverse impacts of load current transient as shown in Fig. 1.30. The first method is to limit the slop of the load current for reducing the harmonics in the transient currents. However, this will reduce the dynamic performance of torque control [CON09]. The second method is to predict the transient load current from the machine parameters and voltage references. The predicted transient currents are then subtracted from the sampled currents [BRI00], [BUG18]. With this technique, the step currents are removed before passing through the BPFs. The third method is to increase the injection frequency for reducing interference between the HF and transient load currents.

However, with higher injection frequency, higher injection voltage is also required for guarantee the amplitude of the position error signal.

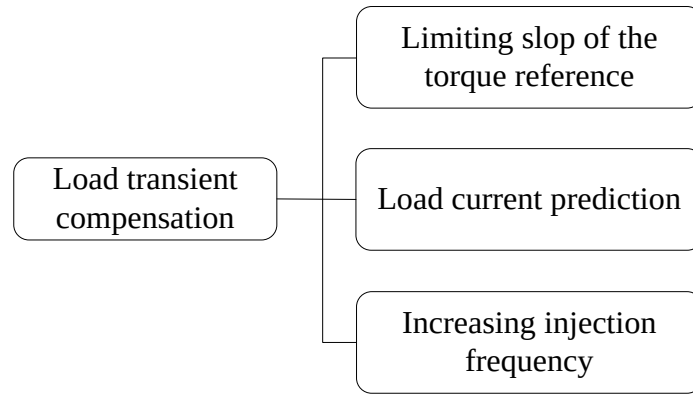


Fig. 1.30. Methods for compensation of load current transient.

Another important issue for applying the sensorless control methods is the switch between different sensorless algorithms. At standstill and low speed, the saliency tracking based sensorless methods have better performance, while at medium and high speeds, the fundamental model based sensorless methods have superiority. Conventionally, according to the number of the closed-loop observers that are used, there are typically two kinds of methods as shown in Fig. 1.31, i.e. single observer and dual observers.

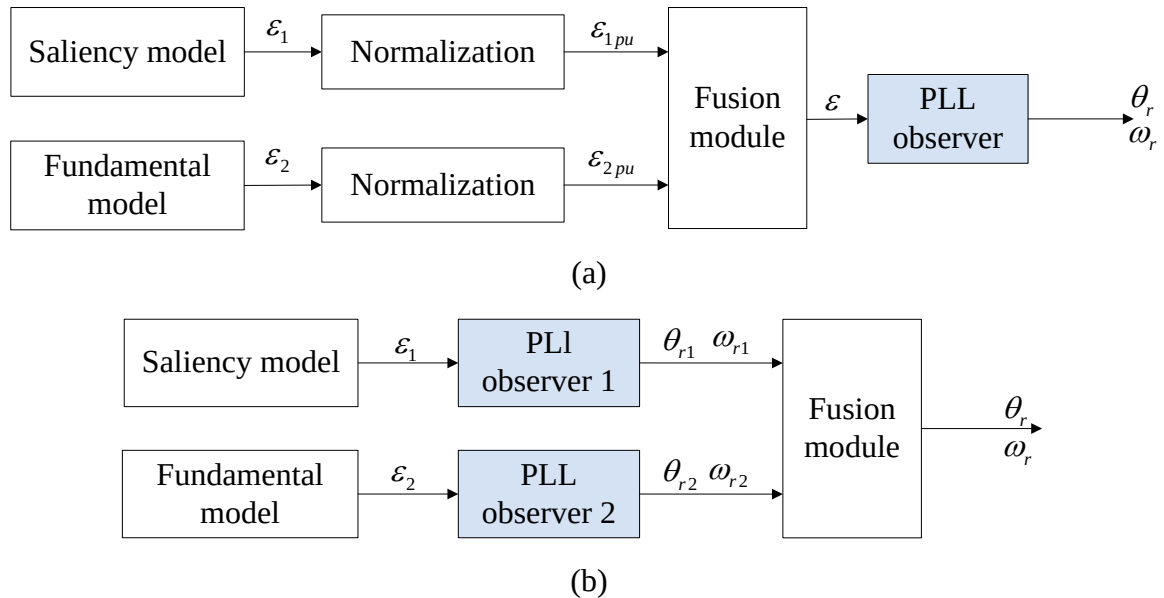


Fig. 1.31. Switchover strategies between saliency and fundamental model based sensorless methods. (a) Single observer. (b) Dual observers.

For dual-observer solution, the positions and speeds are estimated from separate observers and are forged together by weighting functions [IDE06]. The PI parameters for the two

observers should be designed separately. For single observer solution, the position errors from saliency model and fundamental model are normalized separately and then combined by a fusion model to form a hybrid error signal [YAN17]. With single observer, only one set of PI parameters should be tuned. The fusion module is usually a linear weighting function according to the speed, i.e. the saliency model is dominant below the low limit and the fundamental model is dominant above the high limit.

Except the above methods using fusion module to combine different sensorless models, some unified whole speed range sensorless methods are gaining popularity, which do not need to consider the weighting functions and the low limit, high limit of the transition region. In [XIA20], a whole speed sensorless operation method based on extended back EMF is proposed as shown in Fig. 1.32. It makes use of the sufficient large extended back EMF signal at low speed due to q -axis HF voltage injection.

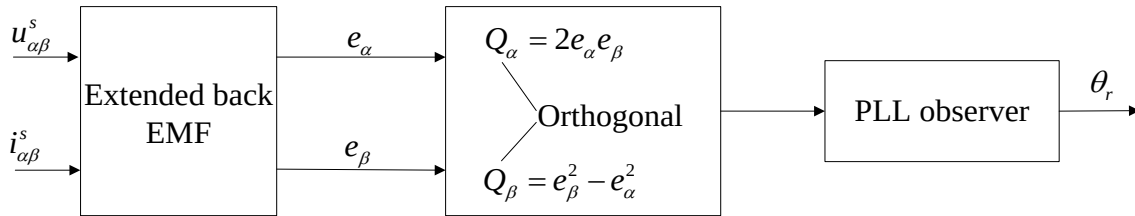


Fig. 1.32. Whole speed range sensorless control based on extended back EMF model.

1.2.4.6 Incremental inductance measurement

According to the previous introduction on the saliency tracking based sensorless control, the incremental inductance values, i.e. L_{dh}, L_{qh}, L_{dqh} , are very important for evaluating the sensorless performance. As these inductances are used in HF signal injection based sensorless control, they are also referred to as HF inductances in the following sections.

There are many methods for obtaining these parameters as shown in Fig. 1.33. The FEA technique can be employed to predict the incremental inductances. However, as the design data are not always available, experimental methods by making use of general purpose inverter drives are required. The HF inductances can be obtained from the flux linkage-current characteristic by post-processing [LI19]. However, the measurement of flux linkages is relatively complicated and requires back electromagnetic force integration operation for standstill tests [HIN17] or spinning the rotor for rotating test [PEL15]. It is preferable if the HF inductances can be measured directly at standstill. The HF inductances can be directly

measured by employing HF signal injection techniques which have been widely used in position sensorless control [XU14], [KUE14], [LI07].

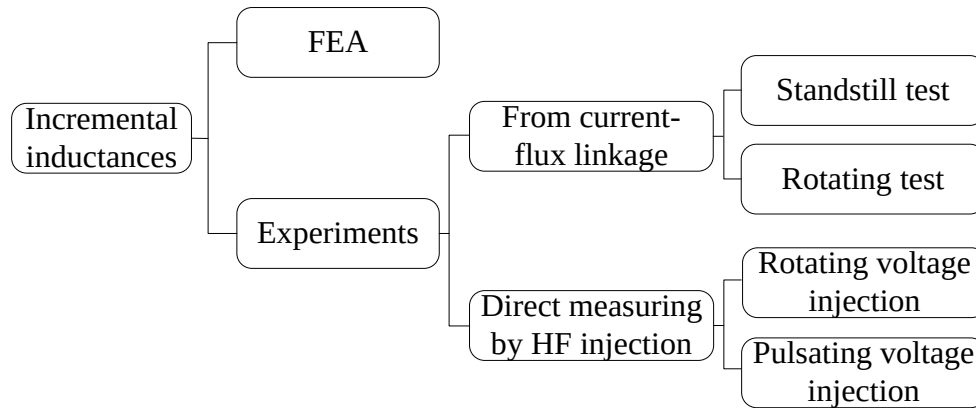


Fig. 1.33. Conventional methods for obtaining the incremental inductances.

In [XU14], the HF sinusoidal rotating voltage is injected to the stationary reference frame for measuring the HF inductances, including the cross-coupling effect. The offset angle $\Delta\theta_m$ caused by the cross-coupling effect is obtained from the difference between the actual rotor position measured from an encoder and the estimated rotor position obtained from a position estimator with the negative sequence current as the input. The dq -axis and cross-coupling HF inductances are then calculated from the offset angle and amplitudes of the positive sequence and negative sequence HF currents. The method in [XU14] requires a position estimator, the design of which is dependent on inductance values and affected by the delay effects.

In [KUE14], the HF pulsating voltage injection technique is used for measuring the HF inductances. The angle offset caused by the cross-coupling effect is also obtained from the difference between the actual rotor position measured by an encoder and the estimated rotor position obtained from a position estimator. The HF inductances are then calculated from the observed angle offset and the measured HF current amplitudes due to pulsating voltage injections. The position estimator is also required and the measured HF current amplitudes are affected by the delay effects which will result in errors in the calculated HF inductances. To overcome the drawbacks of using a position estimator, an open-loop type measuring method is proposed in [LI07] by injecting HF pulsating voltages in the d -axis and q -axis which are obtained from an encoder. The HF inductances are then calculated with the measured HF current amplitudes on the dq -axes. This method does not require a position estimator. However, accuracies of the calculated inductances are also affected by the delay effects. The existing

methods for obtained the HF inductances with HF voltage injection techniques are listed in Table 1.9.

Table 1.9 Comparison of HF inductance measuring methods

Methods		Implementation	Drawbacks
By FEA		Simulation	Require design data
By experiments	Flux linkage-current trace [LI19]	$(\psi(n+1) - \psi(n))/\Delta i$	Low accuracy Complicated
	HF rotating voltage injection [XU14]	$L_h = f(I_p, I_n, \Delta\theta_m)$	Observer required Delay sensitive
	HF pulsating voltage injection [CHE14]	$L_h = g(I_{dh}, I_{qh}, \Delta\theta_m)$	Observer required Delay sensitive

1.3 Sensorless Torque Control

Precise torque control of PMSM is required in many applications, such as robotics, elevators and electric vehicles [ZHU07]. Accurate torque feedback of the machine is essential for both control and monitoring purposes. The torque in IPMSM is composed of permanent magnet torque, reluctance torque and cogging torque as shown in Fig. 1.34. The cogging torque is caused by the interaction of permanent magnet magnetic motive force (MMF) and the air gap permeance harmonics due to slotting [ZHU00]. The cogging torque is not investigated here due to relatively low magnitude.

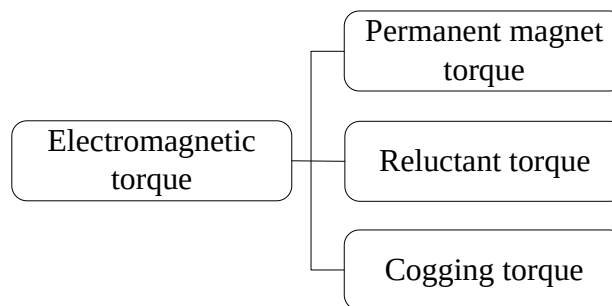


Fig. 1.34. Composition of electromagnetic torque.

There are usually two ways to control torque, i.e. field oriented control (FOC) and direct torque control (DTC). For FOC, the torque is indirectly controlled by controlling the dq -axis currents, while for DTC, the torque is directly controlled by regulating the stator flux linkage and the torque angle by selecting the optimal voltage vector according to the torque and flux

linkage errors [TAK86]. Some general differences are summarised in Table 1.10 according to existing literature.

Table 1.10 Comparison between FOC and DTC in PMSMs

	DTC	FOC
Original paper	[TAK86]	[BLA72]
Controlled objects	Flux linkage	Dq -axis currents
PWM frequency	Variable or fixed (DTVC)	Fixed
Dynamics	High	Medium (depending on bandwidth)
Core modules	Flux linkage observer/ Vector selection	Current regulators
Position dependent	Required	Required
Torque ripple	Medium	Low

Theoretically, the DTC method can achieve faster response compared to the FOC method. For position sensorless control at standstill and low speed, sensorless rotor position estimation is the common requirement for either DTC or FOC. In the thesis, FOC is adopted as the way for investigating the sensorless torque control topic.

Sensorless torque control means that both the rotor position and the torque are obtained without hardware sensors. Hence, it is also essential to estimate the torque, including electromagnetic torque and load torque. The electromagnetic torque is usually calculated using general torque equation [YAM16], while the load torque is estimated with mechanical model as shown in Fig. 1.35.

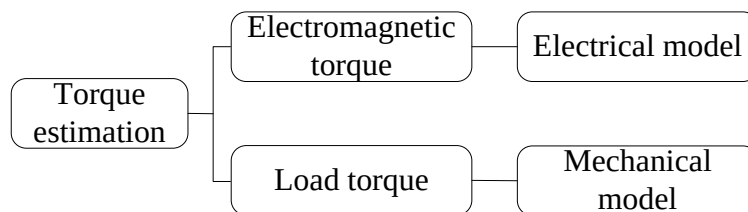


Fig. 1.35. Torque estimation in torque control.

1.3.1 Torque control methods

1.3.1.1 Field oriented control (FOC)

General torque equation in the synchronous rotating reference frame is given by

$$T_e = \frac{3p}{2} [\psi_{pm} i_q + (L_d - L_q) i_d i_q] \quad (1.77)$$

where ψ_{pm} is the permanent magnet flux linkage, L_d, L_q are the dq -axis inductances. The torque is composed of permanent magnet torque and reluctance torque. The torque can be controlled by controlling i_d, i_q respectively, which is the core principle of FOC. The dq -axis currents are then controlled by dq -axis voltages, which are output by closed-loop current controllers as shown in Fig. 1.36.

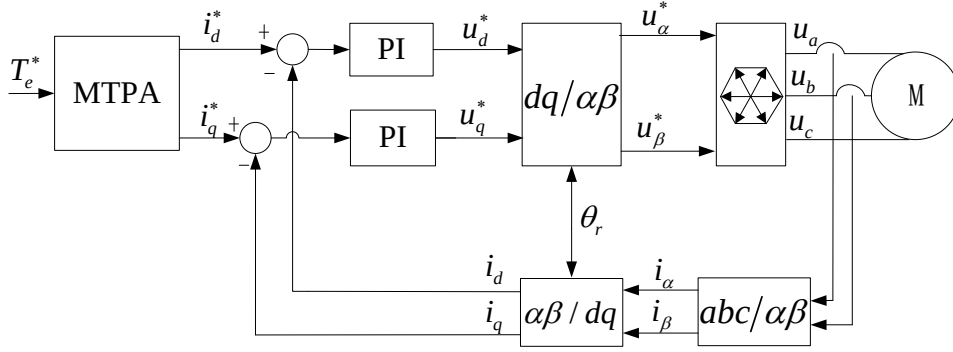


Fig. 1.36. FOC control structure.

To achieve minimum joule loss, i.e. $\sqrt{i_d^2 + i_q^2} * R_s$, the torque to current amplitude ratio as shown in (1.78) should be maintain maximum under a given torque. The trace of the current references which meet such requirement is called maximum torque per ampere (MTPA) trace as shown in Fig. 1.37.

$$\frac{T_e}{\sqrt{i_d^2 + i_q^2}} = \frac{3p}{2\sqrt{i_d^2 + i_q^2}} [\psi_{pm} i_q + (L_d - L_q) i_d i_q] \quad (1.78)$$

The current references can also be selected by producing the minimum stator flux linkage for a given torque as expressed in (1.75). As the applied voltage is approximately proportional to the flux linkage when resistance voltage drop is neglected, the current trace with maximum torque flux ratio is called maximum torque per voltage (MTPV) as shown in Fig. 1.37.

$$\frac{T_e}{\sqrt{(\psi_{pm} + L_d i_d)^2 + (L_q i_q)^2}} = \frac{3p[\psi_{pm} i_q + (L_d - L_q) i_d i_q]}{2\sqrt{(\psi_{pm} + L_d i_d)^2 + (L_q i_q)^2}} \quad (1.79)$$

In actual applications, the available current and voltage are limited by the machine and inverter capabilities expressed in (1.80), where U_{max} , I_{max} are the maximum available voltage and current amplitudes.

$$\sqrt{(\psi_{pm} + L_d i_d)^2 + (L_q i_q)^2} \leq U_{max}, \sqrt{i_d^2 + i_q^2} \leq I_{max} \quad (1.80)$$

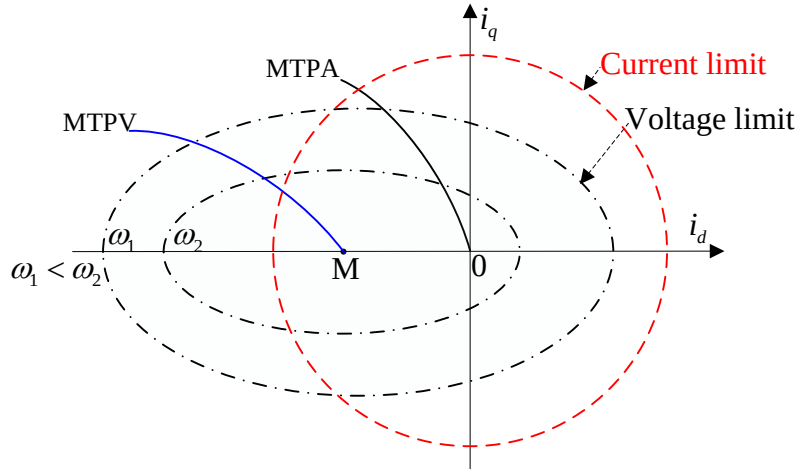


Fig. 1.37. Optimal operating ranges for FOC in the i_d - i_q plane.

For FOC, the torque is controlled in open-loop type according to the MTPA curve, torque feedback is not required. However, the knowledge of the developed torque can be used for improving performance in many applications, such as lift gates, test benches, electronic looms, or simply for diagnostic purposes. The torque feedback can also be used for regulating the torque reference in [CHE10].

The rotor position used in FOC can be obtained by position sensors or sensorless algorithms. As described in the previous section, the saliency model and fundamental model based sensorless methods can be used. For sensorless torque control at standstill and low speed, the saliency model is desirable and has been widely used in industrial applications.

1.3.1.2 Direct torque control (DTC)

DTC technique which was originally designed for induction machines, was proposed in [TAK86]. It has now been widely employed in the control of PMSM due to its fast response and simple structure [TAN04] [BUJ04] [SWI02].

The torque equation can be expressed as the cross-product of the stator flux linkage vector and current vector as

$$T_e = \frac{3p}{2} \vec{\psi}_s \times \vec{I}_s = \frac{3p}{2} \psi_s I_s \sin\left(\varphi + \frac{\pi}{2} - \delta\right) = \frac{3p}{2} \psi_s I_s \cos(\delta - \varphi) \quad (1.81)$$

where δ is angle between the d-axis and the stator flux linkage vector, φ is the angle between the q-axis and the stator current vector as shown in Fig. 1.38.

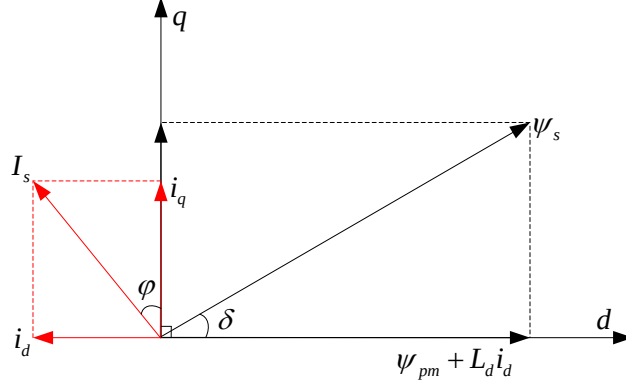


Fig. 1.38. Vector diagram of the PMSM in the synchronous rotating reference frame.

As can be seen, $(\varphi + \frac{\pi}{2} - \delta)$ is the angle between the stator flux linkage vector and the stator current vector. The torque equation can further be expressed as

$$T_e = \frac{3p}{2} \psi_s I_s \sin\left(\varphi + \frac{\pi}{2} - \delta\right) = \frac{3p\psi_s}{4L_d L_q} \left[2\psi_{pm} L_q \sin(\delta) + \psi_s (L_d - L_q) \sin(2\delta) \right] \quad (1.82)$$

For the machine with non-salient pole, i.e. $L_d = L_q$, the torque is proportional to the stator flux linkage ψ_s and $\sin(\delta)$. It means that the when the amplitude of stator flux linkage is fixed, the torque can be controlled by increasing and decreasing the torque angle δ . As the rotor position changes slower than the stator flux linkage vector, the torque can be controlled by fast changing the amplitude or phase of the flux linkage vector. The increment of stator flux linkage vector $\Delta\vec{\psi}_s$ during a control period Δt can be expressed as

$$\Delta\vec{\psi}_s = (\vec{u}_s - \vec{i}_s R_s) \Delta t \quad (1.83)$$

where $\vec{u}_s$ is the applied stator voltage vector, $\vec{i}_s$ is the stator current vector. When the voltage drop on the resistance is neglected, the increment of stator flux linkage vector is along the voltage vector and proportional to the magnitude of the voltage vector. For a three phase inverter, the eight voltage vectors (including zero vectors) can be selected as shown in Fig. 1.39(a). Each voltage vector is corresponding to a combination of the three inverter bridges.

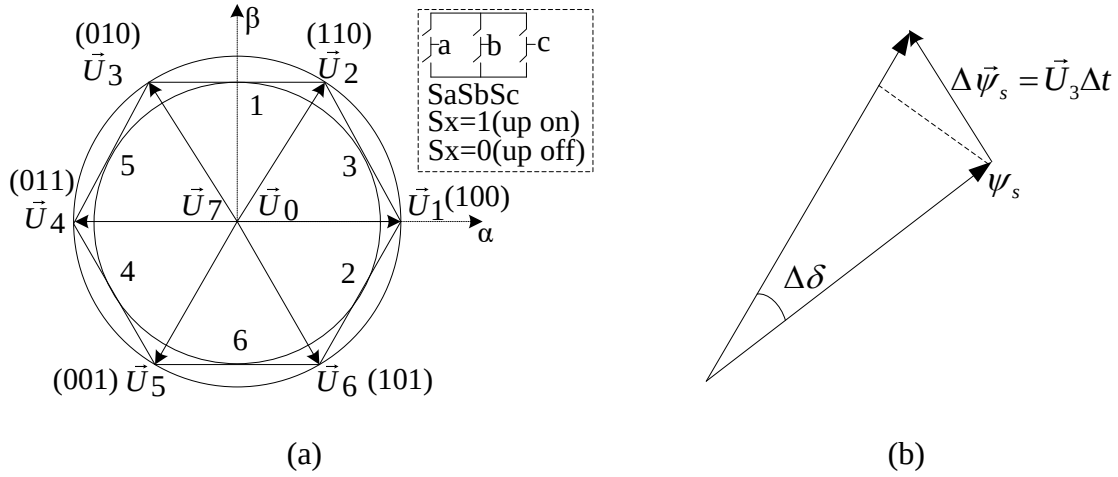


Fig. 1.39. Space vectors in DTC. (a) 7 voltage vectors. (a) Stator flux linkage increment caused by vector $\vec{U}_3$.

These voltage vectors can change the amplitude or direction of the stator flux linkage vector according to the location of the flux linkage. Fig. 1.39(b) shows the flux linkage increment caused by a voltage vector $\vec{U}_3$. Hence, a lookup table can be made to store the optimal voltage vector for different positions of the flux linkage. However, with lookup table (hysteresis type), the PWM switching frequency is not fixed and complexity of the drive circuit design is increased.

Alternatively, the direct torque control with space vector modulation (DTC-SVM) with fixed switching frequency was introduced [BUJ04]. In DTC-SVM, the voltage reference is generated from closed-loop PI controllers [BUJ04] or predictive/dead-beat controllers [LEE11] as shown in Fig. 1.40. The voltage vector is then synthesized by space vector modulation.

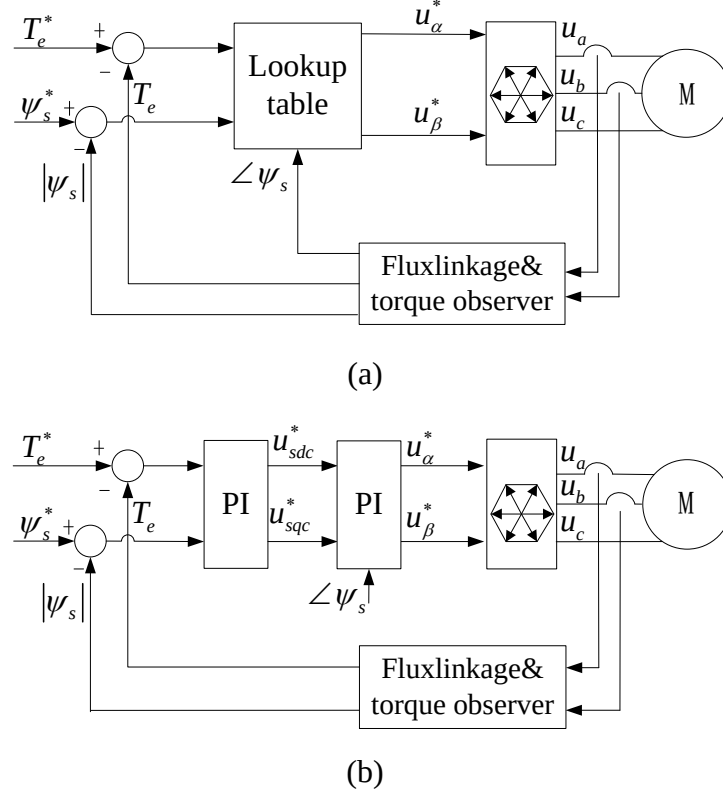


Fig. 1.40. DTC control structure. (a) Lookup table based DTC with variable switching frequency. (b) DTC-SVM-fixed switching frequency.

1.3.2 Torque estimation

1.3.2.1 Electromagnetic torque estimation

Conventionally, the machine torque can be estimated from the product of stator flux linkage and current [BOL08]. The stator flux linkage is usually observed from the terminal voltages and phase currents. There are two kinds of flux linkage observers which are all based on the fundamental electrical models, i.e. voltage and current models [CHE03], [HEJ11], [CHE20].

$$\begin{cases} T_e = \frac{3p}{2} (\psi_\alpha i_\beta - \psi_\beta i_\alpha) & \text{Voltage model} \\ T_e = \frac{3p}{2} [\psi_{pm} i_q + (L_d - L_q) i_d i_q] & \text{Current model} \end{cases} \quad (1.84)$$

The voltage model calculates the stator flux linkage by integrating the back electromagnetic force (EMF) which is the difference between the terminal voltage and resistance voltage drop [CHE03]. This method is effective at medium and high speed ranges and only requires the resistance parameter, but has deteriorated performance at standstill and low speed due to reduced amplitude of the EMF signal [HEJ11] and voltage distortions caused by the inverter

nonlinearities [CHE20]. The current model based flux linkage observer calculates the flux linkage from the summation of the permanent magnet flux and the stator current induced flux in the synchronous rotating reference frame. It is not dependent on the back EMF signal and thus has good performance at standstill and low speed. However, it requires dq-axis apparent inductances [XU14] which vary with currents, and the rotor position which is used to identify the dq -axes.

(a) Torque estimation at standstill and low speed

At standstill and low speed, the conventional parameter identification methods have deteriorated performance due to low amplitude of the back EMF signal. To reduce the dependency of torque estimation on the fundamental model at standstill and low speed, high frequency (HF) signal injection which is superimposed on the fundamental excitations is employed for online inductance parameter identification [KAN21], [MAR20]. However, the identified incremental inductances cannot be directly used for torque estimation in (1.84).

In [MAR20], the dq-axis incremental inductances are online estimated by injecting the HF pulsating current signal and then used to calculate the torque by the empirical equation between torque and incremental inductances. This method can be used without interfering with the normal operation of the machine. However, the injected HF signal in [MAR20] is merely used for identifying the incremental inductances while the rotor position should also be provided by a position sensor. For reducing the hardware cost, the injected HF signal can be used for both sensorless rotor position estimation and online incremental inductance identification simultaneously, which are further used to estimate the torque.

(b) Torque estimation at medium and high speed

As presented before, at medium and high speed, the fundamental model has good performance due to high signal amplitude of the back EMF signal. Hence, the flux linkage observer based on voltage model has high accuracy as

$$\psi_s = \int (u_{\alpha\beta} - i_{\alpha\beta} R_s) dt \quad (1.85)$$

Thus, only the resistance parameter which varies with temperature needs to be online estimated. There are many methods for online resistance identification. In [MOB04], a MARS-based resistance identification method for sensorless control is proposed. In [ICH06], an RLS method was employed for estimating the winding resistance and dq-axis inductances of surface-mounted and interior PMSMs. In [LIU11], a novel method by injecting a short pulse of negative d-axis current is proposed for parameter identification with the measured terminal

currents, voltages and rotor speed. The torque estimation strategies at different speed ranges are shown in Fig. 1.41.

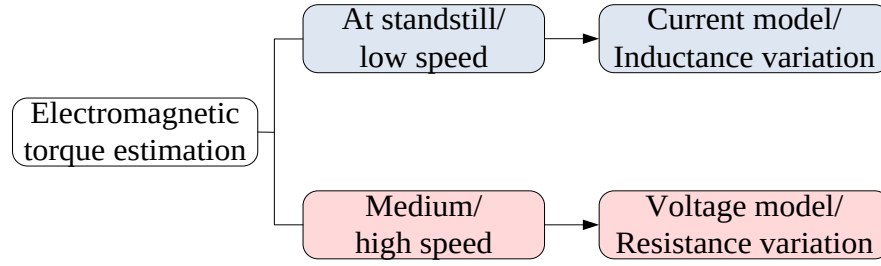


Fig. 1.41. Torque estimation strategies at different speed ranges.

1.3.2.2 Load torque estimation

For improving the control performance and monitoring the load condition, load torque estimation is also very important. Identification of the load torque has many benefits for improving the performance of machine drive systems. In speed control mode, the speed ripple caused by the load disturbance can be suppressed by employing load torque feedforward control technique [RUD13], [KIM10].

Unlike the electromagnetic torque which is calculated based on electrical model, the load torque is determined by the load characteristics. The load torque is usually estimated from the speed response and mechanical parameters of the drive system. The motion equation of the drive system is given by

$$\frac{J}{p} \frac{d\omega_e}{dt} = T_e - T_L \quad (1.86-a)$$

$$\omega_e(s) = \frac{p}{Js} [T_e(s) - T_L(s)] \quad (1.86-b)$$

where ω_e is the electrical speed, p is the pole-pair number, J is the moment of inertia of the motor, T_e is the electromagnetic torque, T_L is the load torque which can be estimated with the observer in Fig. 1.42.

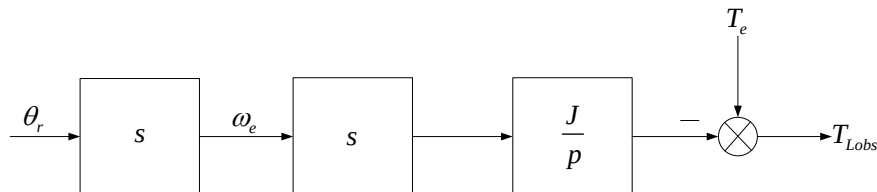


Fig. 1.42. Load torque estimation with basic mechanical model.

The load torque observer in Fig. 1.42 needs to differentiate the position to obtain the speed, and differentiate the speed to obtain the acceleration. The differentiation operation may introduce much noise to the estimation.

To avoid differentiation operation, a separate load torque [YAN12b] observer is used as shown in Fig. 1.43.

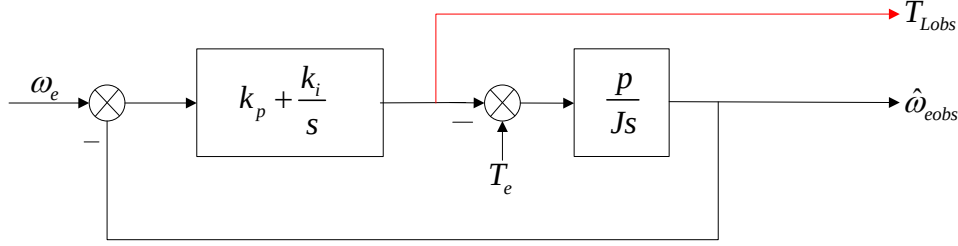


Fig. 1.43. Load torque estimation with basic mechanical model.

As can be seen, the transfer function between the actual speed ω_e and the observed speed $\hat{\omega}_{eobs}$ is

$$\hat{\omega}_{eobs} = \frac{p(k_p s + k_i)}{Js^2 + p(k_p s + k_i)} \omega_e + \frac{ps}{J^2 s^3 + p^2(k_p s + k_i)} T_e \quad (1.87)$$

Combining (1.86), (1.87), the transfer function between T_L and T_{Lobs} is

$$T_{Lobs} = -\left(k_p + \frac{k_i}{s}\right)(\omega_e - \hat{\omega}_{eobs}) = \frac{p(k_p s + k_i)}{Js^2 + p(k_p s + k_i)} T_L \quad (1.88)$$

With this method, the load torque can be estimated with less noise, but the phase delay is introduced. Besides, the bandwidth of the estimation is dependent on the inertia parameter which should be online estimated.

The estimated load torque can be used to suppress the speed ripple in FOC control by load current feedforward control as shown in Fig. 1.44.

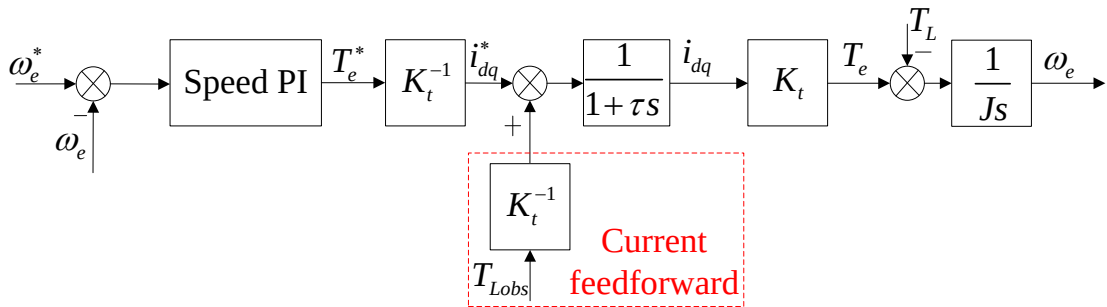


Fig. 1.44. Current feedforward control with the observed load torque.

1.4 Scope of Research and Contributions

Sensorless torque control for IPMSM have been broadly reviewed in this chapter from two aspects, i.e. the sensorless position estimation and the torque control. Fig. 1.45 shows the scope of research for each chapter.

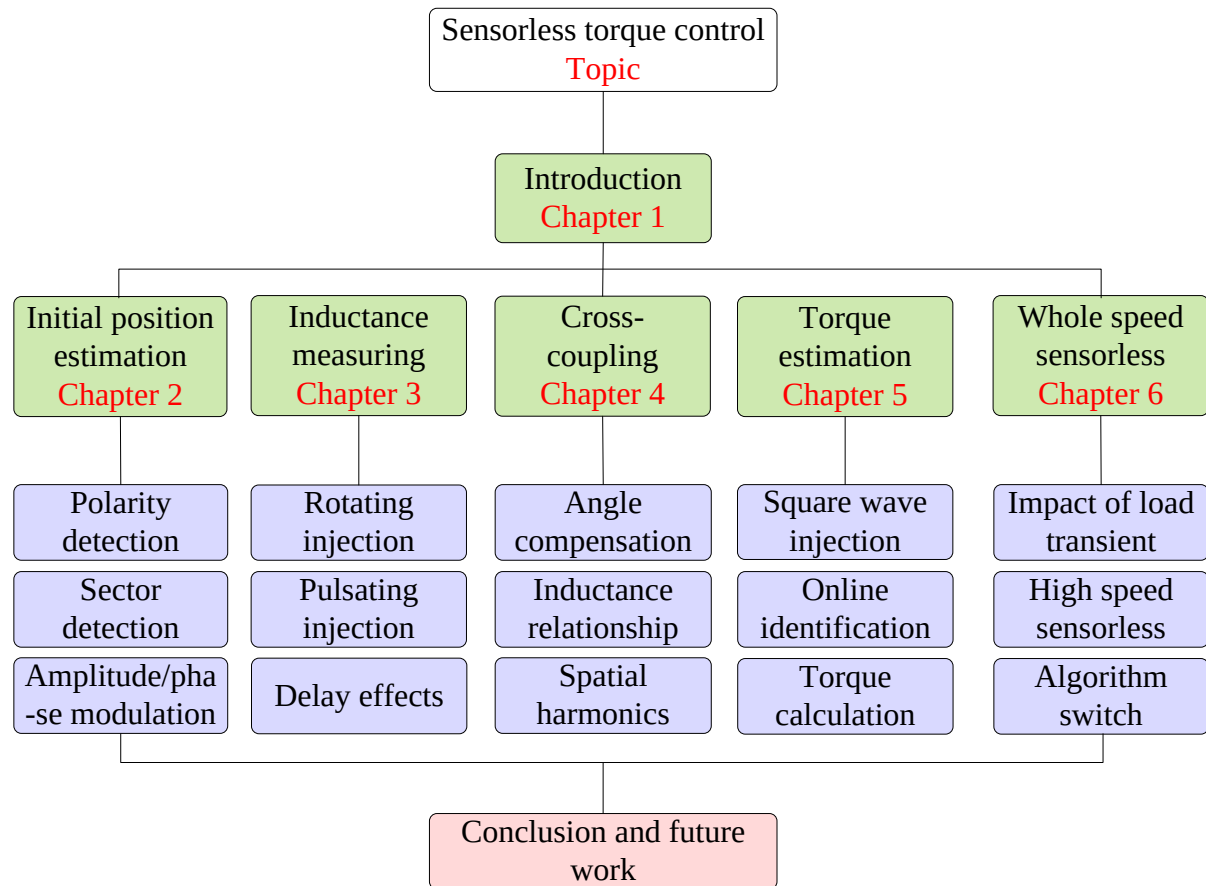


Fig. 1.45. Scope of research in this thesis.

Scope of the research for each chapter is:

Chapter 2 proposes a novel method for sensorless initial position estimation with fast response based on HF sinusoidal rotating voltage injection.

Chapter 3 proposes a simple method for obtaining the incremental inductances of IPMSM based on both rotating and pulsating voltage injection techniques. The method does not require closed-loop observer and is free of the delay effects caused by PWM voltage synthesis and current sampling, filtering.

Chapter 4 compares existing cross-coupling effect compensation in saliency tracking based sensorless control and proposes a novel method for offline measuring the error caused by the cross-coupling effect.

Chapter 5 introduces a simultaneous sensorless rotor position and torque estimation method based on HF square wave voltage injection for IPMSM at standstill and low speed, considering the cross-coupling effect.

Chapter 6 investigates the impacts of load transient on sensorless position estimation. A whole speed range sensorless method is proposed based on HF rotating voltage injection.

Chapter 7 summarizes this research work and gives discussions about future work.

The major contributions of this thesis are summarised as

- 1) Evaluating the relationship between the amplitude modulation and phase modulation in HF rotating and pulsating voltage injection based sensorless control methods in Chapter 2.
- 2) A novel method for fast detecting the initial rotor position based on HF rotating voltage injection is proposed in Chapter 2.
- 3) A method for measuring the incremental inductances without position observer and delay compensation is proposed in Chapter 3.
- 4) A novel method for offline measuring the position error caused by the cross-coupling effect is proposed in Chapter 4.
- 5) A simultaneous position and torque estimation method for IPMSM at standstill and low speed based on HF square wave voltage injection is developed in Chapter 5.
- 6) A whole speed range sensorless operation method for IPMSM is proposed in Chapter 6 with only one close-loop position observer which is shared by the saliency model and fundamental model.

CHAPTER 2

INITIAL ROTOR POSITION ESTIMATION AND STARTUP METHOD BASED ON AMPLITUDES OF HIGH FREQUENCY CURRENTS

2.1 Introduction

Interior permanent magnet synchronous machines (IPMSMs) have been widely used in both domestic and industrial applications due to high torque density and power density [BOL08], [HEJ11], [JAN95]. Sensors such as optical encoders and resolvers are used to provide rotor position, which will increase cost and complexity. Thus, the sensorless control of IPMSMs which estimates the position from stator currents and terminal voltages has been proposed over the past decades.

At medium and high speeds range, back electromotive force (EMF) or flux linkage induced by permanent magnet are observed to calculate rotor position [BOL08], while at zero or low speed range, saliency detection based methods by high frequency (HF) signal injection show better performance [DEG98], [BIA07]. These two kinds of methods can be combined to provide whole speed range sensorless operation [WAN13], [YAN17]. For applications that mainly operate at high speed, fast and reliable sensorless startup is the most important.

Generally, sensorless startup methods can be divided into the open-loop type which generates current reference based on predetermined angle trace [BOL15], [FAT08] and the closed-loop type which obtains rotor position from sensorless algorithms. Open-loop startup methods are employed when the machine has low saliency level and light load [WAN12], [YU18], but they suffer from instability and low adaptability to the load variation. The closed-loop startup has higher reliability as the position is obtained continuously based on real time circuit variables. For all the sensorless start-up methods, the first step is to identify the initial position.

There are mainly two kinds of methods to detect the position at zero or low speed region. The pulse injection method is robust but needs separate injection cycles which decreases bandwidth and results in shrill noise [ZHO20]. The other kind of method injects continuous HF signal to the stator windings, such as sinusoidal rotating signal which is injected to the synchronous stationary reference frame (SSRF) [JAN95] and pulsating signal which is injected

to the synchronous rotational reference frame (SRRF) [LUO16]. Usually, the signal injection into SSRF is preferred. This is because the SRRF is established by the estimated position which is obtained by converging the position estimation error to zero, while the ambiguous convergent point would make the system unstable [WAN19b]. The initial position is obtained by driving the position error signal to zero with a proportional integral (PI) controller, the convergence time of which may be long. The tuning of the PI parameters is dependent on the negative sequence current and machine parameters. In this chapter, a simple initial position estimation method is proposed based on the sinusoidal rotating voltage injection. In the conventional method, the three phase HF current amplitudes are regarded as analog signals, while in the proposed method, the HF current amplitudes are regarded as digital signals. The values of HF current amplitudes are not important. What makes sense is the relation among the three amplitudes. A rough position signal can be obtained fast than conventional method by comparing the amplitudes.

The magnet polarity is required after initial rotor direction is detected. Generally, three methods can be used for polarity detection, i.e. double pulses injection [TAN18], HF current amplitude variation [GON13] and secondary saliency [RAC08]. The pulse injection method is simple but should be performed as independent process. The amplitude variation method which compares HF current amplitudes under different fundamental currents, may produce driving torque if there exists position estimation error. The method utilizing secondary saliency is fast but of low signal to noise ratio (SNR). After the initial position has been identified, the machine can be driven to high speed until the back EMF based sensorless methods take control.

An associated problem with signal injection is the noise caused by HF currents. Generally, lower injection magnitude or lower frequency can reduce the noise at the cost of reducing the SNR and bandwidth [WAN17a]. Hardware measures such as current over-sampling or analog filtering [JUN15] and software methods such as disturbance observer [YAN12b] are introduced to mitigate the impact of reduced amplitude. For square wave pulsating voltage injection, the PWM switching frequency can be selected beyond the maximum limit of human ear perception. But this method is only practical for few wide band gap semiconductor devices [KIM12]. Pseudo-random injection [WAN16], [WAN17b] and variable frequency injection [DU20] can achieve low noise effect by distributing the power spectrum density (PSD) to multiple peaks. In this thesis, a multi-frequency sinusoidal rotating voltage injection based sensorless method is proposed to provide low noise startup. Similarly, as the PSD has multiple peaks, the shrill noise is largely relieved.

The chapter is organized as follows. Firstly, conventional initial position estimation methods based on rotating voltage injection and pulsating voltage injection are reviewed. The phase modulation and amplitude modulation phenomena are analysed. Then, the proposed initial position estimation method including polarity detection is explained in detail. The discrete rotor position at low speed can be obtained by comparing the HF current amplitudes and used for startup. To further improve the startup performance continuous position can also be obtained by using vector tracking observer. Experiments are carried out to verify the effectiveness of the proposed method. Thirdly, the proposed method is used for reducing noise problem during initial position estimation with multi-frequency signals injection. Finally, conclusions are given.

2.2 Conventional Methods for Initial Rotor Position Estimation

Sufficient starting torque at standstill is vital for successful startup of electrical machines. The induction machines can generate considerable starting torque as long as alternative currents are applied on the stator windings according to the torque-slip curve. However, for PMSM, the flux linkages induced by the stator currents should be synchronized with the flux linkages induced by the permanent magnets to produce stable electromagnetic torque. Thus, the rotor position is required for the inverter to apply appropriate currents to the stator windings with optimum phase and amplitudes especially when large startup torque is needed.

The initial position is usually obtained by extracting the stator voltages or currents which contain rotor position. However, at standstill, the widely used back EMF model fails due to low signal magnitude. Instead, the saliency variations of the machine caused by structure or saturation are made use of. The PMSM with non-salient pole is not considered here due to inherently low saliency level at standstill, while the IPMSM with high initial saliency is the main focus of this thesis.

The widely used saliency detection methods with continuous signal injection are high frequency sinusoidal rotating signal injection and pulsating signal injection, which will be investigated later in this chapter.

2.2.1 Saliency model

The voltage equations of PMSM with salient pole in the SRRF are as

$$\begin{bmatrix} u_d^r \\ u_q^r \end{bmatrix} = \begin{bmatrix} R_s + L_d p & -\omega_e L_q \\ \omega_e L_d & R_s + L_q p \end{bmatrix} \begin{bmatrix} i_d^r \\ i_q^r \end{bmatrix} + \omega_e \begin{bmatrix} 0 \\ \psi_{pm} \end{bmatrix} \quad (2.1)$$

where u_d^r, u_q^r and i_d^r, i_q^r are the voltages and currents in the synchronous reference frame, ω_e is the electrical speed, R_s is the stator resistance. By superimposing HF voltages on the fundamental voltages, HF currents will be induced in the stator windings. For easy separation of the HF components with digital filters, the frequency of the injection signal should be higher than the fundamental frequency. For example, for a machine with 50Hz rated frequency, the frequency for HF signal can vary from 200Hz to several kilohertz. The equivalent circuits in the dq -axis reference frame for describing the HF variables are as

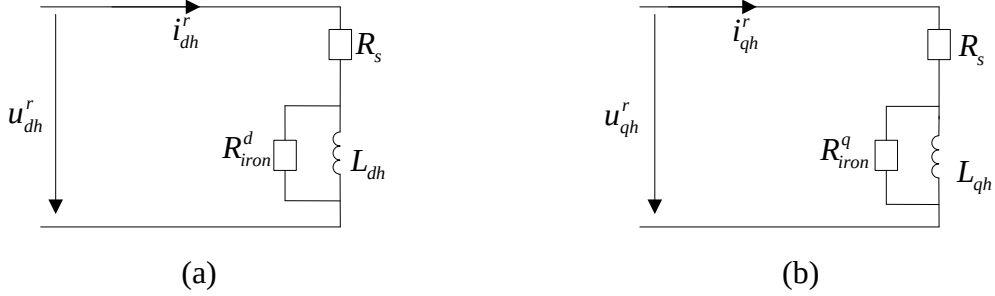


Fig. 2.1. Equivalent HF circuits in the dq -axis. (a) d -axis. (b) q -axis.

where $u_{dh}^r, u_{qh}^r, i_{dh}^r, i_{qh}^r$ are the HF voltages and currents in SRRF, R_{iron}^d, R_{iron}^q are the iron loss reflected series resistances in the dq -axes, L_{dh}, L_{qh} are the incremental inductances which are related to the magnetic saturation levels. The superscript “ r ” means the SRRF in following equations. The back EMF is not included in the HF equivalent circuits as its frequency is relatively lower than the injected HF signal. Cross-coupling effect inductances which describe the impact of magnetic saturation on the quadrature axes, are not considered in this model. Since the reactance of HF equivalent circuit is much higher than the stator resistance and the iron-loss reflected resistance, the resistances in HF model are often neglected. Thus, the HF voltage equations of IPMSM in the SRRF are expressed as

$$\begin{bmatrix} u_{dh}^r \\ u_{qh}^r \end{bmatrix} = \begin{bmatrix} L_{dh} & 0 \\ 0 & L_{qh} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{dh}^r \\ i_{qh}^r \end{bmatrix} \quad (2.2)$$

For better analysing the HF currents, the voltage equations in the two phase stationary reference frame can be obtained by Park transform as

$$\begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} = \begin{bmatrix} L_{\Sigma} - L_{\Delta} \cos(2\theta_r) & -L_{\Delta} \sin(2\theta_r) \\ -L_{\Delta} \sin(2\theta_r) & L_{\Sigma} + L_{\Delta} \cos(2\theta_r) \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{\alpha h}^s \\ i_{\beta h}^s \end{bmatrix} \quad (2.3)$$

where $L_{\Sigma} = \frac{L_{dh} + L_{qh}}{2}$, $L_{\Delta} = \frac{L_{qh} - L_{qh}}{2}$, θ_r is the rotor position. The incremental inductances in the stationary reference frame are varying with rotor position and are coupled with the quadrature axes as shown in Fig. 2.2. Thus, the rotor position can be obtained by detecting the inductances. The saliency detection based sensorless estimation methods are all based on the two models in (2.2) and (2.3). However, it is difficult to directly detect the inductance and extract the rotor position. In next sections, different kinds of signals are injected to the stator winding to indirectly derive the rotor position. The rotor position demodulation methods are different depending on different injection signals.

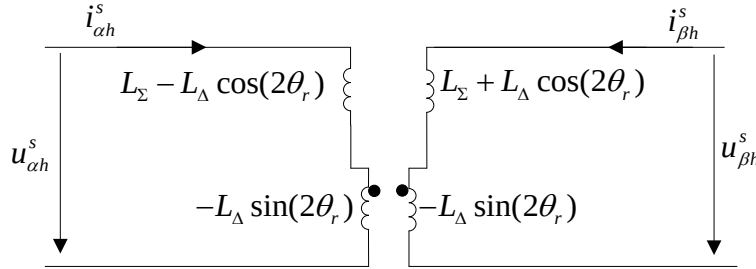


Fig. 2.2. Equivalent HF circuits in the $\alpha\beta$ -axis without considering resistances.

2.2.2 High frequency rotating and pulsating voltage injection methods

The most commonly used methods for position estimation are rotating voltage injection and pulsating voltage injection. In this chapter, their general principles will be presented. The difference and relationship between the two methods are investigated in detail.

2.2.2.1 High frequency rotating voltage injection

HF sinusoidal rotating voltages $u_{\alpha h}^s, u_{\beta h}^s$ are injected to the SSRF as

$$\begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} = \begin{bmatrix} u_h \cos(\theta_h) \\ u_h \sin(\theta_h) \end{bmatrix}, \theta_h = \omega_h t \quad (2.4)$$

where u_h, θ_h, ω_h are the amplitude, angle and frequency of the injected signal. The induced HF currents $i_{\alpha h}^s, i_{\beta h}^s$ at low speed can be solved by substituting (2.4) in (2.3) as

$$\begin{aligned} i_{\alpha h}^s + ji_{\beta h}^s = & \underbrace{I_p e^{j(\theta_h - \pi/2)}}_{\text{Positive sequence current}} + \underbrace{I_n e^{-j(\theta_h - 2\theta_r - \pi/2)}}_{\text{Negative sequence current}} \end{aligned} \quad (2.5)$$

$$I_p = \frac{(L_{dh} + L_{qh})u_h}{2\omega_h L_{dh} L_{qh}}, I_n = \frac{(L_{qh} - L_{dh})u_h}{2\omega_h L_{dh} L_{qh}} \quad (2.6)$$

where I_p, I_n are the current amplitudes of positive and negative sequence currents, ω_e is the rotor electrical rotating frequency. The current responses in (2.5) are composed of positive sequence current and negative sequence current. Conventionally, there are two methods for extracting the rotor position as shown in Fig. 2.3. The heterodyning method moves positive sequence component to high frequency domain and filters it out by low pass filter (LPF). The remaining error signal ε_1 is sent to a phase locked loop (PLL) observer for position estimation. The synchronous frame filter (SFF) method filters off the positive sequence current using a high pass filter (HPF) implemented in a reference frame which is synchronous with the positive sequence current [DEG98]. The error signals ε_1 and ε_2 have the same expression as

$$\varepsilon = I_n \sin(2\theta_r - 2\hat{\theta}_r) = I_n \sin(2\tilde{\theta}_r) \quad (2.7)$$

where $\hat{\theta}_r$ is the estimated rotor position and $\tilde{\theta}_r = \theta_r - \hat{\theta}_r$ is the position estimation error.

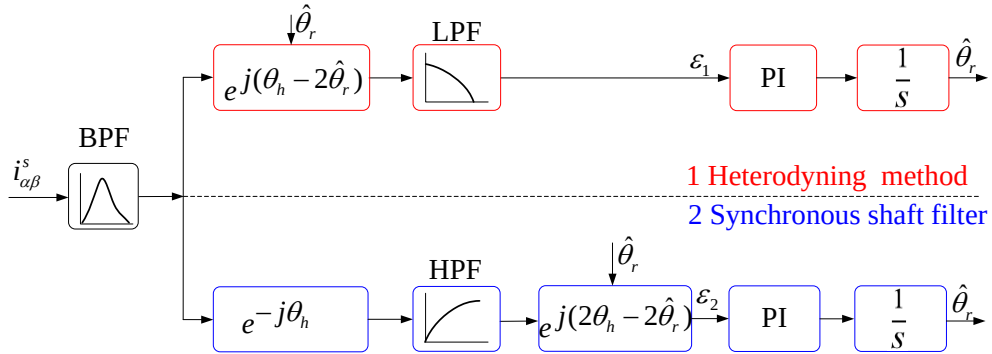


Fig. 2.3 Conventional demodulation methods for rotating HF injection.

The rotor position is contained in the phase of the error signal ε . When $\tilde{\theta}_r$ is small, ε is proportional to $\tilde{\theta}_r$ and sent to a PLL observer as shown in Fig. 2.4(a) where K_p, K_i, K_D are observer parameters, $\hat{\omega}_e, \hat{T}_e$ are the estimated electrical speed and torque, P is the pole-pairs, $\hat{J}$ is the estimated inertia.

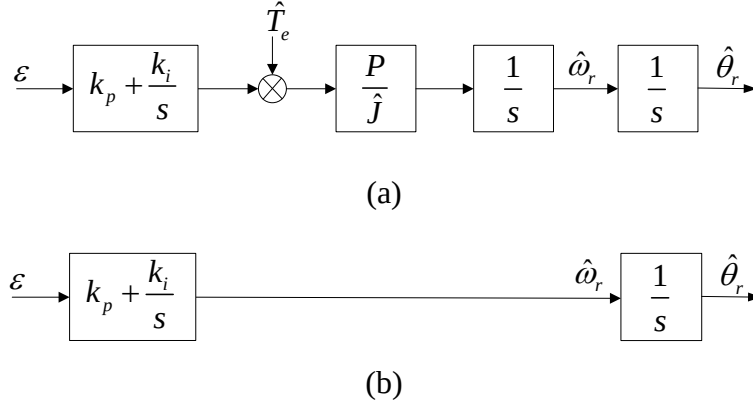


Fig. 2.4 Different PLL observer structures. (a) With mechanical model. (b) Simplified model.

The observer with mechanical model has good disturbance rejection ability to load variation, but is dependent on the estimated torque $\hat{T}_e$ and inertia $\hat{J}$. Instead, the simplified PLL observer in Fig. 2.4(b) is more widely used in actual applications.

2.2.2.2 High frequency pulsating signal injection

For the pulsating voltage injection based sensorless control method, the sinusoidal pulsating voltage as shown in (2.8) is injected to the d -axis of the estimated synchronous rotating reference frame shown in Fig. 2.5.

$$u_{dh}^{\wedge} = u_h \cos(\theta_h), \theta_h = \omega_h t \quad (2.8)$$

where the superscript ' $\wedge$ ' means the estimated synchronous rotating reference frame (ESRRF). If the estimated d -axis is not aligned with the actual d -axis, there will be HF current on the estimated q -axis.

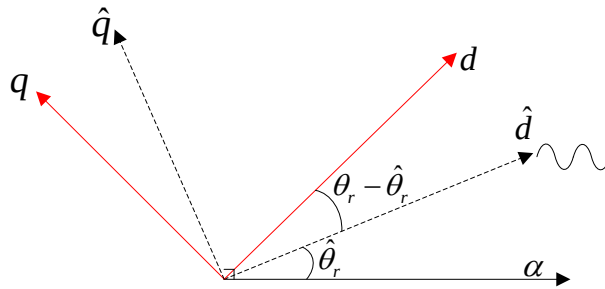


Fig. 2.5 HF pulsating voltage injection in the estimated d -axis.

The HF voltage equations in the estimated reference frame can be expressed as

$$\begin{bmatrix} u_{dh}^{\wedge} \\ u_{qh}^{\wedge} \end{bmatrix} = R^{-1}(\tilde{\theta}_r) \begin{bmatrix} L_{dh} & 0 \\ 0 & L_{qh} \end{bmatrix} R(\tilde{\theta}_r) \frac{d}{dt} \begin{bmatrix} i_{dh}^{\wedge} \\ i_{qh}^{\wedge} \end{bmatrix} \quad (2.9)$$

where $R(\tilde{\theta}_r)$ is the rotating matrix as

$$R(\tilde{\theta}_r) = \begin{bmatrix} \cos(\tilde{\theta}_r) & \sin(\tilde{\theta}_r) \\ -\sin(\tilde{\theta}_r) & \cos(\tilde{\theta}_r) \end{bmatrix} \quad (2.10)$$

By combining (2.8) and (2.9), the HF in the estimated reference frame are solved as

$$\begin{bmatrix} \hat{i}_{dh}^{\hat{r}} \\ \hat{i}_{qh}^{\hat{r}} \end{bmatrix} = \frac{u_h}{\omega_h L_{dh} L_{qh}} \begin{bmatrix} \frac{L_{qh} + L_{dh}}{2} + \frac{L_{qh} - L_{dh}}{2} \cos(2(\theta_r - \hat{\theta}_r)) \\ \frac{L_{qh} - L_{dh}}{2} \sin(2(\theta_r - \hat{\theta}_r)) \end{bmatrix} \sin(\omega_h t) \quad (2.11)$$

It is observed that both the dq -axis HF currents are amplitude modulation waveforms. Their amplitudes can be extracted using synchronous demodulation technique as

$$I_{dh}^{\hat{r}} = LPF(2\hat{i}_{dh}^{\hat{r}} \sin(\omega_h t)) = \frac{u_h (L_{qh} + L_{dh})}{2\omega_h L_{dh} L_{qh}} + \frac{u_h (L_{qh} - L_{dh})}{2\omega_h L_{dh} L_{qh}} \cos(2(\theta_r - \hat{\theta}_r)) \quad (2.12-a)$$

$$I_{qh}^{\hat{r}} = LPF(2\hat{i}_{qh}^{\hat{r}} \sin(\omega_h t)) = \frac{u_h (L_{qh} - L_{dh})}{2\omega_h L_{dh} L_{qh}} \sin(2(\theta_r - \hat{\theta}_r)) \quad (2.12-b)$$

where $I_{dh}^{\hat{r}}, I_{qh}^{\hat{r}}$ are the HF current amplitudes. It is observed and the rotor position is contained in the amplitudes. However, as the d -axis current amplitude has a DC offset $\frac{u_h(L_{dh}+L_{qh})}{2\omega_h L_{dh} L_{qh}}$ which is difficult to remove, the q -axis current amplitude is used for position estimation. The amplitude of the q -axis HF current has the same expression as the error signal in (2.7) for the rotating voltage injection. Thus, the same PLL observers shown in Fig. 2.4 can be used for position estimation.

Although it has been revealed that there is not substantial difference in position estimation performance between the rotating and pulsating injection techniques [BIA07], their position demodulation methods are different. The method for extracting rotor position in rotating injection is phase demodulation, as the rotor information is contained in the phase of the negative sequence current, whereas the method for obtaining rotor information in pulsating voltage injection is amplitude demodulation, as the rotor position is contained in the current amplitude. Comparison of HF rotating and pulsating voltage injection methods is listed in Table 2.1.

Table 2.1 Comparison between HF rotating and pulsating voltage injection methods.

	HF rotating injection	HF pulsating injection
Injection signal	Sinusoidal rotating	Sinusoidal pulsating
Injection frame	Stationary frame	D-axis of the ESRRF
Current responses	Positive & negative sequence carrier currents	Dq-axis HF currents
Demodulation method	Phase demodulation	Amplitude demodulation
Noise level	High	Low

The pulsating voltage injection has relatively lower noise level than the rotating voltage injection as the HF current on the d -axis has less impact on torque ripple. The demodulation methods for the two injection types are different and the internal relationship is investigated in the next section.

2.2.2.3 Phase modulation and amplitude modulation

The signals for rotating and pulsating injections are all sinusoidal waveforms. Why the demodulation methods for rotating and pulsating injections are different? Could phase demodulation also be used for pulsating voltage injection or amplitude demodulation be used for rotating injection? Firstly, to investigate the difference, the pulsating voltage can be divided into forward rotating voltage $\vec{u}_{posi}^{\hat{r}}$ and backward rotating voltage $\vec{u}_{nega}^{\hat{r}}$ as

$$\begin{bmatrix} u_{dh}^{\hat{r}} \\ u_{qh}^{\hat{r}} \end{bmatrix} = u_h \begin{bmatrix} \cos(\theta_h) \\ 0 \end{bmatrix} = \underbrace{\frac{u_h}{2} \begin{bmatrix} \cos(\theta_h) \\ \sin(\theta_h) \end{bmatrix}}_{\vec{u}_{posi}^{\hat{r}}} + \underbrace{\frac{u_h}{2} \begin{bmatrix} \cos(-\theta_h) \\ \sin(-\theta_h) \end{bmatrix}}_{\vec{u}_{nega}^{\hat{r}}} \quad (2.13)$$

The forward and backward rotating voltages can be further transformed to the stationary reference frame as

$$\begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} = \begin{bmatrix} u_{dh}^{\hat{r}} \\ u_{qh}^{\hat{r}} \end{bmatrix} e^{j\hat{\theta}_r} = \underbrace{\frac{u_h}{2} \begin{bmatrix} \cos(\theta_h + \hat{\theta}_r) \\ \sin(\theta_h + \hat{\theta}_r) \end{bmatrix}}_{\vec{u}_{posi}^s} + \underbrace{\frac{u_h}{2} \begin{bmatrix} \cos(\hat{\theta}_r - \theta_h) \\ \sin(\hat{\theta}_r - \theta_h) \end{bmatrix}}_{\vec{u}_{nega}^s} \quad (2.14)$$

$$\vec{u}_{posi}^s = \frac{u_h}{2} e^{j(\theta_h + \hat{\theta}_r)} \quad \vec{u}_{nega}^s = \frac{u_h}{2} e^{j(\hat{\theta}_r - \theta_h)}$$

Thus, the pulsating voltage injection in the estimated d -axis can be seen as two rotating voltage injections which are in opposite directions. Naturally, the HF current responses due to pulsating voltage injection to the estimated d -axis can be calculated by superimposing the current responses by positive sequence voltage and negative sequence voltage injections as

$$i_{\alpha h}^s + j i_{\beta h}^s = \vec{i}_{posi}^s + \vec{i}_{nega}^s = [I_p e^{j\hat{\theta}_r} + I_n e^{j(2\theta_r - \hat{\theta}_r)}] \sin(\theta_h) \quad (2.15)$$

$$\vec{i}_{posi}^s = \frac{I_p}{2} e^{j(\theta_h + \hat{\theta}_r - \frac{\pi}{2})} + \frac{I_n}{2} e^{-j(\theta_h + \hat{\theta}_r - 2\theta_r - \frac{\pi}{2})} \quad (2.16)$$

$$\vec{i}_{nega}^s = \frac{I_p}{2} e^{j(\hat{\theta}_r - \theta_h - \frac{\pi}{2})} + \frac{I_n}{2} e^{-j(\hat{\theta}_r - \theta_h - 2\theta_r - \frac{\pi}{2})} \quad (2.17)$$

where I_p, I_n are defined in (2.6), $\vec{i}_{posi}^s, \vec{i}_{nega}^s$ are the HF currents induced by $\vec{u}_{posi}^s$ and $\vec{u}_{nega}^s$ respectively. From the expressions above, the following conclusions can be drawn.

① The stator HF currents due to pulsating voltage injection in the estimated d -axis are amplitude modulation waveforms according to (2.15).

② The amplitude modulated waveforms in (2.15) are synthesized by the two phase modulated waveforms in (2.16) and (2.17). This is quite similar to the standing wave phenomenon in physics.

③ As the phases of the currents in (2.15) do not contain rotor position, phase demodulation method is not applicable for the pulsating voltage injection based sensorless method.

Thus, for the pulsating voltage injection, only the amplitude demodulation method of the HF currents can be used for position estimation. Whereas for the rotating voltage injection, the stator HF currents are not only phase modulated, but also amplitude modulated which will be revealed and utilized for fast initial position estimation in the next section.

2.3 Proposed Method for Initial Rotor Position Estimation

The aforementioned conventional position estimation method based on rotating voltage injection tracks the negative sequence current phase and are called phase-demodulation method. However, there are several drawbacks. Firstly, it is hard to predetermine the PLL observer parameters as they are related to the saliency ratio and the injection voltage. Secondly, the convergence time may be very long if the observer parameters are set small for stability. Then, due to execution and PWM logic in digital voltage synthesis [WAN19b], there exists a phase delay between actual injection angle and θ_h . This delay is hard to determine since it is related to dead-time effect and current levels. To overcome these problems, a novel initial position estimation method based on the three phase HF current amplitudes is presented in this chapter.

2.3.1 Position estimation by using HF current amplitudes

The commonality of the phase demodulation based methods described in Fig. 2.3 is that they both utilize the phase signal in the negative sequence current. Phase modulated HF currents can also be seen as amplitude modulated signals which is firstly proposed in [L117]. Instead of using two-phase stationary currents, three phase HF currents $i_{abch}^s(i_{ah}^s, i_{bh}^s, i_{ch}^s)$ can be expressed as (2.18) from (2.5) by reverse Clarke transformation.

$$\begin{bmatrix} i_{ah}^s \\ i_{bh}^s \\ i_{ch}^s \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} \\ \frac{1}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i_{\alpha h}^s \\ i_{\beta h}^s \end{bmatrix} = \begin{bmatrix} I_p \sin(\theta_h) + I_n \sin(\theta_h - 2\theta_r) \\ I_p \sin(\theta_h) + I_n \sin(\theta_h - 2\theta_r + 2\pi/3) \\ I_p \sin(\theta_h) + I_n \sin(\theta_h - 2\theta_r - 2\pi/3) \end{bmatrix} \quad (2.18)$$

When the machine is operating at low speed region, the rotor rotating electrical frequency is very low compared to the injection frequency ω_h . Thus, the three phase HF currents i_{abch}^s can be regarded as sinusoidal waveforms of frequency ω_h with slowly changing amplitudes and phases as

$$\begin{bmatrix} i_{ah}^s \\ i_{bh}^s \\ i_{ch}^s \end{bmatrix} = \begin{bmatrix} Amp - i_{ah}^s \sin(\theta_h + \varphi_{ah}) \\ Amp - i_{bh}^s \sin(\theta_h + \varphi_{bh}) \\ Amp - i_{ch}^s \sin(\theta_h + \varphi_{ch}) \end{bmatrix} \quad (2.19)$$

$$\begin{bmatrix} Amp - i_{ah}^s \\ Amp - i_{bh}^s \\ Amp - i_{ch}^s \end{bmatrix} = \begin{bmatrix} \sqrt{I_p^2 + I_n^2 + 2I_p I_n \cos(2\theta_r)} \\ \sqrt{I_p^2 + I_n^2 + 2I_p I_n \cos(2\theta_r + 2\pi/3)} \\ \sqrt{I_p^2 + I_n^2 + 2I_p I_n \cos(2\theta_r - 2\pi/3)} \end{bmatrix} \quad (2.20)$$

where $\varphi_{ah}, \varphi_{bh}, \varphi_{ch}, Amp - \hat{i}_{ah}^s, Amp - \hat{i}_{bh}^s, Amp - \hat{i}_{ch}^s$ are the phases and amplitudes of three phase HF currents respectively. The amplitudes can be calculated by discrete Fourier transform (DFT) from stator currents as

$$Amp - i_{abch}^s = \frac{2}{W_s} \sqrt{[\sum_{k=0}^{W_s-1} i_{abc}^s(k) \cos(\frac{2\pi k T_s}{T_h})]^2 + [\sum_{k=0}^{W_s-1} i_{abc}^s(k) \sin(\frac{2\pi k T_s}{T_h})]^2} \quad (2.21)$$

where W_s is the number of sampling points in a DFT window and T_s, T_h are the periods of sampling and injection signal. The DFT window length $W_s T_s$ should be an integer multiple of

half of the injection period as (2.22) to avoid frequency aliasing. This poses a weak constraint on the selection of injection frequency.

$$W_s T_s = \frac{k T_h}{2} (k = 1, 2, 3 \dots) \quad (2.22)$$

However, as the amplitude is modulated by rotor frequency ω_e , the calculated amplitude will have an angle shift $\Delta\theta_{shift}$ to the actual value due to the delay caused by DFT window. The simulation result in Fig. 2.6 shows that the calculated amplitude has a time delay to the actual amplitude. The delay is about half of the DFT window. It is verified by simulation that $\Delta\theta_{shift}$ is approximated as half of rotor rotating angle during a DFT window as

$$\Delta\theta_{shift} = \frac{W_s T_s}{2} \omega_e = \frac{k\pi}{2\omega_h} \omega_e (k = 1, 2, 3 \dots) \quad (2.23)$$

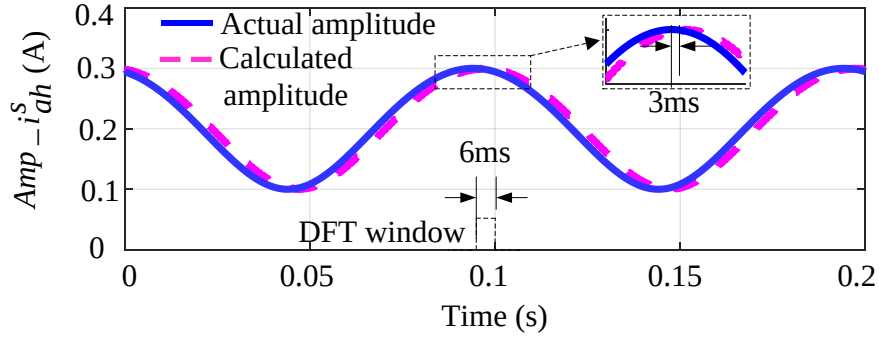


Fig. 2.6. Delay of the calculated amplitude by DFT for $\omega_e = 20\pi \text{ rad/s}$, $\omega_h = 1000\pi \text{ rad/s}$, $T_s = 1e^{-4} \text{ s}$, $W_s = 60$ (6 half injection cycles).

The DFT is carried out in the dSPACE with DS6001 processor board. The program is executed in the Intel@Core™ i7-6820EQ four cores processor with 2.8GHz working frequency. Therefore, the calculation time is negligible compared to the DFT window. It can be seen in Fig. 2.6, the 3ms delay will result in 0.36 degrees in the boundary detection for the speed 60rpm. The angle error caused by DFT increases with the rotor frequency and the window length. However, as the injection frequency ω_h is much higher than ω_e at low speed, the angle shift is relatively small. It is also noted that the calculation in (2.21) is independent of the injection angle θ_h , and thus is immune to the delay effects. The HF current amplitude can also be extracted by calculating the modulus of the current vector which is filtered out by notch filter as shown in Fig. 3.4. However, the calculation requires more calculation resources.

To obtain the rotor position from (2.20), θ_r was calculated by inverse tangent computation as (2.24), which will introduce noise in the position output due to division operation.

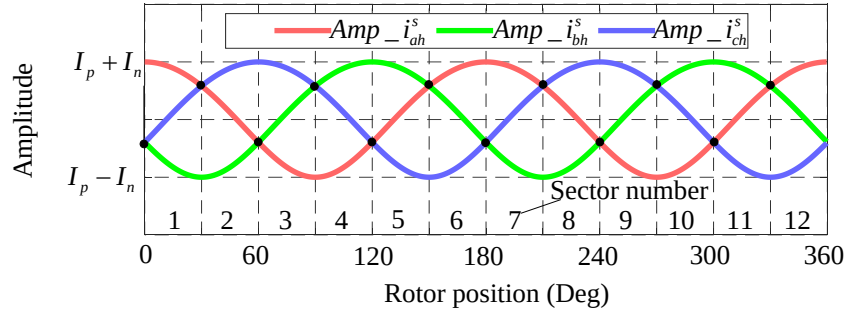
$$\theta_r = 0.5 \tan^{-1} \left[\frac{\sqrt{3}[(Amp_i_{ch}^s)^2 - (Amp_i_{bh}^s)^2]}{2(Amp_i_{ah}^s)^2 - (Amp_i_{bh}^s)^2 - (Amp_i_{ch}^s)^2} \right] \quad (2.24)$$

The aforementioned methods obtain an accurate magnet position which increases the complexity of estimation. In practice, a rotor sector information is already enough for startup. In this chapter, a simple initial sector detection method is proposed based on the HF current amplitudes which is an extension of the phenomenon discovered in [LI17]. The HF current amplitudes are seen as digital signals instead of analog signals. Besides, a multi-frequency injection strategy is utilized to reduce the noise caused by HF injection based on the proposed method.

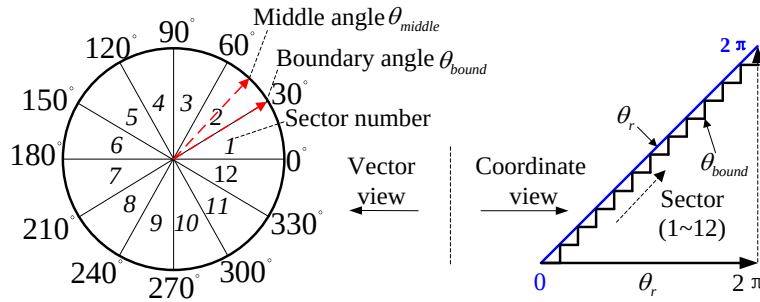
2.3.2 Proposed initial position detection based on HF current amplitudes

2.3.2.1 Fast initial sector detection

The traces of $Amp_i_{ah}^s, Amp_i_{bh}^s, Amp_i_{ch}^s$ with rotor position are plotted in Fig. 2.7(a) according to (2.20). The whole electrical period is evenly divided into 12 sectors by amplitude crossings shown by the black dots. The angles $\theta_{bound}, \theta_{middle}$ are the boundary angle between adjacent sectors and the middle angle in each sector respectively as shown in Fig. 2.7(b). The amplitude waveforms in the range $(180^\circ, 360^\circ)$, i.e. south pole, are identical to those in the range $(0^\circ, 180^\circ)$, i.e. north pole.



(a)



(b)

Fig. 2.7. Variation of three phase amplitudes with rotor position. (a) Variation of phase current amplitudes with position. (b) 12 sectors separated by crossings.

Table 2.2 describes the relationship between the HF current amplitudes and the sector number, which can be used for initial position detection. The rotor position, which refers to the north pole of the magnet, may be located in one sector or on boundary according to the three current amplitudes. If the rotor is on the boundary, the boundary angle θ_{bound} is treated as the initial position and if the rotor is in one sector, the middle angle θ_{middle} is seen as the initial position. The maximum initial position estimation error is $\pm 15^\circ$ which is half of the sector length.

Table 2.2 Sector or boundary judgement according to HF current amplitudes

Relationship	Sector(θ_{middle})	Boundary(θ_{bound})
$Amp_{ah}^s > Amp_{ch}^s > Amp_{bh}^s$	1 or 7 ($\theta_{middle}=\pi/12$)	
$Amp_{ch}^s > Amp_{ah}^s > Amp_{bh}^s$	2 or 8 ($\theta_{middle}=\pi/4$)	
$Amp_{ch}^s > Amp_{bh}^s > Amp_{ah}^s$	3 or 9 ($\theta_{middle}=5\pi/12$)	
$Amp_{bh}^s > Amp_{ch}^s > Amp_{ah}^s$	4 or 10 ($\theta_{middle}=7\pi/12$)	
$Amp_{bh}^s > Amp_{ah}^s > Amp_{ch}^s$	5 or 11 ($\theta_{middle}=3\pi/4$)	
$Amp_{ah}^s > Amp_{bh}^s > Amp_{ch}^s$	6 or 12 ($\theta_{middle}=\pi/12$)	
$(Amp_{bh}^s = Amp_{ch}^s) < Amp_{ah}^s$		0 or π
$(Amp_{ah}^s = Amp_{ch}^s) > Amp_{bh}^s$		$\pi/6$ or $7\pi/6$
$(Amp_{ah}^s = Amp_{bh}^s) < Amp_{ch}^s$		$\pi/3$ or $4\pi/3$
$(Amp_{bh}^s = Amp_{ch}^s) > Amp_{ah}^s$		$\pi/2$ or $3\pi/2$
$(Amp_{ah}^s = Amp_{ch}^s) < Amp_{bh}^s$		$2\pi/3$ or $5\pi/3$
$(Amp_{ah}^s = Amp_{bh}^s) > Amp_{ch}^s$		$5\pi/6$ or $11\pi/6$

2.3.2.2 Polarity detection

The initial position can be identified on the sector boundary or in one specific sector according to Table 2.2. However, there still exists an ambiguity of π as the saliency undergoes two cycles in one electric period. The double voltage pulses injection method is introduced to identify the polarity [HOL08]. The initial position which is θ_{bound} or θ_{middle} is selected as the double pulses injection angle θ_{inj} .

Two opposite voltage pulses with the same magnitude V_{pulse} and width T_{pulse} are injected to the stator windings at position θ_{inj} as shown in Fig. 2.8. The stator current magnitude I_s which is calculated by the two phase currents i_α, i_β in the SSRF as $\sqrt{i_\alpha^2 + i_\beta^2}$ is varying during injection. The stator current magnitude variations caused by the two voltage pulses are recorded as ΔI_{sp} and ΔI_{sn} respectively. If $\Delta I_{sp} > \Delta I_{sn}$, the initial position is θ_{inj} . Otherwise, the initial position is $\theta_{inj} + \pi$. As θ_{inj} only has a maximum error of $\pm 15^\circ$, at least 96.6% ($\cos(15^\circ)$) of the voltage is applied on the real d-axis. This has little impact on the saturation level and the polarity detection

is still effective. However, the current caused by the error voltage on q-axis may produce driving torque during voltage injection.

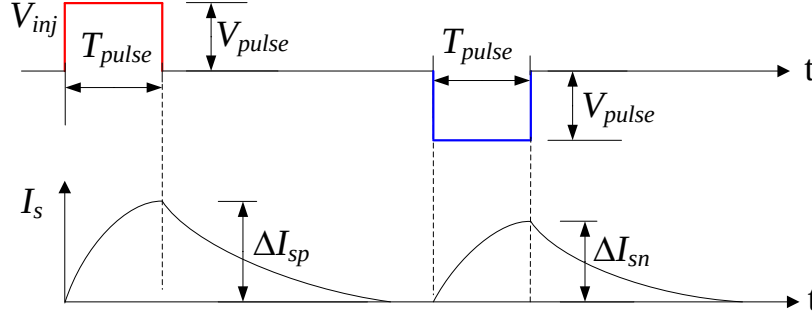


Fig. 2.8. The pulse injection voltage at θ_{inj} and the stator current magnitude for double pulses injection based polarity detection process.

Whether the rotor will be rotated and the movement angle in case of rotating are determined by many other factors, such as the load inertia, the static friction and the injection voltage. The q-axis current with 15° position errors during injection is expressed as

$$i_q(t) = \frac{\sin(15^\circ)V_{pulse}}{R_s} (1 - e^{-t/\tau_q}) \approx \frac{0.26V_{pulse}t}{L_q}, \tau_q = \frac{L_q}{R_s} \quad (2.25)$$

where R_s, L_q, τ_q are the stator resistance, the q-axis inductance and the q-axis time constant. The produced reluctance torque is neglected due to relatively small magnitude and the motion equation of the rotor is expressed as

$$\frac{J}{P} \frac{d\omega_e}{dt} = \frac{3P\psi_{pm}i_q}{2} - T_f \quad (2.26)$$

where P, J, ψ_{pm}, T_f are the pole-pairs, the load moment of inertia, the permanent magnet flux linkage and the load torque, ω_e is the electrical speed. The load torque is composed of static friction torque, cogging torque and the external load torque. According to (2.25) and (2.26), the maximum electrical angle $\Delta\theta_{move}$ caused by rotor movement without considering load torque is approximated as

$$\Delta\theta_{move} = \frac{0.26P^2\psi_{pm}}{4JL_q} V_{pulse} T_{pulse}^3 \quad (2.27)$$

The product of V_{pulse} and T_{pulse} is fixed as it indicates the d-axis flux linkage increment from the initial value to saturation. The DC bus voltage can be fully utilized to generate V_{pulse} . The rotor movement can be largely reduced by decreasing the pulse width to its minimum value.

The inertia J is significantly increased after the load is connected to the rotor. In actual applications, the static friction, the cogging torque and the connected mechanical load can further prevent the rotor from rotating. In this thesis, the rotor will only rotate less than 0.1 degrees when the magnitude and width of the pulse are 12V and 10ms. This is negligible for polarity detection. After the initial position is obtained, the machine can be started. The maximum starting torque is decreased by only 3.4% ($1-\cos(15^\circ)$) at the most which will not seriously affect the startup capability.

Comparison of the proposed initial position estimation method and the conventional methods is shown in Fig. 2.9. Sliding DFT (SDFT) can be introduced to reduce the calculation time. Each new DFT is efficiently computed directly from the results of the previous DFT with only one complex multiply and two real adds per output sample [JAC03]. A rough comparison between the proposed method and the conventional heterodyning method on a Texas Instruments TMS320F28035 digital signal processor developing board shows that the proposed method only requires 3 μ s execution time, 1 μ s longer than the conventional method. This is considered acceptable in industrial applications.

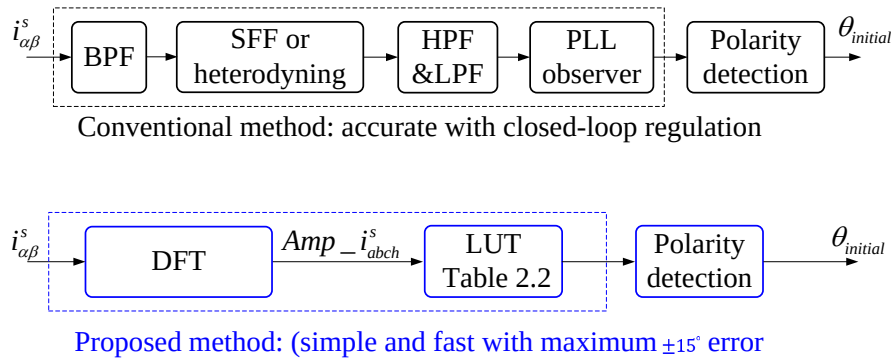


Fig. 2.9. Comparison of conventional and proposed initial position estimation methods.

However, the convergence time for the conventional methods may be longer, as the convergence time is dependent on observer parameters. PI parameters with good performances are not easy to obtain without accurate inverter model and machine parameters. Inappropriate control parameters may result in long setting time and even oscillation. Besides, the time interval between initial position detection and polarity detection is often set long with margin to guarantee reliable convergence. Such time margin further increases the execution time for the conventional methods. Whereas for the proposed method, as the first DFT calculation can be completed after only half of the injection period, the initial position is reliably detected without any additional time delay. The fast initial position detecting property is of practical use in fan motors, wash machines and pumps where consumers are concerned on the starting time.

2.3.3 Discrete and continuous positions output

After the machine is started from the initial position, the rotor sectors can be continuously obtained. From Fig. 2.7(a), the sector number and boundary angle will be updated every 30° if the crossings of the traces can be reliably detected. At each sampling point, $Amp_{i_{abch}^s}$ are updated by SDFT and compared according to Table 2.2. If two values of the amplitude traces are equal, a boundary angle can be identified. However, this is not always the case as the situations of equal amplitudes at sampling points may not be detected due to limited sampling frequency as demonstrated in Fig. 2.10(a). Alternatively, the boundary can be detected by the change of sector numbers between two adjacent sampling instants as shown in Fig. 2.10(b). If the calculated sector number changes, the rotor must be crossing a boundary and the boundary angle can be identified according to Table 2.2.

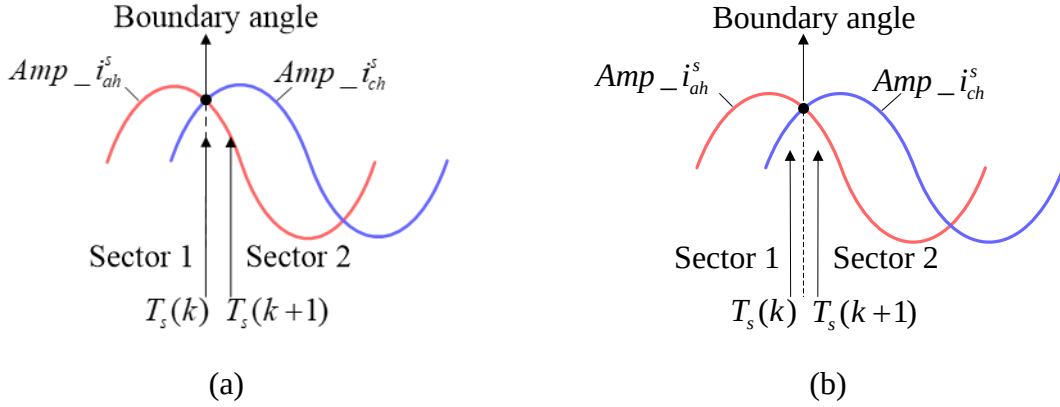


Fig. 2.10. Boundary detection situations. (a) Sampling point aligned with crossing point. (b) Sampling points deviated from crossing point.

The situation in Fig. 2.10(b) will result in a maximum position error of $\omega_e T_s$ due to limited sampling frequency. The sampling points in a fundamental electric period is inversely proportional to the rotor speed which means that the sector detection accuracy is deteriorated at higher speed. For example, for $T_s=10^{-4}s$, $\omega_e=20\pi rad/s$, the maximum discrete position error caused by the detection offset is $\omega_e T_s=0.36^\circ$. It should be noted that if I_n is not large enough, the crossing detection may become ineffective as the crossings may be seriously distorted by noises caused by sampling quantization error, inverter nonlinearity and multi-saliencies. The rotor electrical frequency ω_{cal} can be calculated from discrete position as

$$\omega_{cal} = \frac{\pi}{6\Delta t} \quad (2.28)$$

where Δt is the time interval between two adjacent crossings. When the speed reaches a certain level, the machine switches to operate under back EMF based sensorless control mode.

However, when the rotor is running, the discrete position will result in low frequency harmonics in the phase currents, which would produce torque pulsation, as well as noise. Furthermore, it is not easy to stabilize the EMF based sensorless model with the distorted phase currents at low speed. As an option to the sensorless startup process, it is better to output continuous position from the discrete boundary positions.

The discrete boundary position θ_{bound} can be modeled as space vector $V_{discrete}$ as

$$V_{discrete}(k) = \cos(\theta_{bound}) + j \sin(\theta_{bound}), k = 1 \dots 12 \quad (2.29)$$

The trace of $V_{discrete}$ is a dodecagon and can be modelled as the combination of a rotating fundamental vector V_f and a set of positive and negative rotating spatial harmonics $V_{harmonic}$ as shown in Fig. 2.11.

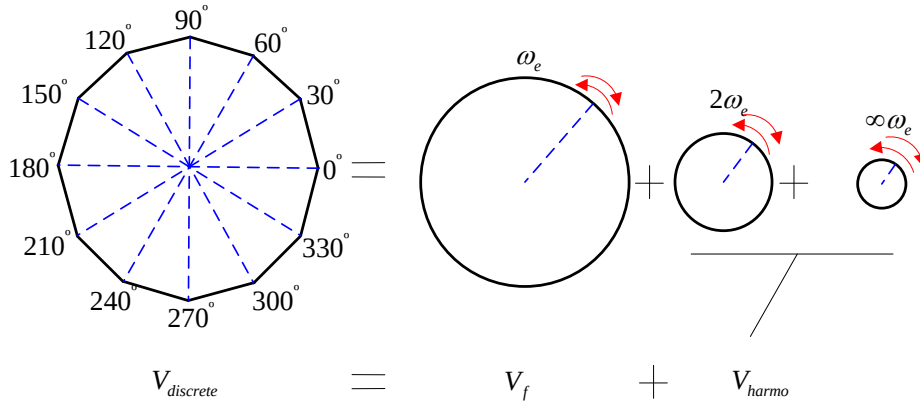


Fig. 2.11. Space vector model of discrete boundary angles.

According to [WAN17a], the spatial Fourier series of $V_{discrete}$ can be set as

$$V_{discrete} = \sum_{k=-\infty}^{\infty} a_k e^{jk\theta_r} \quad (2.30)$$

$$a_k = \frac{1}{2\pi} \sum_{n=0}^{n=12-1} V_{discrete} \int_{\frac{2\pi n}{N}}^{\frac{2\pi(n+1)}{N}} e^{-jk\theta_r} d\theta_r \quad (2.31)$$

The spatial Fourier series of the discrete boundary position can be calculated as

$$V_{discrete} = V_f + V_{harmonic} = e^{j(\theta_r - \varphi)} + \sum_{k=1}^{\infty} \left[-\frac{1}{kN-1} e^{-j((kN-1)\theta_r - \varphi)} + \frac{1}{kN+1} e^{j((kN+1)\theta_r - \varphi)} \right] \quad (2.32)$$

where $N=12$ denotes the sectors in one electrical period, $\varphi = \pm\pi/N$ is the phase delay and of the same sign as ω_r . In [WAN17a], an enhanced vector tracking observer is used to obtain the accurate rotor position from the discrete position. Similar structure is introduced in this thesis to deal with the discrete boundary positions as shown in Fig. 2.12.

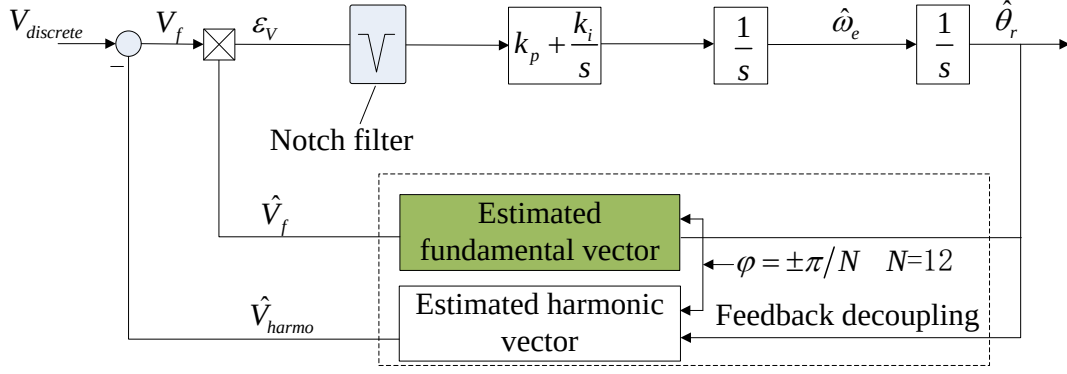


Fig. 2.12. Vector tracking observer for continuous position estimation.

The fundamental and harmonic vectors can be estimated by the estimated position $\hat{\theta}_r$ as

$$\hat{V}_f = e^{j(\hat{\theta}_r - \varphi)} \quad (2.33)$$

$$\hat{V}_{harmono} = \sum_{k=1}^{\infty} \left[-\frac{1}{kN-1} e^{-j((kN-1)\hat{\theta}_r - \varphi)} + \frac{1}{kN+1} e^{j((kN+1)\hat{\theta}_r - \varphi)} \right] \quad (2.34)$$

The position error ε_V is obtained by cross product as

$$\varepsilon_V = V_f \otimes \hat{V}_f \quad (2.35)$$

If the position estimation error is small enough, the estimated position will be converged to rotor position $\hat{\theta}_r$. The notch filter is used to mitigate the harmonic components caused by position estimation error. The phase φ and number N here are not dependent on the estimated speed $\hat{\omega}_r$. The speed observed by vector tracking observer shows less ripple than that calculated by (2.28).

2.3.4 Experimental verification

The proposed sensorless initial position estimation method is verified on a dSPACE platform as shown in Fig. 2.13. The parameters of the test IPMSM are shown in the Appendix 1. An incremental encoder with 1024 pulses is used to provide accurate rotor position. The switching frequency f_{PWM} and system sampling frequency f_s are both set as 10 kHz. A load machine is connected to the machine to provide the load torque. Two hall sensors are used for current sampling and the resolution of the ADC module is 16 bits.

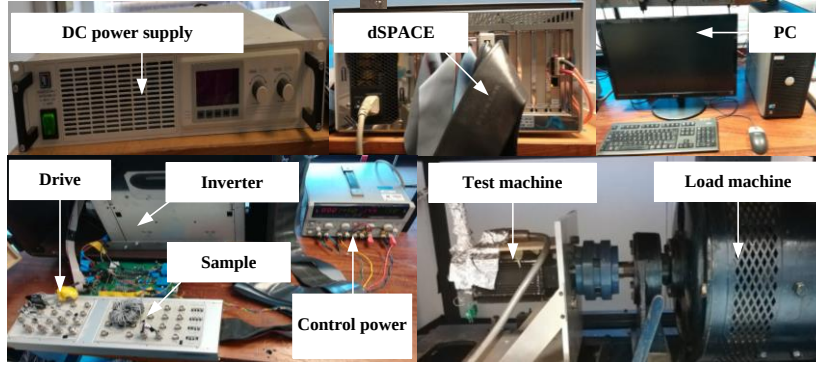


Fig. 2.13. Experimental platform.

2.3.4.1 Initial position estimation performance

In order to evaluate the effectiveness of the proposed initial position detection method, the rotor is located at position $\theta_r=50^\circ$ without fundamental currents. A rotating HF voltage with magnitude 15V and frequency 750Hz is selected considering the saliency level. Both conventional and proposed methods are executed in the software.

Fig. 2.14 shows the comparison of initial position estimation between the conventional method with different PI parameters and the proposed method. The relationship between calculated current amplitudes is $Amp_{i_{ch}^s} > Amp_{i_{ah}^s} > Amp_{i_{bh}^s}$. According to Table 2.2, the rotor position is located in sector 2 or sector 8 and the middle angle 45° is treated as the initial rotor position.

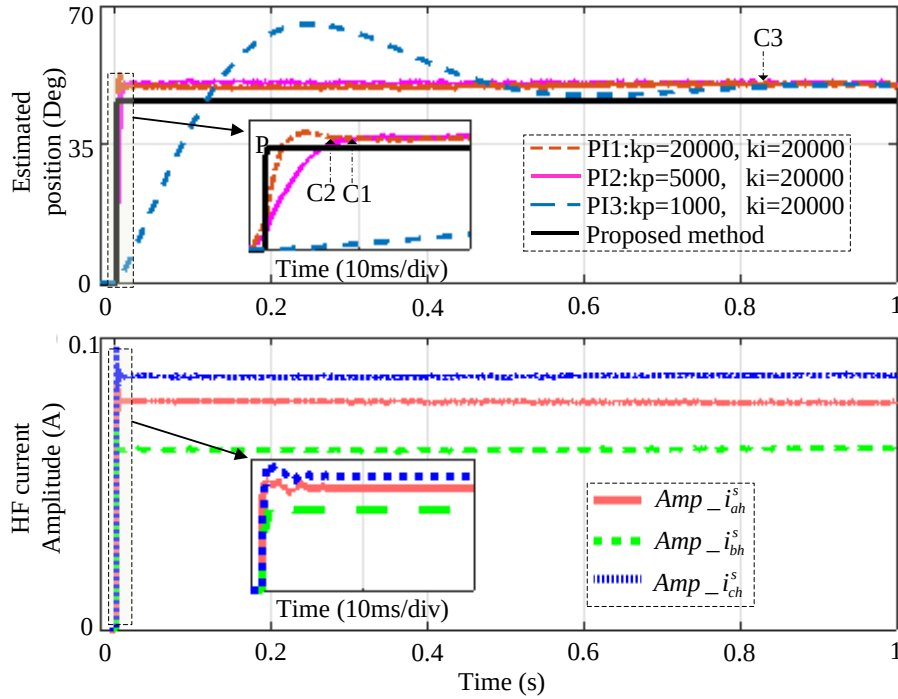


Fig. 2.14. Comparison of proposed and conventional initial position estimation methods with different PI parameters in heterodyning observer.

It is observed that the amplitudes of i_{abch}^s are stable at point P, which is about only 3 injection cycles after initial injection. The convergence time for the conventional method varies with different PI parameters. For PI parameters with overshoot and oscillation, the convergence time can be as long as 1 second as shown at point C3 for PI3. The convergence time can also be very short with well-tuned PI, such as point C1 for PI1 and point C2 for PI2. Whereas the parameter tuning is not only limited by filters but also dependent on machine parameters, it is difficult to predetermine optimal PI parameters. Thus, enough time should be reserved between initial injection and polarity detection to guarantee reliable convergence. This time gap further increases the execution time of conventional method.

To verify the effectiveness of pulse injection based polarity detection even with 15° position error, the pulse injection is performed both at real position 45° and deviated position 30° . The rotor keeps standstill during pulse injection due to large mechanical load inertial. Fig. 2.15 shows the stator current waveforms of double pulses injection at the accurate rotor position and the deviated position. The pulse magnitude and duration are 12V/10ms. It is observed that current variation difference between real position points A and B is 0.711A and the current variation difference between deviated position points C and D is 0.702A. Therefore, the saturation difference between point C and point D is still significant enough and the polarity can be judged as between $[0, \pi]$. Thus, the polarity detection by double pulses injection is still effective even the pulse injection position has a 15° angle offset to the real rotor position.

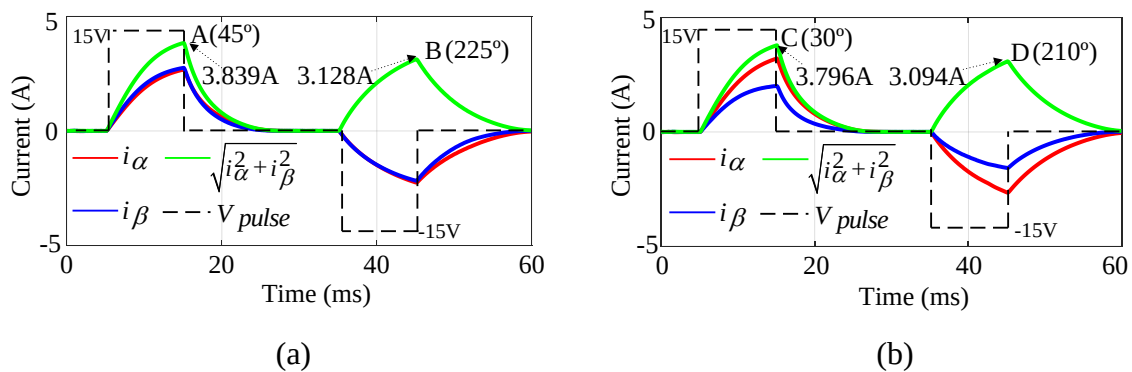


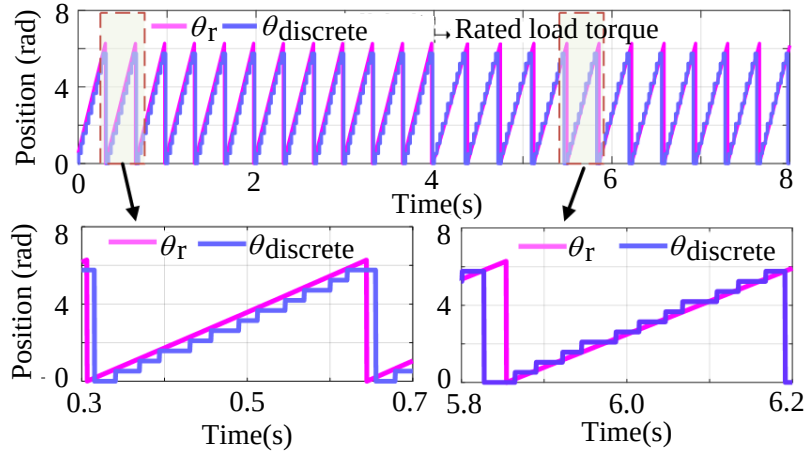
Fig. 2.15. Comparison of double pulses injection at accurate and deviated position. (a) At accurate position ($\theta_r = 45^\circ$). (b) At deviated position with 15° offset.

2.3.4.2 Startup and low speed performances

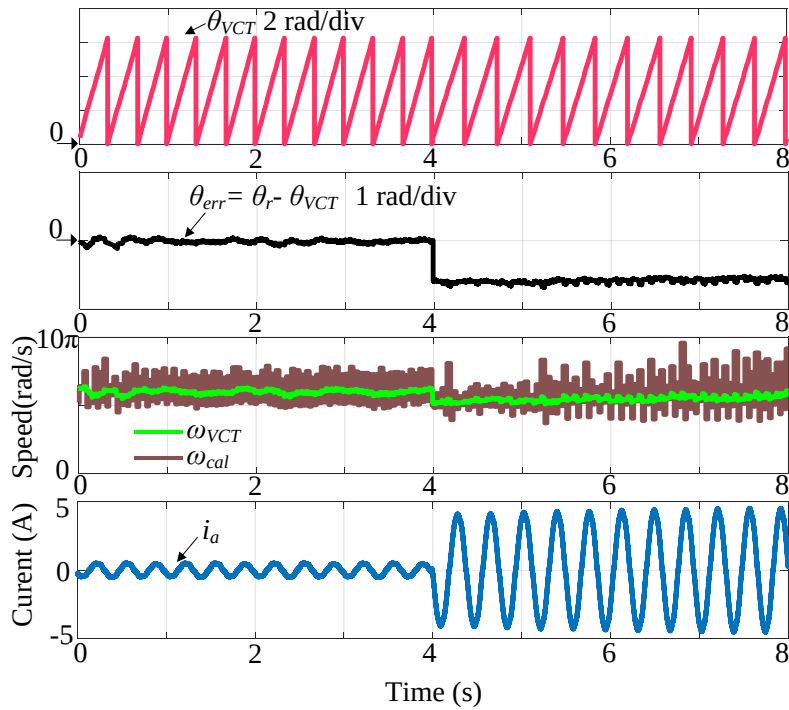
Sensored test is carried out to evaluate the proposed position estimation performance. The machine is controlled at 60rpm (3Hz) with position obtained from the encoder under different load torque. There are 12 discrete positions in one electrical period shown in Fig. 2.16(a). The

discrete position trace is ahead of the real rotor position at rated load, which is due to the cross coupling effect. Compensation of the cross-coupling effect will be investigated in Chapter 4. It is observed that each sector length is not equal and shows periodical characteristic. This is mainly caused by multi-saliencies. As the frequencies of multi-saliencies induced harmonic currents are too close to that of the primary saliency induced current at low speed [CHE14], the resolution of the DFT is not high enough to separate them. The calculated amplitudes will be distorted and result in periodically offsets at the crossing points. The sampling error and limited sampling frequency also affect the detection accuracy of the boundary angle. It should be mentioned that as the proposed method is independent on injection angle, it is less sensitive to phase delay effects and inverter nonlinearities [GUE05].

The smooth position θ_{VCT} is obtained from discrete position $\theta_{discrete}$ by the vector tracking observer as shown in Fig. 2.16(b). The subscript “VCT” means the vector tracking observer. The position estimation error of θ_{VCT} is nearly zero at zero load and increases to -0.6 rad at rated load which is due to cross coupling effect. It shows that the proposed method is influenced by the cross coupling effect which is not the main concern in this chapter [REI08]. The speed ω_{VCT} output by the tracking observer shows less ripple compared to the speed calculated speed ω_{cal} .



(a)



(b)

Fig. 2.16. Sensored operation at 3Hz with different load torque. (a) Actual rotor position and detected discrete position. (b) Continuous position from vector tracking observer.

Dynamic performance of the proposed method are also evaluated. The machine is running in sensorless mode at 2Hz with rated load step T_{load} at 2s. The phase current and estimated position are shown in Fig. 2.17. The speed decreases to near zero at step load and returns to set speed after about 1 second. Compared to the conventional method, the speed and position are not updated at very low speed due to long time interval between adjacent crossings. From the perspective of current feedback, as there exists inherent position estimation error for the proposed method and the estimated position is not updated, the feedback currents are not only inaccurate but also not updated at very low speed. This will affect the dynamic performance of

the proposed method. Even for the conventional method with PI tuning, the dynamic performance also suffers from the problem of estimation error caused by delay and inverter nonlinearities.

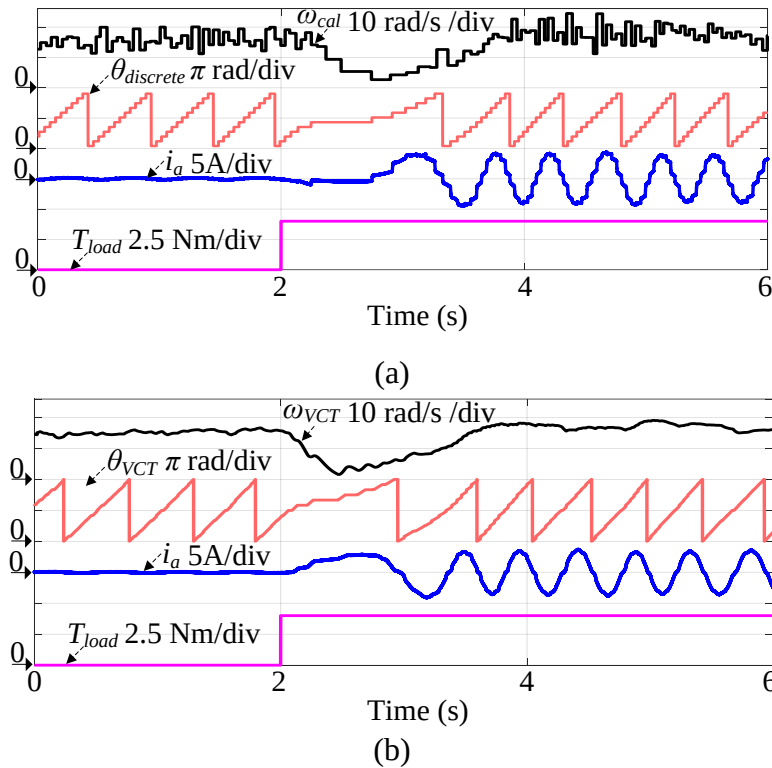
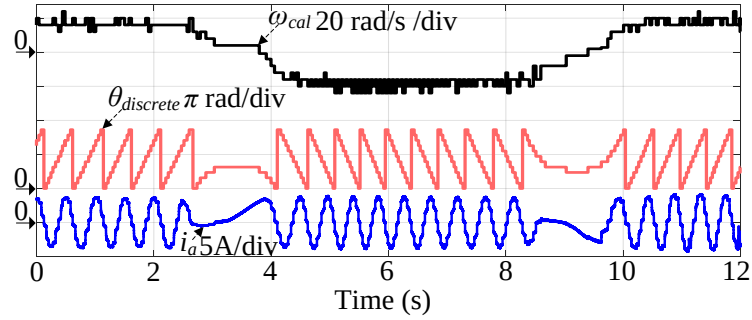


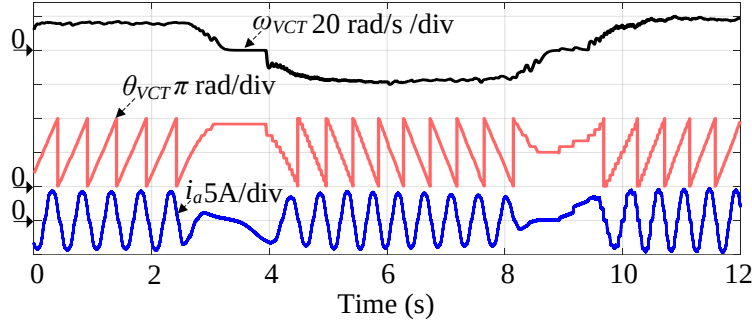
Fig. 2.17. Dynamic performance at load torque step. (a) By discrete position. (b) By continuous position from tracking observer.

It is noted that different from the conventional HF injection based sensorless control, the proposed method is an open loop type estimation. The stability is highly dependent on the sampling and computing accuracy and sensitive to saliency ratio and multi-saliencies. As the DFT calculation is carried out based on multiple samples, the sampling spikes caused by switches or external electric pulse at a few sampling points are averaged. However, for the conventional method, a sharp noise in the current sampling may cause instability of the estimation loop. From this point of view, the proposed method shows advantage in situations with much sampling noise.

Speed reversal experiments are also carried out and shown in Fig. 2.18. The speed order is reversed at 2s and 8s under rated load. It is observed that the speed is not updated at very low speed.



(a)



(b)

Fig. 2.18. Speed reversal experiments at rated load. (a) By discrete position. (b) By continuous position from tracking observer.

Sensorless startup experiments with different load torque are carried out to verify the startup capability. The machine is started from standstill to 2Hz. The phase current, estimated speed and position used in startup are shown in Fig. 2.19.

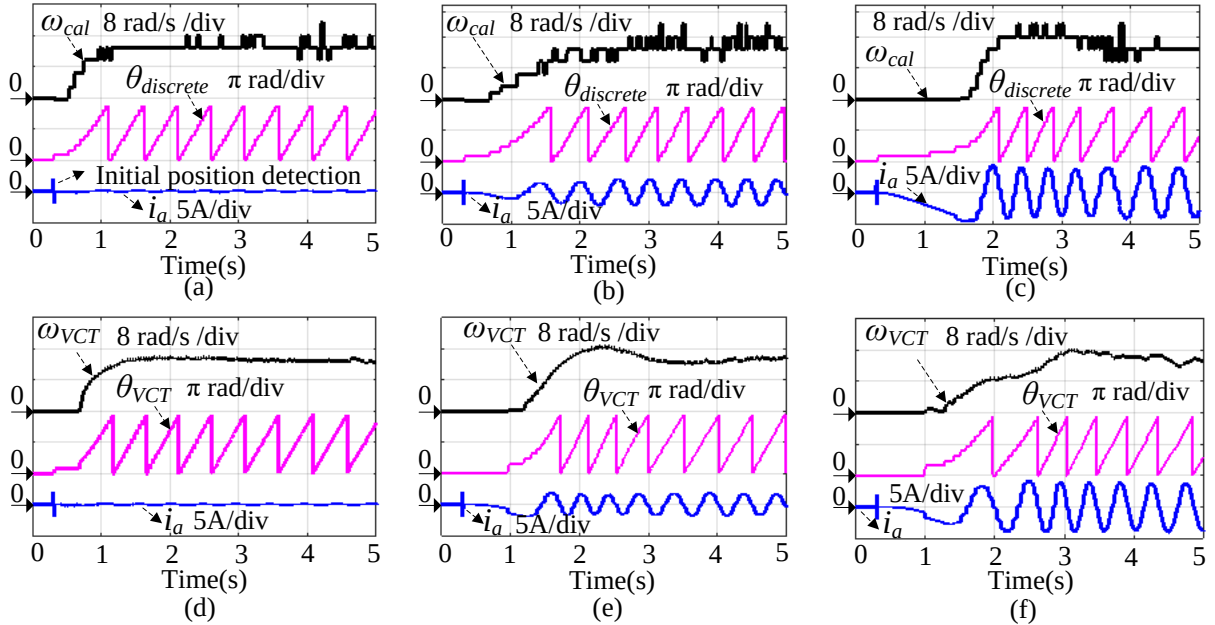


Fig. 2.19. Sensorless startup with various load torque. (a) Discrete position, 0% rated torque. (b) Discrete position, 50% rated torque. (c) Discrete position, 100% rated torque. (d) Continuous position, 0% rated torque. (e) Continuous position, 50% rated torque (f) Continuous position, 100% rated torque.

From Fig. 2.19, it can be seen that the machine is successfully started with both discrete positions and continuous position. The position output by the tracking observer is discrete at startup and becomes continuous with the increase of speed. The speed calculated by (2.22) shows higher ripple compared to the speed output by the tracking observer.

2.4 Application of the Proposed Method in Noise Reduction

2.4.1 Noise reduction by multi-frequency signals injection

In the previous section, simple startup can be achieved by injecting a single frequency HF voltage. However, the noise caused HF currents is shrill to hear. In this section, instead of injecting single frequency signal, noise emission can be reduced by injecting multi-frequency signals as

$$\begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} = \begin{bmatrix} \sum_{i=1}^M u_{hi} \cos(\omega_{hi} t) \\ \sum_{i=1}^M u_{hi} \sin(\omega_{hi} t) \end{bmatrix} \quad (2.36)$$

where M is the number of frequencies contained in the signal, u_{hi}, ω_{hi} are the magnitude and frequency of each frequency signal. The resulting HF currents $i_{\alpha h}^s, i_{\beta h}^s$ are expressed as

$$i_{\alpha h}^s + j i_{\beta h}^s = \sum_{i=1}^M [I_{pi} e^{j(\omega_{hi} t - \frac{\pi}{2})} + I_{ni} e^{-j(\omega_{hi} t - 2\theta_r - \frac{\pi}{2})}] \quad (2.37)$$

$$I_{pi} = \frac{u_{hi}(L_{qh} + L_{dh})}{2\omega_{hi} L_{dh} L_{qh}}, I_{ni} = \frac{u_{hi}(L_{qh} - L_{dh})}{2\omega_{hi} L_{dh} L_{qh}} \quad (2.38)$$

where I_{pi}, I_{ni} are the current magnitudes of the positive sequence and negative sequence currents for each frequency. Conventionally, the currents of each frequency can be separated by different band pass filters and the negative sequence currents of various frequencies are extracted by the methods described in Fig. 2.3. The position error signals from each frequency are combined together to form an enhanced error signal ε_{enp} as (3.39) and used for position estimation [CHE14]. The conventional signal processing process is shown in Fig. 2.20(a).

$$\varepsilon_{enp} = (\sum_{i=1}^M I_{ni}) \sin(\theta_r - \hat{\theta}_r) \quad (2.39)$$

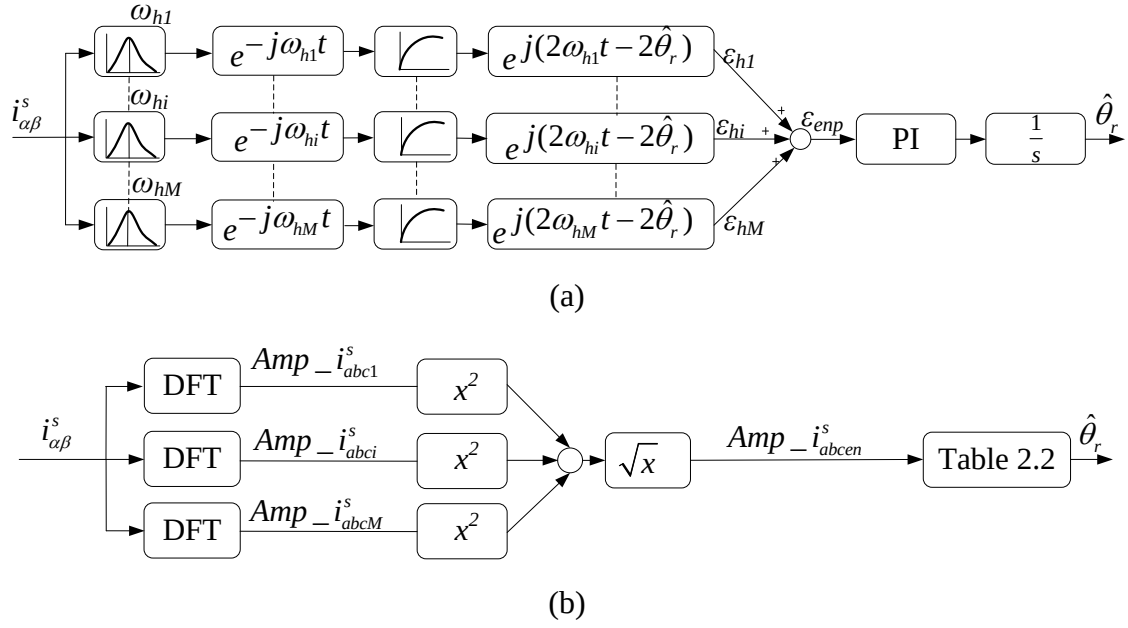


Fig. 2.20. Comparison of demodulation methods for multi-frequency injection. (a) Method based on negative sequence current. (b) Method based on HF current amplitudes.

By selecting the proper values of ω_{hi} and u_{hi} of the multi-frequency signal according to (2.40) and (2.41), ε_{enp} can keep the same amplitude as single frequency injection. The voltage utilization of multi-frequency injection is not increased and the hardware cost is not suffered. Generally, if u_h, ω_h of the single frequency signal are not unknown before selecting multi-frequency signal, the combination of u_{hi}, ω_{hi} should be selected according to actual noise reduction effects and sampling circuits.

$$\sum_{i=1}^M \frac{u_{hi}}{\omega_{hi}} = \frac{u_h}{\omega_h}, \quad \sum_{i=1}^M I_{ni} = I_n \quad (2.40)$$

$$\sum_{i=1}^M u_{hi} \leq u_h \quad (2.41)$$

PSD, which is the Fourier transform of the autocorrelation of a signal, is an important tool in signal analysis. In [WAN16], the PSD calculation is carried out to analyse random HF currents. The same procedures can be introduced to analyse the proposed multi-frequency injection method. According to (2.37), the α -axis currents are composed of a series of sinusoidal waveforms as

$$i_{\alpha h}^s = \sum_{i=1}^M [I_{pi} \cos(\omega_{hi} - \frac{\pi}{2}) + I_{ni} \cos(\omega_{hi} - 2\theta_r - \frac{\pi}{2})] \quad (2.42)$$

The PSD function $S_h(f)$ for $i_{\alpha h}^s$ is calculated according to the definition of PSD as

$$\begin{aligned}
S_h(f) = & \frac{\pi}{2} \sum_{i=1}^M I_{pi}^2 [\delta(f - \frac{\omega_{hi}}{2\pi}) + \delta(f + \frac{\omega_{hi}}{2\pi})] \\
& + \frac{\pi}{2} \sum_{i=1}^M I_{ni}^2 [\delta(f - \frac{\omega_{hi} - 2\omega_e}{2\pi}) + \delta(f + \frac{\omega_{hi} - 2\omega_e}{2\pi})]
\end{aligned} \tag{2.43}$$

where $\delta(\cdot)$ denotes an impulse function. It proves theoretically that the PSD for multi-frequency injection has multiple peaks, each of which is relatively lower than single frequency injection. As a result, the audible noise can be reduced. The number M and the value of each u_{hi}/ω_{hi} are selected considering the balance of computational burden and noise reduction effect.

However, the conventional signal processing flow shown in Fig. 2.20(a) not only requires BPFs and LPFs with different centre frequencies but also needs to take delay effects of different frequencies and tuning of observer parameters into consideration. The position estimation method based on HF current amplitudes provides an easy way to deal with multi-frequency injection as the current amplitude extracting is independent of the carrier signal and delay effects. The resulting three phase HF currents can be seen as the superposition of currents with different amplitudes and frequencies as

$$\begin{bmatrix} i_{ah}^s \\ i_{bh}^s \\ i_{ch}^s \end{bmatrix} = \begin{bmatrix} \sum_{i=1}^M Amp - i_{ahi}^s \sin(\omega_{hi}t + \varphi_{ahi}) \\ \sum_{i=1}^M Amp - i_{bhi}^s \sin(\omega_{hi}t + \varphi_{bhi}) \\ \sum_{i=1}^M Amp - i_{chi}^s \sin(\omega_{hi}t + \varphi_{chi}) \end{bmatrix} \tag{2.44}$$

$$\begin{bmatrix} Amp - i_{ahi}^s \\ Amp - i_{bhi}^s \\ Amp - i_{chi}^s \end{bmatrix} = \begin{bmatrix} \sqrt{I_{pi}^2 + I_{ni}^2 + 2I_{pi}I_{ni} \cos(2\theta_r)} \\ \sqrt{I_{pi}^2 + I_{ni}^2 + 2I_{pi}I_{ni} \cos(2\theta_r + 2\pi/3)} \\ \sqrt{I_{pi}^2 + I_{ni}^2 + 2I_{pi}I_{ni} \cos(2\theta_r - 2\pi/3)} \end{bmatrix} \tag{2.45}$$

The values of $Amp - i_{abchi}^s (i=1, \dots, M)$ for various frequencies are calculated by DFT as (2.46). Sliding DFT can also be introduced to reduce computation time. The sampling point window length $W_s T_s$ should be integer multiples of all the half frequencies. The calculated amplitudes of all frequencies are combined to derive $Amp - i_{abcen}^s$ as (2.47). The process to deal with multi-frequency injection is shown in Fig. 2.20(b).

$$Amp_{-i_{abchi}^s} = \frac{2}{W_s} \sqrt{\left[\sum_{k=0}^{W-1} i_{abc}^s(k) \cos(k\omega_{hi}T_s) \right]^2 + \left[\sum_{k=0}^{W-1} i_{abc}^s(k) \sin(k\omega_{hi}T_s) \right]^2} \quad (2.46)$$

$$Amp_{-i_{abcen}^s} = \begin{bmatrix} Amp_{-i_{aen}^s} \\ Amp_{-i_{ben}^s} \\ Amp_{-i_{cen}^s} \end{bmatrix} = \begin{bmatrix} \sqrt{\sum_{i=1}^M (Amp_{-i_{ahi}^s})^2} \\ \sqrt{\sum_{i=1}^M (Amp_{-i_{bhi}^s})^2} \\ \sqrt{\sum_{i=1}^M (Amp_{-i_{chi}^s})^2} \end{bmatrix} \quad (2.47)$$

$$= \begin{bmatrix} \sqrt{\sum_{i=1}^M (I_{pi}^2 + I_{ni}^2) + \sum_{i=1}^M (2I_{pi}I_{ni}) \cos(2\theta_r)} \\ \sqrt{\sum_{i=1}^M (I_{pi}^2 + I_{ni}^2) + \sum_{i=1}^M (2I_{pi}I_{ni}) \cos(2\theta_r + \frac{2\pi}{3})} \\ \sqrt{\sum_{i=1}^M (I_{pi}^2 + I_{ni}^2) + \sum_{i=1}^M (2I_{pi}I_{ni}) \cos(2\theta_r - \frac{2\pi}{3})} \end{bmatrix}$$

The traces of $Amp_{-i_{abcen}^s}$ are of the same shapes as the three phase HF current amplitudes for single frequency injection. Thus, they can be used for initial position detection and startup according to Table 2.2 as for single frequency injection. As the execution time for multi-frequencies demodulation will be increased, the frequency number M should be chosen considering the computation resources limit. In this chapter, the number M is chosen as 2.

The proposed multi-frequency injection is different from random injection or variable frequency injection in the mode of time occupying. For the proposed multi-frequency injection, the time is occupied by different frequencies simultaneously. It is called frequency division multiplexing (FDM). For pseudo-random injection or variable frequency injection, which is called time division multiplexing (TDM), the injection time is occupied by each frequency alternatively. The effectiveness of the proposed method is verified in following experiments.

2.4.2 Experimental verification

To verify the startup performance and noise reduction effect of multi-frequency injection, two group of signals of different frequencies are injected to the machine at the same time. To compare with previous single frequency injection signal of 15V/750Hz, a combination of 5V/500Hz and 10V/1000Hz is selected. According to (2.34), the negative sequence current level and the voltage utilization ratio keep unchanged. The machine is running at 2Hz in sensorless mode with single frequency injection and two frequencies injection respectively.

The HF phase voltage u_{ah} and HF phase current i_{ah} waveforms are shown in Fig. 2.21. The envelop of the phase current of two frequencies injection is smaller than that of the single frequency injection.

$$\frac{15V}{750Hz} = \frac{5V}{500Hz} + \frac{10V}{1000Hz} \quad (2.48)$$

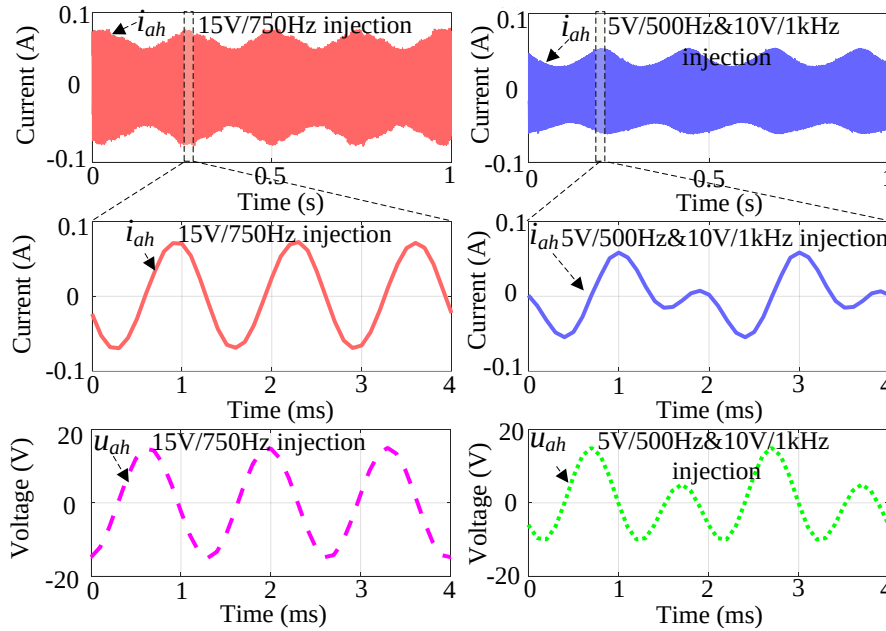


Fig. 2.21. Comparison of HF currents and voltages between single frequency injection and two frequencies injection.

To verify the load capability of multi-frequency injection, the machine is sensorless started to 2Hz with rated load. The position for sensorless control comes from discrete boundary position $\theta_{discrete}$ in Fig. 2.22(a) and continuous position θ_{VCT} from vector tracking observer in Fig. 2.22(b).

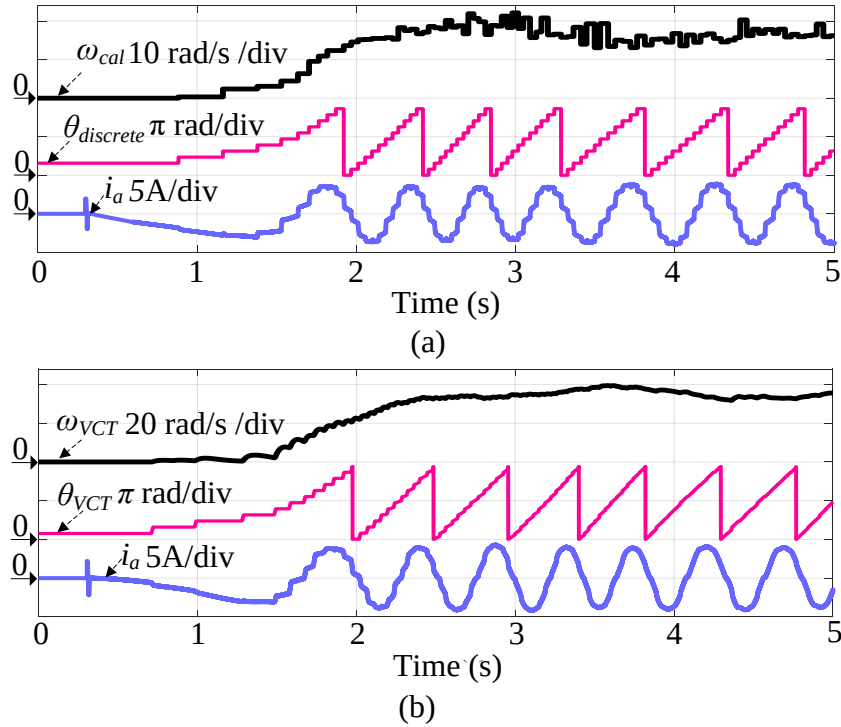


Fig. 2.22. Sensorless startup at rated load with two frequencies injection. (a) By discrete position. (b) By continuous position from tracking observer.

It is verified that the machine can be started both by discrete position and continuous position estimated by the multi-frequency injection based sensorless method at rated load torque. Compared to the sensorless startup method by single frequency injection, the noise emission of the two frequencies injection method is largely relieved. The PSDs of the HF phase currents with single frequency and multi-frequency injection during initial position estimation are compared in Fig. 2.23.

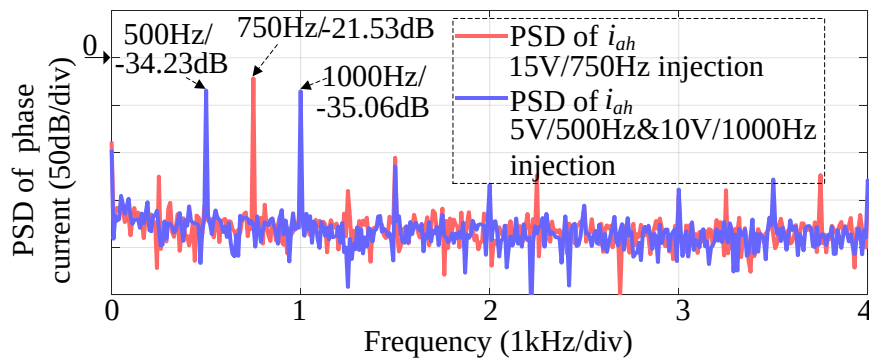


Fig. 2.23. Comparison of PSDs of the high frequency phase current by single frequency injection and two frequencies injection at startup.

The PSD values at 500Hz and 1000Hz are lower than that of the value at 750Hz for single frequency injection. The power of the noise is not concentrated at single point and the shrill acoustic noise is largely reduced. More frequencies could be injected, but it will require more computation resources and longer execution time.

2.5 Conclusion

The basic methods for position estimation at low speed with sinusoidal rotating or pulsating voltage injection are presented. Then, the amplitude modulation and phase modulation in the HF injection based sensorless control methods are analysed.

It is revealed that the HF currents induced by rotating voltage injection are also amplitude modulated waveforms. Based on this, a novel sensorless initial position estimation method by making use of the HF current amplitudes is proposed for IPMSM. The initial position with maximum $\pm 15^\circ$ error and sector information can be detected just by comparing HF current amplitudes without closed-loop observer. Discrete rotor position can be detected by comparing the HF current amplitudes in the low speed range. Smooth position can also be obtained with the help of a vector tracking observer. Sensorless startup and low speed operation can be achieved with both discrete and continuous positions.

To mitigate the noise caused by single frequency HF voltage injection during startup, a multi-frequency injection strategy is employed and the proposed position demodulation method by current amplitudes is employed instead of conventional phase demodulation methods. The shrill noise caused by high frequency voltage injection is largely reduced due to multiple peaks in the current PSDs. The effectiveness of the proposed amplitude based initial position estimation method is verified by experiments.

CHAPTER 3

A NOVEL METHOD FOR MEASURING HIGH FREQUENCY INCREMENTAL DQ-AXIS AND CROSS-COUPLING INDUCTANCES IN INTERIOR PERMANENT MAGNET SYNCHRONOUS MACHINES

3.1 Introduction

Interior permanent magnet synchronous machines (IPMSM) have been widely used in domestic and industrial applications due to high efficiency and power density [ZHU07]. The electrical parameters of IPMSM, such as inductances, resistances and permanent magnet flux linkages are essential for high performance operation [RAH05]. However, it may be difficult for the user to obtain accurate machine parameters due to lack of design data or suitable measuring tools. Therefore, it is necessary for the machine drives to have the capability of measuring machine parameters which is usually carried out using general purpose three phase inverters [WAN14], [ODH14]. The measured results are widely used in control parameter tuning, sensorless position estimation and condition monitoring [ERT20], [MAR20].

Obtaining the inductances by offline measurement is the main focus of this thesis. Inductance can be classified as apparent inductance $L_{apprent}$ or incremental inductance $L_{incremental}$, as graphically illustrated in Fig. 3.1. The apparent inductance at point b is defined as the ratio between the flux linkage increment from zero current to a specific working current point and the current, while the incremental inductance is the slop of the tangent line at a specific working current point on the flux linkage-current characteristic. The apparent inductance information can be used for control parameter tuning [YOO09] and back electromagnetic force based position observer [BER14]. The incremental inductances, also referred to as high frequency (HF) inductances (for convenience, they will be simply designated as HF inductances in this thesis), can be used to predict the position estimation error in HF injection based sensorless control [REI08]. Both the apparent and HF inductances can be obtained from the flux linkage-current characteristic by post-processing [LI19]. However, the measurement of flux linkages is relatively complicated and requires back electromagnetic force integration operation for standstill tests [HIN17] or spinning the rotor for rotating test [PEL15]. It is preferable if the HF inductances can be measured directly at standstill.

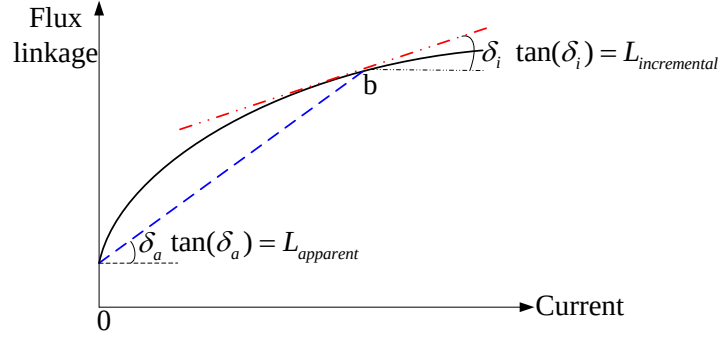


Fig. 3.1. Illustration of apparent and incremental inductances.

The HF inductances can be directly measured by employing HF signal injection techniques which have been widely used in position sensorless control [KAN21], [XU14], [KUE14], [LI07a]. In [KAN21], the dq -axis HF inductances are measured for surface mounted permanent magnet synchronous machine by injecting HF pulsating voltage to the position between the d -axis and the q -axis. However, the method in [KAN21] is only used for surface permanent magnet synchronous machines and the cross-coupling inductance which is not negligible in IPMSM is not considered [LI09], [SHU20]. In [XU14], the HF sinusoidal rotating voltage is injected to the stationary reference frame for measuring the HF inductances, including the cross-coupling effect. The offset angle caused by the cross-coupling effect is obtained from the difference between the actual rotor position measured from an encoder and the estimated rotor position obtained from a position estimator with the negative sequence current as the input. The dq -axis and cross-coupling HF inductances are then calculated from the offset angle and amplitudes of the positive sequence and negative sequence HF currents. The method in [XU14] requires a position estimator, the design of which is dependent on inductance values and affected by the delay effects [GHA13], [XU17a]. In [KUE14], the HF pulsating voltage injection technique is used for measuring the HF inductances. The angle offset caused by the cross-coupling effect is also obtained from the difference between the actual rotor position measured by an encoder and the estimated rotor position obtained from a position estimator. The HF inductances are then calculated from the observed angle offset and the measured HF current amplitudes due to pulsating voltage injections. The position estimator is also required and the measured HF current amplitudes are affected by the delay effects which will result in errors in the calculated HF inductances.

To overcome the drawbacks of using a position estimator, an open-loop type measuring method is proposed in [LI07a] by injecting HF pulsating voltages in the d -axis and q -axis which are obtained from an encoder. The HF inductances are then calculated with the measured HF

current amplitudes on the dq -axes. This method does not require a position estimator. However, accuracies of the calculated inductances are also affected by the delay effects, which will be evaluated later.

A novel HF inductance measuring method which is immune to the delay effects and does not require a position estimator is proposed in this chapter by employing both HF rotating and pulsating voltage injection techniques. The proposed method is composed of HF rotating voltage injection in the stationary reference frame, HF voltage injection in the d -axis, and HF voltage injection in the q -axis. The HF inductances are calculated from the measured HF current amplitudes in each injection stage without using a position estimator. Different from conventional methods which either use rotating voltage injection [XU14], or pulsating voltage injections [KUE14], [LI07a], this method is immune to the delay effects as the delay factor is cancelled out in the process of HF inductance calculation.

This chapter is organized as follows. Firstly, the concept of incremental inductance (called HF inductance in this chapter) and its application in sensorless control are introduced. Then, the conventional methods for obtaining the HF inductances are introduced. Finally, the proposed novel method for measuring the HF incremental inductances is presented in detail. The measured inductances by the proposed method are compared with the simulation results and those of the conventional methods.

3.2 Conventional Methods for Measuring Incremental Inductances

As mentioned in the introduction, the HF inductances are useful in sensorless operation capability evaluation, position estimation error compensation and fault detection. There are many methods for obtaining the HF inductances. In this section, the FEA technique for predicting the HF inductances will firstly be given. Then, the conventional experimental measuring methods by employing either HF rotating voltage injection or HF pulsating voltage injection technique will be presented.

3.2.1 FEA prediction

If the detailed geometric dimensions and material properties of a machine are known, the HF inductances can be obtained by finite element analysis (FEA). The HF inductances $L_{dh}, L_{qh}, L_{dqh}, L_{qdh}$ are defined as the derivatives of flux linkages to currents as

$$L_{dh} = \frac{\partial \psi_d}{\partial i_d}, L_{qh} = \frac{\partial \psi_q}{\partial i_q}, L_{dqh} = \frac{\partial \psi_d}{\partial i_q}, L_{qdh} = \frac{\partial \psi_q}{\partial i_d} \quad (3.1)$$

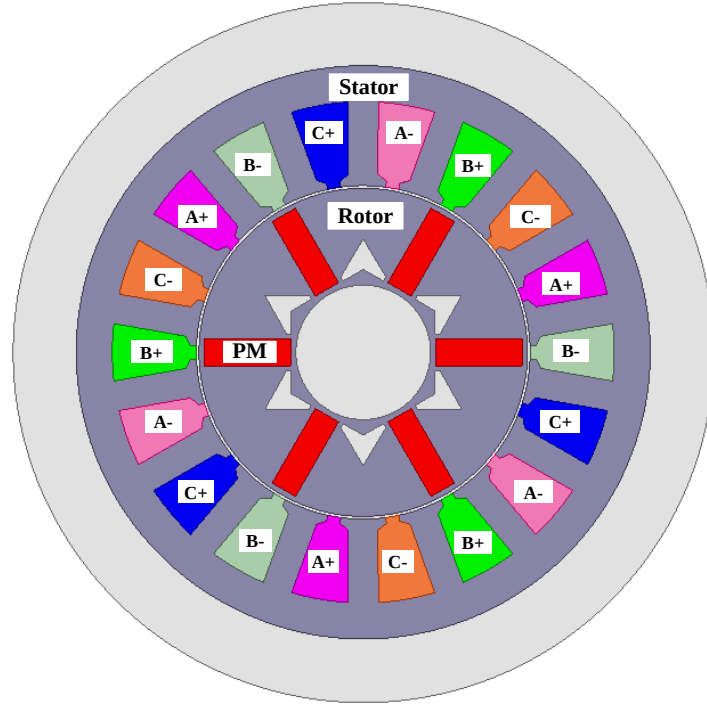
where ψ_d, ψ_q are the dq -axis flux linkages, i_d, i_q are the dq -axis currents. In FEA, the HF inductances at (i_d, i_q) can be calculated as

$$\begin{aligned} L_{dh} &= \left[\psi_d(i_d + \Delta i_d, i_q) - \psi_d(i_d, i_q) \right] / \Delta i_d \\ L_{qh} &= \left[\psi_q(i_d, i_q + \Delta i_q) - \psi_q(i_d, i_q) \right] / \Delta i_q \\ L_{dqh} &= \left[\psi_d(i_d, i_q + \Delta i_q) - \psi_d(i_d, i_q) \right] / \Delta i_q \\ L_{qdh} &= \left[\psi_q(i_d + \Delta i_d, i_q) - \psi_q(i_d, i_q) \right] / \Delta i_d \end{aligned} \quad (3.2)$$

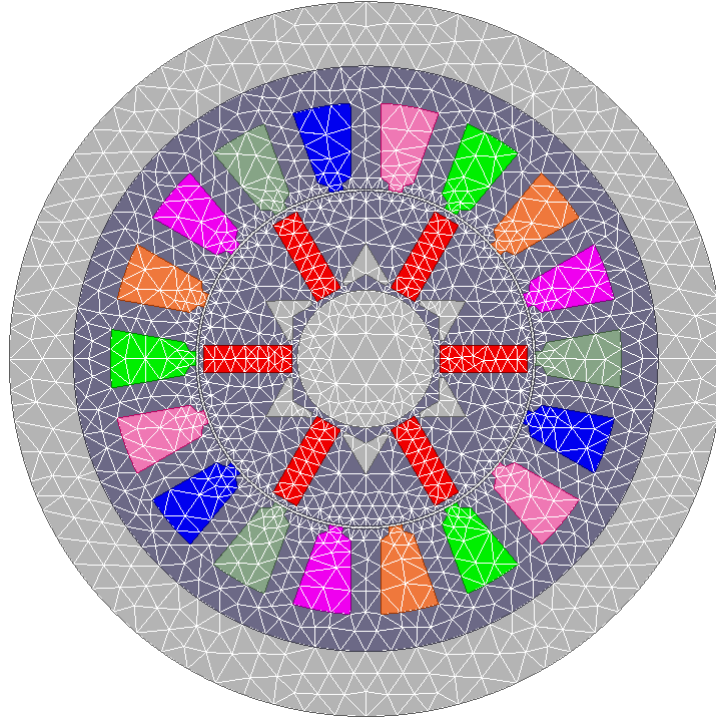
where $\Delta i_d, \Delta i_q$ are the current steps in the simulation. Considering the computation burden and the rated current ($I_N = 4A$), the dq -axis current increments are all chosen as 1A. For calculating HF inductances, there are mainly five steps for calculating the HF inductances.

① Draw machine topology and assign mesh operation.

The cross-section of the machine is shown in Fig. 3.2(a). The detailed machine parameters are listed in Table A1.1 in Appendix 1. There are 6 poles and 18 teeth in the stator. The permanent magnets are embedded in the rotor iron. The airspace outside the stator is added to account for the leakage of flux linkages in the model. The length of the model is $l_m = 0.03m$. The topology after mesh operation is shown in Fig. 3.2(b). Accuracy of the simulation with the mesh in Fig. 3.2(b) is sufficient as the measured HF inductances show a good agreement with the FEA results. If smaller mesh size is adopted, the simulation time is also increased. Therefore, the mesh in Fig. 3.2(b) is appropriate.



(a)



(b)

Fig. 3.2. Schematic of the prototype machine. (a) Cross-section. (b) Cross-section with mesh.

② Setup PM material and current excitations.

The relative permeability of copper and air are approximated as $\mu_r=1$. The relative permeability of permanent magnet is $\mu_r=1.07$, with the remanence $B_r=1.17\text{T}$. The copper conductors in the stator, A+, A-, B+, B-, C+ and C-, are assigned with currents, respectively.

For the prototype machine, each phase winding has $N_w=152$ turns. The three phase currents are assigned according to the rotor position and dq -axis currents as

$$\begin{aligned} i_A &= i_d \cos(\theta_r) - i_q \sin(\theta_r) \\ i_B &= (-0.5i_d + \sqrt{3}/2i_q) \cos(\theta_r) + (\sqrt{3}/2i_d + 0.5i_q) \sin(\theta_r) \\ i_C &= (-0.5i_d - \sqrt{3}/2i_q) \cos(\theta_r) + (-\sqrt{3}/2i_d + 0.5i_q) \sin(\theta_r) \end{aligned} \quad (3.3)$$

③ Solve machine model and obtain three phase flux linkages.

After, the machine topology, materials and currents are constructed in steps 1 and 2, the model is solved to obtain the 3 phase flux linkages. The obtained data are exported after simulation.

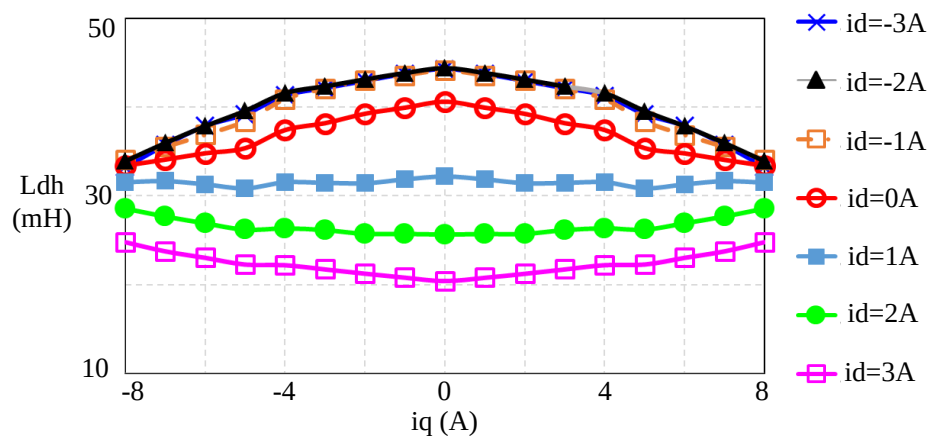
④ Average 3 phase flux linkages with different skew angles.

A mechanical skew angle of 20° is designed for the stator with 5 slices. This helps to reduce flux linkage harmonics. Thus, the flux linkages for different skew angles should be calculated separately and averaged to obtain the overall flux linkages of each phase.

⑤ Calculate the average dq -axis flux linkages in one electrical period.

The dq -axis flux linkages contain ripple components as functions of the rotor position due to slotting effect. The ripple terms repeat periodically every 60° electrical degrees. The flux linkage ripple will induce the 6th harmonics in the back EMF and currents. In the low speed region, the harmonic frequency is much lower than the injected frequency and the current ripples will not affect the injected signal [LI09]. In most cases, the dq -axis flux linkage ripples are neglected. Here, the average values are used.

The FEA predicted HF inductances are listed in Appendix 2 and shown in Fig. 3.3 and the v. The HF cross-coupling inductance is obtained as the average value of the calculated L_{dqh} and L_{qdh} .



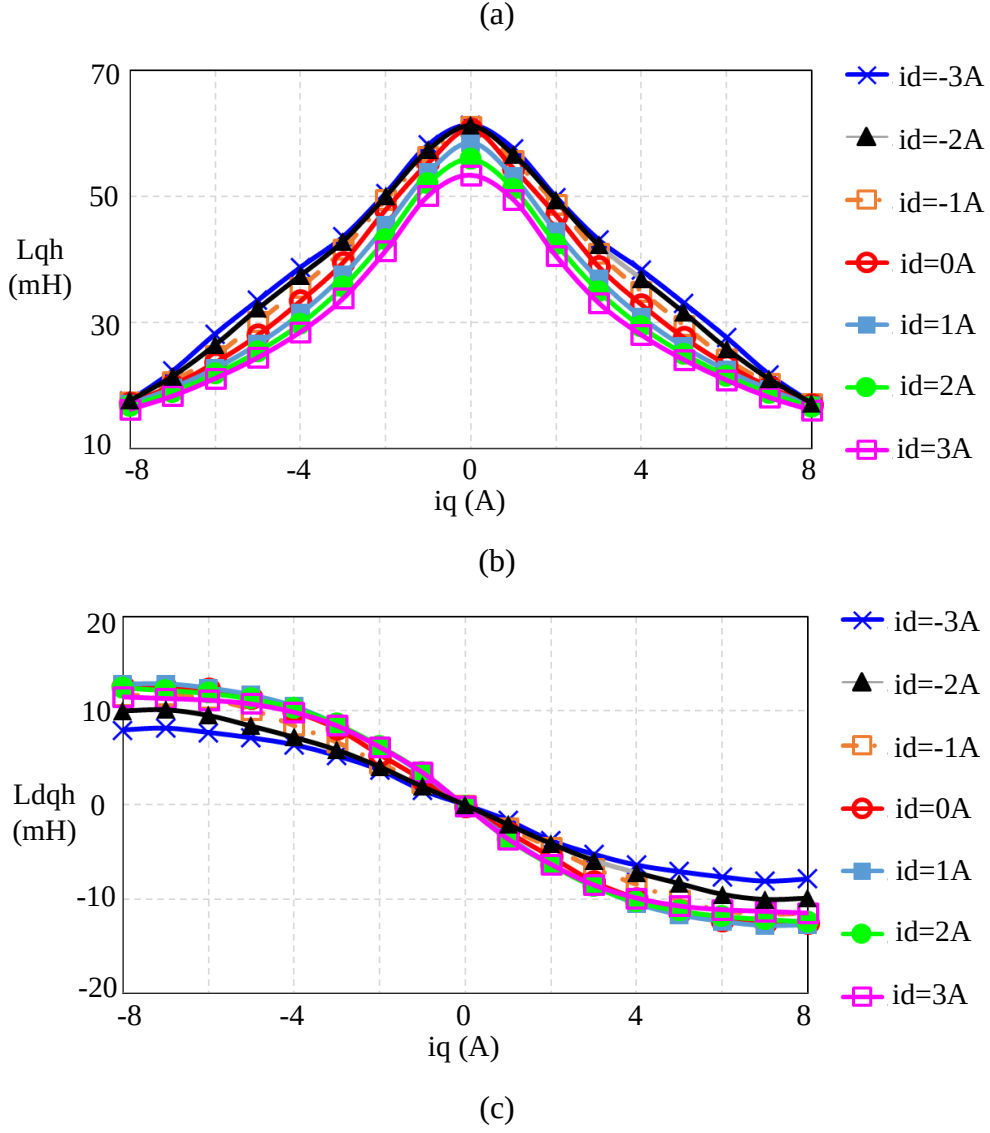


Fig. 3.3. FEA predicted HF inductances. (a) L_{dh} . (b) L_{dh} . (c) L_{dqh} .

Generally, only the HF inductances within the rated current range (4A) are calculated. However, for analyzing the cross-coupling effect and the convergence issue in sensorless control mode under highly saturated condition, the HF inductances within $i_q \in [-8A, 8A]$ are also calculated. For example, at about $(i_d, i_q) = (-1A, 4A)$, L_{dh} is equal to L_{qh} . The convergence of the sensorless position observer is a dynamic process and is not dependent on the specific values of the HF inductances. Therefore, convergence can be achieved even at the zero or reversed saliency point as analysed in [KWO20]. The convergence criteria and corresponding compensation methods will be investigated in the next chapter in detail. In this chapter, the focus is the experimental method for obtaining the HF inductances.

The FEA predicted results in Fig. 3.3 are consistent with those in [LI09] for the same machine. However, in actual applications, the machine design data are not always available especially

when the machine is manufactured by commercial producers and the actual parameters may also drift with time. Thus, experimental methods by making use of general purpose inverter drives are preferred, in which HF voltage injection is commonly employed as follows.

3.2.2 Conventional method based on high frequency rotating voltage injection

Voltage equations of the IPMSM in the synchronous rotating reference frame are given by

$$\begin{bmatrix} u_d^r \\ u_q^r \end{bmatrix} = \begin{bmatrix} R_s + pL_{dh} & -\omega_e L_q + pL_{dqh} \\ \omega_e L_d + pL_{qdh} & R_s + pL_{qh} \end{bmatrix} \begin{bmatrix} i_d^r \\ i_q^r \end{bmatrix} + \omega_e \begin{bmatrix} 0 \\ \psi_{pm} \end{bmatrix} \quad (3.4)$$

where R_s is the stator resistance, ω_e, ψ_{pm} are the electrical angular speed and the PM flux linkage, $u_d^r, u_q^r, i_d^r, i_q^r$ are the dq-axis voltages and currents, “p” denotes the differential operator, the superscript “r” stands for the synchronous rotating reference frame. L_d, L_q are the apparent d-axis and q-axis inductances, L_{dh}, L_{qh} are the incremental d-axis and q-axis inductances, L_{dqh}, L_{qdh} are the incremental mutual inductances between dq-axes and their values are equal. The incremental inductances represent the dynamic behavior of the machine, while the apparent inductances represent the amounts of the dq-axis flux linkages at specific dq-axis currents.

For measuring the inductances of IPMSM at standstill with HF voltage injection technique, the speed dependent terms and the resistance voltage drop are negligible compared to the HF reactance. Therefore, the HF voltage equations in the synchronous rotating reference frame can be obtained from (3.4) as

$$\begin{bmatrix} u_{dh}^r \\ u_{qh}^r \end{bmatrix} = \begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{qh} \end{bmatrix} p \begin{bmatrix} i_{dh}^r \\ i_{qh}^r \end{bmatrix} \quad (3.5)$$

where $u_{dh}^r, u_{qh}^r, i_{dh}^r, i_{qh}^r$ are the HF voltages and currents in the synchronous rotating reference frame, $L_{dh}, L_{qh}, L_{dqh}, L_{qdh}$ are the HF inductances and the mutual HF inductances are equal. Different from the apparent inductance which describes the ratio between flux linkage increment and current, the HF inductance describes the change rate of flux linkage to current.

The HF voltage equations in the stationary reference frame can then be obtained from (3.5) by coordinate transformation as [LIU14]

$$\begin{bmatrix} i_{\alpha h}^s \\ i_{\beta h}^s \end{bmatrix} = R^{-1}(\theta_r) \begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{qh} \end{bmatrix} R(\theta_r) \int \begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} dt, R(\theta_r) = \begin{bmatrix} \cos(\theta_r) & \sin(\theta_r) \\ -\sin(\theta_r) & \cos(\theta_r) \end{bmatrix} \quad (3.6)$$

The HF inductances, i.e. L_{dh}, L_{qh}, L_{dqh} , can be measured by injection HF rotating voltage in the stationary reference frame as

$$u_{\alpha h}^s + ju_{\beta h}^s = V_h e^{j\omega_h t}, \theta_h = \omega_h t \quad (3.7)$$

where V_h, ω_h, θ_h are the amplitude, angular frequency and injection angle of the injected HF voltage. The resulting HF currents can be solved from (3.6), (3.7) as

$$i_{\alpha h}^s + ji_{\beta h}^s = I_p e^{j(\theta_h - \pi/2)} + I_n e^{-j[\theta_h - 2(\theta_r - \Delta\theta_m) - \pi/2]} \quad (3.8)$$

$$I_p = \frac{V_h L_\Sigma}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)}, I_n = \frac{V_h \sqrt{L_\Delta^2 + L_{dqh}^2}}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)} \quad (3.9)$$

$$L_\Sigma = (L_{dh} + L_{qh})/2, L_\Delta = (L_{qh} - L_{dh})/2 \quad (3.10-a)$$

$$\Delta\theta_m = 0.5 \tan^{-1} \left(\frac{2L_{dqh}}{L_{qh} - L_{dh}} \right) \quad (3.10-b)$$

where $\Delta\theta_m$ is the angle offset caused by the cross-coupling effect [LI09], [SHU20], I_p, I_n are the amplitudes of the positive and negative sequence currents, and can be extracted by the method in Fig. 3.4.

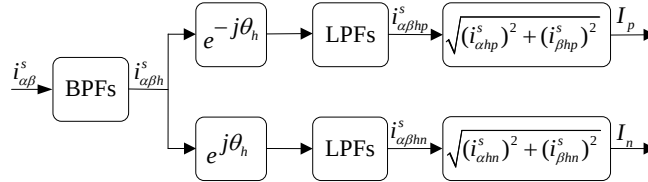


Fig. 3.4. Demodulation of the amplitudes of positive and negative sequence HF currents due to HF rotating voltage injection.

The positive and negative sequence currents are firstly transferred to low frequency domain and filtered out as $i_{\alpha\beta hp}^s, i_{\alpha\beta hn}^s$, i.e.

$$i_{\alpha\beta hp}^s = I_p e^{j(-\pi/2)} \quad (3.11-a)$$

$$i_{\alpha\beta hn}^s = I_n e^{j(2\theta_r - 2\Delta\theta_m + \pi/2)} \quad (3.11-b)$$

According to (3.9), I_p, I_n are positive and their values can be obtained from the moduli of $i_{\alpha\beta hp}^s, i_{\alpha\beta hn}^s$. The calculation of modulus is not dependent on the phase information. Hence, it is not affected by the phase distortion caused by the delay effects, including filtering delay

and voltage synthesis delay. The angle $\Delta\theta_m$ can be obtained by measuring the difference between the actual rotor position obtained from a position encoder and the estimated rotor position obtained by sensorless position estimation technique incorporating heterodyning demodulation as shown in Fig. 3.5(a).

The angle $\Delta\theta_m$ can be obtained by measuring the difference between the actual rotor position obtained from a position encoder and the estimated rotor position obtained by sensorless position estimation technique. The heterodyning demodulation based sensorless position estimation method shown in Fig. 3.5(a) can be used to estimate the rotor position.

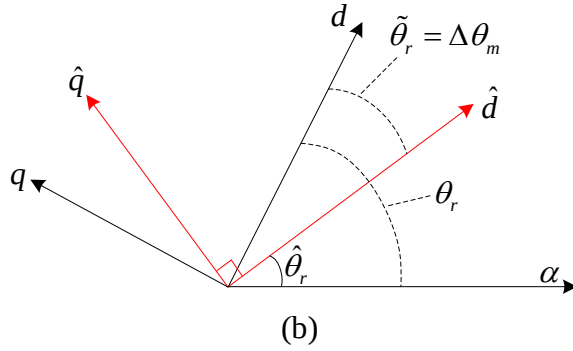
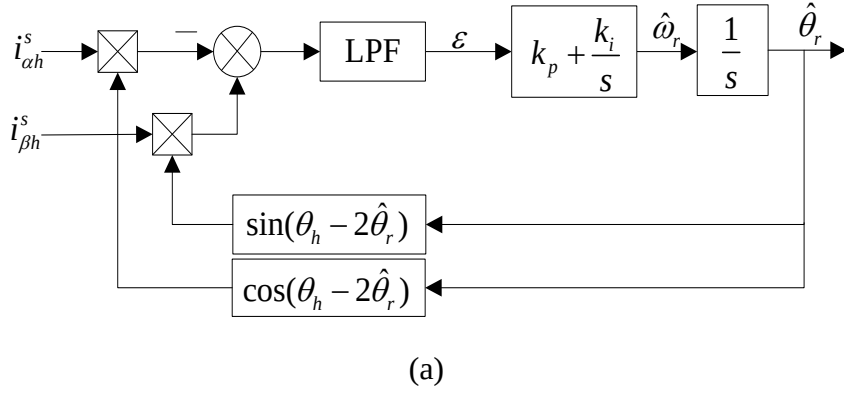


Fig. 3.5. Sensorless position estimation based on HF rotating voltage injection. (a) Position estimator. (b) Estimated position and error.

where $\hat{\theta}_r$ is the estimated rotor position, k_p, k_i are the estimator parameters. The low pass filter (LPF) in Fig. 3.5(a) is used to filter out the HF component after heterodyning operation. The error signal ε is obtained from (3.8) as

$$\varepsilon = I_n \sin\left(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r\right) \quad (3.12)$$

The estimated rotor position is obtained by controlling $\varepsilon=0$, i.e.

$$\hat{\theta}_r = \theta_r - \Delta\theta_m \quad (3.13)$$

As can be seen in (3.13), there is a position estimation error as shown in Fig. 3.5(b). The angle $\Delta\theta_m$ can be obtained from the difference between the rotor position θ_r measured by encoder and the estimated position $\hat{\theta}_r$. With the measured I_p, I_n and $\Delta\theta_m$, the HF inductances can be solved from (3.9), (3.11) as

$$L_{dh} = \frac{V_h \left(I_p - I_n / \sqrt{1 + \tan^2(2\Delta\theta_m)} \right)}{\omega_h (I_p^2 - I_n^2)} \quad (3.14)$$

$$L_{qh} = \frac{V_h \left(I_p - I_n / \sqrt{1 + \tan^2(2\Delta\theta_m)} \right)}{\omega_h (I_p^2 - I_n^2)} \quad (3.15)$$

$$L_{dqh} = \frac{V_h I_n \tan(2\Delta\theta_m)}{\omega_h (I_p^2 - I_n^2) \sqrt{1 + \tan^2(2\Delta\theta_m)}} \quad (3.16)$$

Obviously, this method is dependent on the estimated position, and the accuracy of the estimated position is affected by the delay effects [GHA13], [XU17a]. The actual applied HF voltage has delay to the reference HF voltage as denoted by φ_v in Fig. 3.6 due to voltage synthesis and pulse-width-modulation (PWM) update delay [BAE03]. The HF currents which are used for rotor position estimation also have delays to the actual HF currents due to sampling and digital filters as denoted by φ_i in Fig. 3.6. Therefore, the angle θ_h used for position demodulation in Fig. 3.5 has a delay to the desired demodulation angle by $(\varphi_v + \varphi_i)$, which will result in an error of $0.5(\varphi_v + \varphi_i)$ in the estimated rotor position if not compensated [GHA13]. However, the accurate delay, i.e. $(\varphi_v + \varphi_i)$ is hard to determine as it is related to the current levels, sampling hardware, and the injection frequency [XU17a]. The power angle of the HF circuit is almost 90 degrees as the resistance is relatively small compared to the reactance. Hence, the phase delay caused by the resistance is neglected in this thesis.

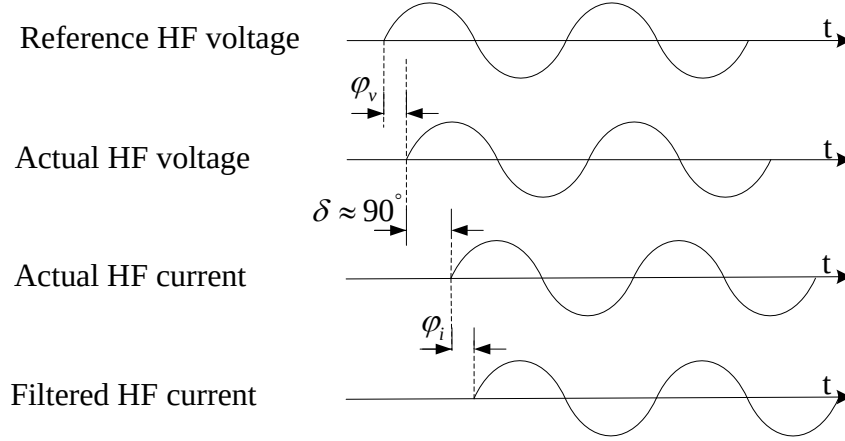


Fig. 3.6. Delay effects in HF voltage injection based sensorless control.

Besides, tuning of the estimator parameters is dependent on the saliency information which is not known before measuring [ERT20]. Therefore, the HF rotating voltage injection based measuring method requires a closed-loop position estimator and is affected by the delay effects.

3.2.3 Conventional method based on high frequency pulsating voltage injection

The HF pulsating voltage injection technique can also be used for measuring the HF inductances [KUE14], [LI07a]. In this section, two kinds of HF inductances measuring methods will be introduced, which are all based on pulsating voltage injection. One measuring method requires a position estimator, while the other one does not. They are both affected by the delay effects, as will be presented later.

(a) Measuring method requiring a position estimator based on pulsating voltage injections in the estimated d- and q-axis. (conventional method 2a)

For the first measuring method that requires a position estimator, the HF inductance matrix in (3.5) is symmetrical and its diagonalization is given by [KUE14]

$$\begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{qh} \end{bmatrix} = R(\Delta\theta_m) \begin{bmatrix} X_1 & 0 \\ 0 & X_2 \end{bmatrix} R^{-1}(\Delta\theta_m) \quad (3.17)$$

where X_1, X_2 are the two characteristic roots as

$$X_1 = L_\Sigma - \sqrt{L_\Delta^2 + L_{dqh}^2}, X_2 = L_\Sigma + \sqrt{L_\Delta^2 + L_{dqh}^2} \quad (3.18)$$

As can be seen from (3.17), the HF inductances can be calculated by measuring X_1, X_2 and $\Delta\theta_m$ respectively. $\Delta\theta_m$ can be measured by injecting HF pulsating voltage as (3.19) to the estimated d-axis. The HF current amplitudes on the estimated dq-axes can be solved as (3.20).

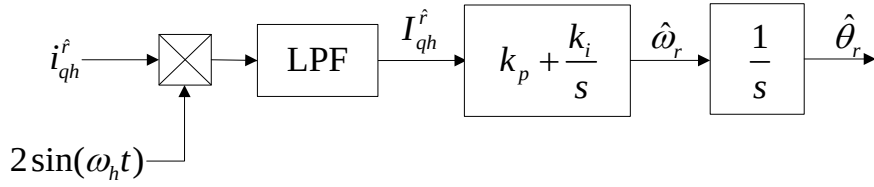
$$u_{dh}^{\hat{r}} = V_h \cos(\omega_h t) \quad (3.19)$$

$$i_{dh}^{\hat{r}} = I_{dh}^{\hat{r}} \sin(\omega_h t), i_{qh}^{\hat{r}} = I_{qh}^{\hat{r}} \sin(\omega_h t) \quad (3.20)$$

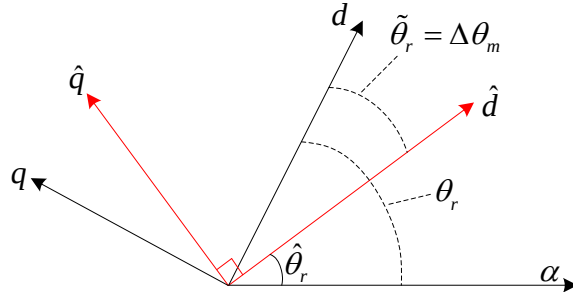
where “ $\hat{r}$ ” denotes the estimated synchronous rotating reference frame, $I_{dh}^{\hat{r}}, I_{qh}^{\hat{r}}$ are the amplitudes of the HF currents and can be extracted by synchronous demodulation [TAN17] shown in Fig. 3.7, i.e.

$$I_{dh}^{\hat{r}} = \frac{V_h (X_1 + X_2)}{2\omega_h X_1 X_2} + \frac{V_h (X_2 - X_1)}{2\omega_h X_1 X_2} \cos(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \quad (3.21)$$

$$I_{qh}^{\hat{r}} = \frac{V_h (X_2 - X_1)}{2\omega_h X_1 X_2} \sin(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \quad (3.22)$$



(a)



(b)

Fig. 3.7. Sensorless position estimator based on HF pulsating voltage injection. (a) Position estimator. (b) Estimated position and error.

The estimated position $\hat{\theta}_r$ can be obtained by controlling $I_{qh}^{\hat{r}} = 0$, as shown in Fig. 3.7(b), i.e.

$$\hat{\theta}_r = \theta_r - \Delta\theta_m \quad (3.23)$$

which is the same as (3.13) for HF rotating voltage injection based sensorless control. $\Delta\theta_m$ can be obtained from the difference between θ_r and $\hat{\theta}_r$. X_1 can then be calculated from the measured d-axis current amplitude at steady state as

$$X_1 = \frac{V_h}{\omega_h I_{dh}^{\hat{r}}} \quad (3.24)$$

The value of X_2 can be calculated in a similar way by tracking the estimated q -axis. When HF pulsating voltage is injected to the estimated q -axis, the HF currents on the estimated dq -axes are

$$I_{dh}^{\hat{r}} = \frac{V_h (X_2 - X_1)}{2\omega_h X_1 X_2} \sin(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \quad (3.25)$$

$$I_{qh}^{\hat{r}} = \frac{V_h (X_1 + X_2)}{2\omega_h X_1 X_2} - \frac{V_h (X_2 - X_1)}{2\omega_h X_1 X_2} \cos(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \quad (3.26)$$

As can be seen, by controlling $I_{dh}^{\hat{r}} = 0$, the estimated position can also be obtained as $\hat{\theta}_r = \theta_r - \Delta\theta_m$, and the value of X_2 can be obtained from $I_{qh}^{\hat{r}}$ at steady state as

$$X_2 = \frac{V_h}{\omega_h I_{qh}^{\hat{r}}} \quad (3.27)$$

Therefore, in order to obtain X_1, X_2 , two separate injections are necessary [KUE14]. The HF inductances can then be calculated from X_1, X_2 and $\Delta\theta_m$ with (3.17). Similar to the method in [XU14] with rotating voltage injection, this method also requires a position estimator. The delay effects will not affect the estimated rotor position $\hat{\theta}_r$ and the observed $\Delta\theta_m$, but may cause error in the measured HF current amplitude $I_{dh}^{\hat{r}}$, i.e.

$$LPF \left[2 \sin(\omega_h t) I_{dh}^{\hat{r}} \sin(\omega_h t + \varphi_{delay}) \right] = I_{dh}^{\hat{r}} \cos(\varphi_{delay}) \quad (3.28)$$

where φ_{delay} is the actual phase delay between the filtered HF current and the demodulation signal $\sin(\omega_h t)$, and can also be described by the voltage synthesis delay and current filtering delay in Fig. 3.6. As can be seen, from (3.28), there will be a scaling factor $1/\cos(\varphi_{delay})$ in the measured HF inductances with pulsating voltage injections if the delay is not compensated. However, the delay is difficult to determine accurately as it is dependent on the current level, sampling hardware and the injection frequency [XU17].

(b) Measuring method without a position estimator based on pulsating voltage injections in the actual d - and q -axis. (conventional method 2b)

To avoid the use of a position estimator which requires complicated parameter tuning, a measuring method is proposed in [LI07a] by injecting pulsating voltage $V_h \cos(\omega_h t)$ in the actual d -axis and q -axis, respectively. The HF inductances can be calculated from the measured HF current amplitudes as

$$\begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{dh} \end{bmatrix} = \frac{V_h}{\omega_h} \begin{bmatrix} I_{dh}^I & I_{dh}^{II} \\ I_{qh}^I & I_{qh}^{II} \end{bmatrix}^{-1} \quad (3.29)$$

where I_{dh}^I, I_{qh}^I are the HF current amplitudes on the dq -axes due to HF voltage injection in d -axis, I_{dh}^{II}, I_{qh}^{II} are the HF current amplitudes on the dq -axes due to HF voltage injection in q -axis, as extracted by synchronous demodulation as

$$I_{dh}^I = LPF \left(2 \sin(\omega_h t) i_{dh}^I \right), I_{qh}^I = LPF \left(2 \sin(\omega_h t) i_{qh}^I \right) \quad (3.30-a)$$

$$I_{dh}^{II} = LPF \left(2 \sin(\omega_h t) i_{dh}^{II} \right), I_{qh}^{II} = LPF \left(2 \sin(\omega_h t) i_{qh}^{II} \right) \quad (3.30-b)$$

This method features an open-loop type implementation as the sensorless position estimator is not required. However, the measured results are also affected by the delay effects described in Fig. 3.6. Due to the delay effects as analysed in (3.28), there will be a scaling factor in the calculated results as

$$\begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{dh} \end{bmatrix} = \frac{1}{\cos(\varphi_{delay})} \frac{V_h}{\omega_h} \begin{bmatrix} I_{dh}^I & I_{dh}^{II} \\ I_{qh}^I & I_{qh}^{II} \end{bmatrix}^{-1} \quad (3.31)$$

As can be seen from [XU14], [KUE14], [LI07a], the existing measuring methods with either HF rotating or pulsating injections are all suffered from the delay effects and some of them [XU14], [KUE14] require position estimators which increase the complexity of the measuring methods. In this thesis, a novel measuring method which is immune to the delay effects and does not require a position estimator will be proposed and described in the next section.

3.3 Proposed Method Based on both Rotating and Pulsating HF Voltage Injections

In this section, the proposed measuring method is firstly presented based on both rotating and pulsating high frequency voltage injection techniques. Compared with conventional measuring methods based on either rotating or pulsating voltage injection, the proposed method does not require a position estimator and is immune to the delay effects. Then, the measuring principle and the impacts of delay effects are analysed. Finally, comparisons between the proposed and conventional methods are made.

3.3.1 Proposed method

The control diagram of the proposed method is shown in Fig. 3.8.

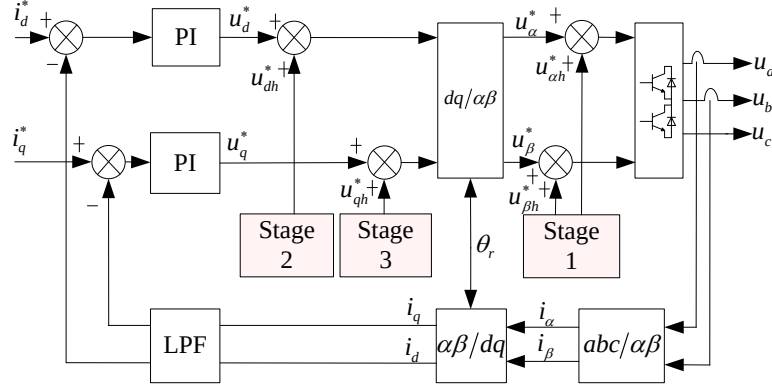


Fig. 3.8. Control diagram of the proposed measuring method.

The machine is controlled in current mode to provide stable working point for the estimation. During the estimation process, the rotor is locked to a fixed position to prevent movement. The proposed method is mainly composed of three injection stages as shown in Fig. 3.9, i.e. rotating voltage injection, HF pulsating voltage injection in d -axis, and HF pulsating voltage injection in q -axis, respectively. The positions of d -axis and q -axis are obtained from a position encoder. The three injection stages as shown in Fig. 3.9 are repeated for each current point (i_d, i_q) .

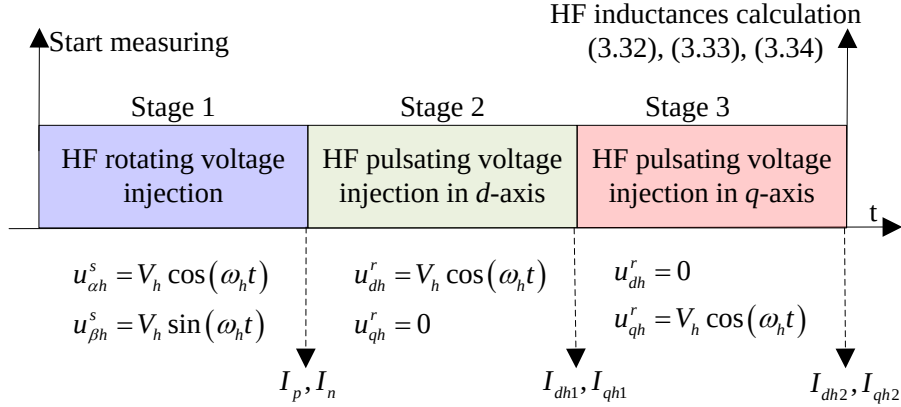


Fig. 3.9. Proposed measuring method.

V_h, ω_h are the amplitude and angular frequency of the injected HF voltages. I_p, I_n are the measured amplitudes of the positive and negative sequence HF currents due to rotating voltage injection. I_{dh1}, I_{qh1} are the measured dq -axis HF current amplitudes due to HF pulsating voltage injection in the d -axis, I_{dh2}, I_{qh2} are the measured dq -axis HF current amplitudes due to HF pulsating voltage injection in the q -axis. The applied sequence of the three injection stages is not important. The HF inductances can be calculated as

$$L_{dh} = \frac{2V_h I_p I_{qh2}}{\omega_h (I_p^2 - I_n^2) (I_{dh1} + I_{qh2})} \quad (3.32)$$

$$L_{qh} = \frac{2V_h I_p I_{dh1}}{\omega_h (I_p^2 - I_n^2) (I_{dh1} + I_{qh2})} \quad (3.33)$$

$$L_{dqh} = -2V_h I_p \frac{I_{qh1} \text{ or } I_{dh2}}{\omega_h (I_p^2 - I_n^2) (I_{dh1} + I_{qh2})} \quad (3.34)$$

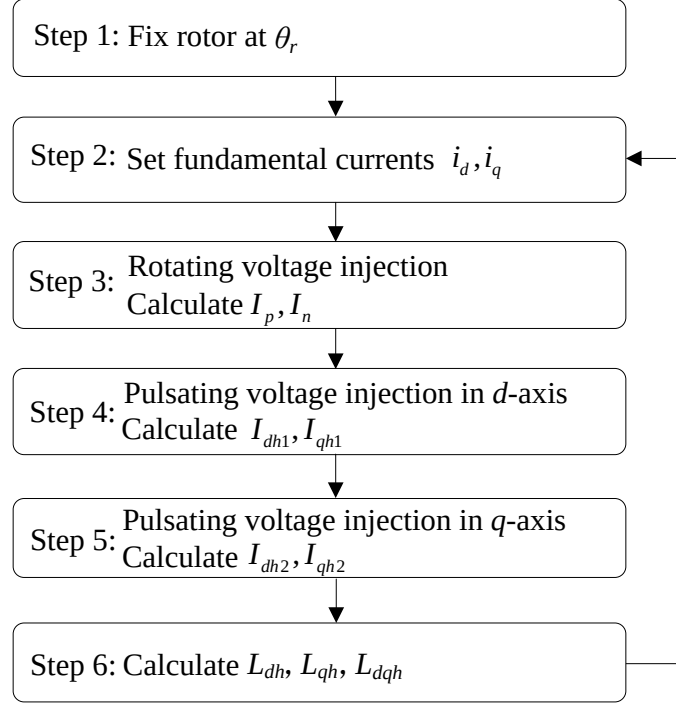


Fig. 3.10. Measuring procedure for the proposed method.

Step 1: the rotor is fixed at a known position. If the rotor is disconnected with the mechanical load, the rotor can be parked at a known position by applying a direct current to the stator windings. The rotor is then mechanically locked to prevent any movement during test. Whereas, if the rotor is connected to an external load and is not permitted to rotate or deviate from its original position, the initial position can be estimated by HF injection based sensorless algorithms [WAN14].

Step 2: the d-axis and q-axis current references are assigned to the current controllers. After the feedback dq-axis fundamental currents reach the reference values.

Step 3: HF rotating voltage is injected to the stationary frame, and the HF current amplitudes I_p, I_n are extracted.

Step 4: HF pulsating voltage is injected to the actual d-axis, and the dq-axis HF current amplitudes I_{dh1}, I_{qh1} are extracted.

Step 5: HF pulsating voltage is injected to the actual q-axis, and the dq-axis HF current amplitudes I_{dh2}, I_{qh2} are extracted.

Step 6: the HF inductances are calculated. Then, new current references are assigned and the step 2 to step 6 are repeated until all the i_d, i_q points are measured. The measured results will be shown in the experiments.

The proposed method is immune to the delay effects as will be verified later.

3.3.2 Delay effect immune property analysis

Then, how the HF inductances can be calculated from (3.32)-(3.34) will be explained. Firstly, The d -axis and q -axis HF inductances can be equivalently expressed as

$$L_{dh} = \frac{L_{dh} + L_{qh}}{2} \left(1 - \frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}} \right) \quad (3.35)$$

$$L_{qh} = \frac{L_{dh} + L_{qh}}{2} \left(1 + \frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}} \right) \quad (3.36)$$

As can be seen from (3.35), (3.36), the values of L_{dh}, L_{qh} can be obtained by measuring the values of $(L_{dh} + L_{qh})$ and $\frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}}$, respectively. According to (3.9), the value of $(L_{dh} + L_{qh})$ can be obtained from the measured HF current amplitudes I_p, I_n in the HF rotating voltage injection stage as

$$L_{dh} + L_{qh} = \frac{2V_h I_p}{\omega_h (I_p^2 - I_n^2)} \quad (3.37)$$

The value of $\frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}}$ can be obtained from I_{dh1}, I_{qh2} . I_{dh1} is the d-axis HF current amplitude due to d-axis HF voltage injection in stage 2 and can be solved from (3.5) as (3.38). I_{qh2} is the q-axis current amplitude due to q-axis HF voltage injection in stage 3 and can be solved from (3.5) as (3.39).

$$I_{dh1} = \frac{V_h L_{qh}}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)} \quad (3.38)$$

$$I_{qh2} = \frac{V_h L_{dh}}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)} \quad (3.39)$$

As can be seen, $\frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}}$ can be obtained from (3.38), (3.39) as

$$\frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}} = \frac{I_{dh1} - I_{qh2}}{I_{dh1} + I_{qh2}} \quad (3.40)$$

By substituting $(L_{dh} + L_{qh})$ and $\frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}}$ in (3.35), (3.36) with (3.37), (3.40), L_{dh}, L_{qh} can be calculated as (3.32), (3.33) for the proposed method. It is also noted that in the measurement of $\frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}}$, the delay φ_{delay} is cancelled in the division operations as

$$\frac{I_{dh1} \cos(\varphi_{delay}) - I_{qh2} \cos(\varphi_{delay})}{I_{dh1} \cos(\varphi_{delay}) + I_{qh2} \cos(\varphi_{delay})} = \frac{I_{dh1} - I_{qh2}}{I_{dh1} + I_{qh2}} \quad (3.41)$$

Therefore, the measured results with the proposed method are immune to the delay effects. After L_{dh}, L_{qh} are calculated, L_{dqh} can be obtained from $-\frac{I_{qh1}}{I_{dh1}}$ which is the ratio between L_{dqh} and L_{qh} as used in [LI09b], i.e.

$$\frac{L_{dqh}}{L_{qh}} = -\frac{I_{qh1}}{I_{dh1}} \quad (3.42)$$

Actually, L_{dqh} can also be obtained from $-\frac{I_{dh2}}{I_{qh2}}$ which is the ratio between L_{dqh} and L_{dh} as (3.43). L_{dqh} can then be calculated from the measured L_{dh} and (3.43).

$$\frac{L_{dqh}}{L_{dh}} = -\frac{I_{dh2}}{I_{qh2}} \quad (3.43)$$

3.3.3 Comparison with conventional methods

The proposed measuring method includes three injection stages which are sinusoidal rotating voltage injection, pulsating voltage injection in d-axis, and pulsating voltage injection in q-axis, while the conventional methods use either rotating or pulsating voltage injection as shown in Fig. 3.11.

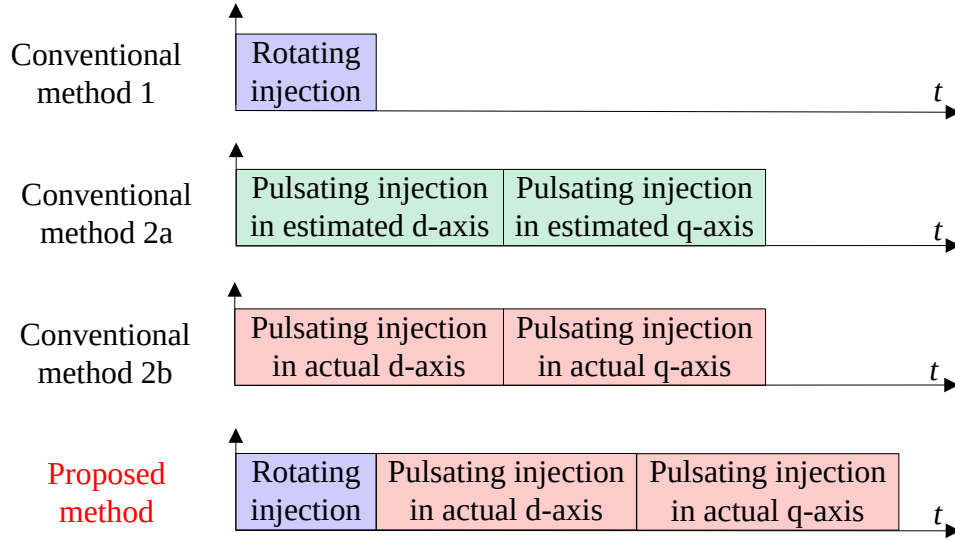


Fig. 3.11. Comparison of injection sequences between the proposed and conventional methods.

Selection of the injection voltage

Selection of the injection frequency should take the estimation time and voltage utilization into consideration. The estimation time can be reduced with higher injection frequency as it usually takes several injection periods for extracting the HF current amplitudes. However, with higher injection frequency, higher voltage amplitude is also required to maintain the amplitudes of the HF currents, which decreases the available voltage for fundamental currents. The optimum voltage amplitude is dependent on the resolution of the sampling circuit and the machine characteristics, and it is difficult to find analytical expressions between the accuracy requirement and applied voltages from (3.32)-(3.34). A feasible way to find a suitable voltage amplitude is to gradually increase the injection voltage amplitude under the same injection frequency until the estimated results are almost unchanged. According to the above guidelines, 20V/500Hz are chosen for both HF rotating and pulsating voltages under a dc bus voltage of 150V.

Resolution of the current sampling circuit is important for improving accuracy of the estimation. For the maximum current sampling range $[-I_{max}, I_{max}]$ with sampling resolution N bits, the sampled current resolution is

$$I_{LSB} = \frac{2I_{max}}{2^N} \quad (3.44)$$

The measurement error of the current due to quantization is $0.5I_{LSB}$ [25]. In (3.32)-(3.34), the term $\frac{2V_h I_p}{\omega_h(I_p^2 - I_n^2)(I_{dh1} + I_{qh2})}$ is assumed to be fixed due to relative large value, and the estimation error caused by quantization can be approximated as

$$\frac{\Delta L_{dh}}{L_{dh}} = \frac{0.5 I_{LSB}}{I_{qh2}} * 100\% \quad (3.45-a)$$

$$\frac{\Delta L_{qh}}{L_{qh}} = \frac{0.5 I_{LSB}}{I_{dh1}} * 100\% \quad (3.45-b)$$

$$\frac{\Delta L_{dqh}}{L_{dqh}} = \frac{0.5 I_{LSB}}{I_{dh2} \text{ or } I_{qh1}} * 100\% \quad (3.45-c)$$

These equations provide a simple way for evaluating the quantization error and choosing the required sampling resolution at different working point by using the measured HF current amplitudes. For example, for the estimation at $i_d = 0A, i_q = 2A$ with 16 bits sampling resolution in the experiment, i.e. $I_{LSB} = 0.224mA, I_{qh2} = 55mA$, the L_{dh} estimation error caused by quantization is $\frac{0.5*0.224}{55} 100\% = 0.20\%$. If the sampling circuit resolution is 12 bits which is commonly used in commercial inverters, the L_{dh} estimation error caused by sampling quantization is increased to 3.2%.

Table 3.1 compares the proposed method with conventional measuring methods. The proposed method does not require a position estimator and is immune to the delay effects. Therefore, it is a choice for measuring the HF inductances.

Table 3.1
COMPARISON OF HF INDUCTANCE MEASURING METHODS

	Rotating	Pulsating		Proposed
	[XU14]	[KUE14]	[LI07a]	
Injection stages	Rotating	Pulsating, Pulsating	Pulsating, Pulsating	Rotating, Pulsating, Pulsating
Independent of estimator	No	No	Yes	Yes
Immune to delay	No	No	No	Yes

Application of the proposed method

The proposed method provides a method for estimating the HF inductances by injection three HF voltages. It can also be used for online HF inductances estimation as that in [KAN21], but some requirements are essential as: 1) the machine should be operated at standstill and low speed to reduce the interference between the fundamental currents and the HF currents [XU14], 2) the dq -axis positions should be obtained from either position sensor or sensorless algorithms. As the HF inductances vary with the saturation level, the fundamental currents should be

controlled stable during a single estimation. If the current command changes within this interval, the estimated results are not reflections of the saturation level at the current point. Besides, the changing fundamental currents may also produce interference with the injected HF currents, which further distorts the estimated results [BRI00].

The PM magnetization strength has impact on the estimated HF inductances. The actual HF inductances may change due to aging effects of the PM after the machine is used for a period [FER20]. Therefore, the HF inductances of the machine should be estimated regularly to keep up to date. For example, the proposed estimation method can be carried out each time before start. Meanwhile, the PM magnetization strength can also be assessed from the estimated HF inductance data by setting up an empirical model between the HF inductances and the PM strength [MAR20].

As only the d -axis and q -axis positions are required for injecting the HF voltages, the proposed method can theoretically be applied to different kinds of machines which can be modelled in the dq -axis frame. However, for some specific machines such as fractional slot machines with surface mounted PMs, the estimated results may have low accuracy due to relatively low inductances. The accuracy can be improved by choosing higher injection voltage and higher sampling resolution accordingly.

For high performance machine control, the apparent inductances are also very important. Although with the proposed method, the apparent inductances cannot be directly estimated, their values can be obtained by integrating the estimated HF inductances [XU14] as

$$L_d = \frac{1}{i_d} \int_0^{i_d} L_{dh} di = \frac{1}{i_d} \sum_{k=1}^{i_d/\Delta i_d} L_{dh}(k) \Delta i_d \quad (3.46-a)$$

$$L_q = \frac{1}{i_q} \int_0^{i_q} L_{qh} di = \frac{1}{i_q} \sum_{k=1}^{i_q/\Delta i_q} L_{qh}(k) \Delta i_q \quad (3.46-b)$$

$$L_{dq} = \frac{1}{i_q} \int_0^{i_q} L_{dqh} di = \frac{1}{i_q} \sum_{k=1}^{i_q/\Delta i_q} L_{dqh}(k) \Delta i_q \quad (3.46-c)$$

where $\Delta i_d, \Delta i_q$ are the increments of the fundamental currents. For improving accuracy of the estimated apparent inductances, the integration intervals $\Delta i_d, \Delta i_q$ should be very small which results in very large sampling point and increases the complexity of the measurement. Besides, the discrete integration form in (3.46-a)-(3.46-c) will inevitably introduce discretization errors. There are still other apparent inductances estimation methods which are not dependent on HF voltage injections and need to rotate the machine, such as measuring the flux linkages by integrating the back EMF [PE15], injecting short negative d -axis current pulse [LIU11] and those based on Kalman filters and recursive least square method [ZHU21].

3.3.4 Experimental results

To verify effectiveness of the proposed measuring method, finite element analysis (FEA) is firstly carried out. The FEA predicted HF inductances provide references for the measured results. Then, the proposed measuring method is verified on a dSPACE platform DS1006 as shown in Fig. 3.12. A commercial inverter which receives PWM signals from the dSPACE is used to drive the test machine. The parameters of the machine are shown in Appendix 1. Two external hall-effect based current sensors are installed on an analog conditioning board for sampling phase currents. The resolution of the analog to digital converter for sampling current is 16bits. An incremental encoder of 1024 pulses per revolution is used to obtain the rotor position. The PWM and sampling frequency are all set as 10kHz. For measuring the HF inductances with the proposed method, the amplitudes and frequencies of the rotating sinusoidal voltage and pulsating sinusoidal voltage are all chosen as 20V/500Hz considering sampling accuracy and saliency level [ZHU11].

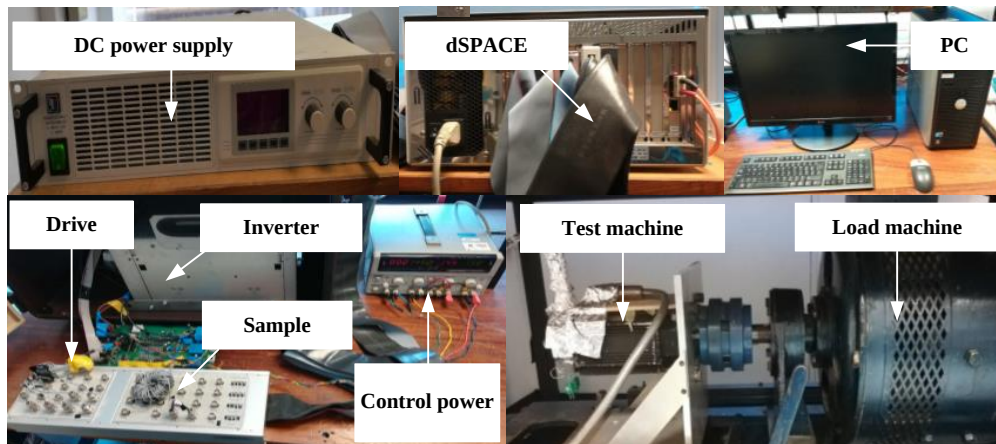


Fig. 3.12. Experimental setup.

A. Measured results

Fig. 3.13 shows the dq-axis currents for the HF inductance measurement at $i_d = 0A, i_q = 2A$.

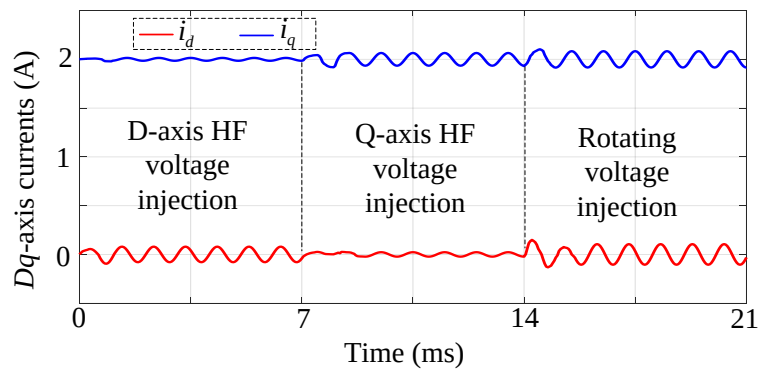


Fig. 3.13. Dq -axis current sequences with the proposed measuring method at $i_d = 0A, i_q = 2A$.

As can be seen, the proposed measuring procedure at one specific point is composed of three injection stages, i.e. d-axis pulsating voltage injection, q-axis pulsating voltage injection and rotating voltage injection. The q-axis (d-axis) HF current during d-axis (q-axis) voltage injection is due to the cross-coupling effect. When HF rotating voltage is injected to the stationary reference frame, both dq-axes will contain HF currents. The sequence of the three injection stages is not important for HF inductance measuring as long as the HF current amplitudes in each injection stage can be measured. The duration for each injection stage should be long enough for the HF currents to be stable and for the measuring strategy to extract the HF current amplitudes. In this measurement, the duration for each injection stage is 14ms, i.e. 7 injection cycles for 500Hz injection frequency.

Fig. 3.14 shows the measured HF inductances at different rotor positions for the working point $i_d = 0A, i_q = 2A$. It is found that the measured HF inductances vary with the rotor position.

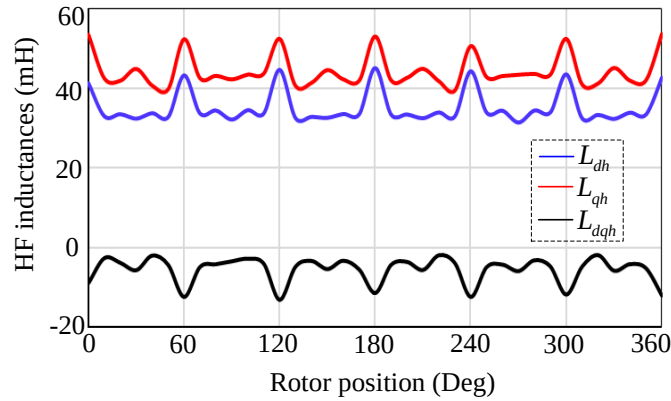


Fig. 3.14. Measured HF inductances with the proposed method at different rotor positions for $i_d = 0A, i_q = 2A$.

As can be seen, the measured HF inductances contain sixth harmonics which can be caused by the inverter nonlinearities and spatial harmonics [KWO20]. The concept of equivalent HF resistance [GON11] can be used to modelling the HF voltage losses caused by the inverter nonlinearities. The equivalent HF resistances are dependent on the phase currents and only exist near the zero current regions. As the three phase currents cross zero six times during one electrical period revolution of the rotor position, there will be sixth harmonics in the HF voltage losses which will result in ripples in the demodulated current amplitudes. The ripples can be reduced by increasing the injection voltage amplitude at the cost of reduced voltage margins. The spatial inductance harmonics also contribute to the ripples in the measured inductances. In

this thesis, to obtain the average HF inductances, the measured HF currents amplitudes of the three injection stages during one electrical period are averaged and used for calculation.

Fig. 3.15(a) shows the measured amplitudes of the positive and negative sequence HF currents I_p, I_n due to rotating voltage injection of the proposed method at different i_d, i_q . Fig. 3.15(b) shows the HF current amplitudes I_{dh1}, I_{qh1} on the dq-axes due to pulsating voltage injection in d-axis, while Fig. 3.15(c) shows the dq-axis HF current amplitudes I_{dh2}, I_{qh2} due to pulsating voltage injection in q-axis of the proposed method.

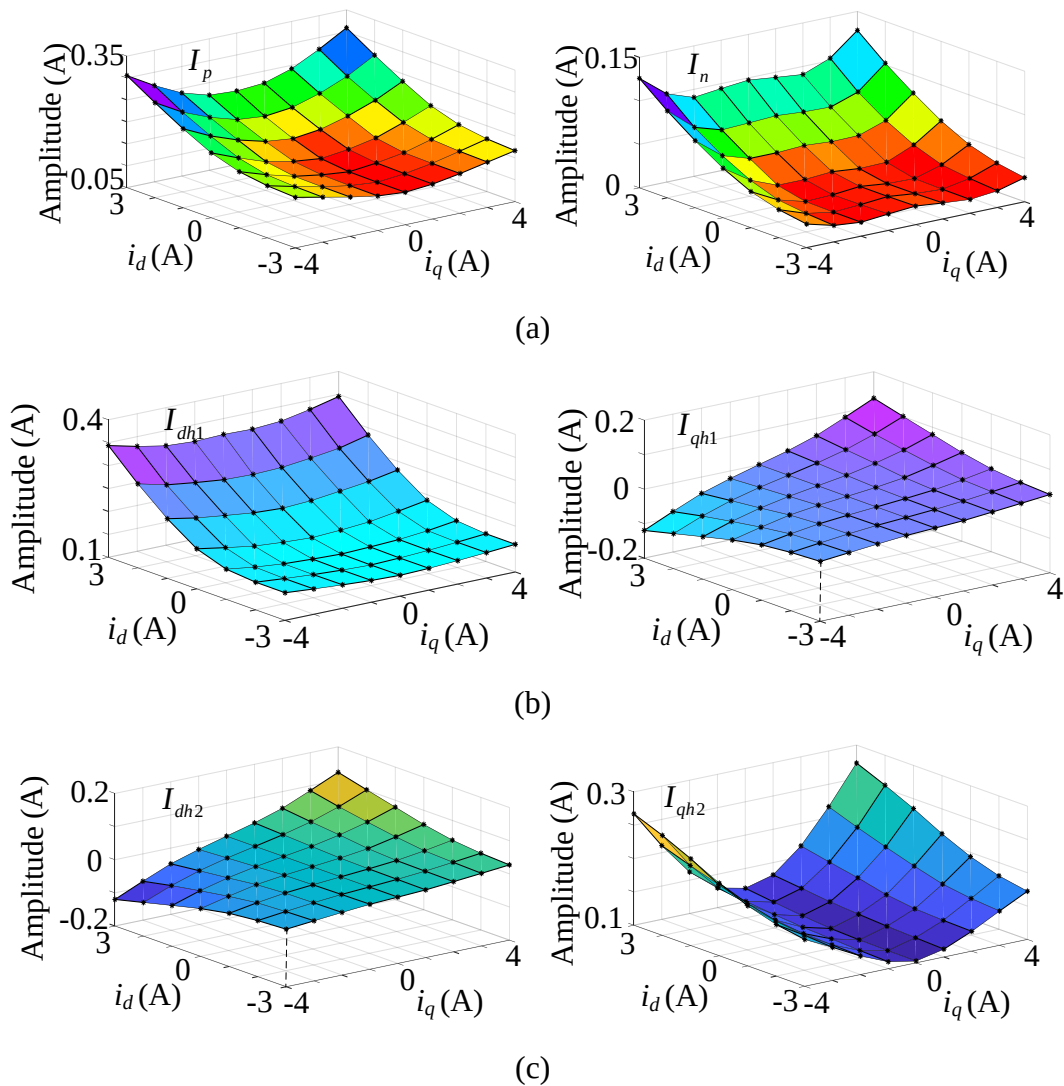


Fig. 3.15. Measured HF current amplitudes for calculating HF inductances with the proposed method at different injection stages. (a) I_p, I_n due to rotating voltage injection. (b) I_{dh1}, I_{qh1} due to d-axis voltage injection. (c) I_{dh2}, I_{qh2} due to q-axis voltage injection.

The HF inductances can be calculated with the measured HF currents using (3.32), (3.33) and (3.34). Finite element analysis is also carried out to obtain the HF inductances. Comparison of

the FEA predicted results and measured results by the proposed method are shown in Fig. 3.16 and included in appendices 2, 3.

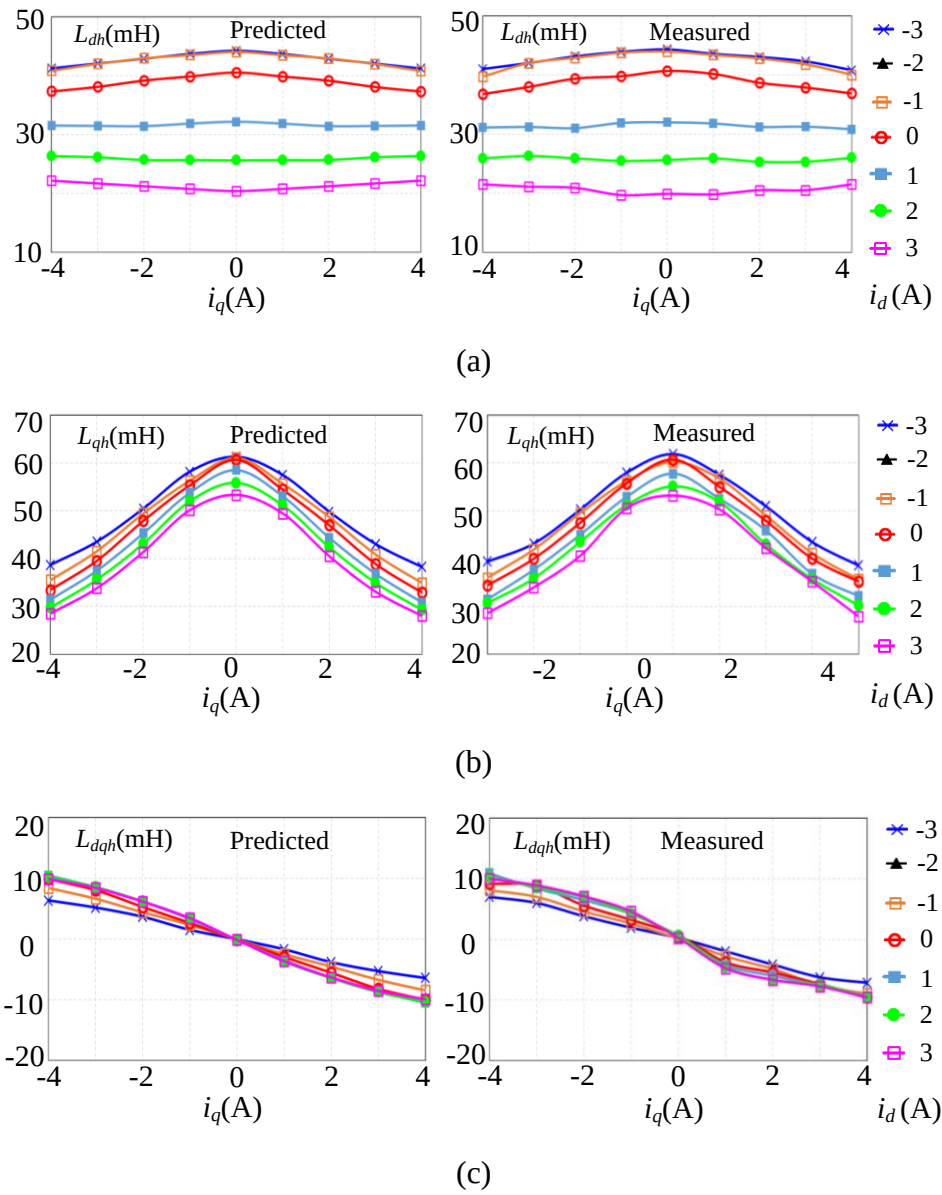


Fig. 3.16. Comparison of FEA predicted and measured HF inductances with the proposed method. (a) L_{dh} . (b) L_{qh} . (c) L_{dqh} .

As can be seen that the measured results show a good agreement with the FEA predicted values. The d-axis HF inductance changes slightly with the variation of q-axis current. The d-axis and q-axis HF inductances are almost symmetrical along the current $i_q=0$ A. It is also noted that the proposed method does not require a closed-loop position estimation and only makes use of the HF current amplitudes.

B. Comparison with Conventional Methods

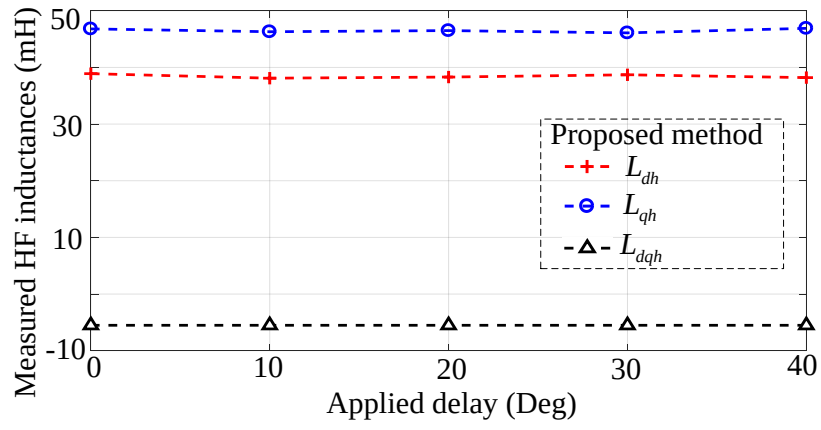
In this section, the proposed method and the conventional estimation methods are both carried out for comparison at the current point $(i_d, i_q) = (0A, 2A)$. The rotor is locked at $\theta_r = 10^\circ$ during the estimation process. To evaluate the delay immune property of the proposed method, a delay φ_{comp} is subtracted from the HF angle θ_h in the demodulation of the HF current amplitudes.

As analyzed in section 3.3.2, the proposed method is immune to the delay effects as the factor relating to the actual delay φ_{delay} is cancelled in the division operation of the HF current amplitudes. To evaluate the delay immune property of the proposed method, a delay φ_{comp} is applied for the proposed method. The applied delay will affect the measured HF current amplitudes $I_{dh1}, I_{qh1}, I_{dh2}, I_{qh2}$ as

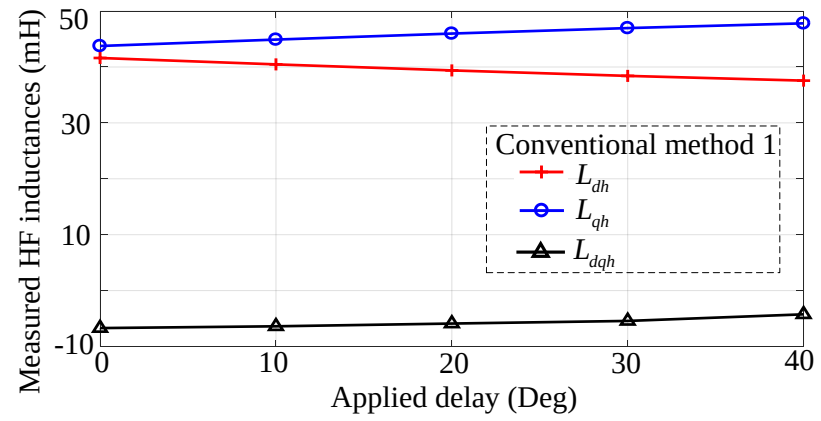
$$LPF \left[2 \sin(\omega_h t + \varphi_{comp}) I \sin(\omega_h t + \varphi_{delay}) \right] = I \cos(\varphi_{delay} - \varphi_{comp}) \quad (3.47)$$

where φ_{delay} is the actual delay in the system. As can be seen from (3.47), the applied phase delay φ_{comp} will result in changes in the measured HF current amplitudes. However, as the proposed method only requires the ratios between the HF current amplitudes $I_{dh1}, I_{qh1}, I_{dh2}, I_{qh2}$ as analyzed in section 3.3.2, the calculated HF inductances will not be affected by the applied delay φ_{comp} no matter whether it is equal or different to the actual delay φ_{delay} in the system. In other words, the proposed method is immune to the delay effects. In Fig. 3.17(a), it can be seen that the measured HF inductances with the proposed method are nearly unchanged with the applied delay in the demodulation signal, and are about 38mH for L_{dh} , 47mH for L_{qh} , -5.5mH for L_{dqh} , which are in accordance with the FEA predicted values.

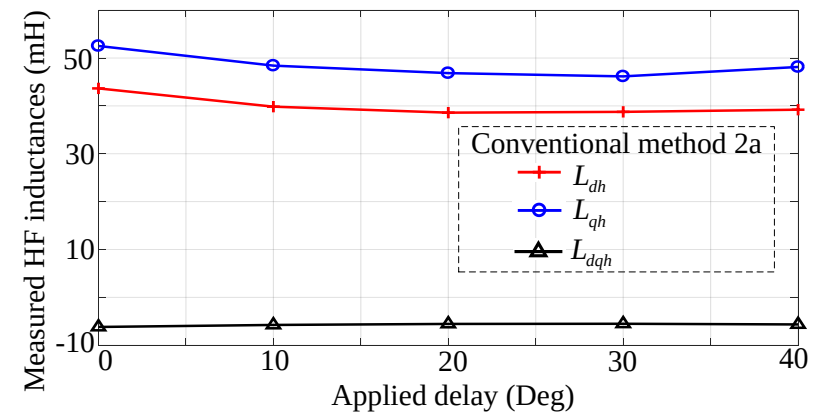
To evaluate the superiority of the proposed method in terms of the delay effects, the conventional measuring method 1 in [XU14] based on rotating voltage injection and the conventional methods 2a in [KUE14], 2b in [LI07a] based on pulsating voltage injections are carried out for comparison. The injection voltages for the conventional methods are all chosen as the same of the proposed method, i.e. 20V/500Hz.



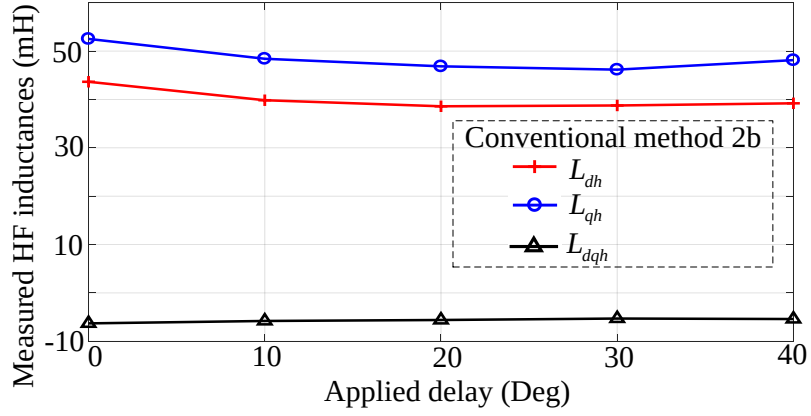
(a)



(b)



(c)



(d)

Fig. 3.17. Comparison of the measured HF inductances with the conventional and proposed methods at $(i_d, i_q) = (0A, 2A)$ with different applied delay. (a) Proposed method. (b) Conventional method 1 with rotating voltage injection. (c) Conventional method 2a with voltage injections in the estimated dq-axes. (d) Conventional method 2b with voltage injections in the actual dq-axes.

For the conventional measuring method 1 with only rotating voltage injection, the delay effects will cause error in the observed angle $\Delta\theta_m$ which will result in errors in the calculated HF inductances. In the experiment, a phase delay φ_{comp} is added to the injection angle θ_h for compensating the delay effects. As described in Fig. 3.6, the actual delay is hard to determine accurately, although it is usually approximated as the delay over 1.5 times of the PWM periods, i.e. $1.5 * 2\pi * 500 * 10^{-4} \text{rad} = 27$ degrees. Therefore, in this test, the applied delay is varied from 0 to 40 degrees to evaluate the dependency of the conventional method 1 on the delay effects.

The measured dq-axis HF inductances with conventional method 1 under different applied delays at $(i_d, i_q) = (0A, 2A)$ are shown in Fig. 3.17(b). When the applied delay is zero, which means the delay effects are not accounted, the measured results ($L_{dh} \approx 42\text{mH}$, $L_{qh} \approx 44\text{mH}$, $L_{dqh} \approx -6.7\text{mH}$) are deviated from the FEA predicted values. When the applied delay changes, the measured results with the conventional method 1 also vary. To obtain HF inductances without errors, accurate applied delay is required. However, as it is related to the injection frequency and the hardware, the delay is difficult to determine accurately. Therefore, compared with the conventional method 1, the proposed method is not affected by the delay effects and produces negligible errors in the measured HF inductances.

For the conventional measuring method 2a in [KUE14] with pulsating voltage injections in the estimated dq-axes, the delay effects will affect the measured HF current amplitudes. For example, an actual delay of 18 degrees will result in 5% decrease, i.e. $(1/\cos(18^\circ) - 1)$, in the measured HF current amplitudes as shown in (3.28), and ultimately 5% increase in the calculated inductances as shown in (3.31). In the experiment, a phase φ_{comp} is added to the injection angle θ_h in the program for compensating the delay effects, i.e. $\sin(\omega_h t + \varphi_{comp})$. The measured results for conventional method 2a with different applied delays are shown in Fig. 3.17(c). As can be seen, the estimated inductances with the conventional method 2a are decreased with the applied delay. This is due to that with the increase of φ_{comp} , $\cos(\varphi_{comp} - \varphi_{delay})$ in (3.44) is approaching unity, which results in reduced estimation errors. If φ_{comp} is larger than φ_{delay} , the estimation errors will be increased due to decrease of the factor $\cos(\varphi_{comp} - \varphi_{delay})$. Similar to the conventional method 1, accurate delay is required for compensating the errors, i.e. $\varphi_{comp} = \varphi_{delay}$. However, it is difficult to determine as it is related to hardware and the injection frequency. Therefore, compared to the proposed method, the conventional method 2a [KUE14] is also sensitive to the delay effects.

The conventional method 2b in [LI07a] with pulsating voltage injections in the actual dq-axes is similar to conventional method 2a in [KUE14] as they both use the dq-axis HF current amplitudes due to pulsating voltage injections to calculate the HF inductances and are affected by the delay effects by introducing a scaling factor $1/\cos(\varphi_{delay})$ to the measured HF inductances. Therefore, the measured results for the conventional method 2b with different applied delays in Fig. 3.16(d), show similar waveforms to the conventional method 2a in Fig. 3.17(c). Compared to the proposed method, the conventional method 2b is also affected by the delay effects which are difficult to compensate accurately.

Therefore, compared to all the conventional methods with either rotating voltage injection or pulsating voltage injections, the proposed method is immune to the delay effect and does not require a position estimator.

3.4 Conclusion

In this chapter, the concept of incremental inductance (called HF inductance in this chapter) and its application in sensorless control are firstly introduced. Secondly, the conventional methods for obtaining the HF inductances are introduced, including FEA simulation and experimental methods by HF voltage injection techniques. Finally, a novel method for

measuring the HF incremental inductances is proposed for interior permanent magnet synchronous machine. It mainly includes three HF voltage injection stages, which are rotating voltage injection in the stationary frame, pulsating voltage injection in the d -axis, and pulsating voltage injection in the q -axis of the synchronous rotating reference frame. Compared with conventional methods based on either rotating or pulsating voltage injection, the proposed method does not require position estimator and is immune to the delay effects which are caused by voltage synthesis and digital filters. Effectiveness of the proposed method has been verified by FEA simulation and experiments and also comparing with the conventional methods.

CHAPTER 4

IMPROVED CROSS-COUPLING EFFECT COMPENSATION METHOD BASED ON HIGH FREQUENCY VOLTAGE INJECTION

4.1 Introduction

Accurate rotor position is essential in high performance permanent magnet synchronous machine (PMSM) drives. For reducing the hardware cost and ensuring system safety, sensorless control as a main source of rotor position information or as a backup in case of encoder failure, has been investigated for decades [WAN20]. At medium and high speeds, back electromotive force (EMF) based sensorless control methods are widely used. However, they suffer from performance deterioration at standstill and low speed due to low amplitude of the back EMF signal [WAN13], [HEJ11]. For sensorless operation at standstill and low speed, magnetic anisotropy, which is caused by machine structure or magnetic saturation, is utilized [XIE16], [GON13], [JAN03]. The interior permanent magnet synchronous machine (IPMSM) exhibits high saliency ratio, whereas surface-mounted permanent magnet synchronous machine (SPMSM) has less or almost negligible structural saliency ratio, which is not easy for position estimation at low speed except utilizing the magnetic saturation introduced saliency.

Magnetic anisotropy can be exploited by injecting high frequency (HF) voltages which are superimposed on the fundamental excitations [TAN17], [ALM17], [HOL08]. The HF sinusoidal rotating voltage injection based sensorless control method is performed in the synchronous stationary reference frame (SSRF) [TAN17], while the HF sinusoidal pulsating voltage injection based sensorless control method is usually carried out in the estimated synchronous rotating reference frame (SRRF) [ALM17]. The induced HF currents which contain the rotor information can be extracted by digital filters [HOL08]. The phase locked loop (PLL) based closed-loop observer is usually adopted for extracting the rotor position [LI20]. Although there are differences in the injection signal types, frames and demodulation algorithms for these methods, they all suffer from magnetic saturation, multi-saliencies and cross-coupling effect which are inherent magnetic properties of the machines.

Magnetic saturation can reduce the saliency level and deteriorates the position estimation accuracy [GAR07]. However, for an IPMSM that has high initial saliency level, it usually does not have major impact. Multi-saliencies may introduce ripples in the estimation results, but can

be neglected if the machine is properly designed and the magnitudes are relatively small [CHE14], [CHE16]. Cross-coupling effect which can be described as the mutual inductance between d-axis and q-axis has been identified as the major source of estimation error [BAR17], [GUG06]. Many researches [MUR10], [REI08], [WAN19b], [LI09] have been carried out on the cross-coupling effect, its impact on sensorless control and the corresponding compensation methods.

For compensating the cross-coupling effect in HF rotating voltage injection based sensorless control, the position error can be compensated by adding the online calculated [REI08], [MOR21] or offline measured error angle [FRE10], [WAN19], [XU14] to the estimated position. Such compensation method is designated as “angle compensation”. In [REI08], the incremental inductances and the error caused by the cross-coupling effect are obtained from the amplitudes of HF currents based on the assumption of fixed d-axis (or q-axis) incremental inductance when the q-axis (or d-axis) incremental inductance is calculated. The assumption of constant d-axis (or q-axis) incremental inductance in [REI08] is only feasible for some specific machines as the incremental inductances usually vary with the dq-axis currents due to magnetic saturation and cross-coupling effect. In [MOR21], the error caused by the cross-coupling effect considering spatial harmonics is online calculated with the measured current derivatives and voltage vectors, but it requires high current sampling accuracy and the impact of magnetic saturation is not considered. In [FRE10], [WAN19], the errors caused by the cross-coupling effect are curve-fitted as functions of the dq-axis currents and the cross-coupling effect can be compensated according to the fundamental currents. For obtaining the coefficients of the fitted-curves, offline measurements are usually required. In [XU14], the errors caused by the cross-coupling effect are offline measured by the difference between the actual rotor position measured by an encoder and the estimated position obtained by a closed-loop position observer, and stored in a lookup table (LUT). The offline measured errors can be used for cross-coupling effect compensation irrespective of the HF injection signal that is used for sensorless position estimation as only the fundamental currents are required for looking up the LUT. In this thesis, the first contribution is that to reduce the offline measuring complexity, a simple method for offline measuring the cross-coupling induced errors is proposed by injecting HF voltages in d-axis and q-axis, respectively.

For HF pulsating voltage injection based sensorless control, the method in [YOU18] compensates the cross-coupling effect by demodulating the estimated q-axis high frequency flux linkage based on the offline measured current-flux linkage characteristics. However, the

process for offline measuring the current-flux linkage characteristics is usually very complicated. In [WAN20], the error caused by the cross-coupling effect is online identified and compensated by simply varying the direction of the current vector for a given torque reference. However, the implementation in [WAN20] is only applicable for machines with negligible reluctance torque. In [LI09], a compensation method is proposed by adding a current which is the product of the cross-coupling factor and d-axis HF current amplitude to the input of the position observer. This method is designated as “current compensation” [LI09]. The cross-coupling factor which is the ratio between the mutual and q-axis HF inductances can be offline measured by injecting HF voltage to the d-axis and measuring the corresponding dq-axis HF currents for different working points. The method in [LI09] using cross-coupling factor is restricted only to the HF pulsating voltage injection based sensorless control. In this thesis, the second contribution is that a method is proposed to extend the offline measured cross-coupling factor for cross-coupling effect compensation in HF rotating voltage injection based sensorless control by angle compensation.

Both angle compensation and current compensation can be used for cross-coupling effect compensation. However, seldom comparison has been made between the two methods in terms of the dynamic performance. In this thesis, the dynamic performances of the two cross-coupling compensation methods using cross-coupling factor will be compared. The third contribution of this thesis, as will be confirmed by experiments, is that the better dynamic performance of angle compensation can be achieved and explained by the fact that angle compensation is not dependent on the proportional-integral (PI) and integral modules of the position observer which may introduce delay to the estimation for current compensation.

For sensorless operation at highly saturated conditions, position estimation with HF voltage injection may diverge due to absence of convergence point [MAN20], [KWO19], [LEE20]. In [MAN20], it is shown that the angle compensation could be insufficient to make robust sensorless drive at highly saturated machines due to saturation and spatial harmonics, and the current compensation should also be implemented. In [KWO19], sensorless operation is extended to high torque by tilting the HF voltage vector and compensating the observer input. In [LEE20], the current reference is tilted so that the controller sees more desirable side of the inductance and the sensorless operation region can be extended. In this thesis, stability and convergence of sensorless operation with the proposed cross-coupling effect compensation methods at highly saturated conditions are also analyzed and experimentally evaluated.

This chapter is organized as follows. Firstly, in section 4.2, the impact of cross-coupling effect on sensorless control and the existing compensation methods are presented, together with their advantages and disadvantages. Then, in section 4.3, an improved compensation method based on high frequency voltage injection is proposed by offline measuring the cross-coupling induced error with a simple method. Finally, experiments are carried out for verification and comprehensive comparison.

4.2 Existing Cross-coupling Effect Compensation Methods

In this section, the cross-coupling effect and its impact on HF voltage injection based sensorless position estimation are introduced. Then, several methods for compensating the cross-coupling induced position errors are presented.

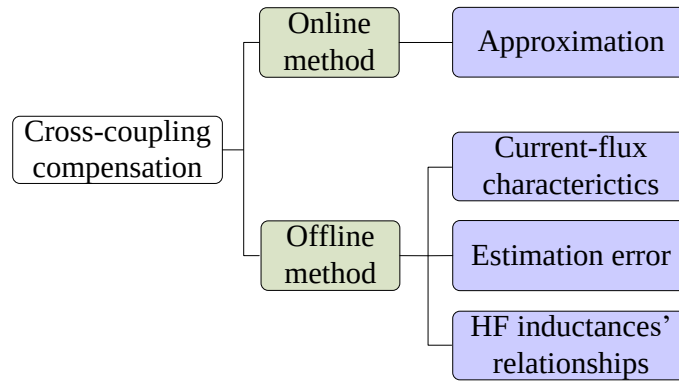
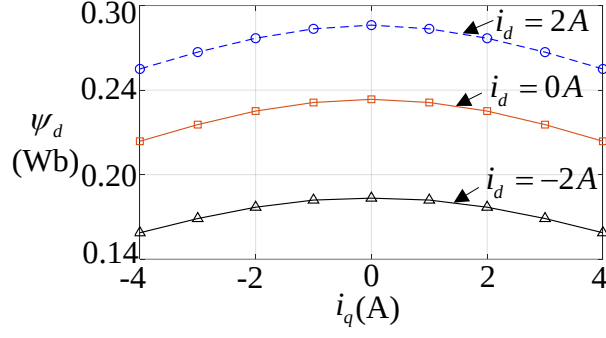


Fig. 4.1. Category of cross-coupling effect compensation methods.

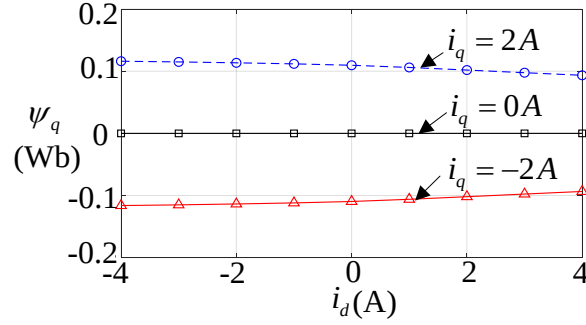
As can be seen in Fig. 4.1, the compensation methods can be divided as online and offline methods. The online methods obtain the cross-coupling information from the online measured HF currents or voltages without any pre-measured-knowledge of the machine characteristics. The offline methods use the offline measured machine data to online compensate the cross-coupling effect. In the next section, the impact of the cross-coupling effect on sensorless control will firstly be introduced.

4.2.1 Impact of cross-coupling effect on sensorless control

The cross-coupling phenomenon in electrical machines has been subjected to detailed scrutiny in the last two decades [LEV99] [LEV00] [BOJ03]. Fig. 4.2 shows the dq -axis flux linkage at different dq -axis currents for the prototype machine in Appendix 1.



(a)



(b)

Fig. 4.2. Illustration of the cross-coupling effect between dq -axes. (a) $\psi_d - i_q$. (b) $\psi_q - i_d$.

It can be seen that the saturation level of d -axis (or q -axis) has impact on the q -axis (or d -axis) flux linkage. Such effect becomes significant with the increase of currents. The circuit model accounting for the cross-coupling effect is given by

$$\begin{bmatrix} u_d^r \\ u_q^r \end{bmatrix} = \begin{bmatrix} R_s + pL_d & pL_{dq} - \omega_e L_q \\ pL_{qd} + \omega_e L_d & R_s + pL_q \end{bmatrix} \begin{bmatrix} i_d^r \\ i_q^r \end{bmatrix} + \omega_e \begin{bmatrix} 0 \\ \psi_f \end{bmatrix} \quad (4.1)$$

where “ p ” is the differential operator, L_{dq} , L_{qd} are the cross-coupling inductances which are equal according to the reciprocal theory [MIN17]. The impacts of cross-coupling effect on sensorless position estimation at medium and high speeds have been intensively investigated [LEE17], [LI07b]. This chapter mainly focuses on HF voltage injection based sensorless control at standstill and low speed. Hence, the speed-dependent term is neglected and the voltage model in (4.1) is simplified as

$$\begin{bmatrix} i_{\alpha h}^s \\ i_{\beta h}^s \end{bmatrix} = R^{-1}(\theta_r) \begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{qh} \end{bmatrix}^{-1} R(\theta_r) \int \begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} dt \quad (4.2)$$

$$R(\theta_r) = \begin{bmatrix} \cos(\theta_r) & \sin(\theta_r) \\ -\sin(\theta_r) & \cos(\theta_r) \end{bmatrix} \quad (4.3)$$

where θ_r is the rotor position, $u_{\alpha h}^s, u_{\beta h}^s, i_{\alpha h}^s, i_{\beta h}^s$ are the HF voltages and currents in the stationary frame, $L_{dh}, L_{qh}, L_{dqh}, L_{qdh}$ are the incremental inductances in the synchronous rotating reference frame (SRRF), which are also referred to as HF inductances. The mutual inductances L_{dqh} and L_{qdh} are equal according to the reciprocity theorem [MIN17]. The superscript “s” means the stationary reference frame. The stator resistance and iron loss reflected resistance in the HF model are neglected due to relatively small magnitude compared to the HF reactance. The general HF voltage injection based sensorless methods and the impacts of the cross-coupling effect will be analysed as follows.

(a) HF rotating voltage injection based sensorless control

For HF rotating voltage injection based sensorless control method, sinusoidal voltages $u_{\alpha h}^s, u_{\beta h}^s$ are injected to the stationary reference frame as

$$u_{\alpha h}^s + ju_{\beta h}^s = V_h e^{j\omega_h t}, \theta_h = \omega_h t \quad (4.4)$$

where V_h, ω_h, θ_h are the amplitude, frequency, angle of the injection signal. According to (4.2), the HF currents $i_{\alpha\beta h}^s$ can be solved as

$$i_{\alpha h}^s + ji_{\beta h}^s = I_p e^{j(\theta_h - \pi/2)} + I_n e^{-j[\theta_h - 2(\theta_r - \Delta\theta_m) - \pi/2]} \quad (4.5-a)$$

$$\begin{aligned} i_{\alpha h}^s &= I_p \sin(\theta_h) + I_n \sin[\theta_h - 2(\theta_r - \Delta\theta_m)] \\ i_{\beta h}^s &= -I_p \cos(\theta_h) + I_n \cos[\theta_h - 2(\theta_r - \Delta\theta_m)] \end{aligned} \quad (4.5-b)$$

$$I_p = \frac{V_h L_{\Sigma}}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)}, I_n = \frac{V_h \sqrt{L_{\Delta}^2 + L_{dqh}^2}}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)} \quad (4.6)$$

$$\Delta\theta_m = 0.5 \tan^{-1} \left(\frac{2L_{dqh}}{L_{qh} - L_{dh}} \right) \quad (4.7)$$

where $L_{\Sigma} = \frac{L_{dh} + L_{qh}}{2}$, $L_{\Delta} = \frac{L_{qh} - L_{dh}}{2}$. The HF currents are composed of positive and negative sequence currents and their amplitudes I_p, I_n are determined by the injection voltage and HF inductances as shown in (4.6). The angle $\Delta\theta_m$ is caused by the cross-coupling effect. As L_{dh}, L_{qh} and L_{dqh} vary with the magnetic saturation levels, $i_{\alpha h}^s, i_{\beta h}^s$ will show different shapes

under different i_d, i_q as shown in Fig. 4.3. The rotor position is fixed at 30° and the injection voltage is chosen as 15V/500Hz, i.e. $V_h = 15, \omega_h = 2\pi * 500$.

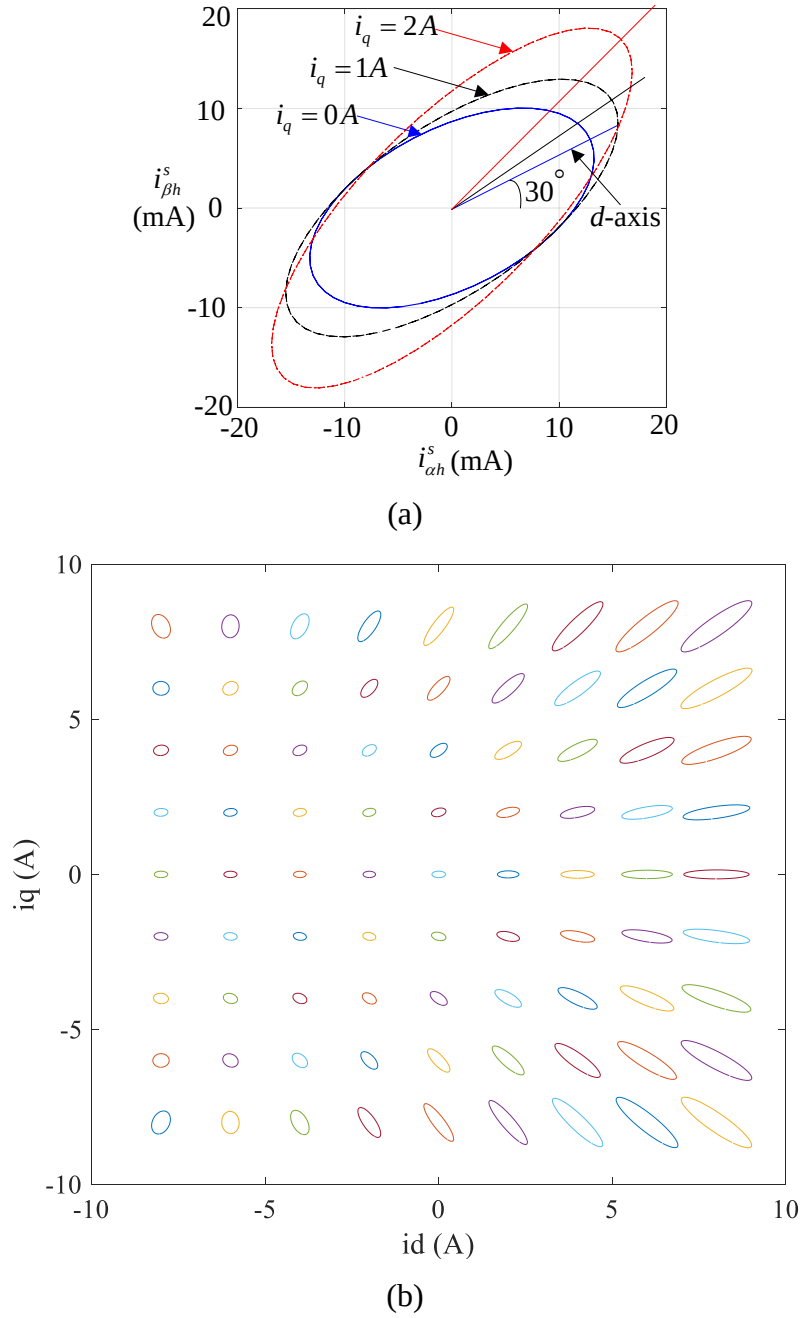


Fig. 4.3. Traces of HF currents due to HF rotating voltage injection in the stationary frame at different currents when $\theta_r = 30^\circ$. (a) Fixed $i_d = 0A$. (b) Traces at different i_d, i_q .

As can be seen from Fig. 4.3(a), the traces of the HF currents are ellipses. When i_d, i_q are zero, the long axis of the ellipse is aligned with the d -axis. With the increase of i_q , the long axis gradually deviates from the d -axis, which is caused by the cross-coupling effect. This deviation will result in position estimation error in HF voltage injection based sensorless control as follows. The shapes of the HF currents vary with the incremental inductances and

the fundamental currents as shown in Fig. 4.3(b). Generally, the offset increases with the q-axis current.

It can be seen from (4.5-b) that the length of the long axis is $(I_p + I_n)$, and the length of the short axis is $(I_p - I_n)$. The values of I_p, I_n can be calculated from the injection voltage and the FE predicted HF inductances in Fig. 3.3. The difference between the long axis and the short axis, i.e. $2I_n$, indicates the level of the saliency signal. This means that the area of the ellipse does not represent the capability of sensorless control. If the shape of the HF current trace becomes circular, the saliency level disappears. This corresponds to the situation of $L_{dh} = L_{qh}$ and $L_{dqh} = 0mH$, which is also called the critical point. It can be seen from Fig. 4.3(b), the ellipse gradually shrinks to a circle with the increase of load current and finally becomes a circle at the point $(i_d, i_q) = (-5A, 7A)$. Fortunately, as verified in [KWO19], the sensorless position observer can achieve stable convergence at the critical point in the dynamic process.

Generally, in HF rotating voltage injection based sensorless control, a PLL observer incorporating heterodyning demodulation [JAN95] is used as shown in Fig. 4.4. HF currents $i_{\alpha h}^s, i_{\beta h}^s$ superimposed on the fundamental currents are extracted by band pass filters (BPFs).

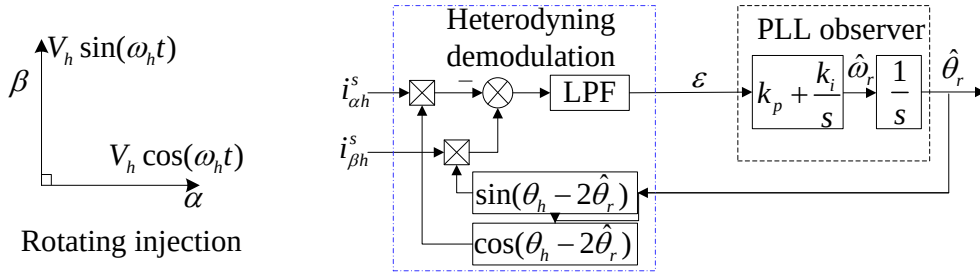


Fig. 4.4. Position demodulation for HF rotating voltage injection based sensorless control.

The error signal ε in Fig. 4.4 is

$$\varepsilon = I_n \sin(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \quad (4.8)$$

where $\hat{\theta}_r$ is the estimated rotor position, which can be obtained by controlling $\varepsilon = 0$ with the PLL observer, i.e.

$$\hat{\theta}_r = \theta_r - \Delta\theta_m \quad (4.9)$$

As can be seen, there is a position estimation error $\Delta\theta_m$ in HF rotating voltage injection based sensorless control caused by the cross-coupling effect.

(b) HF pulsating voltage injection based sensorless control

For HF pulsating voltage injection based sensorless control method, HF model used in the estimated synchronous rotating reference frame (SRRF) is given by

$$\begin{bmatrix} \hat{i}_{dh}^r \\ \hat{i}_{qh}^r \end{bmatrix} = R^{-1}(\theta_r - \hat{\theta}_r) \begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{qh} \end{bmatrix}^{-1} R(\theta_r - \hat{\theta}_r) \int \begin{bmatrix} \hat{u}_{dh}^r \\ \hat{u}_{qh}^r \end{bmatrix} dt \quad (4.10)$$

where $\hat{u}_{dh}^r, \hat{u}_{qh}^r, \hat{i}_{dh}^r, \hat{i}_{qh}^r$ are the HF voltages and currents, the superscript “ $\hat{\cdot}$ ” means the estimated synchronous rotating reference frame. When HF pulsating voltage $\hat{u}_{dh}^r = V_h \cos(\omega_h t)$ is injected to the estimated d -axis, the estimated dq -axis HF currents can be expressed from (4.10) as

$$\hat{i}_{dh}^r = I_{dh}^r \sin(\omega_h t), \hat{i}_{qh}^r = I_{qh}^r \sin(\omega_h t) \quad (4.11)$$

where I_{dh}^r, I_{qh}^r are the amplitudes of the HF currents and can be extracted by synchronous demodulation [JAN03] as

$$I_{dh}^r = LPF[2\hat{i}_{dh}^r \sin(\omega_h t)] = I_p + I_n \cos(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \quad (4.12-a)$$

$$I_{qh}^r = LPF[2\hat{i}_{qh}^r \sin(\omega_h t)] = I_n \sin(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \quad (4.12-b)$$

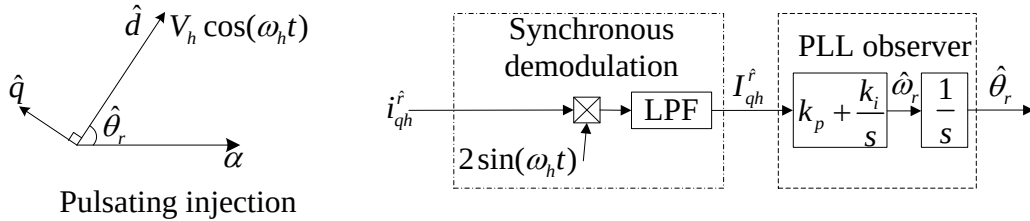


Fig. 4.5. Position demodulation for HF pulsating voltage injection based sensorless control.

The estimated position $\hat{\theta}_r$ can be obtained by controlling $I_{qh}^r = 0$, as shown in Fig. 4.5, i.e.

$$\hat{\theta}_r = \theta_r - \Delta\theta_m \quad (4.13)$$

which is the same as (4.9) for HF rotating voltage injection based sensorless control.

As can be seen from Figs. 4.4 and 4.5, due to existence of the cross-coupling effect, there is a position estimation error $\Delta\theta_m$ for both HF rotating and pulsating voltage injection based sensorless control methods. It has been verified that the error is inherent to the machine and not dependent on specific injection signals [GUG06]. If such error is not compensated, it will result in significant performance degradation especially at heavy load conditions. The maximum available torque may be limited due to incorrect rotor position estimation. For

example, a position error of 45° will result in approximately 30% ($1-\cos(45^\circ)$) decrease of the torque. The position estimation error also has adverse impact on other advanced machine control techniques, such as MTPA, flux weakening control. Thus, for high performance sensorless control, the cross-coupling effect should be compensated.

For compensating the cross-coupling effect, a preferable solution is to design a machine with negligible cross-coupling effect [MUR12]. However, this is not always feasible as the machine design is confined by many other design objectives, such as efficiency, cost and manufacturability. Instead, software based compensation methods are widely used.

Solo online compensation is quite challenging due to the nonlinear properties of the cross-coupling effect. For reducing the computation complexity, some approximations are made for online compensation, e.g. fixed d -axis HF inductance [MUR10]. Some other methods compensate the cross-coupling effect compensation by using offline measured data of the machine. The offline data can be position errors [FRE10], [WAN19b], current-flux linkage characteristics [YOU18], or HF inductances' relationships [LI09], [LI19]. These methods differ in offline measuring procedures, universality and compensation strategies, which will be presented in detail.

4.2.2 By online calculating the HF inductances

The online compensation method calculates the HF inductances and the error caused by the cross-coupling effect from the HF currents and voltages. For HF sinusoidal rotating voltage injection based sensorless control, the amplitudes of the positive sequence current (PSC) and negative sequence current (NSC) are used to calculate the HF inductances as

$$I_p = \frac{V_h L_\Sigma}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)}, I_n = \frac{V_h \sqrt{L_\Delta^2 + L_{dqh}^2}}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)} \quad (4.14)$$

where $L_\Sigma = \frac{L_{dh} + L_{qh}}{2}$, $L_\Delta = \frac{L_{qh} - L_{dh}}{2}$. However, I_p, I_n only provide two equations for the three knowns L_{dh}, L_{qh}, L_{dqh} . Another condition is required for solving the three HF inductances. Below are two methods for calculating the HF inductances by assuming approximations.

① In [MUR10], the d -axis HF inductance L_{dh} is assumed to be fixed as i_d is controlled to remain constant. The value of L_{dh} can be solved from I_{p0}, I_{n0} which are the amplitudes of PSC and NSC at zero load current. When the load current is zero, the magnetic material is not saturated and $L_{dqh} = 0$, i.e. (4.14) can be expressed as

$$I_{p0} = \frac{V_h (L_{dh0} + L_{qh0})}{2\omega_h L_{dh0} L_{qh0}}, I_{n0} = \frac{V_h (L_{qh0} - L_{dh0})}{2\omega_h L_{dh0} L_{qh0}} \quad (4.15)$$

From (4.15), the initial dq -axis HF inductances can be solved as

$$L_{dh0} = \frac{V_h}{\omega_h (I_{p0} + I_{n0})}, L_{qh0} = \frac{V_h}{\omega_h (I_{p0} - I_{n0})} \quad (4.16)$$

According to the assumption in [MUR10], L_{dh} is a constant and not changed with load current. At other i_d, i_q , the HF inductances L_{qh}, L_{dqh} can be solved from (4.14), (4.16) as

$$L_{qh} = \frac{2V_h I_p}{\omega_h (I_p^2 - I_n^2)} - L_{dh0}, \quad (4.17)$$

$$L_{dqh} = \pm \sqrt{L_{dh0} \left(\frac{2V_h I_p}{\omega_h (I_p^2 - I_n^2)} - L_{dh0} \right) - \frac{V_h^2}{\omega_h^2 (I_p^2 - I_n^2)}}$$

Then, the position error caused by the cross-coupling effect can be calculated as

$$\Delta\theta_m = 0.5 \tan^{-1} \left(\frac{2L_{dqh}}{L_{qh} - L_{dh0}} \right) \quad (4.18)$$

This method does not take the variation of L_{dh} into consideration. Its value changes with i_d, i_q due to saturation and cross-coupling. Besides, in [MUR10], the method for determining the sign of L_{dqh} is not given.

② In [REI08], a different method for calculating the HF inductances was proposed by making use of the positive sequence current as

$$L_{dh} = -\frac{V_h L_{qh0}}{V_h - 2I_p \omega_h L_{qh0}}, L_{qh} = -\frac{V_h L_{dh0}}{V_h - 2I_p \omega_h L_{dh0}} \quad (4.19)$$

where L_{dh0}, L_{qh0} are the HF inductances for the no load condition which can be determined before start with (4.16). After L_{dh}, L_{qh} are obtained, L_{dqh} can be calculated from I_p . The calculations in (4.19) are based on the assumption that when L_{dh} (or L_{qh}) is calculated, L_{qh} (or L_{dh}) is assumed to be L_{qh0} (or L_{dh0}). This assumption does not take the variations of dq -axis HF inductances into consideration.

As can be seen, the existing online compensation methods are only used in HF rotating voltage injection based sensorless control and require approximations which are only applicable for specific machines. To enable cross-coupling effect compensation for different

machines, other methods such as offline method can be used as presented in the next section.

4.2.3 By offline measured current-flux characteristics

A cross-coupling effect compensation method by making use of the offline measured current-flux linkage characteristics was proposed in [YOU18] based on HF pulsating voltage injection in the estimated d -axis.

When a HF pulsating voltage $u_c \cos(\omega_h t)$ is injected to the estimated d -axis, the HF currents can be obtained from (4.12) as

$$\begin{bmatrix} \hat{i}_{dh} \\ \hat{i}_{qh} \end{bmatrix} = k_h \begin{bmatrix} L_\Sigma + L_\Delta \cos(2\tilde{\theta}_r) + L_{dqh} \cos(2\tilde{\theta}_r) \\ L_\Delta \sin(2\tilde{\theta}_r) - L_{dqh} \cos(2\tilde{\theta}_r) \end{bmatrix} \sin(\omega_h t) \quad (4.20)$$

where $k_h = \frac{u_c}{\omega_h(L_{dh}L_{qh} - L_{dqh}^2)}$, $L_\Sigma = \frac{L_{dh} + L_{qh}}{2}$, $L_\Delta = \frac{L_{qh} - L_{dh}}{2}$. The dq -axis flux linkages $\hat{\psi}_{di}, \hat{\psi}_{qi}$ can be obtained from $\hat{i}_d, \hat{i}_q$ by looking up the offline measured current-flux linkage characteristics and the HF flux linkages $\hat{\psi}_{dhi}, \hat{\psi}_{qhi}$ can be extracted by band pass filters from $\hat{\psi}_{di}, \hat{\psi}_{qi}$ as

$$BPF \left(\begin{bmatrix} \hat{\psi}_{di} \\ \hat{\psi}_{qi} \end{bmatrix} \right) = \begin{bmatrix} \hat{\psi}_{dhi} \\ \hat{\psi}_{qhi} \end{bmatrix} = \begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{dh} \end{bmatrix} \begin{bmatrix} \hat{i}_{dh} \\ \hat{i}_{qh} \end{bmatrix} \quad (4.21)$$

where the subscript 'i' means the flux linkages are obtained using current-flux characteristics.

Replacing $\hat{i}_{dh}, \hat{i}_{qh}$ in (4.21) with (4.20), $\hat{\psi}_{dhi}, \hat{\psi}_{qhi}$ are expressed as

$$\begin{bmatrix} \hat{\psi}_{dhi} \\ \hat{\psi}_{qhi} \end{bmatrix} = k_h \begin{bmatrix} L_{dh}L_\Delta + L_{dh}L_\Delta \cos(2\tilde{\theta}_r) + L_{dqh}L_\Sigma \sin(2\tilde{\theta}_r) \\ L_{dqh}L_\Sigma (1 - \cos(2\tilde{\theta}_r)) + (L_{qh}L_\Delta - L_{dqh}^2) \sin(2\tilde{\theta}_r) \end{bmatrix} \sin(\omega_h t) \quad (4.22)$$

When the position error $\tilde{\theta}_r$ is small, i.e. $(1 - \cos(2\tilde{\theta}_r)) \approx 0$, the q -axis HF flux linkage $\hat{\psi}_{qhi}$ can be approximated as

$$\hat{\psi}_{qhi} \approx k_h (L_{qh}L_\Delta - L_{dqh}^2) \sin(2\tilde{\theta}_r) \sin(\omega_h t) \quad (4.23)$$

By controlling the amplitude of $\hat{\psi}_{qhi}$ to zero with the observer shown in Fig. 4.6, i.e. $k_h(L_{qh}L_\Delta - L_{dqh}^2) \sin(2\tilde{\theta}_r) = 0$, the position can be estimated.

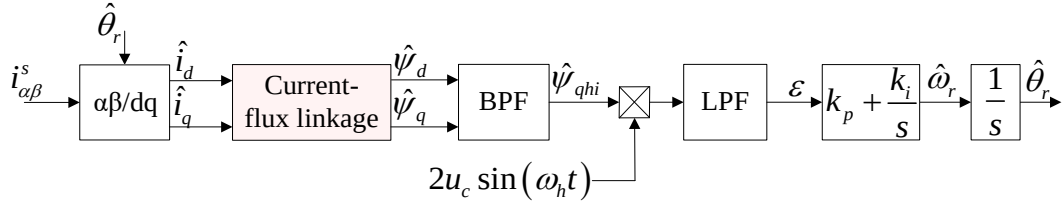


Fig. 4.6. Cross-coupling effect compensation with offline measured current-flux linkage characteristics.

The estimated position finally converges at

$$\hat{\theta}_r = \theta_r \quad (4.24)$$

As can be seen, the error caused by the cross-coupling effect can be mitigated. However, this method is only applicable for HF pulsating voltage injection and the measuring of current-flux linkage characteristics is relatively complicated [HIN17], [ARM13]. In the next section, a simple compensation strategy which can be applied to different HF voltage injection based sensorless control methods, will be introduced.

4.2.4 By offline measured errors

The errors caused by cross-coupling effect at different working points can be offline measured and used for online compensation. For HF rotating and pulsating voltage injection based sensorless control, the compensation methods are shown in Fig. 4.7.

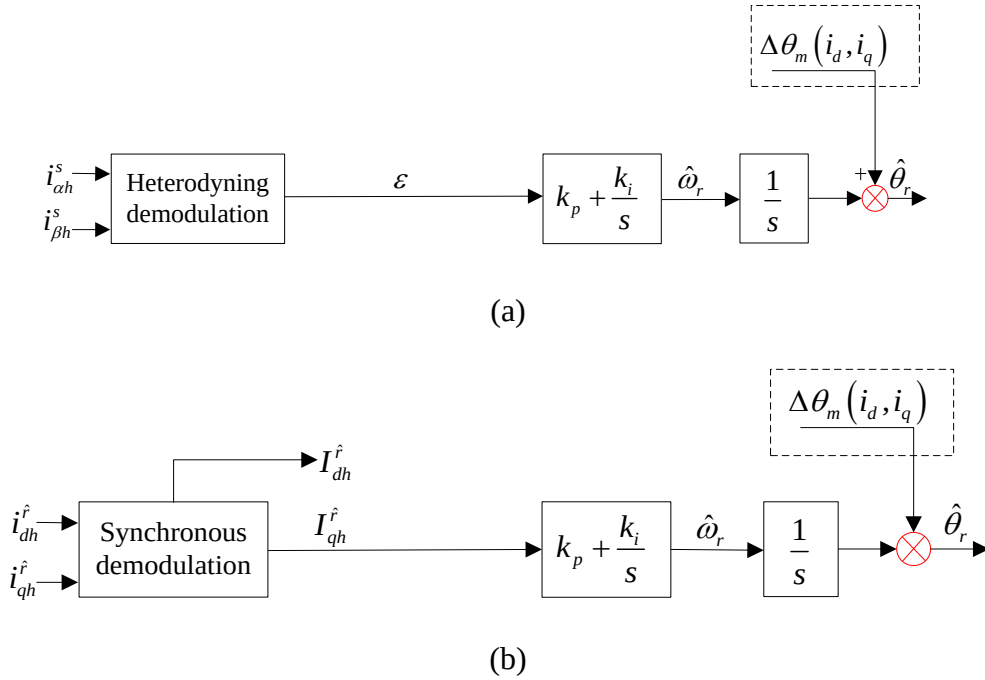


Fig. 4.7. Cross-coupling effect compensation with offline measured error. (a) Rotating voltage injection. (b) Pulsating voltage injection.

where $\Delta\theta_m(i_d, i_q)$ is the offline measured error caused by the cross-coupling effect. The errors shown in Fig. 4.8 can be obtained from the finite element analysis with the method introduced in chapter 3.

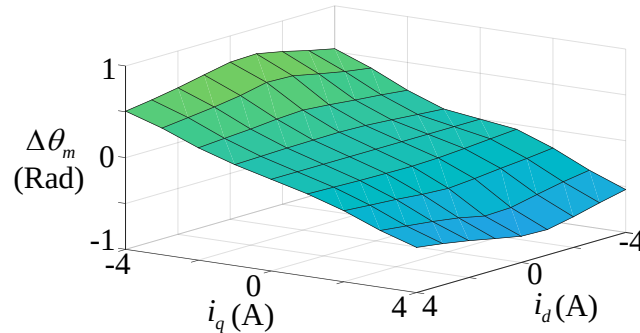


Fig. 4.8. FEA predicted errors caused by cross-coupling effect.

However, FEA is not always available as the machine design data are usually not disclosed by the manufacturers. Thus, experimental method is still required. Generally, the errors can be measured with sensorless position estimation technique as introduced in chapter 3. For example, the estimated rotor position can be estimated by injecting pulsating voltage in the estimated d -axis [KUE14]. The error caused by the cross-coupling effect can be obtained from the difference between the actual rotor position and the estimated rotor position. The measured errors can be stored in the memory of the microprocessor. In sensorless operation, the position error caused by the cross-coupling effect is online corrected by adding the error as

$$\hat{\theta}_r = \hat{\theta}_{PLLoutput} + \Delta\theta_m(\hat{i}_d, \hat{i}_q) \quad (4.25)$$

As can be seen, this compensation method is not dependent on specific injection signal and can be applied to different HF injection based sensorless methods. However, for offline measuring the error, a position observer is required to estimate the rotor position which increases the offline measuring complexity.

For reducing the memory requirement for storing the offline measured errors, the error can be curve-fitted with i_d, i_q . In [WAN19b], the error is approximately linear function of the q -axis current as

$$\Delta\theta_m = k_c i_q \quad (4.26)$$

where k_c can be fitted by least square method. In [FRE10], the error is approximated by using a function that contains a term proportional to i_q and a term proportional to the product of i_d, i_q as

$$\Delta\theta_m = k_1 i_q + k_2 i_d i_q \quad (4.27)$$

4.2.5 By offline measured ratio between L_{dqh} and L_{qh}

Instead of directly using the offline measured errors for compensation, in this section, the offline measured ratios between the HF inductances are used for cross-coupling effect compensation in HF pulsating voltage injection based sensorless control as follows.

For pulsating voltage injection based sensorless control, the pulsating voltage as (4.28) is injected to the estimated d -axis.

$$\hat{u}_{dh} = V_h \cos(\omega_h t) \quad (4.28)$$

The resulting HF currents in the estimated synchronous rotating reference frame are amplitude modulation waveforms and their amplitudes can be obtained as

$$\begin{bmatrix} \hat{I}_{dh} \\ \hat{I}_{qh} \end{bmatrix} = \frac{V_h}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)} \begin{bmatrix} L_{\Sigma} + L_{\Delta} \cos(2\tilde{\theta}_r) + L_{dqh} \cos(2\tilde{\theta}_r) \\ L_{\Delta} \sin(2\tilde{\theta}_r) - L_{dqh} \cos(2\tilde{\theta}_r) \end{bmatrix} \quad (4.29)$$

where $\tilde{\theta}_r = \theta_r - \hat{\theta}_r$ is the different between the actual rotor and the estimated position. The rotor position can be estimated by controlling $\hat{I}_{qh} = 0$ with a closed-loop position observer. Due to the cross-coupling effect, there is a position error $\Delta\theta_m$.

In [LI09], a novel cross-coupling effect compensation method was proposed by making use of the dq -axis HF current amplitudes. The observer input error signal ε is obtained by linear combination of $\hat{I}_{dh}$ and $\hat{I}_{qh}$ as shown in Fig. 4.9, which is different from conventional method.

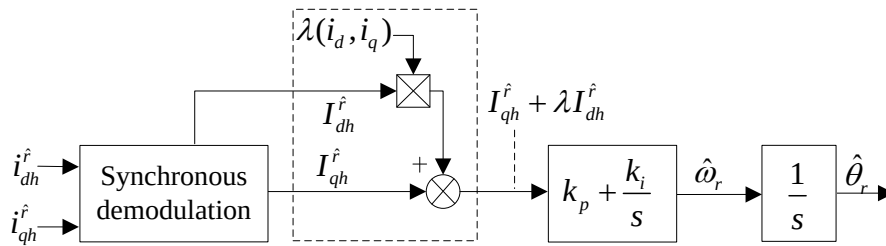


Fig. 4.9. Cross-coupling effect compensation with offline measured $\lambda = \frac{L_{dqh}}{L_{qh}}$ in [LI09b].

where ε, λ are defined as

$$\varepsilon = \hat{I}_{qh} + \lambda \hat{I}_{dh} \quad (4.30)$$

$$\lambda = \frac{L_{dqh}}{L_{qh}} \quad (4.31)$$

By substituting $\hat{I}_{dh}$, $\hat{I}_{qh}$ and λ in (4.30) with (4.29), the modified error signal ε is derived as

$$\varepsilon = \hat{I}_{qh} + \lambda \hat{I}_{dh} = \left(L_{\Delta} + \frac{L_{dqh}^2}{L_{qh}} \right) \sin(2\tilde{\theta}_r) + \frac{2L_{\Sigma}L_{dqh}}{L_{qh}} \sin^2(\tilde{\theta}_r) \quad (4.32)$$

When the position error $\tilde{\theta}_r$ is sufficiently small, the error signal ε can be simplified as

$$\varepsilon = k_{\Delta} \sin(2\tilde{\theta}_r), k_{\Delta} = \frac{V_h (L_{\Delta} + L_{dqh}^2/L_{qh})}{\omega_h (L_{dh}L_{qh} - L_{dqh}^2)} \quad (4.33)$$

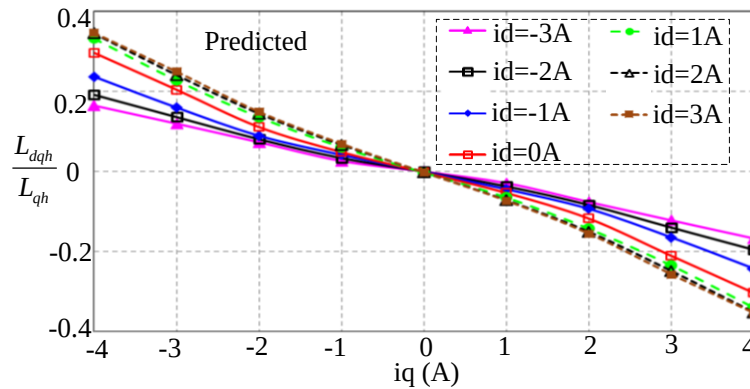
As can be seen that the cross-coupling effect induced error $\Delta\theta_m$ is not included in the error signal. Thus, by controlling $\varepsilon=0$, the cross-coupling effect can be compensated without the need of calculating the accurate error angle $\Delta\theta_m$.

The coefficient λ can be predicted by FEA or offline measurement. In [LI09], a simple measuring method by injecting HF voltage in the actual d -axis is proposed. The value of λ is obtained from the ratio between i_{dh} , i_{qh} as

$$i_{dh} = \frac{V_h L_{qh} \sin(\omega_h t)}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)}, i_{qh} = \frac{-V_h L_{dqh} \sin(\omega_h t)}{\omega_h (L_{dh} L_{qh} - L_{dqh}^2)} \quad (4.34)$$

$$\lambda = -\frac{i_{qh}}{i_{dh}} = \frac{L_{dqh}}{L_{qh}} \quad (4.35)$$

The measured results for the prototype machine in Appendix 1 are shown in Fig. 4.10(b) which show a good agreement with the FEA predicted results in Fig. 4.10(a).



(a)

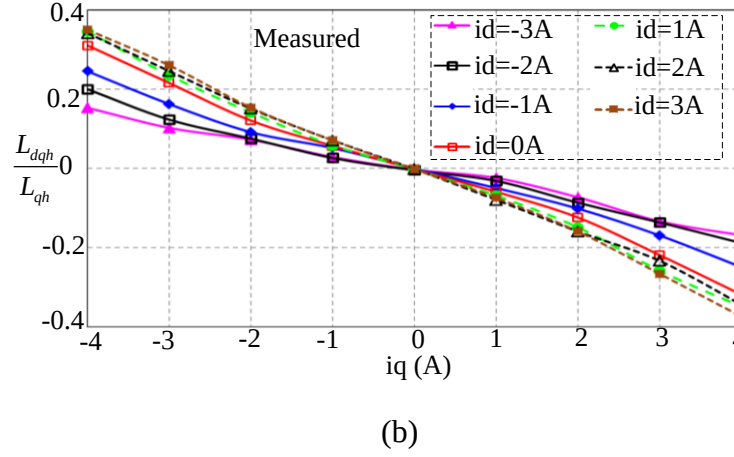


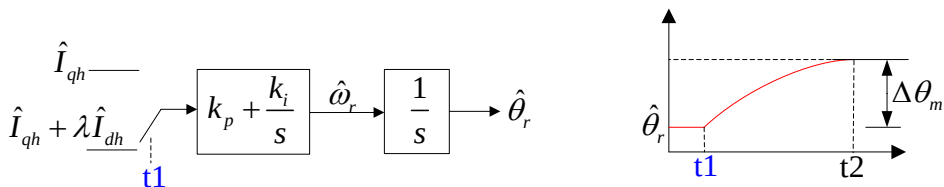
Fig. 4.10. Comparison of measured and FEA predicted L_{dqh}/L_{dh} . (a) Predicted. (b) Measured with 20V/500Hz pulsating voltage injection in d -axis.

According to the analysis above, the cross-coupling effect compensation method by utilizing the offline measured ratio L_{dqh}/L_{qh} does not need to calculate the error angle. In Appendix 4, another cross-coupling effect compensation method by making use of the offline measured ratio between L_{dh} and L_{qh} is proposed. It is an original research idea. However, its universality is not satisfactory, i.e. only applicable in rotating voltage injection based sensorless control. The compensation methods with offline measured $\frac{L_{dqh}}{L_{qh}}$ and $\frac{L_{dh}}{L_{qh}}$ are compared in detail in Appendix 4.

Although, the compensation method with offline measured ratio L_{dqh}/L_{qh} can compensate the cross-coupling effect without obtaining the accurate error, it has some drawbacks as

- ① Decreased dynamic performance compared to direct error compensation.

The method in this section compensates the cross-coupling effect by modifying the input of the position observer. Therefore, under load transient, the dynamic response of the observer output is limited by the proportional-integral (PI) and integration modules as shown in Fig. 4.11(a) with the compensation applied at t_1 . Whereas for direct error compensation method as shown in Fig. 4.11(b), the estimated position is directly corrected at the output of the position observer. The method is carried out in a feedforward compensation way and has fast dynamic response. Experiments will be carried later to compare their dynamic performances.



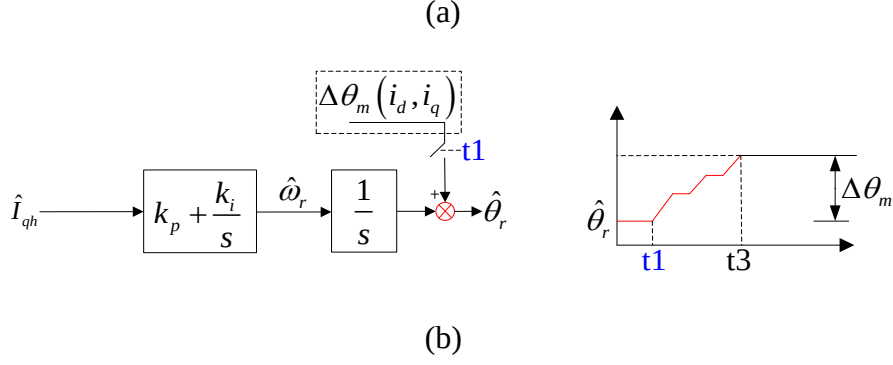


Fig. 4.11. Demonstration of dynamic responses after different compensation methods are applied. (a) With λ before the observer. (b) Feedforward compensation with $\Delta\theta_m$.

② Universality.

The method in this section requires dq -axis HF currents and is used in pulsating voltage injection based sensorless control. Hence, the universality of this method is poor. In the next section, applying $\lambda(L_{dqh}/L_{qh})$ in other HF voltage injection based sensorless methods will be investigated.

4.3 Improved Cross-coupling Effect Compensation Method

In this section, applying offline measured L_{dqh}/L_{qh} for direct error compensation in pulsating voltage injection based sensorless control will firstly be investigated. Then, the proposed method with feedforward compensation is presented. A novel method for offline measuring the errors is proposed. Finally, comprehensive comparison between the compensation methods with L_{dqh}/L_{qh} and $\Delta\theta_m$ will be made in terms of dynamic performance and universality.

4.3.1 Investigation of feedforward compensation with offline measured $\lambda = L_{dqh}/L_{qh}$

As presented in Fig. 4.11, the feedforward compensation method has better dynamic performance than the compensation by modifying the observer input with λ . Therefore, it is desirable if the error can be directly calculated with the offline measured λ in pulsating voltage injection based sensorless control as shown in Fig. 4.12.

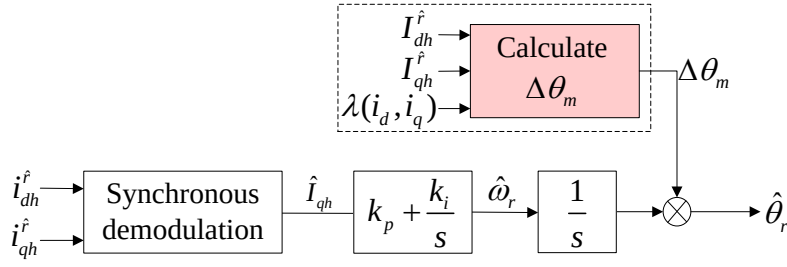


Fig. 4.12. Feedforward cross-coupling effect compensation method with offline measured $\lambda = L_{dqh}/L_{qh}$ in pulsating voltage injection based sensorless control.

As marked by red in Fig. 4.12, the key point for feedforward compensation with λ is to calculate $\Delta\theta_m$ from I_{dh}^* , I_{qh}^* which are expressed as (4.36).

$$\begin{bmatrix} \hat{I}_{dh} \\ \hat{I}_{qh} \end{bmatrix} = \frac{V_h}{\omega_h (L_{dh}L_{qh} - L_{dqh}^2)} \begin{bmatrix} L_\Sigma + \sqrt{L_\Delta^2 + L_{dqh}^2} \cos(2\tilde{\theta}_r - 2\Delta\theta_m) \\ \sqrt{L_\Delta^2 + L_{dqh}^2} \sin(2\tilde{\theta}_r - 2\Delta\theta_m) \end{bmatrix} \quad (4.36)$$

Disappointedly, it is not possible to calculate $\Delta\theta_m$ from λ , I_{dh}^* , I_{qh}^* in sensorless control due to insufficient information. Therefore, it is not feasible to performance feedforward compensation with offline measured λ in pulsating voltage injection based sensorless control. Whereas in sinusoidal rotating voltage injection based sensorless control, as will be evaluated later, $\Delta\theta_m$ can be calculated from offline measured λ and the online measured I_p , I_n .

4.3.2 Applying $\lambda = L_{dqh}/L_{qh}$ in rotating voltage injection

Actually, the offline measured cross-coupling factor $\lambda = L_{dqh}/L_{qh}$ can also be extended for use in HF rotating voltage injection based sensorless control by feedforward compensation. With HF rotating voltage injection, the rotor position error $\Delta\theta_m$ caused by cross-coupling effect can be directly calculated with I_p , I_n and λ as

$$\Delta\theta_m = 0.5 \tan^{-1} \left[\frac{\lambda [I_p + \sqrt{(1 + \lambda^2)I_n^2 - \lambda^2 I_p^2}]}{\sqrt{(1 + \lambda^2)I_n^2 - \lambda^2 I_p^2} - I_p \lambda^2} \right] \quad (4.37)$$

With (4.37), the cross-coupling effect in rotating voltage injection can be compensated with the calculated error $\Delta\theta_m$, as illustrated by the dashed line in Fig. 4.13, where $g(\lambda, I_p, I_n)$ is the function between $\Delta\theta_m$ and λ , I_p , I_n as defined by (4.37).

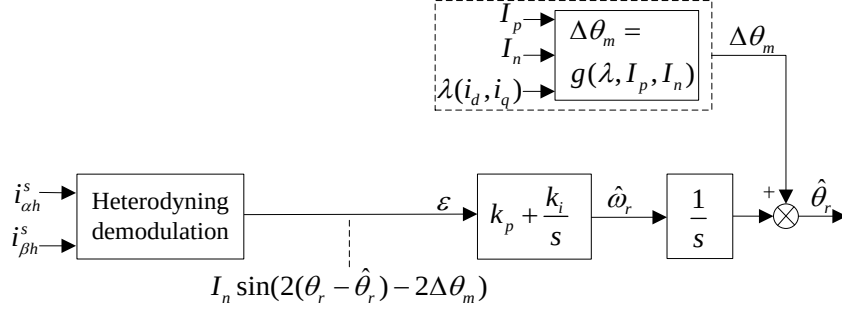


Fig. 4.13. Applying offline measured λ for cross-coupling effect compensation in sinusoidal rotating voltage injection based sensorless control.

4.3.3 Proposed method and comprehensive comparison

As can be seen, the equation $g(\lambda, I_p, I_n)$ for calculating $\Delta\theta_m$ is still very complicated. In this chapter, a cross-coupling effect compensation method with offline measured error $\Delta\theta_m$ is proposed as shown in Fig. 4.14. The error $\Delta\theta_m$ is offline measured with a simple method.

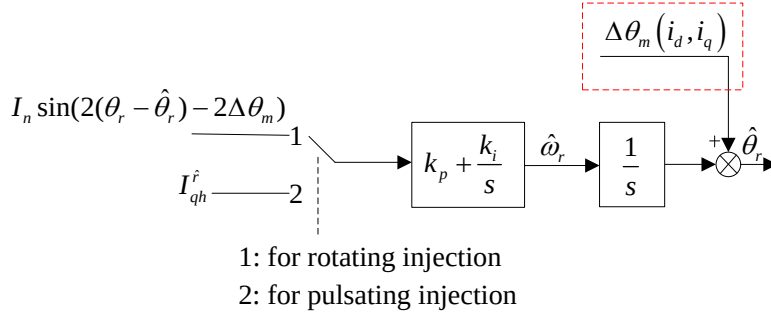


Fig. 4.14. Proposed cross-coupling effect compensation with offline measured $\Delta\theta_m$.

The key point for the proposed method is to offline obtain the error $\Delta\theta_m$. Actually, $\Delta\theta_m$ can be measured in a very similar way as measuring L_{dqh}/L_{qh} as follows.

Firstly, by injecting HF voltage (or HF current i_{dh1}) to the real d -axis which is identified from an encoder, and measuring the corresponding q -axis HF current i_{qh1} , the ratio between L_{dqh} and L_{qh} can be obtained as

$$\lambda_1 = \frac{L_{dqh}}{L_{qh}} = -\frac{i_{qh1}}{i_{dh1}} \quad (4.38)$$

Then, by injecting HF voltage (HF current i_{qh2}) to the real q -axis which is identified from an encoder, and measuring the corresponding d -axis HF current i_{dh2} , the ratio between L_{dqh} and L_{dh} can be obtained as

$$\lambda_2 = \frac{L_{dqh}}{L_{dh}} = -\frac{i_{dh2}}{i_{qh2}} \quad (4.39)$$

Finally, $\Delta\theta_m$ at different working points can be obtained from λ_1, λ_2 , i.e.

$$\Delta\theta_m = 0.5 \tan^{-1} \left(\frac{2L_{dqh}}{L_{qh} - L_{dh}} \right) = 0.5 \tan^{-1} \left(\frac{2\lambda_1\lambda_2}{\lambda_2 - \lambda_1} \right) \quad (4.40)$$

The measured errors are stored in a LUT and can be used for online cross-coupling effect compensation. As can be seen from Fig. 4.13 and Fig. 4.14, the offline measured λ and $\Delta\theta_m$ can both be used for cross-coupling effect compensation. The two methods both have fast dynamic performance due to feedforward compensation as will be confirmed by experiments. However, there are some differences between the cross-coupling effect compensation methods with λ and $\Delta\theta_m$.

Table 4.1 compares the differences between offline measured $\Delta\theta_m$ and λ used for cross-coupling effect compensation. As compensation with offline measured $\Delta\theta_m$ does not need any HF current information, it can be applied to other saliency tracking based sensorless control methods, such as pulse injection based method [FRE10], zero sequence voltage demodulation based method, etc. Whereas, the compensation method with offline measured λ is relying on the HF currents, i.e., I_p, I_n for rotating voltage injection, or $I_{dh}^{\hat{r}}, I_{qh}^{\hat{r}}$ for pulsating voltage injection. Therefore, the universality of the compensation method with offline measured $\Delta\theta_m$ is better compared to λ .

Table 4.1

Comparison of Cross-coupling Effect Compensation Methods with λ and $\Delta\theta_m$

Compensation methods	Injection signal	Compensation without $\Delta\theta_m$	Compensation by $\Delta\theta_m$
Requiring HF currents	HF pulsating	$I_{dh}^{\hat{r}}, I_{qh}^{\hat{r}}, \lambda$ [23] Fig. 2	Not applicable
	HF rotating	Not applicable	Calculated with I_p, I_n [REI08]
			Calculated with current derivatives [MOR21]
			Calculated with I_p, I_n and λ /Eq. 4.37

Not requiring HF currents	Irrespective of the injected signal	Not applicable	With curve fitting techniques [FRE10], [WAN19b]
			By varying reference current direction [WAN20b]
			Looked up from $\Delta\theta_m$ in LUT $\Delta\theta_m$ is measured with [XU14] or the method in (4.40)
Contributions	Extend λ from pulsating to rotating		A simple method for offline measuring $\Delta\theta_m$

4.3.4 Convergence analysis of the proposed method

In sensorless operation, there are differences between the reference dq -axis currents and the actual dq -axis currents due to position error as shown in Fig. 4.15.

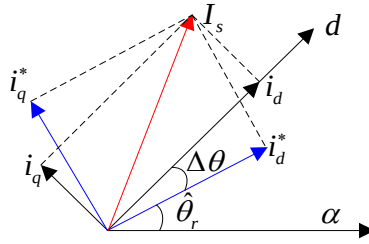


Fig. 4.15. Relationship between reference i_d^*, i_q^* and actual i_d, i_q in case of position error $\Delta\theta$.

In real applications, the inductances vary with the working points. Therefore, the looked up $\Delta\theta_m$ from the LUT may not be equal to the required value as

$$\Delta\theta_m(i_d^*, i_q^*) \neq \Delta\theta_m(i_d, i_q) \quad (4.41)$$

As can be seen from (4.41), due to variation of the actual $\Delta\theta_m(i_d, i_q)$ with position error, the estimated position may diverge especially in high load current region [MAN20]. For analysing stability and convergence with angle $\Delta\theta_m$ compensation, it is essential to analyse the input signal of the position observer with respect to the position error $\Delta\theta$, i.e. ε for rotating voltage injection and $I_{qh}^{\hat{r}}$ for pulsating voltage injection which have the same expressions as

$$\varepsilon = I_{dh}^{\hat{r}} = I_n \sin(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \quad (4.42)$$

For compensation the cross-coupling effect by adding $\Delta\theta_m(i_d^*, i_q^*)$ at the output of the observer, it is equivalent to subtract $\Delta\theta_m(i_d^*, i_q^*)$ from the estimated position $\hat{\theta}_r$ in (4.42). Hence, the observer input signal can be equivalently expressed as

$$\varepsilon = I_n \sin(2\Delta\theta - 2\Delta\theta_m(i_d, i_q) + 2\Delta\theta_m(i_d^*, i_q^*)) \quad (4.43)$$

In (4.43), $\Delta\theta_m(i_d^*, i_q^*)$ is fixed by assuming constant current references i_d^*, i_q^* , while $I_n, \Delta\theta_m(i_d, i_q)$ vary with the position error $\Delta\theta$. To guarantee reliable convergence, the signal ε should have positive slop zero-crossings with respect to the position error $\Delta\theta$ at the convergence point according to the analysis in [KWO19].

For the IPMSM in this thesis, the current references are assigned according to the maximum torque per ampere (MTPA) law.

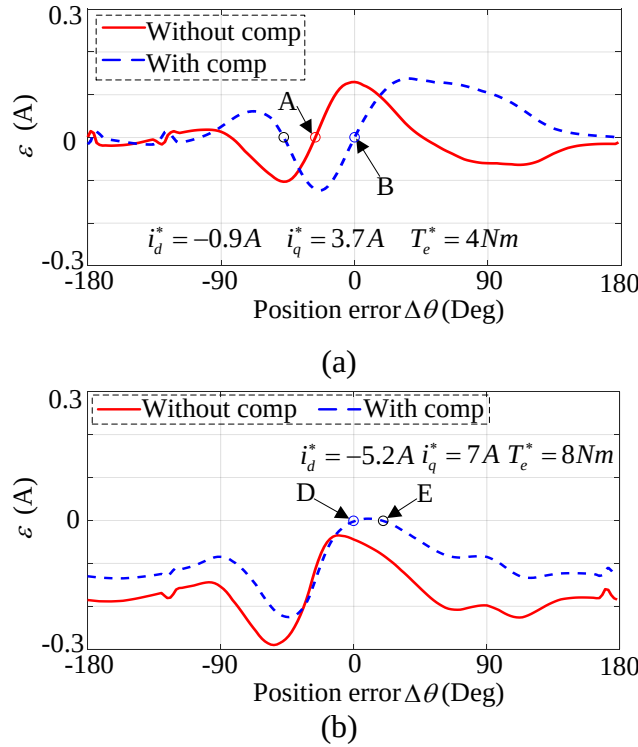


Fig. 4.16. Relationship between ε and $\Delta\theta$ without/with proposed compensation at different torque references ($T_N = 4Nm$). (a) $T_e^* = 4Nm$. (b) $T_e^* = 8Nm$.

Fig. 4.16 shows the ε - $\Delta\theta$ traces based on the offline measured current-flux linkage characteristics without/with angle $\Delta\theta_m$ compensation at different torque references. The injection voltage is chosen as 20V/500Hz. As can be seen from Fig. 4.16(a), for 4Nm (T_N) torque reference, the convergence point before compensation is located at point A ($\Delta\theta \approx -30^\circ$), while after compensation the convergence point is B ($\Delta\theta = 0^\circ$). Stable compensation can be achieved for 4Nm reference no matter whether the proposed compensation is applied.

Whereas for 8Nm ($2T_N$) reference, it is seen in Fig. 4.16(b), there does not exist positive slop zero-crossing before the angle compensation is applied. After compensation, there is a zero-

crossing at point D with $\Delta\theta = 0^\circ$. However, it may slide to the unstable point E, as the convergence interval between D and E is relatively small [KWO19]. Fig. 4.17(a) shows the sensorless current control results with angle $\Delta\theta_m$ compensation. The rotor is locked during test. The current references are ramped up to $i_d^* = -5.2\text{A}$, $i_q^* = 7\text{A}$. As can be seen, the estimation diverges with only $\Delta\theta_m$ compensation. This may be caused by spatial harmonics and inverter nonlinearities accounting for the small stability margin at this point.

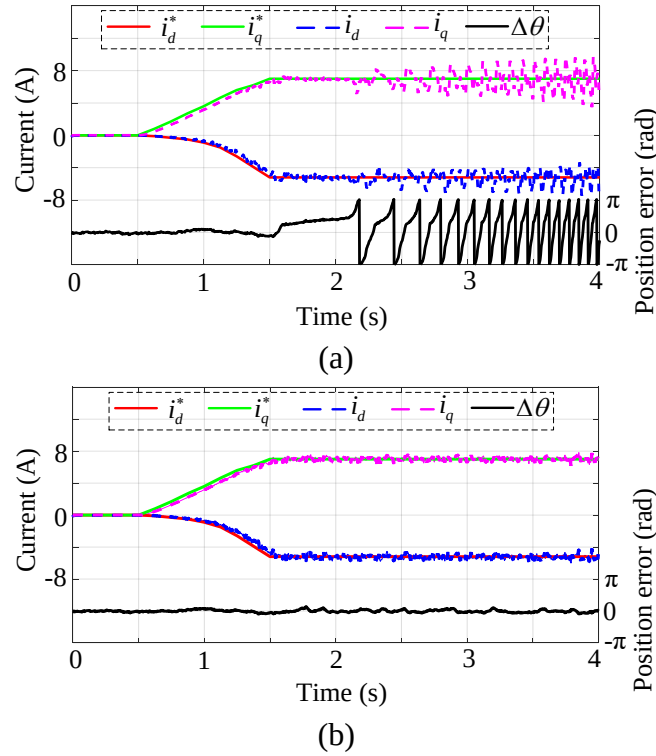


Fig. 4.17. Sensorless operation at highly saturation condition ($T_e^* = 8\text{Nm}$). (a) Divergence with only angle compensation. (b).Convergence with angle and current compensation.

To extend the sensorless operating region, the compensation method by employing both angle compensation and current compensation can be used. The current compensation is carried out by adding a current to the error signal ε , by which stable positive slop zero-crossing can be achieved at $\Delta\theta = 0^\circ$. With the guidelines given in [KWO19], Fig. 4.17(b) shows the sensorless current control results with both angle $\Delta\theta_m$ compensation and current compensation. Compared with only using angle compensation, convergence can be reached at $\Delta\theta = 0^\circ$.

As can be seen, angle compensation and current compensation can be combined to extend sensorless operation to highly saturation condition [MAN20]. Determination of the compensation values are highly dependent on the machine characteristics. For reliable sensorless control of IPMSM at low torque region, compensation the cross-coupling effect with angle $\Delta\theta_m$ is still effective for many applications.

4.4 Experimental Verification

Experiments are carried out on a dSPACE platform as shown in Fig. 4.18 to verify effectiveness of the proposed method. Parameters of the test IPMSM are listed in Appendix 1. The load torque is provided by a direct current generator connected to the test machine. Two external Hall-effect sensors are installed on an analog conditioning board for sampling the phase currents. The resolution of the analog to digital converter for sampling current is 16bits. An incremental encoder of 1024 pulses per revolution is used to obtain the rotor position θ_r . The inverter switching frequency and system sampling frequency are both set as 10 kHz. The HF sinusoidal voltage of 15V/500Hz, which is selected according to the guidelines in [ZHU11], is injected to the estimated d -axis or the stationary frame. The position observer parameters are chosen as $k_p = 1000, k_i = 1000$ considering dynamic performance.

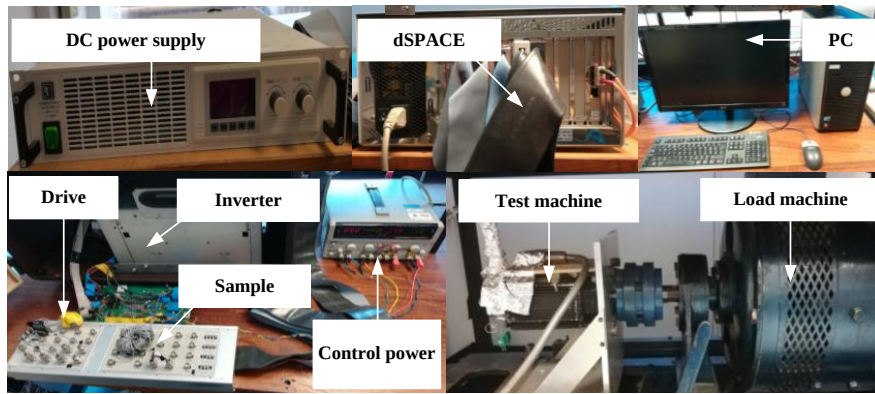


Fig. 4.18. Experimental setup.

4.4.1 Steady state performance

(a) Offline measured $\Delta\theta_m$ with the proposed measuring method

First of all, $\Delta\theta_m$ is offline measured according to the method given in Section 4.3.3. Ideally with sinusoidal distributed flux linkages, the dq-axis inductances and $\Delta\theta_m$ do not vary with the rotor position. However, due to flux linkage harmonics, there are position dependent harmonics in the measured $\Delta\theta_m$. The spatial harmonics can result in ripples in the estimated position and may even lead to instability in the position estimation at high load currents [MAN20]. Fig. 4.19 shows the measured $\Delta\theta_m$ under at different positions.

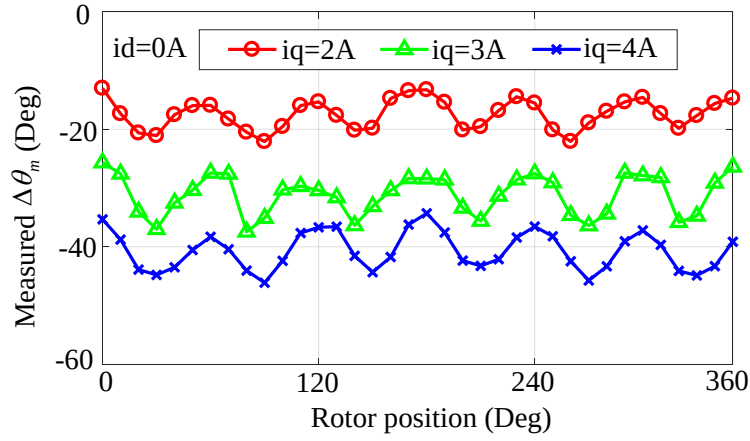


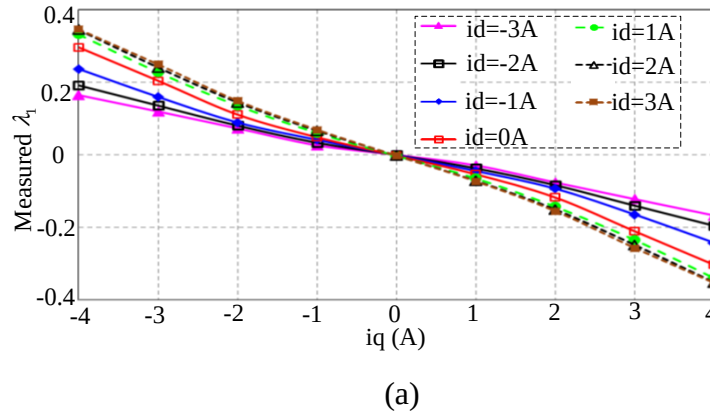
Fig. 4.19. Measured $\Delta\theta_m$ at different rotor positions.

As can be seen, the measured $\Delta\theta_m$ contain 6th harmonics and the peak to peak value is 10 degrees which will result in ± 5 degrees ripple in the compensated position. To mitigate the ripples, the measured $\Delta\theta_m$ at different positions can be stored and used for online cross-coupling compensation. However, this will largely increase the scale of the LUTs. As a common practice in industrial applications, the measured $\Delta\theta_m$ can be modelled as

$$\Delta\theta_m = \Delta\theta_{dc} + \Delta\theta_{ac} \cos(6\theta_r + \varphi) \quad (4.44)$$

where $\Delta\theta_{dc}$, $\Delta\theta_{ac}$ are the average and alternating values of the measured angles, φ is the phase shift for different working point.

The measured results (average values during one electrical period) for d -axis current varies from -3A to 3A and q -axis currents from -4A to 4A, are shown in Fig. 4.20.



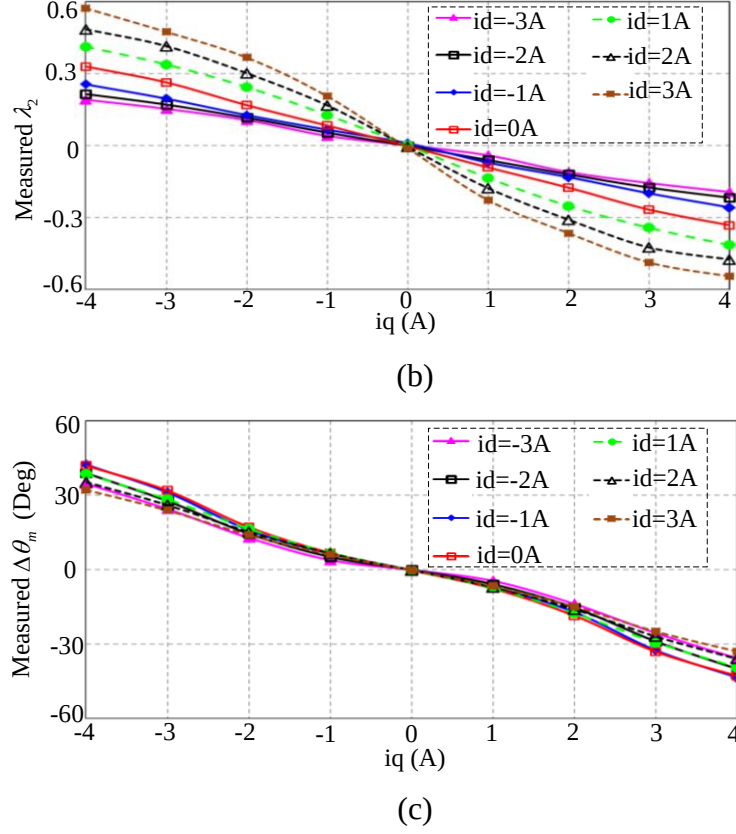


Fig. 4.20. Offline Measured results. (a) λ_1 . (b) λ_2 . (c) $\Delta\theta_m$.

(b) Pulsating voltage injection based sensorless control without/with the proposed compensation

Fig. 4.21 compares the estimated position, the estimation error and the phase current without/with cross-coupling effect compensation in HF pulsating voltage injection based sensorless control, with the proposed compensation method applied at 0.9s. After the proposed compensation method, using offline measured error $\Delta\theta_m$, is applied, the estimation error is reduced from about -30 degrees to approximately zero. The amplitude of the phase current under rated load torque and maximum torque per ampere (MTPA) control is also decreased when the position error is nearly zero.

The method in [LI09b], which is also based on HF pulsating voltage injection, using offline measured ratio λ for cross-coupling effect compensation, is executed for comparison. The pulsating voltage is chosen as 15V/500Hz to guarantee the same frequency-amplitude ratio as the proposed method. The PLL parameters, the LPF cut-off frequency in position demodulation, the speed and current PI parameters are set as the same as the proposed method, shown in Table 4.2. The cross-coupling compensation strategy using λ is applied at 0.9s. The estimated position, estimation error and phase current are shown in Fig. 4.21(a). Before compensation is applied,

the position error is about -30 degrees. After compensation, the error is also reduced to nearly zero which is the same as the proposed method. Therefore, the proposed method with feedforward compensation has similar steady state performance as the method using λ . However, it is noted in Fig. 4.21(a) and (b) that the two methods have different dynamic performances, as will be further compared in detail in the next section.

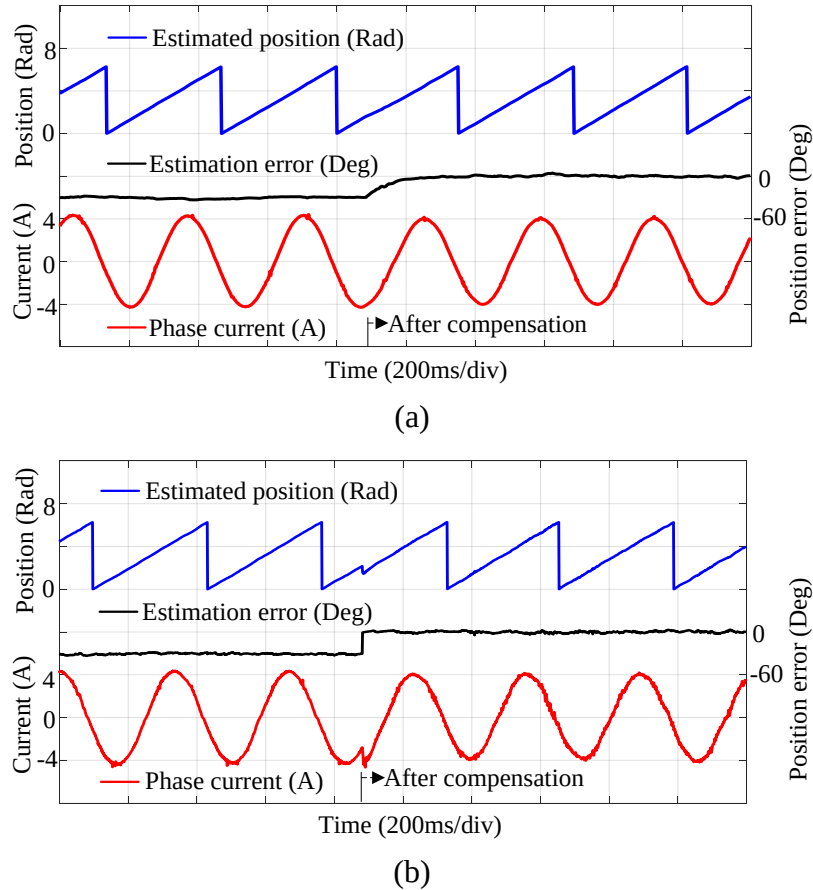


Fig. 4.21. Comparison of steady-state performance of HF pulsating voltage injection sensorless control before and after cross-coupling effect compensation, 60rpm (3Hz) and rated load torque. (a) With conventional method using λ . (b) With proposed method using $\Delta\theta_m$.

The cross-coupling effect can result in position estimation error which is unwanted in sensorless position estimation. However, existence of the cross-coupling effect also extends the feasible region for sensorless operation. Theoretically, at the point of $L_{dh}=L_{qh}$, the saliency variation disappears, i.e. $I_n = 0$. Due to cross-coupling effect, I_n is not zero at this point and estimation convergence can also be achieved.

Table 4.2
LPF, PLL, SPEED PI AND CURRENT PI PARAMETERS

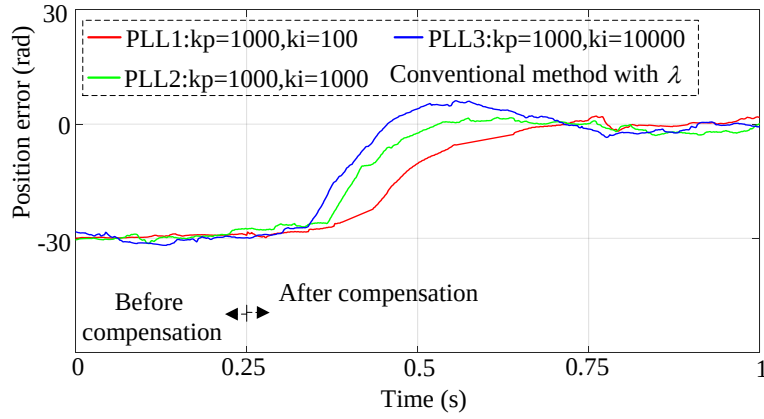
	Injection frequency	LPF	PLL		Speed PI		Current PI	
			k_p	k_i	k_p	k_i	k_p	k_i
Rotating	500Hz	200Hz	1000	1000	0.2	0.2	80	1000
Pulsating	500Hz							

4.4.2 Dynamic performance

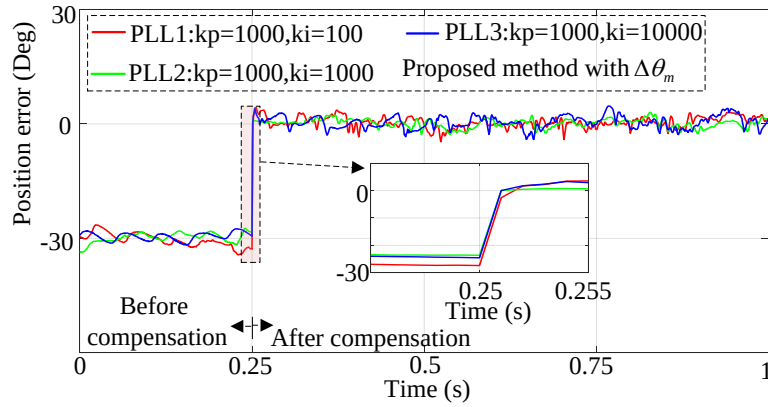
Experiments are carried out to compare the dynamic performances of the conventional method using λ and the proposed method with feedforward compensation $\Delta\theta_m$ in pulsating voltage injection based sensorless control, as shown in Fig. 4.22.

Compensation performances with different PLL parameters for the proposed method under rated load torque are compared in Fig. 4.22(b). The k_p parameter of the observer is fixed as 1000 and the k_i parameter is set as 100, 1000 and 10000, respectively. It is interesting to find that the response time is about 1ms (ten PWM control cycles) and nearly fixed for different PLL observer parameters.

The compensation method in [LI09], which uses offline measured cross-coupling factor λ for cross-coupling effect compensation is also executed for comparison. The pulsating voltage is chosen as 15V/500Hz and the other control parameters are the same as the as the proposed method, shown in Table 4.2. The compensation strategy is applied under the rated load torque at 60rpm (3Hz) with different PLL parameters as shown in Fig. 4.22(a). It is observed that the response time (about 200ms) is longer than that of the proposed method (1ms) with the same PLL parameters. Although the response time can be reduced with higher integral parameter, overshoot may also occur and results in ripples in the estimated position. The varying response time is explained that the compensation in [LI09] is achieved by modifying the input of the PLL and the response time is dependent on the PLL bandwidth. Comparatively, the proposed method compensates the cross-coupling effect by correcting the estimated position with the offline measured error and is therefore faster and less sensitive to the PLL parameters.



(a)



(b)

Fig. 4.22. Comparison of dynamic performance of HF pulsating voltage injection sensorless control with different PLL parameters. (a) With conventional method using λ . (b) With proposed method using $\Delta\theta_m$.

Sensorless start performance, i.e. phase current and estimated position, with the proposed method and the conventional method in pulsating voltage injection based sensorless control are also compared in Fig. 4.23. The PLL parameters, the LPF cut-off frequency, speed and current PI parameters are also the same for the two methods, as listed in Table 4.2. The speed reference is increased from 0rpm to 60rpm over 2s under rated load torque with MTPA strategy employed. The startup times are found to be almost the same for the two methods. However, it is noted that the maximum error during startup with the conventional method is about 30 degrees, while the maximum error with the proposed method is only about 16 degrees. The startup current with the conventional method in [LI09] is also higher than that with the proposed method as a result of higher position error. Thus, the proposed method has better dynamic performance than the conventional method with offline measured λ in pulsating voltage injection based sensorless control when the same PLL and control parameters are used.

After startup to 60r/min, the estimation errors for the two methods are all nearly zero and the phase current amplitudes are the same. This means that the method with feedforward compensation and the method with λ have almost the same compensation performance at steady state.

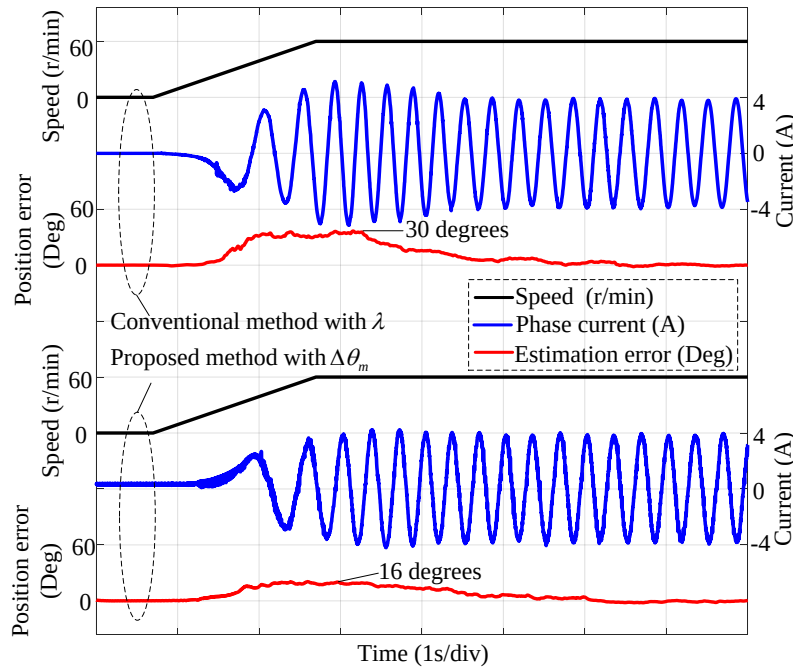


Fig. 4.23. Comparison of HF pulsating voltage injection sensorless startup performance between conventional method with λ and proposed method with $\Delta\theta_m$.

Performances of the proposed method and the conventional method in [LI09] are compared under load torque transients with HF pulsating voltage injection. The machine is sensorless operated at 60rpm (3Hz) with the proposed method and the conventional method, respectively. All their control parameters are set as the same as shown in Table 4.2. The load torque is increased to 4Nm (rated load torque) at 0.7s and then decreased to zero at 3.7s. The speeds, position errors, and q -axis currents with the two methods are shown in Fig. 4.24, respectively. It is observed that the maximum position error with the proposed method during load torque transients are below ± 18 degrees, while the maximum error with the conventional method can reach up to ± 30 degrees. A position error of 18 degrees will result in about only 5% ($1 - \cos(18^\circ)$) torque reduction, while 30 degrees of error will result in nearly 13% ($1 - \cos(30^\circ)$) torque reduction. Besides, the response time for the conventional method to suppress the error to zero is longer than that for the proposed method. Overall, the proposed method with feedforward compensation $\Delta\theta_m$ has better dynamic performance under load torque transients compared to the compensation method with λ .

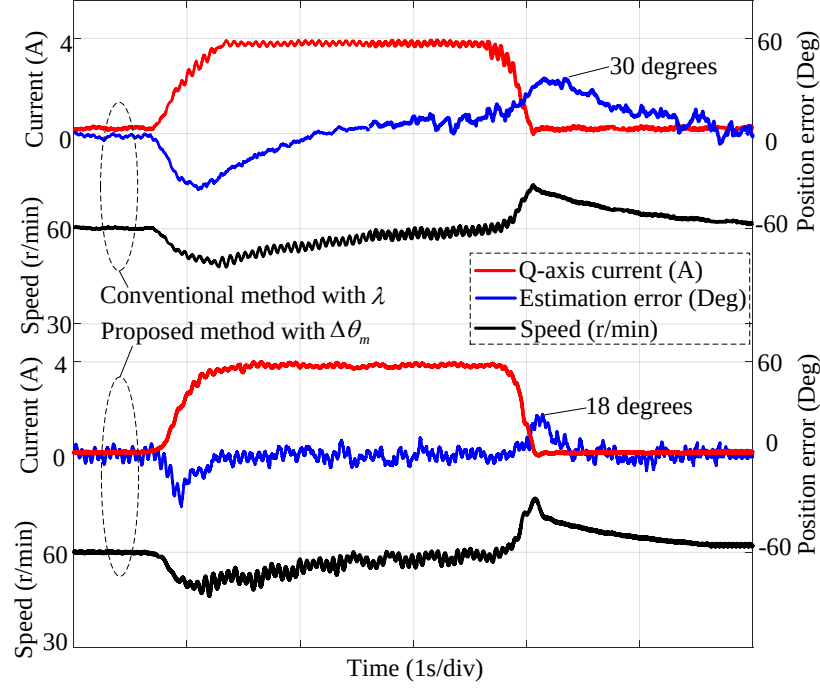


Fig. 4.24. Comparison of HF pulsating voltage injection sensorless performance under load transients between conventional method with offline measured λ and proposed method with offline measured $\Delta\theta_m$.

4.4.3 Compensation performances in rotating voltage injection based sensorless control with offline measured λ

The offline measured λ and $\Delta\theta_m$ are both applicable for cross-coupling effect compensation in rotating voltage injection based sensorless control by feedforward compensating the error. Fig. 4.25 compares the estimated position, the estimation error and the phase current without/with cross-coupling effect compensation in HF rotating voltage injection based sensorless control, with the compensation methods using offline measured λ and $\Delta\theta_m$ applied at 2s, respectively. The HF rotating voltage is chosen as 15V/500Hz and the control parameters are set the same as shown in Table 4.2.

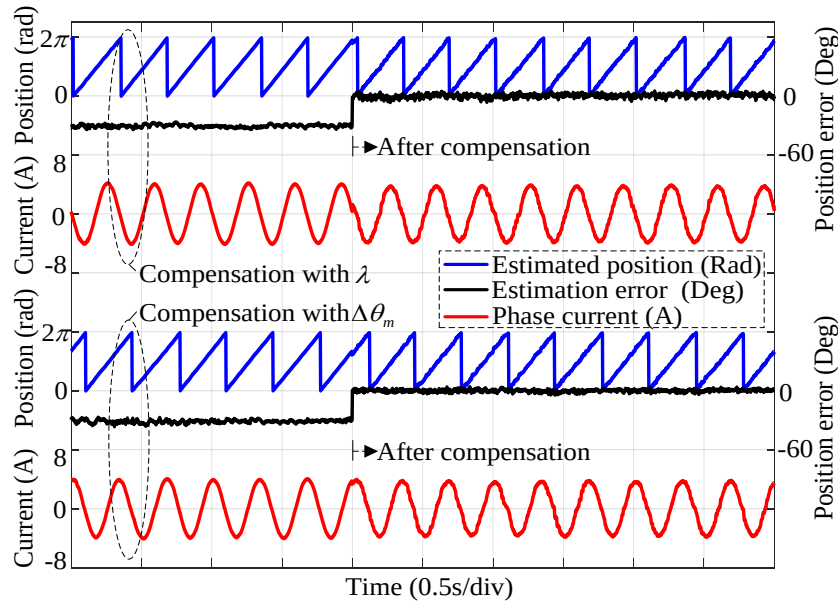


Fig. 4.25. Comparison of steady-state performance of HF rotating voltage injection sensorless control before and after cross-coupling effect compensation, 60rpm (3Hz) and rated load torque with offline measured λ and $\Delta\theta_m$.

As can be seen the position errors after compensation are changing from -30 degrees to nearly zero. The compensation methods with λ and $\Delta\theta_m$ in rotating voltage injection based sensorless control have almost the same steady state performances.

Fig. 4.26 shows the sensorless startup performance with the cross-coupling effect compensation methods using λ and $\Delta\theta_m$ respectively in rotating voltage injection sensorless control. The control parameters for the two methods are set the same as shown in Table 4.2.

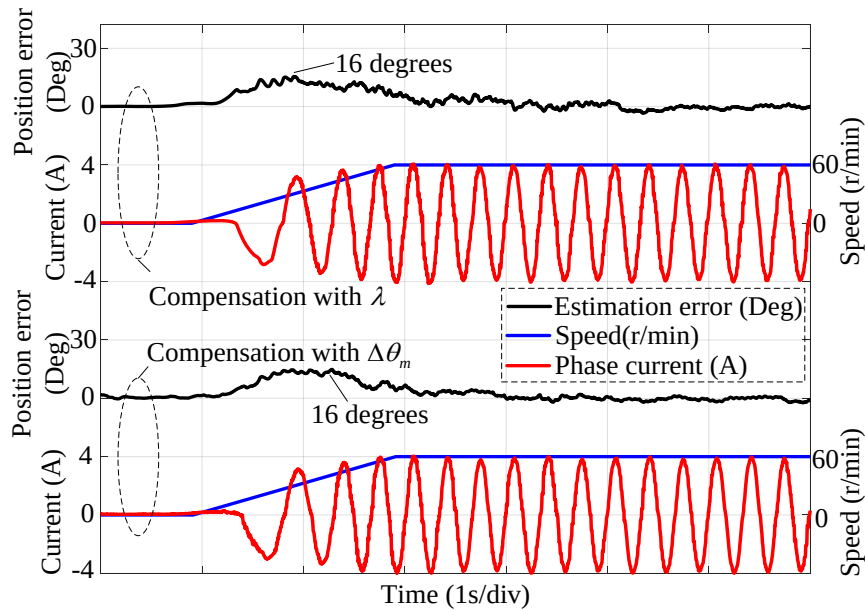


Fig. 4.26. Comparison of HF rotating voltage injection based sensorless startup performance between using λ and $\Delta\theta_m$.

It is observed in Fig. 4.26 that the startup times, the maximum errors (16°) and the startup currents are almost the same. The compensation methods with λ and $\Delta\theta_m$ can both be used to compensate the cross-coupling effect and have similar dynamic performance.

Performances of the compensation methods with λ and $\Delta\theta_m$ are compared under load torque transients with HF rotating voltage injection. The control parameters are set as the same shown in Table 4.2. Rated load torque is applied at 2s and decreased to zero at 6s. The speeds, position errors, and q -axis currents with the two methods are shown in Fig. 4.27, respectively.

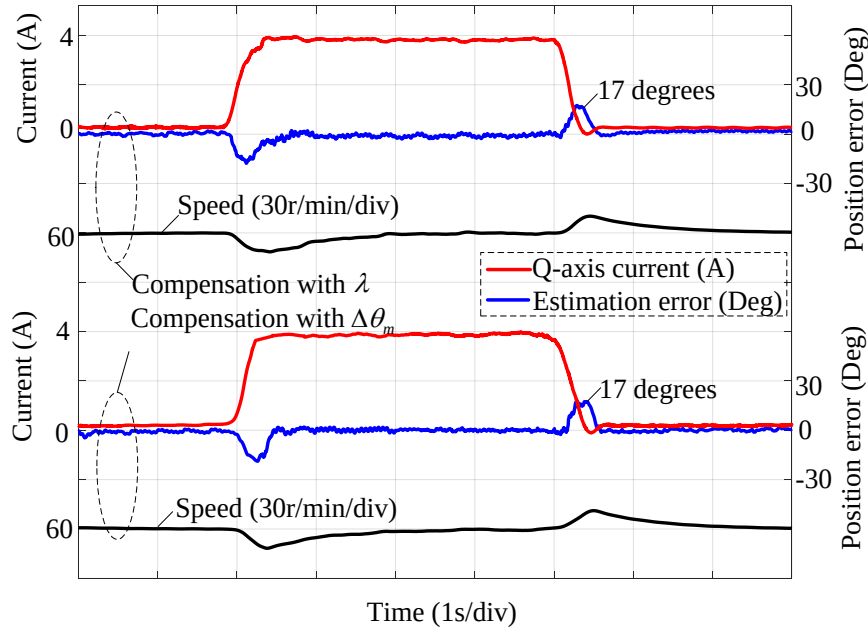


Fig. 4.27. Comparison of HF rotating voltage injection sensorless performance under load transients between using λ and $\Delta\theta_m$.

It is observed in Fig. 4.27 that the compensation methods with λ and $\Delta\theta_m$ in rotating voltage injection have nearly the same speed deviations, estimation errors, which are also quite close to those of the compensation method with $\Delta\theta_m$ in pulsating voltage injection. This can be explained that they all achieve the compensation by feedforward compensating the errors. Comparatively, the compensation method with λ in pulsating voltage injection in Fig. 4.24 has decreased dynamic performance as confirmed by the previous experiments.

4.5 Conclusion

In this chapter, the impacts of cross-coupling effect on sensorless control are firstly analysed. Then, existing cross-coupling effect compensation methods for HF voltage injection based sensorless control have been introduced, together with their advantages and disadvantages.

Thirdly, an improved cross-coupling effect compensation method is proposed for sensorless control of interior permanent magnet synchronous machines with high frequency voltage injection. In the proposed method, the error caused by the cross-coupling effect is offline measured by injecting HF voltages (or HF currents) to the d -axis and q -axes, respectively, and used for online cross-coupling compensation in high frequency voltage injection based sensorless control. Compared with the conventional method based on the offline measured ratio between the mutual and q -axis high frequency inductances, for pulsating voltage injection based sensorless control only, the proposed method has extended it for use in rotating voltage injection, and also has simple implementation, better universality, and better dynamic performance, as confirmed by experiments.

CHAPTER 5

SIMULTANEOUS SENSORLESS ROTOR POSITION AND TORQUE ESTIMATION FOR IPMSM AT STANDSTILL AND LOW SPEED BASED ON HIGH FREQUENCY SQUARE WAVE VOLTAGE INJECTION

5.1 Introduction

Permanent magnet synchronous machines (PMSMs) have been widely used in domestic and industrial applications due to both high torque density and efficiency [ZHU07]. Precise torque control of PMSM is required in many applications, such as robotics, elevators and electric vehicles [CHA08]. Accurate torque feedback of the machine is essential for both control and monitoring purposes. Hardware transducers such as strain gauges [ITO17] and torsional displacement [HEI11] can be used to measure the machine torque. However, they are usually very expensive and require extra installation space and cables. Therefore, torque estimation without sensor is a preferable solution.

Conventionally, the machine torque can be estimated from the product of stator flux linkage and current [BOL08]. The stator flux linkage is usually observed from the terminal voltages and phase currents. There are two kinds of flux linkage observers which are all based on the fundamental electrical models, i.e. voltage and current models [CHE03], [HEJ11], [CHE20]. The voltage model calculates the stator flux linkage by integrating the back electromagnetic force (EMF) which is the difference between the terminal voltage and resistance voltage drop [CHE03]. This method is effective at medium and high speed ranges and only requires the resistance parameter, but has deteriorated performance at standstill and low speed due to reduced amplitude of the EMF signal [HEJ11] and voltage distortions caused by the inverter nonlinearities [CHE20]. The current model based flux linkage observer calculates the flux linkage from the summation of the permanent magnet flux and the stator current induced flux in the synchronous rotating reference frame. It is not dependent on the back EMF signal and thus has good performance at standstill and low speed. However, it requires dq-axis apparent inductances [XU14] which vary with currents, and the rotor position which is used to identify the dq-axes. A hybrid flux observer is introduced to provide smooth switch between the two

models, i.e. the current model dominates below the crossover frequency and the voltage model dominates above the crossover frequency [YOO09]. Although the inductance variations are taken into account by correcting the flux with the outputs of current regulators, complexity of the control algorithm is increased due to tuning of the regulators. Besides, the delay introduced by the low pass filter [YOO09] may result in deterioration of the torque estimation during load transients [TIN15].

To increase the robustness of flux linkage observer, different parameter identification methods can be used. In [ICH06], the winding resistance and dq-axis inductances are identified with recursive least square (RLS) method for improving the sensorless performance with extended EMF model. In [LIU11], a novel method by injecting a short pulse of negative d-axis current is proposed for parameter identification with the measured terminal currents, voltages and rotor speed. The methods in [ICH06], [LIU11] have satisfactory performances at medium and high speeds, but may lose effect at standstill and low speed due to reduced amplitudes of the back EMF signals. Therefore, for the purpose of torque estimation at standstill and low speed, conventional parameter identification methods based on the fundamental electrical models are inappropriate.

To reduce the dependency of torque estimation on the fundamental model, high frequency (HF) signal injection which is superimposed on the fundamental excitations is employed for online inductance parameter identification [KAN21], [MAR20]. In [KAN21], the dq-axis HF inductances are online measured by injecting an HF voltage between the d- and q-axes for self-sensing capability monitoring. However, the method in [KAN21] is intended only for surface mounted permanent magnet machines (SPMSM) without considering the cross-coupling effect and a position sensor is required for identifying the dq-axis positions. In [MAR20], the dq-axis HF inductances are online estimated by injecting the HF pulsating current signal and then used to calculate the torque. This method can be used without interfering with the normal operation of the machine and has better dynamic performance due to HF injection operation. However, the injected HF signal in [MAR20] is merely used for identifying the HF inductances while the rotor position should also be provided by a position sensor. In this chapter, for reducing the hardware cost, the injected HF signal is used for both sensorless rotor position estimation and online HF inductance identification simultaneously, which are further used to estimate the torque.

The saliency tracking method by injecting HF signal is widely used for sensorless control of PMSM at standstill and low speed [WAN20]. The interior permanent magnet synchronous

machine (IPMSM) which has high saliency level is considered in this thesis due to good sensorless performance at standstill and low speed [BAR17], [CHE17]. Conventionally, HF sinusoidal and square wave signals can be injected to the machine for sensorless rotor position estimation [WAN20], [MUR10]. As the injection frequency of sinusoidal signal is usually lower than that of square wave voltage injection due to limited sampling points during one injection cycle [LEI08], [YOO11], it is desirable to achieve simultaneous rotor position estimation and HF inductance identification with square wave voltage injection [YOO11], [LIU14c]. However, with conventional square wave voltage injection in the estimated d-axis, as will be verified later, only the d-axis HF inductance can be identified at steady state from the d-axis HF current variation.

To enable simultaneous HF inductances online identification for torque estimation and rotor position estimation with HF square wave voltage injection, a new method by injecting square wave voltages in the estimated dq-axes is proposed in this thesis. The dq-axis HF inductances can be calculated from the measured current variations and used for torque estimation by curve-fitting the dq-axis flux linkages. Besides, the impacts of cross-coupling effect on both position and torque estimation are analysed and compensated.

This chapter is organized as follows. In section 5.2, the conventional HF square wave voltage injection based sensorless method is introduced and its capability for online identification of HF inductances is investigated. In section 5.3, the proposed injection method is described and the method for estimating the torque with HF inductances is presented. The impacts of cross-coupling effect on rotor position and torque estimation are discussed and compensated. Simulations are carried out to verify the accuracy of the torque estimation with HF inductances. In section 5.4, experiments are given to verify the effectiveness of the proposed method. Finally, conclusion is given.

5.2 Conventional Sensorless Rotor Position Estimation Method

The torque T_e is expressed as the cross-product of flux linkages and stator currents. The key point for torque estimation is the estimation of the stator flux linkage. The voltage model based flux linkage observer has deteriorated the performance at standstill and low speed due to low amplitude of the back EMF signal [HEJ11]. Hence, at low speed and standstill, a current model based flux linkage observer is used.

However, in the current model, two factors should be considered: rotor position and inductance variations. Firstly, accurate rotor position is required to obtain i_d^r, i_q^r . In sensorless control, the rotor position is provided by various sensorless algorithms. Secondly, the dq -axis inductances vary with i_d^r, i_q^r . If the inductance variations are not considered, there will be errors in the calculated dq -axis flux linkages in the current model.

Therefore, for calculating the dq -axis flux linkages for torque estimation, the variation of inductances should be considered. However, at standstill and low speed, the fundamental model based inductance estimation strategy has a deteriorated accuracy due to low back EMF signal, Therefore, in this chapter, the HF voltage injection technique is investigated for both sensorless rotor position estimation and inductance identification which is further used for torque estimation.

5.2.1 HF model

Voltage equations of the IPMSM in the synchronous rotating reference frame (SRRF) can be expressed as

$$\begin{bmatrix} u_d^r \\ u_q^r \end{bmatrix} = \begin{bmatrix} R_s + pL_d & -\omega_e L_q \\ \omega_e L_d & R_s + pL_q \end{bmatrix} \begin{bmatrix} i_d^r \\ i_q^r \end{bmatrix} + \omega_e \begin{bmatrix} 0 \\ \psi_{pm} \end{bmatrix} \quad (5.1)$$

where R_s is the stator resistance, L_d, L_q are the apparent d -axis and q -axis inductances, ω_e, ψ_{pm} are the electrical angular speed and the permanent magnet flux linkage, $u_d^r, u_q^r, i_d^r, i_q^r$ are the dq -axis voltages and currents, “ p ” denotes the differential operator, the superscript “ r ” stands for the SRRF.

For sensorless control of IPMSM with HF voltage injection, by neglecting the speed dependent terms and the resistance, the HF voltage equations in the SRRF are given by

$$\begin{bmatrix} u_{dh}^r \\ u_{qh}^r \end{bmatrix} = \begin{bmatrix} L_{dh} & 0 \\ 0 & L_{qh} \end{bmatrix} p \begin{bmatrix} i_{dh}^r \\ i_{qh}^r \end{bmatrix} \quad (5.2)$$

where $u_{dh}^r, u_{qh}^r, i_{dh}^r, i_{qh}^r$ are the HF voltages and currents, L_{dh}, L_{qh} , are the dq -axis HF inductances which vary with the fundamental currents. Different from the apparent inductances which describe the ratio between flux linkage and current, the HF inductance describes the change rate of flux linkage to current [XU14].

5.2.2 Conventional sensorless method based on HF square wave voltage injection

For pulsating square wave voltage injection based sensorless control [YOO11], the pulsating square wave is injected to the estimated d -axis as

$$u_{dh}^{\hat{r}}(t) = \begin{cases} V_h, t \in [2kT_s, (2k+1)T_s] \\ -V_h, t \in [(2k+1)T_s, (2k+2)T_s] \end{cases}, k = 0, 1, 2, \dots \quad (5.3)$$

where T_s is the duration for the positive or negative voltage, V_h is the injection amplitude as shown in Fig. 5.1, “ $\hat{r}$ ” denotes the estimated SRRF.

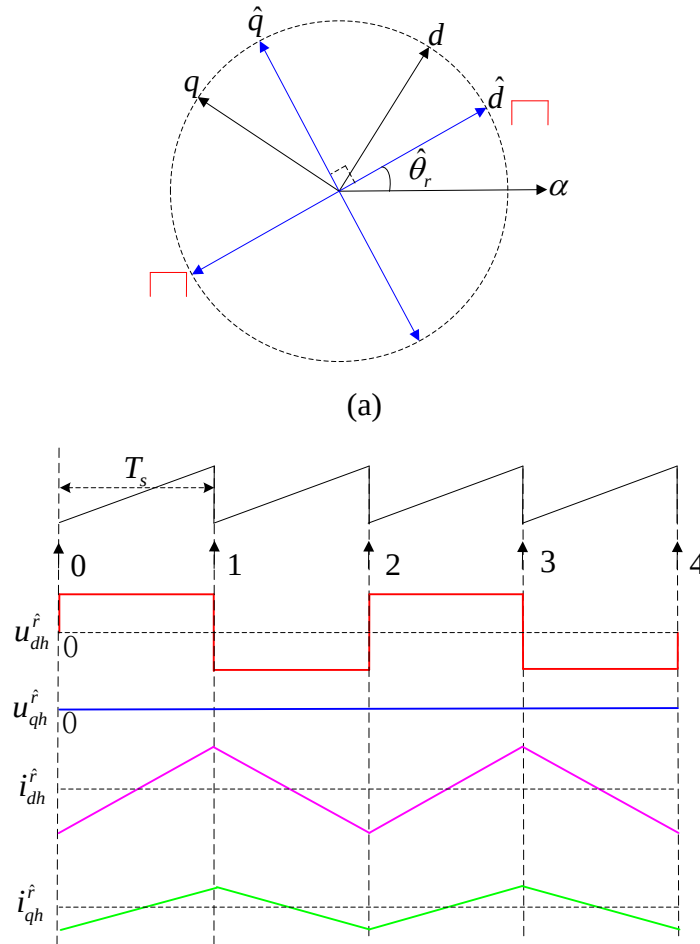


Fig. 5.1. Conventional sensorless rotor position estimation method with square wave voltage injection in the estimated d -axis. (a) Vector diagram. (b) Injection sequence.

The voltage equations in the estimated synchronous rotating reference frame are derived from (5.2) as

$$\begin{bmatrix} \dot{i}_{dh}^{\hat{r}} \\ \dot{i}_{qh}^{\hat{r}} \end{bmatrix} = R^{-1}(\theta_r - \hat{\theta}_r) \begin{bmatrix} L_{dh} & 0 \\ 0 & L_{qh} \end{bmatrix}^{-1} R(\theta_r - \hat{\theta}_r) \int \begin{bmatrix} u_{dh}^{\hat{r}} \\ u_{qh}^{\hat{r}} \end{bmatrix} dt \quad (5.4)$$

$$R(x) = \begin{bmatrix} \cos(x) & \sin(x) \\ -\sin(x) & \cos(x) \end{bmatrix} \quad (5.5)$$

where $\hat{\theta}_r$ is the estimated rotor position. Substituting (5.3) into (5.4), the resulting HF current variations during one injection interval in the estimated frame due to HF square wave voltage injection can be solved as

$$\begin{bmatrix} \Delta \hat{i}_{dh}^r(kT_s) \\ \Delta \hat{i}_{qh}^r(kT_s) \end{bmatrix} = \frac{V_h T_s (-1)^{k-1}}{2L_{dh}L_{qh}} \begin{bmatrix} L_{qh} + L_{dh} + (L_{qh} - L_{dh}) \cos(2(\theta_r - \hat{\theta}_r)) \\ (L_{qh} - L_{dh}) \sin(2(\theta_r - \hat{\theta}_r)) \end{bmatrix} \quad (5.6)$$

The rotor position can be estimated by controlling $\Delta \hat{i}_{qh}^r(-1)^{k+1} = 0$, i.e. $\hat{\theta}_r = \theta_r$, with a closed-loop observer, as shown in Fig. 5.2.

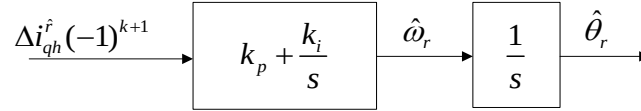


Fig. 5.2. Position estimation with conventional HF square wave voltage injection based sensorless method.

5.2.3 Investigation of capability of obtaining HF inductances

In HF voltage injection based sensorless control, knowledge of the HF inductances is not important. What important for reliable convergence of the position observer is the difference between L_{qh} and L_{dh} . Although accurate values of the HF inductances are not required for sensorless control, they can be used for monitoring the operation conditions, self-sensing capability evaluation [KAN21] and torque estimation [MAR20] if they can be identified online. In this chapter, the HF inductances will be used for online torque estimation. Therefore, in this section, the capability of online HF inductance identification for HF square wave voltage injection based sensorless control will be firstly investigated. As will be shown in Appendix 6, the HF sinusoidal rotating voltage injection technique can be used for simultaneous sensorless rotor and torque estimation. However, due to relatively complicated signal processing for sensorless position estimation and lower injection frequency compared to the HF square wave voltage injection, the sinusoidal rotating voltage injection has lower dynamic performance than square wave voltage injection as will be verified by experiments.

For conventional square wave voltage injection based sensorless rotor position estimation method, the d-axis HF inductance can be calculated from the d-axis current variations $\Delta \hat{i}_{dh}^r$ in (5.6) at steady state as

$$L_{dh}(\hat{\theta}_r = \theta_r) = \frac{V_h (-1)^{k-1}}{\Delta \hat{i}_{dh}^r(kT_s)} = L_{dh}, k = 0, 1, 2, \dots \quad (5.7)$$

Under load transients, i.e. $\hat{\theta}_r \neq \theta_r$ in (5.6), there will be error in the calculated d-axis HF inductance as

$$L_{dh}(\hat{\theta}_r \neq \theta_r) = \frac{V_h T_s (-1)^{k-1}}{\Delta \hat{i}_{dh}^r(kT_s)} = \frac{1}{1 - \frac{L_{qh} - L_{dh}}{L_{qh}} \sin^2(\Delta \theta_r)} L_{dh} \quad (5.8)$$

where $\Delta \theta_r = \theta_r - \hat{\theta}_r$ is the position error. Comparing (5.7) and (5.8), the calculated d-axis HF inductance is larger than the real value under load transient when the position error exists for the conventional square wave injection sensorless method. Besides, from the current variations in (5.6), the q-axis HF inductance cannot be identified at both steady and transient states as $\Delta \hat{i}_{qh}^r$ is controlled to zero for position estimation.

Based on the analyses above, with conventional sensorless method with HF square wave voltage injection in the estimated d-axis, only the d-axis HF inductance can be measured at steady state. In this thesis, for the purpose of simultaneous sensorless rotor position estimation and torque estimation using online identified HF inductances with square wave voltage injection, a new method by injecting HF square wave voltages in the estimated dq-axes is proposed.

5.3 Proposed Method for Simultaneous Rotor Position and Torque Estimation with HF Square Wave Voltage Injection

5.3.1 Proposed injection sequence

The proposed method injects HF square wave voltages in the estimated dq -axes alternatively as shown in Fig. 5.3.

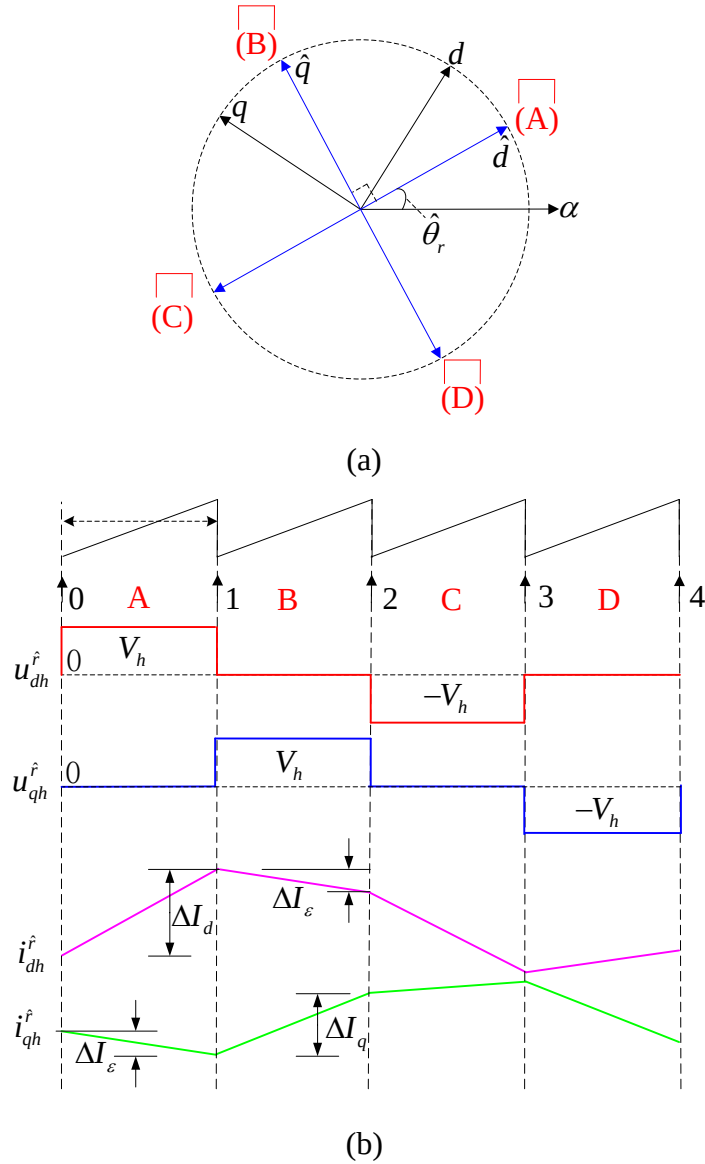


Fig. 5.3. Proposed method with square wave voltage injection in the estimated dq -axes. (a) Vector diagram. (b) Injection sequence.

One injection cycle of the proposed method is composed of four injection stages, i.e. A, B, C and D shown in Fig. 5.3(a) and (b), i.e.

$$\begin{bmatrix} u_{dh}^{\hat{r}}(t) \\ u_{qh}^{\hat{r}}(t) \end{bmatrix} = \begin{bmatrix} V_h & 0 & -V_h & 0 \\ 0 & V_h & 0 & -V_h \end{bmatrix} \begin{matrix} t \in A \\ t \in B \\ t \in C \\ t \in D \end{matrix} \quad (5.9)$$

where $t \in A, B, C, D$ means the time in stages A, B, C, D respectively. At stages of A and C, square wave voltages are injected to the estimated positive and negative d -axis respectively, while at stages of B and D, square wave voltages are injected to the estimated positive and negative q -axis respectively. The current variations in each injection interval can be obtained from (5.4) and (5.9) as

$$\begin{bmatrix} \Delta \hat{i}_{dh}^r(kT_s) \\ \Delta \hat{i}_{qh}^r(kT_s) \end{bmatrix} = \begin{bmatrix} \Delta I_d & \Delta I_\varepsilon & -\Delta I_d & -\Delta I_\varepsilon \\ \Delta I_\varepsilon & \Delta I_q & -\Delta I_\varepsilon & -\Delta I_q \\ t \in A & t \in B & t \in C & t \in D \end{bmatrix} \quad (5.10)$$

where

$$\begin{aligned} \Delta I_d &= \frac{V_h T_s}{2L_{dh}L_{qh}} \left[L_{qh} + L_{dh} + (L_{qh} - L_{dh}) \cos(2(\theta_r - \hat{\theta}_r)) \right], \\ \Delta I_q &= \frac{V_h T_s}{2L_{dh}L_{qh}} \left[L_{qh} + L_{dh} - (L_{qh} - L_{dh}) \cos(2(\theta_r - \hat{\theta}_r)) \right], \\ \Delta I_\varepsilon &= \frac{V_h T_s}{2L_{dh}L_{qh}} (L_{qh} - L_{dh}) \sin(2(\theta_r - \hat{\theta}_r)) \end{aligned} \quad (5.11)$$

5.3.2 Signal processing for sensorless rotor position estimation and HF inductances identification

It can be seen from Fig. 5.3 that the proposed injection sequence is similar to a HF rotating voltage injection. The injected HF voltage can be seen as a HF rotating voltage $V_h e^{j(\omega_h t + \hat{\theta}_r)}$ that is injected into the stator winding at four specific sampling points, i.e. $\omega_h t = 0, \frac{\pi}{2}, \pi, \frac{3\pi}{2}$ where ω_h is the equivalent injection frequency and can be calculated as $\frac{\pi}{2T_s}$. The selected injection interval should be small enough to guarantee high dynamic performance.

The proposed injection voltage is synthesized by a space vector pulse width modulation (PWM) waveform which has a fixed switching frequency. The minimum injection interval is half of the PWM period if the PWM registers can be updated at both the peak and bottom of the PWM waveform. Higher injection frequency also requires higher injection amplitude. As the PWM registers can only take effect at the start of the next PWM period, there is a delay between the voltage reference and the actual generated voltage. Such delay is compensated by triggering current sampling about 1.5 PWM cycles after the square wave voltage is injected.

With the proposed injection method, the rotor position can be estimated by controlling $\Delta I_\varepsilon = 0$ with the demodulation method in Fig. 5.4. The input selection unit selects the signal ΔI_ε according to different injection stages.

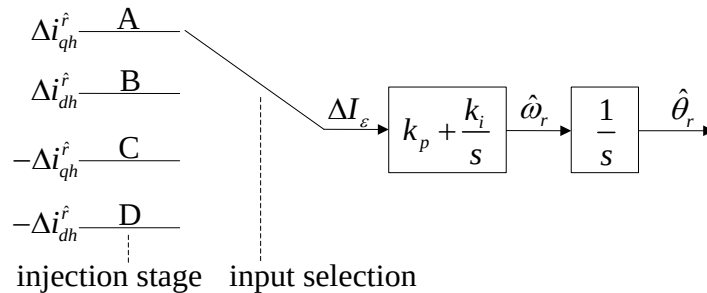


Fig. 5.4. Position demodulation for the proposed injection method.

The rotor position observer input signal ΔI_ε in Fig. 5.4 for the proposed method has the same expression as $\Delta i_{qh}^{\hat{r}}(-1)^{k+1}$ in (5.6) for the conventional method when their injection interval and voltage amplitudes are equal. At each injection stage, the error signal ΔI_ε to the position observer can be updated, which is also the same as the conventional method. The input error will directly switch to a different one if the injection state changes. There will be discontinuity in the input error if the actual and estimated rotor positions change significantly during a single injection stage. At standstill and low speed, the rotor position movement is relatively small due to a short injection interval. The error selection stage acts as a discrete sampling block and can be modelled by a zero-order holder.

Tuning of the position observer parameters should take the sampling interval and injection amplitude into consideration by substituting the input selection stage with (15). Generally, a shorter sampling interval will result in high bandwidth due to a smaller delay introduced by the selection block. It is observed from (5.11) that the estimation error caused by the cross-coupling effect still exists for the proposed method, and the compensation method by using fitted error can be used as the cross-coupling effect is not dependent on the specific injection signal type.

At the same time, with the measured current variations $\Delta I_d, \Delta I_q, \Delta I_\varepsilon$, the dq -axis HF inductances when cross-coupling effect is not considered (which will be considered later in section 5.3.4), can be obtained from (5.11) as

$$L_{dh} = f_1(\Delta I_d, \Delta I_q, \Delta I_\varepsilon), L_{qh} = f_2(\Delta I_d, \Delta I_q, \Delta I_\varepsilon) \quad (5.12)$$

where f_1, f_2 are the functions of $\Delta I_d, \Delta I_q, \Delta I_\varepsilon$ obtained from (5.11) as

$$\begin{aligned} f_1 &= \frac{V_h T_s}{\frac{\Delta I_d + \Delta I_q}{2} + \sqrt{\left(\frac{\Delta I_d - \Delta I_q}{2}\right)^2 + (\Delta I_\varepsilon)^2}} \\ f_2 &= \frac{V_h T_s}{\frac{\Delta I_d + \Delta I_q}{2} - \sqrt{\left(\frac{\Delta I_d - \Delta I_q}{2}\right)^2 + (\Delta I_\varepsilon)^2}} \end{aligned} \quad (5.13)$$

It can be seen from (5.13) that the obtained HF inductances L_{dh}, L_{qh} are not dependent on the position error $\Delta \theta_r$. Therefore, the calculated results will have higher accuracy compared with those of conventional square wave voltage injection method when there is error between the actual rotor position θ_r and the injection position $\hat{\theta}_r$.

It is also noted that the torque ripple may be increased due to additional injection stages in the estimated q -axis compared to the conventional injection method in the estimated d -axis. The pseudo-random square wave injection method [YPP11] can be introduced to distribute the spectrum characteristics of the torque ripple to multiple peaks by changing the injection amplitude and period randomly on the estimated dq -axes. The noise level caused by HF torque ripples can also be reduced. Detailed research for noise reduction may be carried out as future work.

Calculation of the HF inductances is sensitive to current sampling noise as current differences are used. To reduce the distortions caused by sampling noise, higher injection amplitude and higher sampling resolution are preferred. However, with a higher injection amplitude, the noise level is also increased and the voltage margin is reduced. The current sampling resolution is also confined by the hardware and cost. A feasible solution for improving the robustness to sampling noise is to use the average values of the current differences in several injection intervals. The averaging cycle is chosen considering the noise level and dynamic performance requirement. In this thesis, it is chosen as 3

5.3.3 Torque estimation with HF inductances

The obtained HF inductances are reflections of the saturation levels and flux linkages, and can be used to estimate the torque [MAR20]. Conventionally, the torque is calculated with apparent inductances. However, at standstill and low speed, it is difficult to obtain the apparent inductances from the fundamental model due to low amplitudes of the back EMF signals. As presented in the last section, the HF inductances can be obtained based on HF voltage injection. However, it is difficult to find accurate analytical expressions between the HF inductances and the dq -axis flux linkages. The apparent inductances can be calculated from the measured incremental inductances. A feasible way for obtaining the apparent inductances is to integrate the estimated incremental inductances.

Therefore, in this thesis, the polynomial approximations of both d -axis and q -axis flux linkages are adopted considering a trade-off between the accuracy of approximation and complexity of algorithm [TIN15].

(a) D-axis flux linkage and d-axis HF inductance

For the d -axis flux linkage, as the machine is typically operated with $i_d < 0$ from the maximum torque per ampere strategy, the d -axis flux linkage shows good linearity with current and can be approximated with the measured d -axis HF inductance as

$$\psi_d^{\hat{r}} \approx \psi_{pm} + L_{dh} i_d^{\hat{r}} \quad (5.14)$$

where $i_d^{\hat{r}}$ is the d -axis currents, the superscript “ $\hat{r}$ ” here means the current and flux linkages are obtained in rotor position sensorless control.

(b) Q -axis flux linkage and q -axis HF inductance

The q -axis flux linkage shows higher nonlinearity compared to that of d -axis. A third order polynomial expansion is a good compromise between accuracy and implementation simplicity [TIN15] as

$$\psi_q^{\hat{r}} \approx k_{q1} i_q^{\hat{r}} + k_{q2} (i_q^{\hat{r}})^3 \quad (5.15)$$

where k_{q1} , k_{q2} are the coefficients of the fitted curve. The obtained q -axis HF inductance is the derivative of the q -axis flux linkage as

$$L_{qh} = \frac{\partial \psi_q^{\hat{r}}}{\partial i_q^{\hat{r}}} = k_{q1} + 3k_{q2} (i_q^{\hat{r}})^2 \quad (5.16)$$

From the initial q -axis HF inductance, i.e. $L_{qh0}(i_q^{\hat{r}} = 0)$, and the online measured q -axis HF inductance in (5.12), k_{q1} , k_{q2} can be solved from (5.16) as

$$k_{q1} = L_{qh0}, k_{q2} = \frac{L_{qh} - L_{qh0}}{3(i_q^{\hat{r}})^2} \quad (5.17)$$

Substituting k_{q1} , k_{q2} in (5.16) with (5.17), the q -axis flux linkage can be approximated with q -axis HF inductances as

$$\psi_q^{\hat{r}} = \frac{(L_{qh} + 2L_{qh0})i_q^{\hat{r}}}{3} \quad (5.18)$$

(c) Estimation of torque from dq -axis HF inductances

Substituting (5.14), (5.18) into the general torque equation (5.19) where ψ_d , ψ_q , i_d , i_q are the dq -axis flux linkages and currents [CHE14], the torque can be approximated by the dq -axis HF inductances as (5.20).

$$T_e = \frac{3p}{2} (\psi_d i_q - \psi_q i_d) \quad (5.19)$$

$$T_e = \frac{3p}{2} \left[\psi_{pm} i_q^{\hat{r}} + \left(L_{dh} - \frac{L_{qh} + 2L_{qh0}}{3} \right) i_d^{\hat{r}} i_q^{\hat{r}} \right] \quad (5.20)$$

5.3.4 The impact of cross-coupling effect on position and torque estimation

In this section, the impacts of the cross-coupling effect on sensorless rotor position estimation and torque estimation will be analyzed and the corresponding compensation method will be introduced. Then, simulations are carried out to verify effectiveness of the proposed torque estimation method accounting for the cross-coupling effect.

① Impacts of the cross-coupling effect on position and torque estimation

The cross-coupling effect is the mutual coupling between the dq -axes. The voltage equations in the estimated reference frame considering the cross-coupling effect is given by [LIU14c]

$$\begin{bmatrix} \dot{i}_{dh}^{\hat{r}} \\ \dot{i}_{qh}^{\hat{r}} \end{bmatrix} = \frac{1}{(L_{\Sigma h} + L_{nh})(L_{\Sigma h} - L_{nh})} \begin{bmatrix} L_{\Sigma h} + L_{nh} \cos(2\Delta\theta_r - 2\Delta\theta_m) & L_{nh} \sin(2\Delta\theta_r - 2\Delta\theta_m) \\ L_{nh} \sin(2\Delta\theta_r - 2\Delta\theta_m) & L_{\Sigma h} - L_{nh} \cos(2\Delta\theta_r - 2\Delta\theta_m) \end{bmatrix} \int \begin{bmatrix} u_{dh}^{\hat{r}} \\ u_{qh}^{\hat{r}} \end{bmatrix} dt \quad (5.21)$$

where $\Delta\theta_m$ is the phase shift caused by the cross-coupling effect, L_{dqh} is the mutual HF inductance,

$$\begin{aligned} L_{\Sigma h} &= \frac{L_{dh} + L_{qh}}{2}, L_{nh} = \sqrt{\left(\frac{L_{dh} - L_{qh}}{2}\right)^2 + L_{dqh}^2}, \\ \Delta\theta_m &= 0.5 \tan^{-1} \left(\frac{2L_{dqh}}{L_{qh} - L_{dh}} \right) \end{aligned} \quad (5.22)$$

With the proposed injection sequence, the error signal ΔI_ε considering the cross-coupling effect is obtained from (5.9), (5.10), (5.21) as

$$\Delta I_\varepsilon = \frac{V_h T_s L_{nh}}{L_{dh} L_{qh} - L_{dqh}^2} \sin(2\Delta\theta_r - 2\Delta\theta_m) \quad (5.23)$$

The position can also be estimated by controlling $\Delta I_\varepsilon = 0$ with the position observer in Fig. 5.4, i.e.

$$\hat{\theta}_r = \theta_r - \Delta\theta_m \quad (5.24)$$

As can be seen, there is an estimation error $\Delta\theta_m$ caused by the cross-coupling effect. This error will result in errors in the current control. To eliminate the position error, an offline measurement is widely used [LI09], [WAN19b] as presented in chapter 4. By simulation and experiments, the estimation error $\Delta\theta_m$ is approximately linear function of the q -axis current as [WAN19]

$$\Delta\theta_m = k_c i_q \quad (5.25)$$

where k_c is the coefficient. Thus, by offline measuring the position error with different load conditions using an encoder and then fitting the compensation curve, the position estimation error can be compensated. The measured data for the test machine shown in Table A1.1 Appendix 1 are plotted in Fig. 5.5 under different load with maximum torque per ampere strategy used. According to (5.25), a compensation angle $-0.16i_q$ can be added to the estimated position, i.e. $k_c = -0.16$.

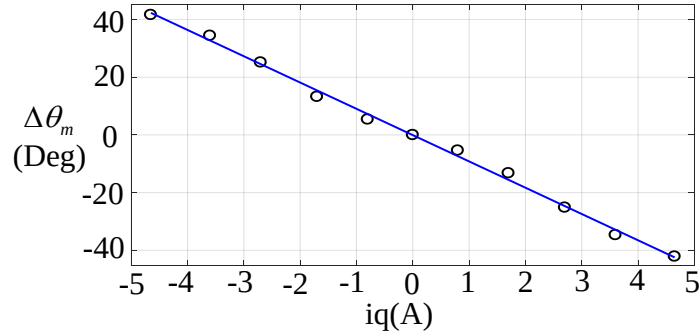


Fig. 5.5. Measured position error at different q -axis current.

The cross-coupling effect not only affects the position estimation, but also causes errors in the calculated HF inductances. The HF inductances considering the cross-coupling effect can be obtained from the current variations $\Delta I_d, \Delta I_q, \Delta I_\varepsilon$ according to (5.13), (5.21), (5.22) as

$$\begin{aligned} L_{dh} &= \frac{1}{2} \left(f_1 + f_2 - \frac{f_2 - f_1}{\sqrt{1 + \tan^2(2k_c i_q)}} \right) \\ L_{qh} &= \frac{1}{2} \left(f_1 + f_2 + \frac{f_2 - f_1}{\sqrt{1 + \tan^2(2k_c i_q)}} \right) \\ L_{dqh} &= \frac{\tan(2k_c i_q)(f_2 - f_1)}{2\sqrt{1 + \tan^2(2k_c i_q)}} \end{aligned} \quad (5.26)$$

where f_1, f_2 have been defined in (5.13) and are calculated with the measured current variations. The additional torque component caused by cross-coupling effect should also be considered especially at high load torque, and can be expressed with mutual HF inductance as [LEE17]

$$T_{cross} = \frac{3p}{2} L_{dqh} (i_q^2 - i_d^2) \quad (5.27)$$

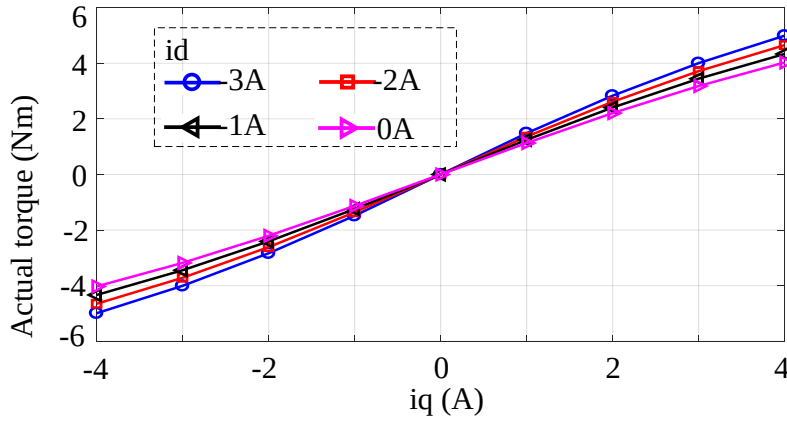
Therefore, from (5.20) and (5.27), the torque, accounting for the cross-coupling effect, can be calculated from the HF inductances by

$$T_{est} = \frac{3p}{2} \left[\psi_{pm} \hat{i}_q + \left(L_{dh} - \frac{L_{qh} + 2L_{qh0}}{3} \right) \hat{i}_d \hat{i}_q + L_{dqh} (i_q^2 - i_d^2) \right] \quad (5.28)$$

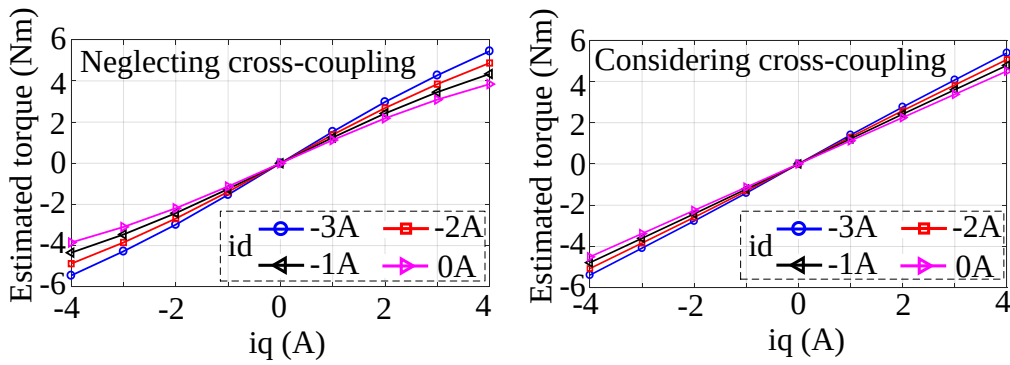
② Simulation evaluation

FE simulation is carried out to verify accuracy of the torque estimation using (5.28). The HF inductances are predicted from FE simulation [LI09]. The actual torque T_e accounting for the cross-coupling effect [CHE14] can be predicted from the general torque equation (5.19), and the estimated torque T_{est} are calculated with (5.28) from the proposed method. The actual torque T_e , estimated torque T_{est} and estimation error ε_{est} defined in (5.29) without and with considering the cross-coupling effect are shown in Fig. 5.6.

$$\varepsilon_{est} = \frac{T_{est} - T_e}{T_e} * 100\% \quad (5.29)$$



(a)



(b)

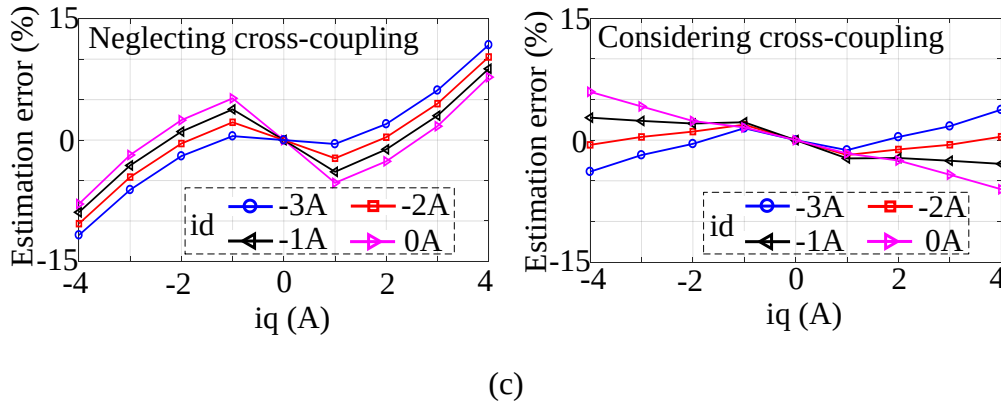


Fig. 5.6. Torque performance with the proposed method using FE predicted HF inductances. (a) Actual torque. (b) Estimated torque. (c) Estimation error.

As can be seen from Figs. 5.6(b) and (c), the torque estimation errors without considering the torque contributed by the cross-coupling effect can rise up to 12%, and can be significantly reduced accounting for the cross-coupling effect although minor errors (below 5%) still remain due to imperfections in the fitted curve between the HF inductances and flux linkages. From the comparison above, the cross-coupling effect is not negligible for torque estimation and by accounting for the cross-coupling effect, the proposed method with HF inductances is capable of estimation the torque for this machine.

5.4 Experimental Verification

Experiments are carried out on a dSPACE platform with a sampling board DS2004, as shown in Fig. 5.7, to verify the effectiveness of the proposed method. Parameters of the test IPMSM are listed in Appendix I. A direct current generator, which is connected to the test machine, acts as a load machine by outputting the generated voltage to an external adjustable resistor. The torque is measured with a torque sensor installed between the test and load machines. An incremental encoder of 1024 pulses per revolution is used to obtain the rotor position θ_r . The dc-bus voltage is 150V, and the injected voltage amplitude is chosen as 20V with an injection interval of 500 μ s considering the current sampling accuracy and the saliency level. The PWM switching frequency is 10kHz. Two external Hall-effect current sensors are installed on an analog conditioning board for sampling the phase currents. Resolution of the analog to digital converter for sampling current is 16bits and the conversion time is 800ns. The current sampling is triggered at 1.5 PWM cycles after the voltage pulse is injected to compensate for the voltage synthesis delay. An anti-aliasing filter of 5kHz bandwidth is used to filter the sampling noise.

The current control loop bandwidth is set as $\omega_{cb} = 200 * 2\pi$ rad/s. The current PI coefficients are then set as $k_p = 50.27, k_i = 7539$ based on the set current loop bandwidth and the machine parameters in Appendix 1 [WAN14]. The position observer parameters for position estimation are chosen as $k_p = 500, k_i = 1000$.

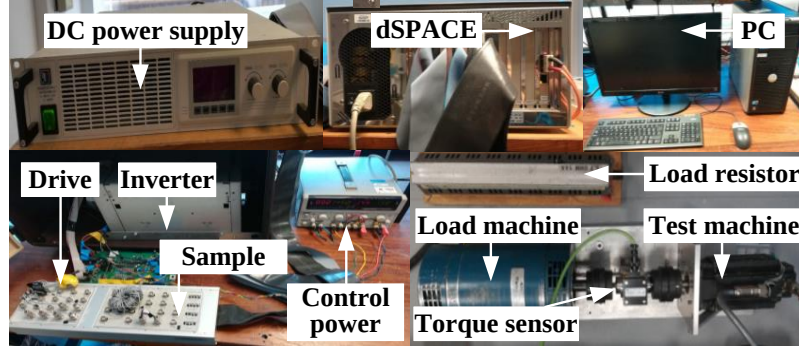
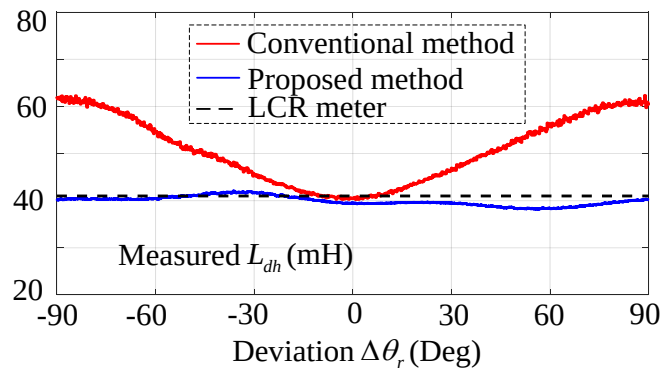


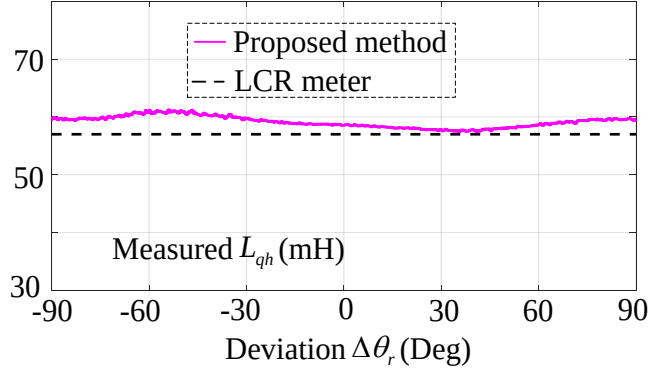
Fig. 5.7. Experimental platform.

5.4.1 Steady state performance

The proposed square wave voltage injection method can measure the d -axis and q -axis HF inductances independently of the rotor position as shown by (5.12), (5.13), while the conventional method can only measure the d -axis HF inductance when there is no offset between the injection position and the actual rotor position as shown by (5.7), (5.8). A simple experiment is carried out to compare the measured results with the two methods under zero load current when the cross-coupling inductance is negligible. The machine is operated at 3Hz (60rpm) with the actual rotor position provided by an encoder. The HF inductances are measured by injecting the conventional and proposed square wave voltages respectively to the position that is deviated from the actual rotor position by $\Delta\theta_r$. The measured d - and q -axis HF inductances for the two methods are shown in Figs. 5.8(a) and 5.8(b), respectively.



(a)



(b)

Fig. 5.8. Comparison of the measured HF inductances with the conventional and proposed methods under different deviations to the actual rotor position. (a) L_{dh} . (b) L_{qh} .

As can be seen from Fig. 5.8(a), the measured d -axis HF inductance with the conventional method varies with the deviation between the injected and actual rotor positions, while the measured d -axis HF inductance with the proposed method does not vary with the deviation and is approximately 40mH except some minor ripples caused by the spatial harmonics [KWO20]. The d -axis HF inductance is also measured [WAN14] by a LCR meter as shown by the dashed line in Fig. 5.8(a), which can be a reference to verify the proposed method. Detailed derivation of measuring the initial dq -axis inductances with LCR meter are attached in Appendix 5.

With the proposed method, the q -axis HF inductance can also be measured, independent of the error between the injected and actual rotor positions as shown in Fig. 5.8(b), which shows a good agreement with the measured result by LCR meter (about 57mH). Whereas with the conventional method, the q -axis HF inductance cannot be measured as analyzed in section II. C. Therefore, the proposed method with HF square wave voltage injection in the estimate dq -axes is capable of measuring both the dq -axis HF inductances independently of the error between the injected and actual rotor positions.

To verify the effectiveness of the proposed method at steady state, the torque estimation performance with position sensor is firstly carried out. The machine is operated at 60rpm (3Hz) under sensor mode with the rotor position provided by the encoder. The current references are generated from the maximum torque per ampere control strategy. HF square wave voltages are injected into the real d -axis and q -axis for online measuring the HF inductances under different load torque. The online identified HF inductances accounting for the cross-coupling effect show a good agreement with the FE predicted values as demonstrated in Fig. 5.9(a). The estimated torque using (5.28) with the identified HF inductances are shown in Fig. 5.9(b),

together with the torque estimation error. The PM induced flux linkage, i.e. 0.26Wb is assumed to be fixed in this thesis. The real torque is measured with the torque sensor. As can be seen, the torque estimation error with the online identified HF inductances is below 5%, which verifies the effectiveness of the proposed torque estimation strategy.

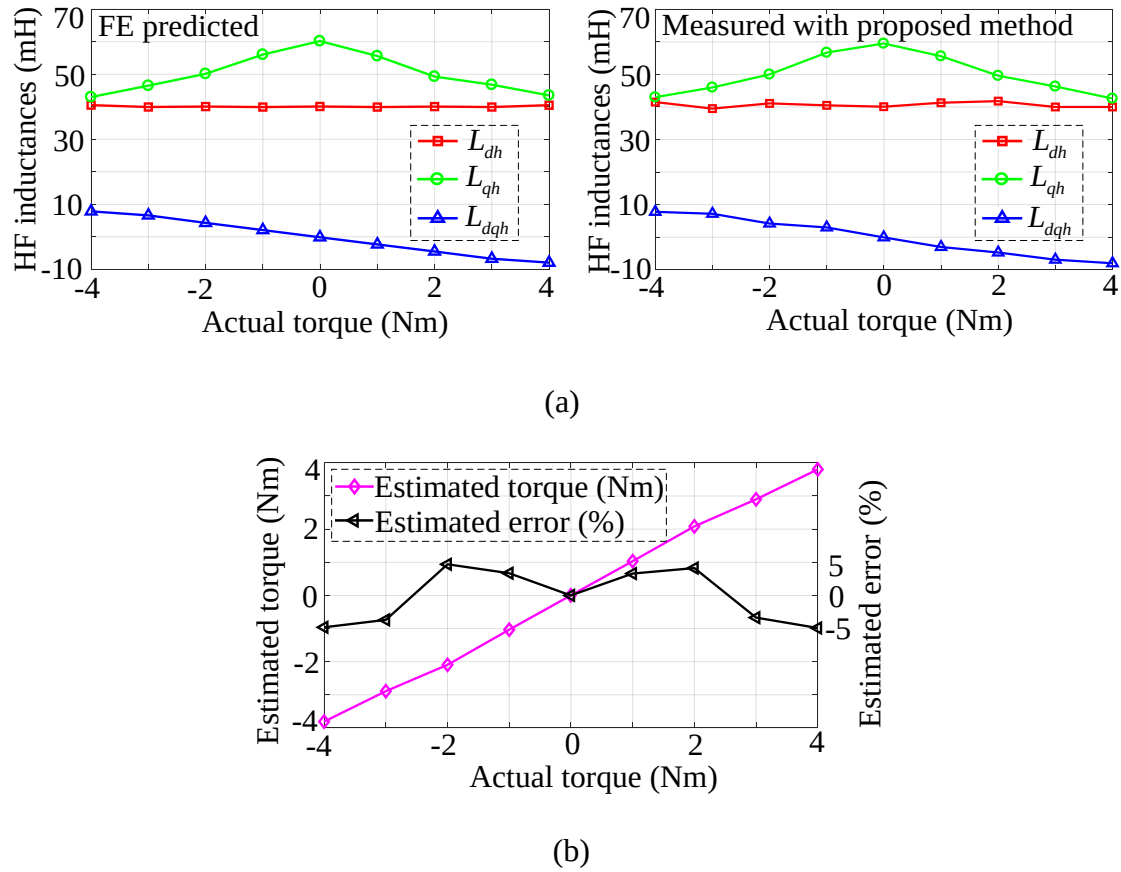


Fig. 5.9. Torque estimation performance from identified HF inductances with position sensor under different torque. (a) HF inductances identification performance. (b) Torque estimation performance.

Then, the sensorless rotor position estimation performance of the proposed method is evaluated. The machine is sensorless operated at 60rpm under different load torque with cross-coupling effect compensation. The position error is obtained from the difference between the actual rotor position and the estimated position as shown by the blue line in Fig. 5.10(b). As can be seen, the position errors under different load torques are below 5° , which verifies the sensorless rotor position estimation performance of the proposed method. The measured HF inductances under sensorless mode are shown in Fig. 5.10(a), which show a good agreement with the results measured with position sensor in Fig. 5.9(a). The torque estimation errors with the sensorless identified HF inductances are shown in Fig. 5.10(b).

For comparison, the conventional torque estimation method by assuming fixed dq -axis inductances, i.e. the initial inductances listed in the machine parameters, are also carried out. The torque estimation error for the conventional method without considering inductance variations can reach up to 10% at high current, while the estimated torque error for the proposed method is below 5%. Therefore, the proposed method shows better performance in torque estimation with less than 5% error and position estimation with less than 5° error under different load torque.

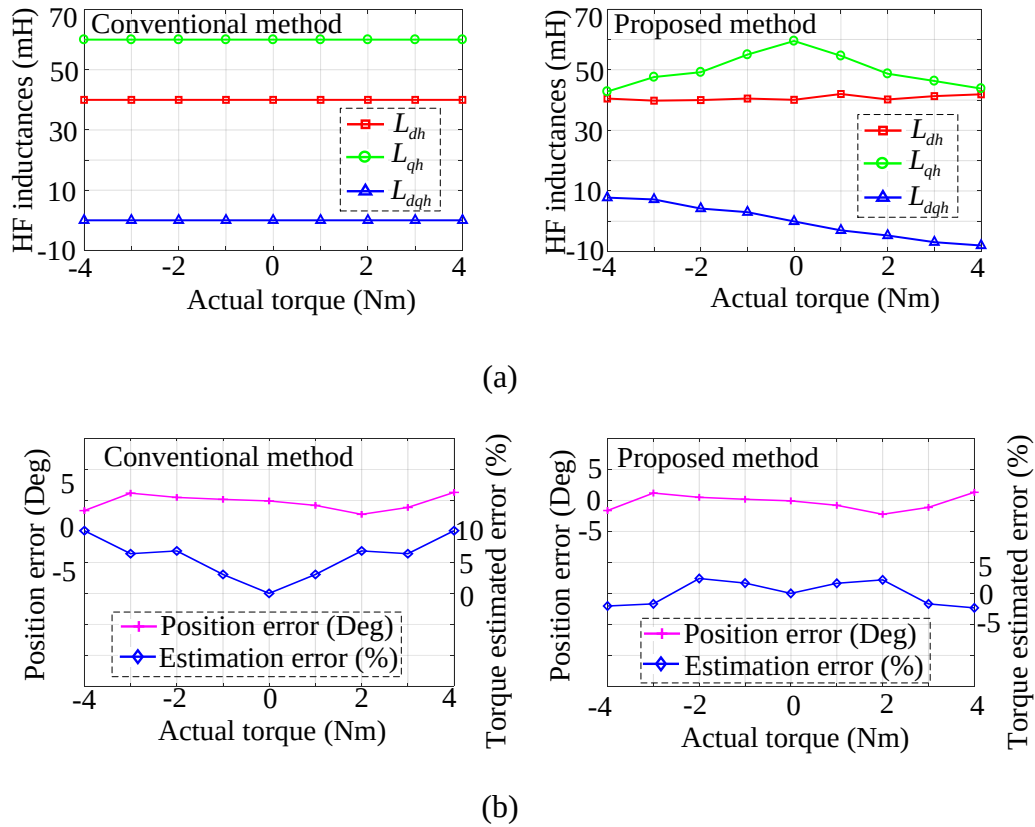


Fig. 5.10. Sensorless rotor position estimation performance of the proposed method and comparison of the torque estimation performances between the conventional and proposed methods. (a) HF inductances for estimating torque. (b) Position and torque estimation errors.

The sensorless torque estimation performance with the proposed method at different i_d and i_q currents are also evaluated. The HF inductances can be sensorless identified with the proposed method and used for estimating the torque. The online sensorless identified HF inductances at different i_d ($i_d \leq 0$), i_q are shown in Fig. 5.11. The identified results are also compared with the incremental inductances predicted by FEA in Fig. 3.3. It can be seen that the online identified HF inductances show a good agreement with the FE predicted values.

However, there are still some minor discrepancies which may be caused by inaccuracies in the FE model or properties of the magnetic materials.

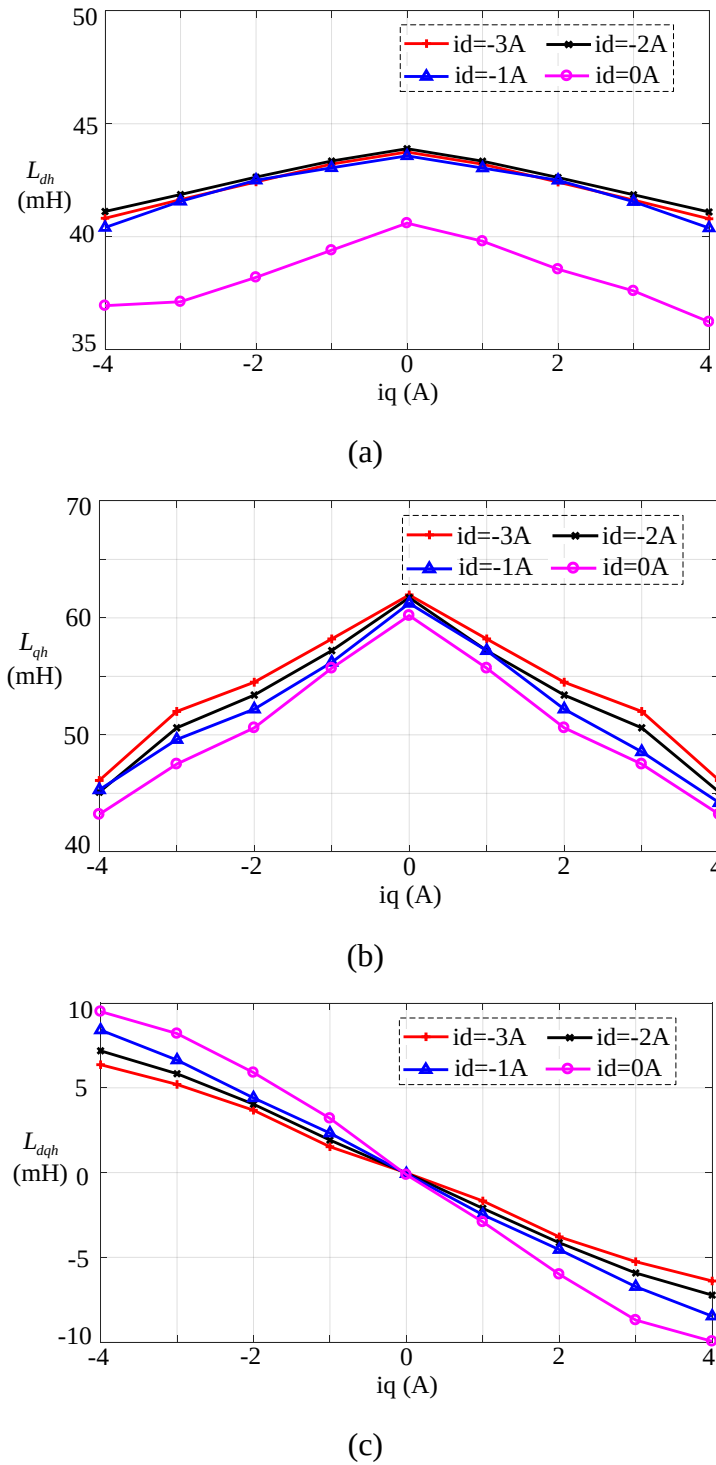


Fig. 5.11. Sensorless online identified HF inductances with the proposed method. (a) L_{dh} . (b) L_{qh} . (c) L_{dqh} .

For comparison, the inductances used for estimating the torque at $i_d = 0A$ with the conventional method and the proposed method are also shown in Fig. 5.12(a). Figs. 5.12(b)

and 5.12(c) show the estimated torque and errors for the conventional method without using identified HF inductances and the proposed method.

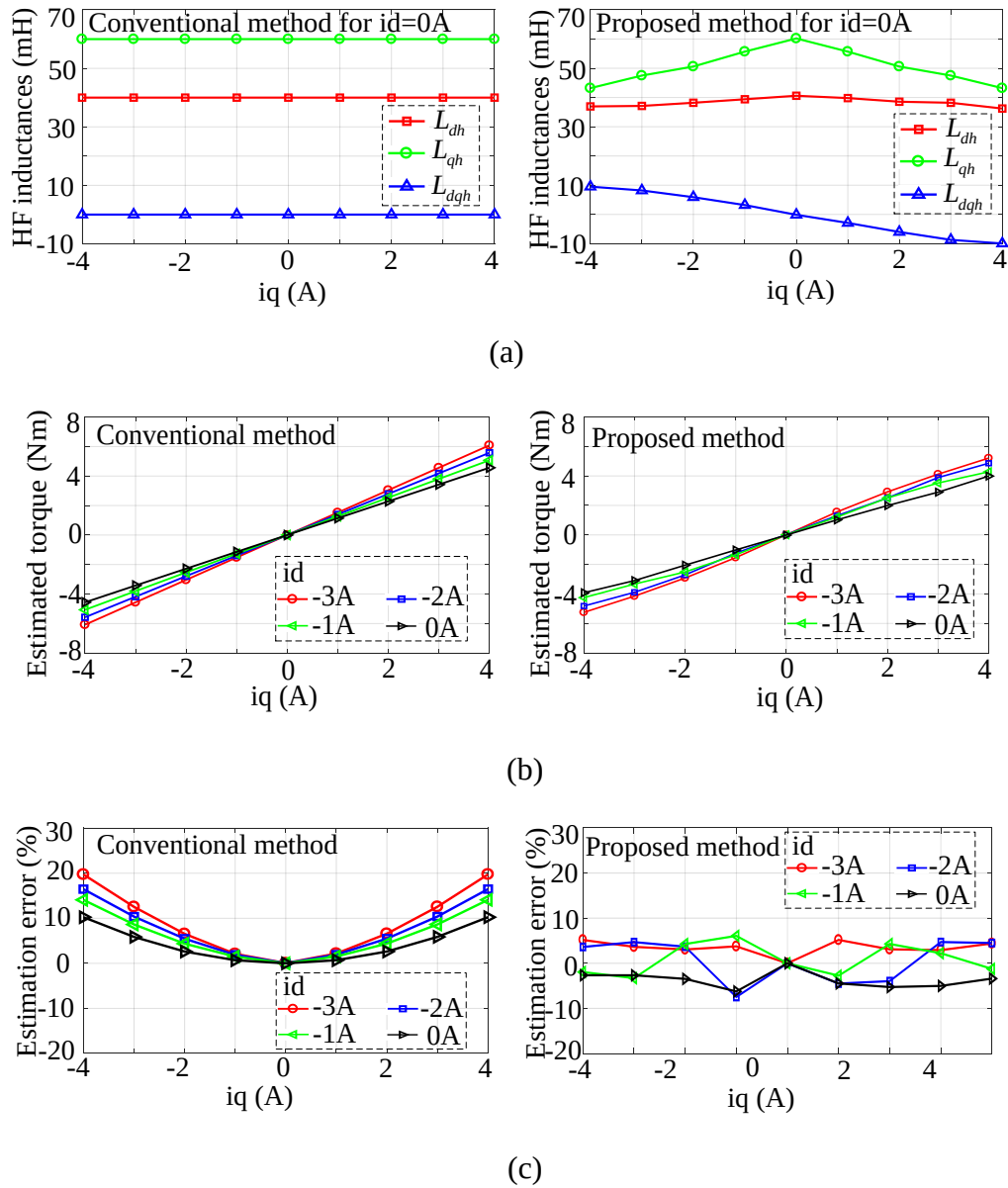


Fig. 5.12. Comparison of steady state torque estimation performances without and with the proposed method. (a) HF inductances. (b) Estimated torque. (b) Estimated torque error.

As can be seen the torque estimation errors are significantly reduced for the proposed method. However, it is also noted that the proposed torque estimation method based on HF inductances only shows significant superiority at high load currents as at low currents the conventional method without HF inductances identification also has less than 5% error as shown in the low current region. i.e., -2A to 2A, in Fig. 5.12(c) for the conventional method. This is due to that at high currents, the magnetic materials are saturated and the flux linkages and the HF

inductances show higher nonlinearities with the currents, and thus, the method with online HF inductances identification can improve torque estimation accuracy.

5.4.2 Dynamic performance

Sensorless startup with the proposed method is tested. The machine is sensorless started from standstill at 1s with the proposed method. The speed reference is ramped up to 60rpm over 1s. The load torque generated by the dc generator is gradually increased with speed. The speed, phase current, estimated position and position error are shown in Fig. 5.13(a). The identified HF inductances with the proposed method are shown in Fig. 5.13(b), together with the HF inductances used for torque estimation with the conventional method. The estimated torque and estimated errors for the conventional method without considering the inductance variations and the proposed method using online identified HF inductances are shown in Fig. 5.13(c).

As shown in Fig. 5.13(c), the actual load torque is increased with the rotor speed as the output voltage of the dc generator is proportional to the rotation speed. After a transient rise to about 0.2Nm during the startup, the torque estimation error for the proposed method is decreased to approximately zero after startup. The torque estimation error for conventional method without using the identified HF inductances is increased to about 0.4Nm, i.e. 10% of the rated load torque and stays at this level after startup.

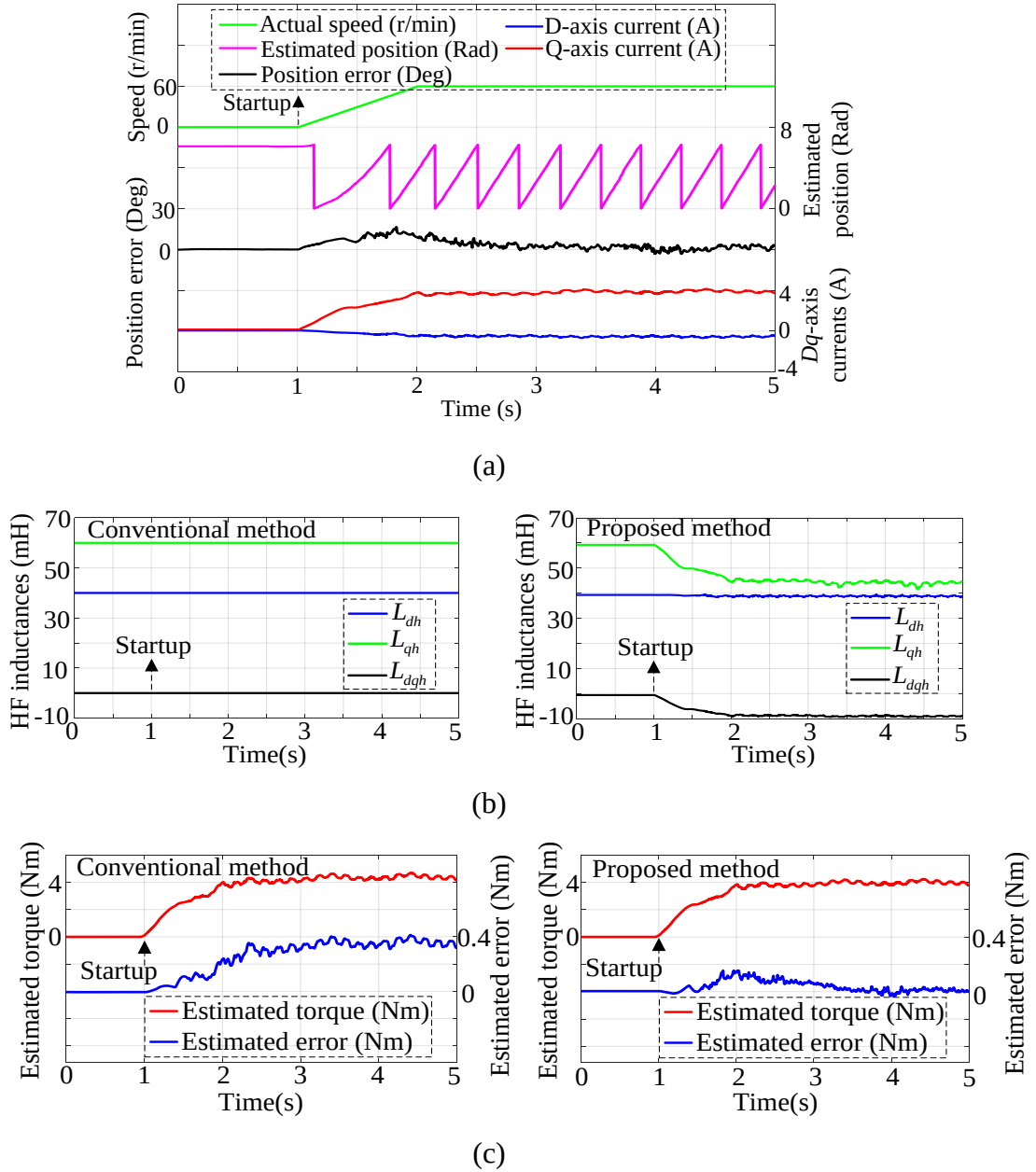


Fig. 5.13. Sensorless startup with the proposed method by experiment. (a) Speed, position estimation performance, and dq -axis currents. (b) Inductances for conventional and proposed torque estimation methods. (c) Estimated torque and estimated errors for the conventional and proposed methods.

As the load torque of the test rig is proportional to the speed of the test machine and dc generator, the load torque can only be increased gradually with speed, but not be applied as a step input at standstill. Therefore, simulation is carried to verify sensorless start performance with rated load torque at standstill as shown in Fig. 5.14.

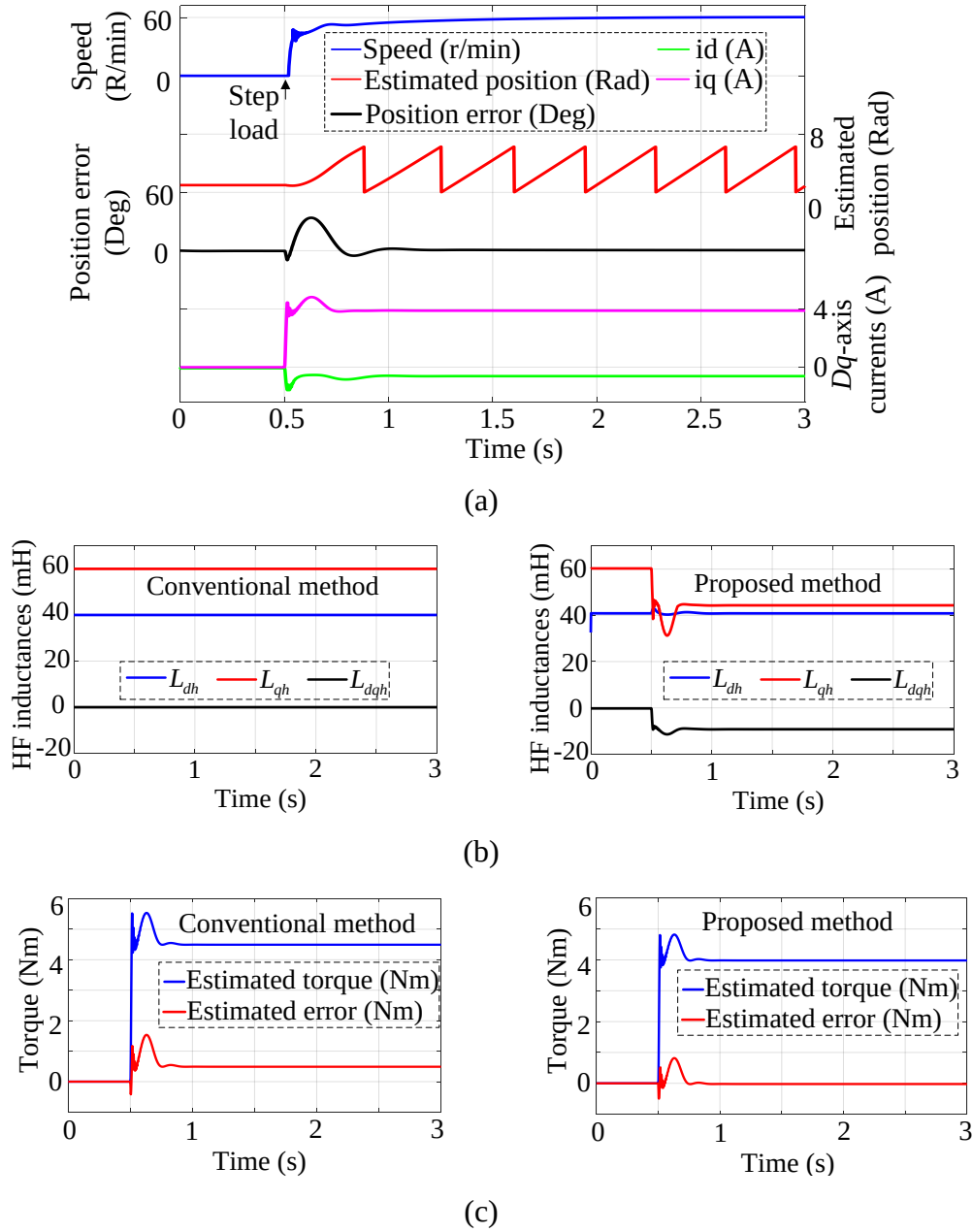
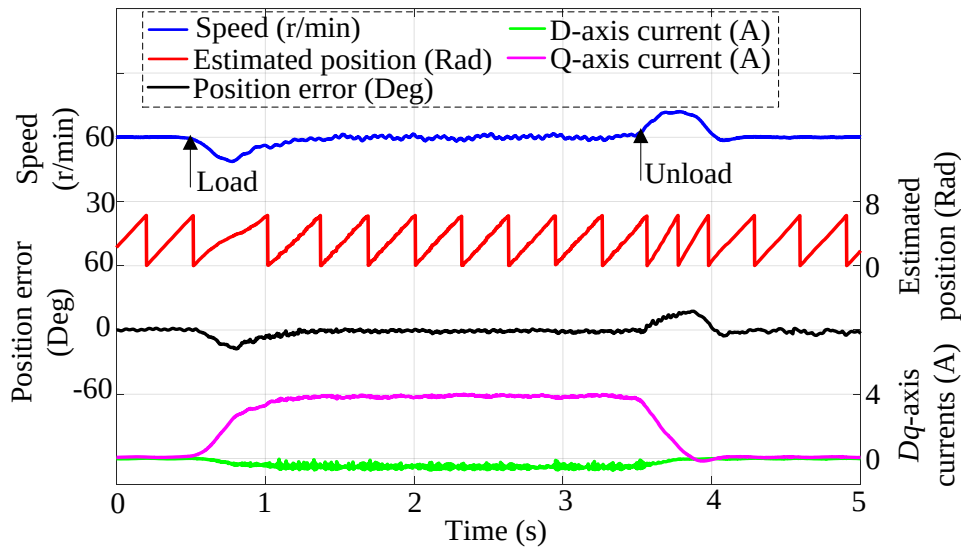


Fig. 5.14. Sensorless startup with the proposed method by simulation. (a) Speed, estimated position, position error, dq -axis currents. (b) Inductances for conventional and proposed torque estimation methods. (c) Estimated torque and estimated error for the conventional and proposed methods.

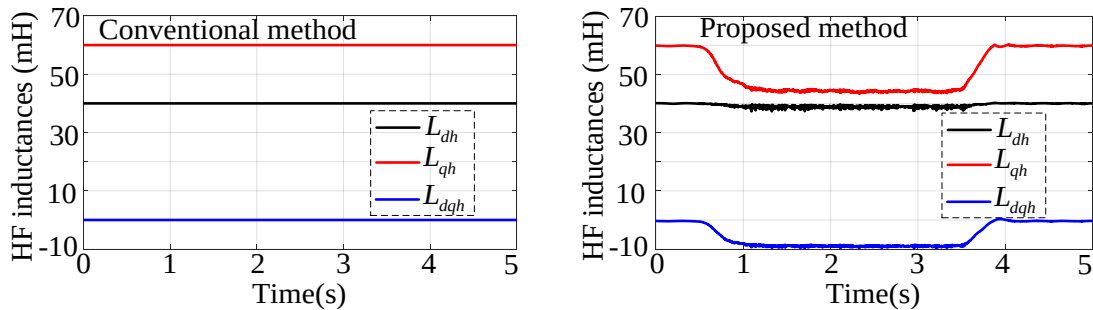
In the simulation, the machine is operated with the proposed sensorless position and torque estimation method. Step load torque (4Nm) and step speed reference (60rpm) are applied simultaneously at 0.5s. The control parameters are set as the same as the experiment in section 5.4.2, i.e. current PI: $k_p = 50.27, k_i = 7539$, position observer: $k_p = 500, k_i = 1000$. As can be seen from Fig. 5.14(a), the position error during startup can reach up to 30 degrees under rated load torque. The torque estimation errors for both the conventional and proposed methods

are also significantly increased with the position error. However, for the conventional method, the maximum torque estimation error can be 1.5Nm, while the maximum error for the proposed is about 0.8Nm. At steady state, the torque error is zero for the proposed method, and 0.5Nm for the conventional method. Therefore, compared to torque estimation without using online identified HF inductances, the torque estimation accuracy is improved with the proposed method.

Load transient experiments are also carried out. The machine is sensorless operated at 60rpm (3Hz) under no load with the proposed method. The load torque is increased to 4Nm at 0.5s, and after several seconds, is unloaded to zero at 3.5s. The speed, estimated position, position error, dq -axis currents are shown in Fig. 5.15(a). With the proposed method, the position estimation errors during load transients are below 18 degrees. The sensorless identified HF inductances with the proposed method are shown in Fig. 5.15(b), together with the HF inductances used for torque estimation with the conventional method.



(a)



(b)

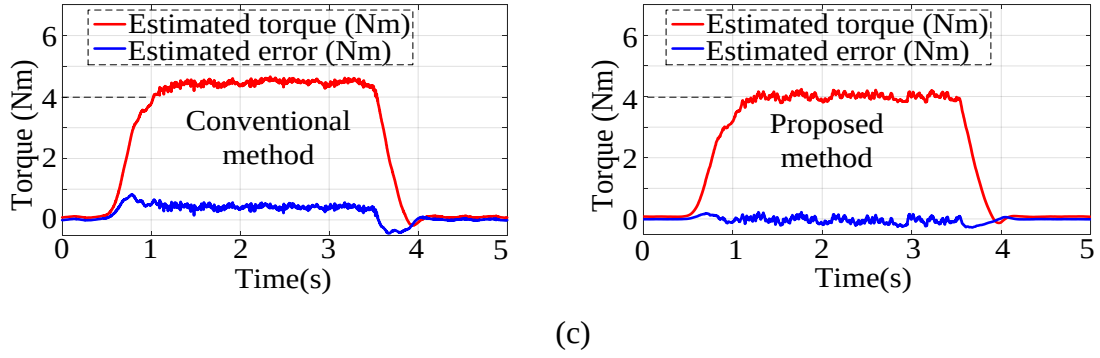


Fig. 5.15. Sensorless position and torque estimation performance under load transients (a) Speed, estimated position, position error and dq-axis currents. (b) HF Inductances for conventional and proposed torque estimation methods. (c) Estimated torque and estimated errors for the conventional and proposed methods.

The maximum torque estimation errors for the proposed method are less than 0.3Nm during the transients, while the torque estimation errors for conventional method without using the identified HF inductances can rise up to about 0.8Nm during the load transients, i.e. 20% of the rated load torque. At steady state, the torque estimated error is nearly zero for the proposed method, and about 0.4Nm for the conventional method. From the above experiments, the proposed method is capable of estimating torque with improved accuracy at standstill and low speed. It is also noted that with the increase of speed to medium and high speed ranges, the rotor position estimation performance of the proposed method may be deteriorated due to difficulty in separation of the HF and fundamental currents. Then, the fundamental model based sensorless rotor position estimation methods can take over with various parameter identification methods [YOO09], [ICH06], [LIU11] based on the fundamental model which can be employed for guaranteeing the torque estimation performance.

A step speed response experiment is carried out to evaluate the speed bandwidth. A step speed reference of 60rpm is applied to the machine. The speed and position estimation performance are shown in Fig. 5.16(a). The HF inductances and torque estimation performance are shown in Figs. 5.16(b), (c).

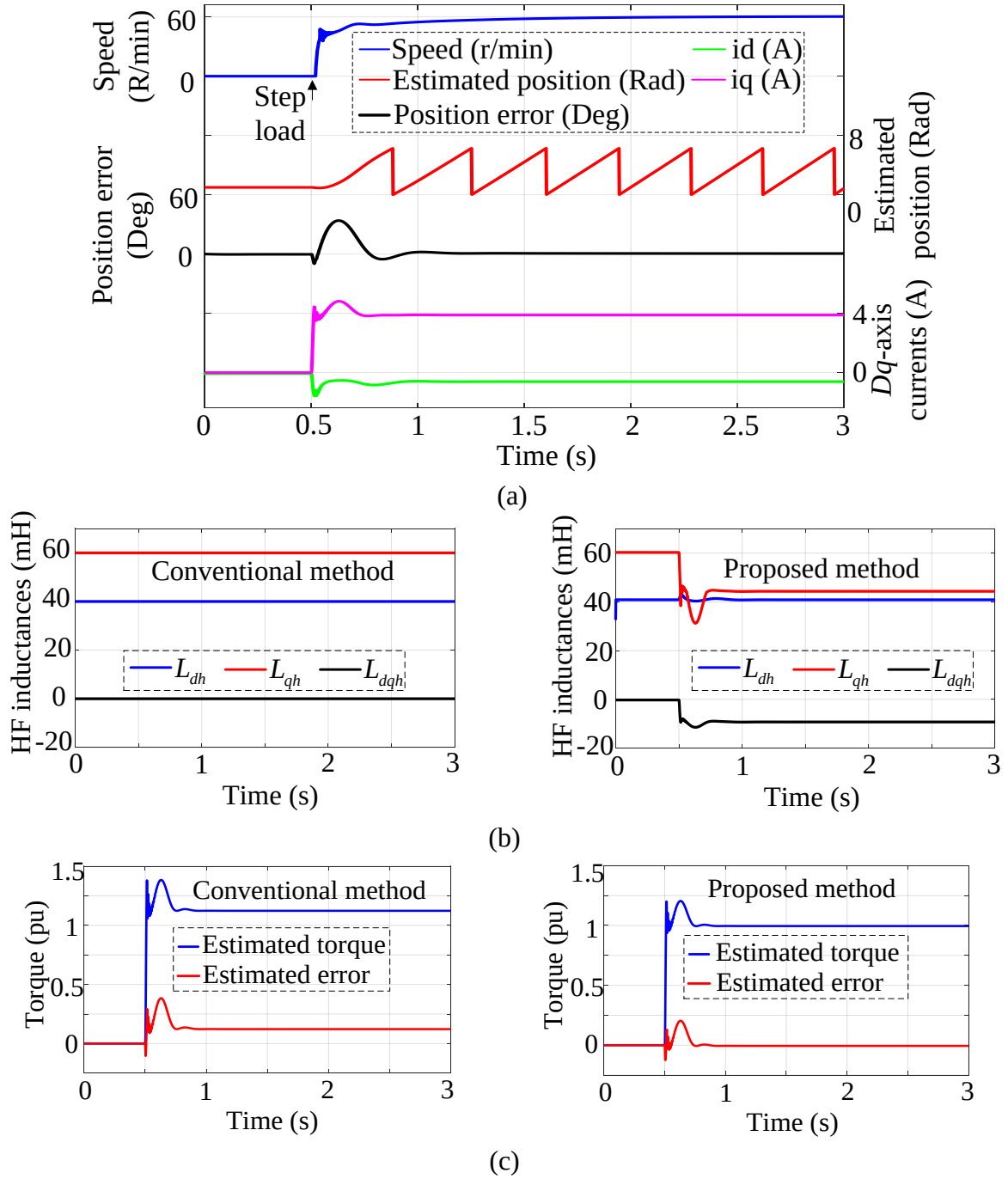


Fig. 5.16 Sensorless position and torque estimation performance under speed step. (a) Speed, estimated position, position error and dq-axis currents. (b) HF Inductances for conventional and proposed torque estimation methods. (c) Estimated torque and estimated errors for the conventional and proposed methods.

It can be seen that the rise time is about 0.4s and the speed band width is approximated as 10Hz. The actual speed bandwidth may change with different injection frequency, load inertia and control parameters.

5.5 Conclusion

In this chapter, the capability of simultaneous sensorless rotor position and torque estimation for the conventional high frequency square wave voltage injection technique is investigated. To enable torque estimation at standstill and low speed with identified high frequency inductances, a method of simultaneous sensorless position and torque estimation is proposed for interior permanent magnet synchronous machine at standstill and low speed by injecting high frequency square wave voltages in the estimated dq -axes. The proposed method estimates the torque using high frequency inductances which are identified from the current variations due to square wave voltage injection. The proposed method is not dependent on the fundamental model, allowing for low cost and fast torque estimation at low speed, and is complimentary with the methods based on the fundamental model which are mainly for torque estimation at medium and high speeds. Accuracy of proposed torque estimation method by HF inductances is verified by simulation. Experiments are also carried out to verify both steady state and dynamic performances of the proposed method.

CHAPTER 6

SENSORLESS PERFORMANCE IMPROVEMENT UNDER LOAD

TRANSIENT AND WHOLE SPEED RANGE SENSORLESS

OPERATION

6.1 Introduction

Position sensors are widely used to provide accurate rotor position for machine control. However, these hardware based sensors increase the system complexity and cost, such as wiring cables and installation accessories. Although the quality of these sensors has been largely improved recently, failures of these sensors still pose big threat to the system safety. Thus, for the purpose of reducing the cost and ensuring the system safety, sensorless control has been investigated for decades [WAN20].

There are generally two kinds of sensorless control methods. For medium and high speeds, back EMF voltages induced by permanent magnets are utilized to track the rotor position, but may become ineffective at standstill and low speed due to deteriorated signal to noise ratio (SNR) of the back EMF signal. At standstill and low speed, saliency variation caused by structure geometry or magnetic saturation is utilized for position tracking. To detect saliency information, continuous high frequency (HF) signal injection is commonly used, including rotating sinusoidal voltage injection [JAN95], pulsating sinusoidal voltage injection [WAN13], and pulsating square wave voltage injection [LAR18]. The rotating voltage injection was proposed first which injects balanced rotating voltage signals into the stationary reference frame (SRF) and extracts the negative sequence current for rotor position estimation. The pulsating voltage injection based sensorless method injects voltage signal to the estimated d-axis and the q-axis HF current amplitude is extracted for position estimation.

The HF currents due to HF voltage injection are usually extracted by the band pass filters. Ideally, there are not interference between the HF current and the fundamental currents. However, under load transient, the step load currents will contain spread frequency spectrum which may be coincided with the injected HF currents. The interference may distort the injected HF currents and result in position estimation error. Hence, for improving the sensorless performance, the impacts of transient load currents should be considered.

A feasible solution is to limit the rising rate of current reference [CON09]. However, it will decrease the response speed. Another method is to separate the load currents from the total current responses by load current prediction technique. In this chapter, the load current prediction technique in HF injection based sensorless control method will be investigated.

The capability of whole speed range sensorless operation is another important issue in actual applications. At standstill and low speed, sensorless methods based on saliency model are used, while at medium and high speed, sensorless methods based on fundamental model are used. The key point for whole speed range sensorless operation is to deal with the switch between different sensorless methods at low speed and high speed. Generally, there are two separate position observers for saliency model and fundamental model. To achieve smooth switchover between different speed ranges, weighting functions are widely used to fuse the estimated speeds and positions from different sensorless algorithms. In such case, two different sets of closed-loop parameters should be tuned, which increases complexity of the sensorless control [IDE06].

To avoid the complexity of two separate position observers, single observer structure which is shared by saliency model and (for low speed) fundamental model (for high speed) is desirable as only one set of closed-loop parameters is required. The input of the shared closed-loop observer is the fusion of the position error signals from saliency model at low speed and fundamental model at medium and high speeds. To share a same PI, the position errors signals from different models should be normalized. In [YAN17], a speed-dependent function is used to blend two position error signals from the saliency model and the back EMF model to form a blended position error signal. A scale factor k_{sal} which is used to normalize the position error signal from the saliency model. However, this factor varies with the working points and is difficult to determine accurately. In this chapter, a novel whole speed sensorless operation method with one closed-loop observer and unity-length position error signals is proposed based on HF rotating voltage injection.

This chapter is organized as follows. Firstly, the impact of transient load currents on sensorless control is analysed and the load current prediction technique is introduced and tested. Then, existing methods for switchover between different sensorless control methods are presented. Finally, the proposed whole speed sensorless operation method is presented and tested.

6.2 Impacts of Load Transient and Load Current Compensation

Conventionally, for HF injection based sensorless control methods, the HF currents are separated by band pass filters. However, this separation is present, at best, for steady-state operation. Ideally, there does not exist interfere between fundamental load currents and HF currents. However, transient current due to sudden change of load often contains spectrum over the entire frequency spectrum. Some of these components will coincide with the HF currents induced by HF voltage injection. In this section, the impacts of the load current transients on HF injection based sensorless control will be investigated and corresponding mitigating methods will be presented.

6.2.1 Impacts of load transient on sensorless position estimation

In this section, the impacts of load current transients on the HF signal injection based sensorless control are analysed.

(a) HF rotating voltage injection based sensorless control

The HF model in the stationary reference frame is given by

$$\begin{bmatrix} i_{\alpha h}^s \\ i_{\beta h}^s \end{bmatrix} = R^{-1}(\theta_r) \begin{bmatrix} L_{dh} & 0 \\ 0 & L_{qh} \end{bmatrix}^{-1} R(\theta_r) \int \begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} dt \quad (6.1)$$

$$R(\theta_r) = \begin{bmatrix} \cos(\theta_r) & \sin(\theta_r) \\ -\sin(\theta_r) & \cos(\theta_r) \end{bmatrix} \quad (6.2)$$

For a HF rotating voltage injection based sensorless control method, sinusoidal voltages $u_{\alpha h}^s, u_{\beta h}^s$ are injected to the stationary reference frame as

$$u_{\alpha h}^s + ju_{\beta h}^s = V_h e^{j\omega_h t}, \theta_h = \omega_h t \quad (6.3)$$

where V_h, ω_h, θ_h are the amplitude, frequency, angle of the injection signal. According to (6.2), the HF currents $i_{\alpha h}^s, i_{\beta h}^s$ can be solved as

$$i_{\alpha h}^s + ji_{\beta h}^s = I_p e^{j(\theta_h - \pi/2)} + I_n e^{-j(\theta_h - 2\theta_r - \pi/2)} \quad (6.4)$$

$$I_p = \frac{V_h (L_{dh} + L_{qh})}{2\omega_h L_{dh} L_{qh}}, I_n = \frac{V_h (L_{qh} - L_{dh})}{2\omega_h L_{dh} L_{qh}} \quad (6.5)$$

The HF currents are composed of positive and negative sequence currents. The synchronous shaft filter (SSF) based position demodulation technique shown in Fig. 6.1 is used to extract the negative sequence HF current. When the stator currents $i_{\alpha\beta}^s$ have step changes due to load transients, there will be negative sequence current output even when the HF voltage is not injected. Such unwanted negative sequence current induced by step change of load current will interfere with the HF currents induced by HF rotating voltage injection.

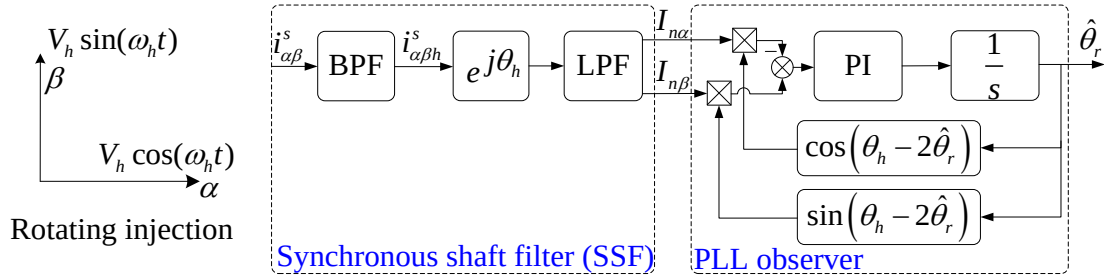


Fig. 6.1. SSF based position estimation method for rotating voltage injection.

In Fig. 6.1, $I_{n\alpha}, I_{n\beta}$ are the negative sequence components in the two phase stationary reference frame as

$$\begin{aligned} I_{n\alpha} &= I_n \cos(2\theta_r + \pi/2) \\ I_{n\beta} &= I_n \sin(2\theta_r + \pi/2) \end{aligned} \quad (6.6)$$

As can be seen, ideally $I_{n\alpha}, I_{n\beta}$ are quadrature signals. Under load current transient, they may be distorted which result in position error.

To evaluate the impact of load current step, simulation is carried out as shown in Fig. 6.2. The HF stator currents are shown in (6.4). The cross-coupling effect is not considered in the simulation for simplicity.

$$\begin{aligned} i_{\alpha h}^s &= I_p \sin(\omega_h t) + I_n \sin(\omega_h t - 2\omega_e t) \\ i_{\beta h}^s &= -I_p \cos(\omega_h t) + I_n \cos(\omega_h t - 2\omega_e t) \end{aligned} \quad (6.7)$$

where the amplitude I_p is set as 0.1A, the NSC amplitude I_n is 0.06A. The injection frequency is 200Hz, i.e. $\omega_h = 2\pi * 200$ rad/s, and the rotor rotating frequency is set as 2Hz, i.e. $\omega_e = 2\pi * 2$ rad/s. In the simulation, the HF currents are sent to the synchronous shaft filter and then the PLL observer for position estimation. The passband of the BPF in the synchronous shaft filter is [150Hz, 250Hz]. The parameters of the PI controller are $k_p = 1000, k_i = 20000$.

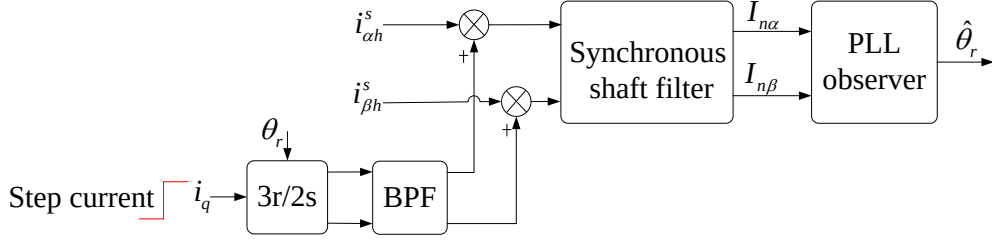


Fig. 6.2. Impact of load current step on sensorless control with HF rotating injection.

In the simulation, different levels of step q -axis current are applied at 0.3s when the rotor position is about 0 degree. The negative sequence currents $I_{n\alpha}$, $I_{n\beta}$ and the position estimation error are shown in Fig. 6.3.

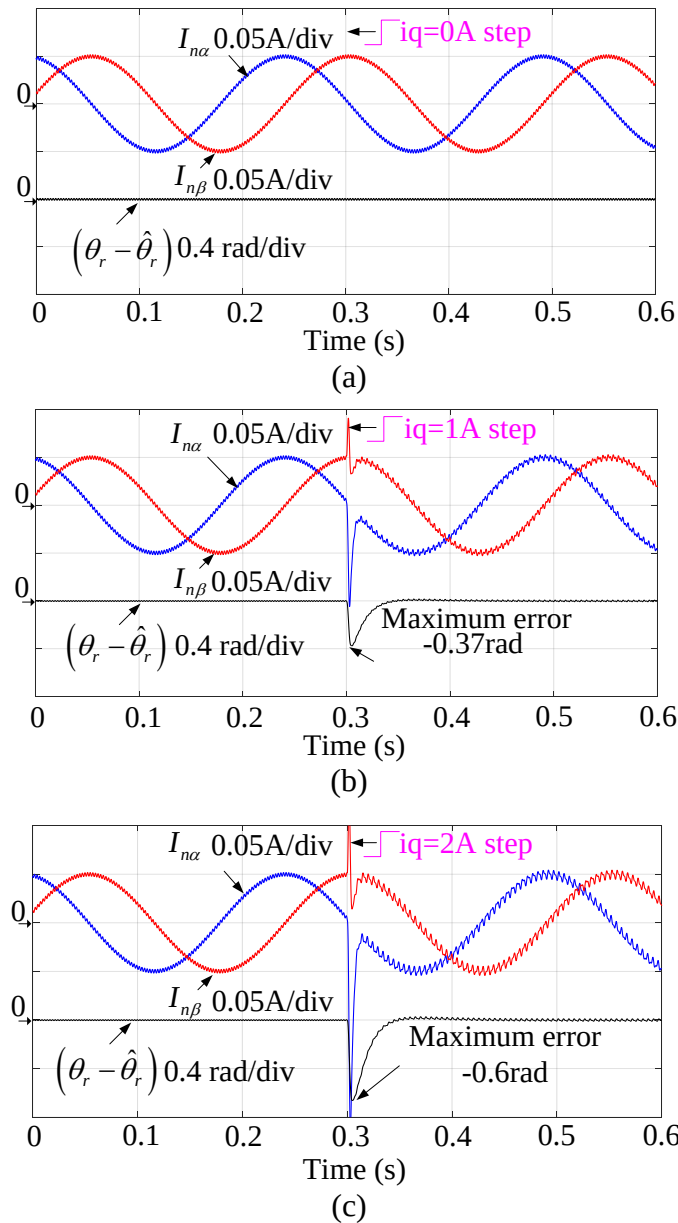


Fig. 6.3. Impact of step load current on rotating voltage injection based sensorless control. (a) No step current. (b) 1A step current on q -axis. (c) 2A step current on q -axis.

It is observed that there is a transient position estimation error after the step load current i_q is applied. Although the lasting time of the transient error is very short ($<0.05s$), it can result in significant decrease in electromagnetic torque, i.e. $(1-\cos(0.6\text{rad}))=18\%$ decrease in the torque for 0.6 rad position error when i_d is set to zero. Therefore, it is vital to suppress the position estimation error during load transients.

In the simulation, it is interesting to find that the maximum position error is dependent on the rotor position when the load current step occurs. In Fig. 6.4, the maximum position errors under 1A step i_q at different rotor positions are plotted. The maximum position error during step current varies with the rotor position and reaches its peak at 60, 180, 300 degrees. This has explained that the HF negative sequence current varies with the rotor position as expressed in (6.4).

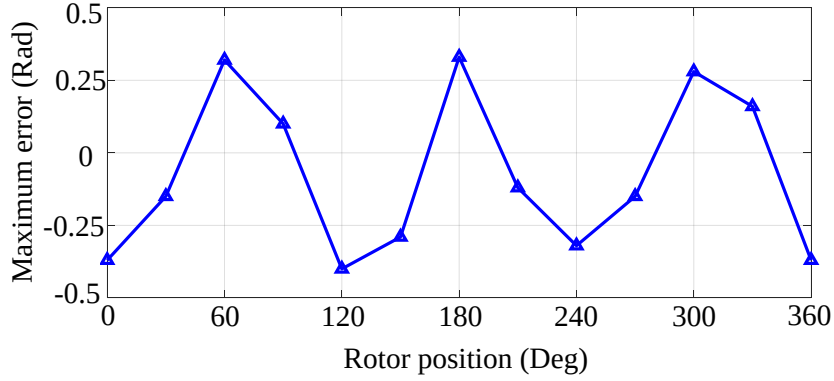


Fig. 6.4. Maximum position error under 1A i_q step at different positions with HF rotating voltage injection.

(b) HF pulsating voltage injection based sensorless control

For the HF pulsating voltage injection based sensorless control method, the HF model used in the estimated synchronous rotating reference frame (SRRF) is given by

$$\begin{bmatrix} \hat{i}_{dh} \\ \hat{i}_{qh} \end{bmatrix} = R^{-1}(\theta_r - \hat{\theta}_r) \begin{bmatrix} L_{dh} & L_{dqh} \\ L_{qdh} & L_{qh} \end{bmatrix}^{-1} R(\theta_r - \hat{\theta}_r) \int \begin{bmatrix} \hat{u}_{dh} \\ \hat{u}_{qh} \end{bmatrix} dt \quad (6.8)$$

where $\hat{u}_{dh}, \hat{u}_{qh}, \hat{i}_{dh}, \hat{i}_{qh}$ are the HF voltages and currents, the superscript “ $\hat{\cdot}$ ” means the estimated synchronous rotating reference frame. When the HF pulsating voltage $\hat{u}_{dh} = V_h \cos(\omega_h t)$ is injected to the estimated d -axis, the estimated dq -axis HF currents can be expressed from (6.8) as

$$\hat{i}_{dh} = I_{dh}^{\hat{r}} \sin(\omega_h t), \hat{i}_{qh} = I_{qh}^{\hat{r}} \sin(\omega_h t) \quad (6.9)$$

where $I_{dh}^{\hat{r}}, I_{qh}^{\hat{r}}$ are the amplitudes of the HF currents and can be extracted by synchronous demodulation [JAN03] as

$$\begin{aligned} I_{dh}^{\hat{r}} &= LPF \left[2\hat{i}_{dh}^{\hat{r}} \sin(\omega_h t) \right] = I_p + I_n \cos(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \\ I_{qh}^{\hat{r}} &= LPF \left[2\hat{i}_{qh}^{\hat{r}} \sin(\omega_h t) \right] = I_n \sin(2(\theta_r - \Delta\theta_m) - 2\hat{\theta}_r) \end{aligned} \quad (6.10)$$

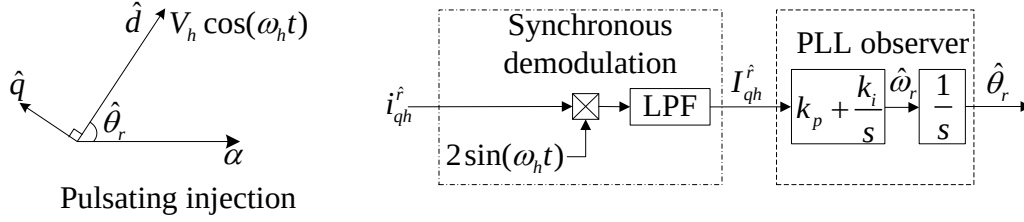


Fig. 6.5. Position demodulation for HF pulsating voltage injection based sensorless control.

The impact of load current step on sinusoidal pulsating voltage injection based sensorless control is analysed with the simulation model in Fig. 6.6. Due to limited bandwidth of the BPF, the step q -axis current cannot be totally filtered out. This will result in distortion of the q -axis HF current amplitude I_{qh} and can bring estimation error.

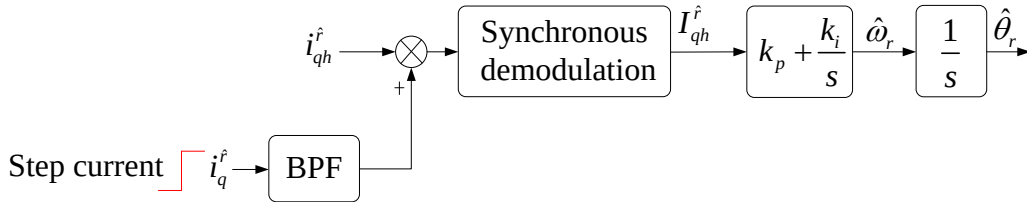
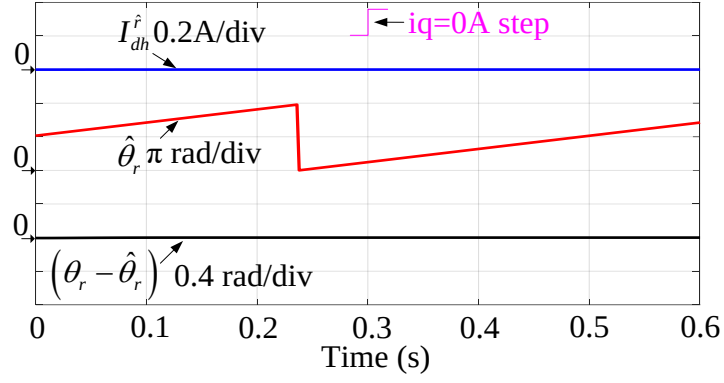
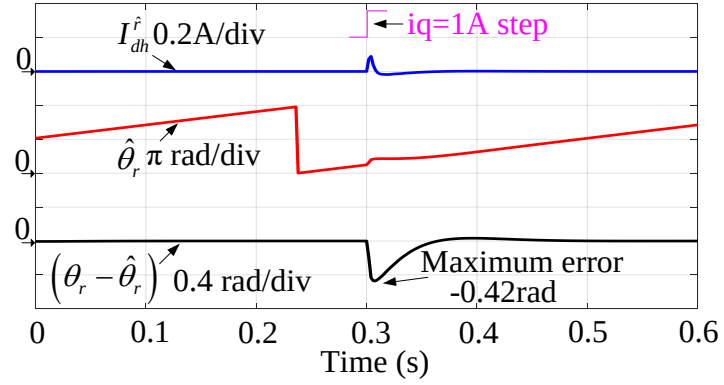


Fig. 6.6. Impact of load current step on sensorless control with HF pulsating injection.

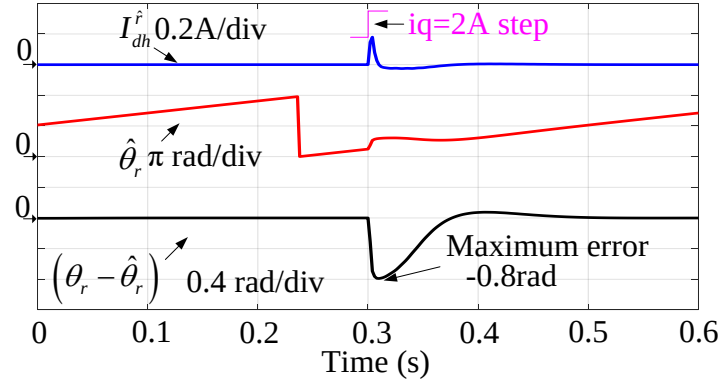
Simulations are carried out to evaluate the impacts of load current transients on pulsating voltage injection based sensorless control. Step load current i_q is applied at 0.3s. The machine is rotating at 2Hz. The position estimation performances under different step currents are shown in Fig. 6.7.



(a)



(b)

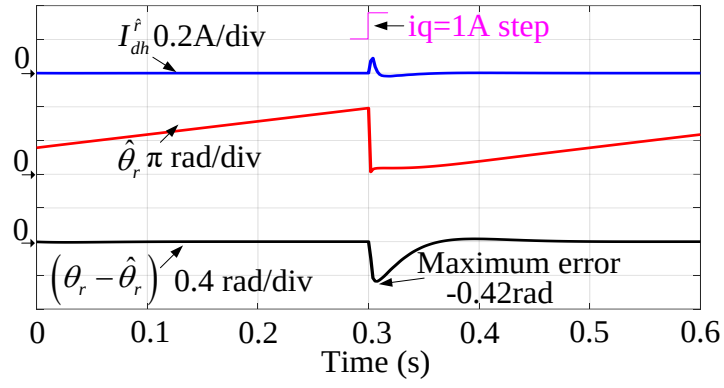


(c)

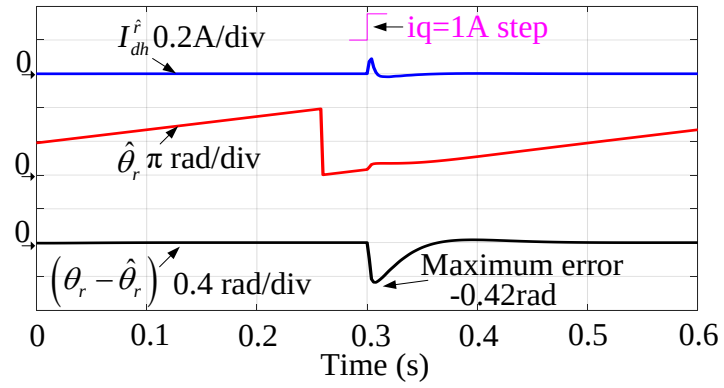
Fig. 6.7. Impact of step load current on pulsating voltage injection based sensorless control.

(a) No step current. (b) 1A step current on q -axis. (c) 2A step current on q -axis.

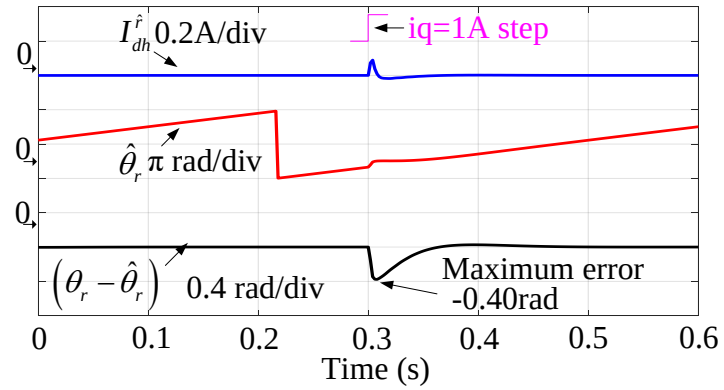
As can be seen, the position estimation error is increased during the load current transients. Different from rotating voltage injection, the maximum position error for pulsating voltage injection based sensorless control during step current transient does not vary with the rotor position, as shown by the simulation in Fig. 6.8.



(a)



(b)



(c)

Fig. 6.8. Impact of step load current on pulsating voltage injection based sensorless control at different rotor positions. (a) $\theta_r = 0 \text{ deg}$. (b) $\theta_r = 30 \text{ deg}$. (c) $\theta_r = 60 \text{ deg}$.

As can be seen, the position error caused by load current transient is not dependent on the rotor position. This is due to the fact that the pulsating voltage based sensorless method only requires the q-axis HF current.

6.2.2 Methods for mitigating the impact of load transient

If the load transient load current is predicted and subtracted from the sampled current, the interference between the HF currents induced by HF voltage injection and the transient load current can be largely reduced.

(a) HF rotating voltage injection based sensorless control

For sinusoidal rotating voltage injection, a fundamental current observer was proposed in [BRI00]. The fundamental currents are observed based on the voltage references and circuit parameters, i.e. resistance and inductances. The observed currents are then subtracted from the sampled currents before being sent to the band pass filters. Thus, the impact of step current on position estimation is mitigated. The control diagram with load current observer for rotating voltage injection based sensorless method is shown in Fig. 6.9.

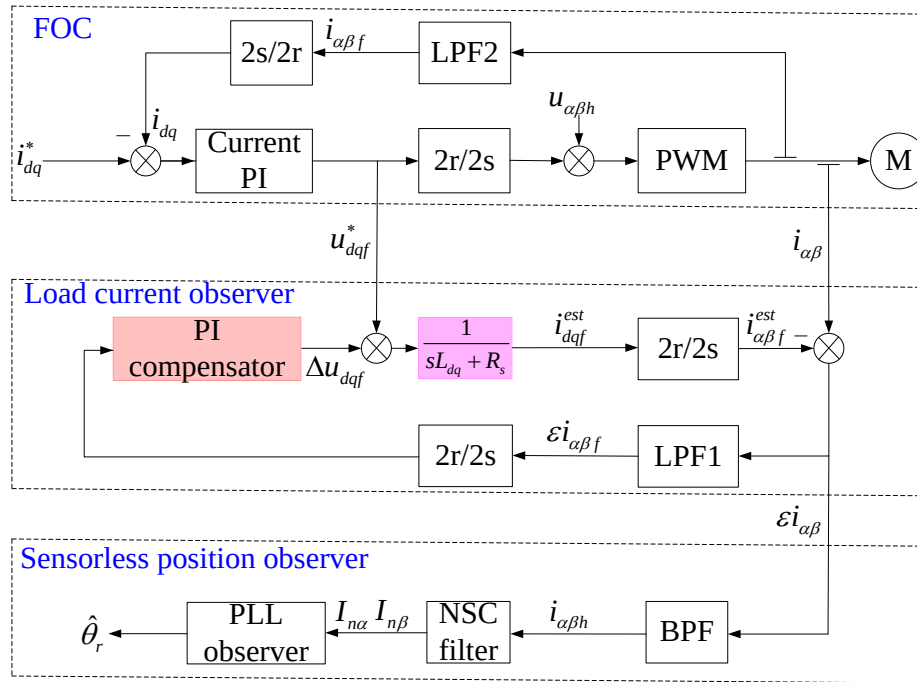


Fig. 6.9. Transient load current compensation in sinusoidal rotating voltage injection based sensorless control.

The fundamental currents i_{dqf}^{est} in the dq -axis frame are estimated using the 1st order linear transfer function as

$$i_{dqf}^{est}(s) = \frac{1}{sL_{dq} + R_s} [u_{dqf}^*(s) + \Delta u_{dqf}(s)] \quad (6.11)$$

where u_{dqf}^* are the voltages generated by the fundamental current control loop, Δu_{dqf} are the voltages generated by the PI compensator which are used to compensate the voltage distortion and parameter mismatch, i.e. L_d, L_q, R_s . The observed fundamental currents i_{dqf}^{est} are then transformed to the stationary frame as $i_{\alpha\beta f}^{est}$. The remaining currents $\varepsilon i_{\alpha\beta}$ which are the differences between the sampled currents $i_{\alpha\beta}$ and the observed currents $i_{\alpha\beta f}^{est}$, are constituted by the fundamental current error $\varepsilon i_{\alpha\beta f}$ and the HF currents $i_{\alpha\beta h}$ as

$$i_{\alpha\beta} - i_{\alpha\beta f}^{est} = \varepsilon i_{\alpha\beta} = \varepsilon i_{\alpha\beta f} + i_{\alpha\beta h} \quad (6.12)$$

The remaining currents $\varepsilon i_{\alpha\beta}$ are then sent to the BPF to extract the HF currents for position estimation. If the observed currents $i_{\alpha\beta f}^{est}$ track the transient load current well, the remaining current $\varepsilon i_{\alpha\beta}$ will only contain negligible transient load current, and the filtered HF currents $i_{\alpha\beta h}$ only contain the components contributed by the HF voltage injection. As a result, the impacts of the transient load currents on HF injection based sensorless control can be largely mitigated.

Due to voltage distortions caused by inverter nonlinearities and variations of the machine parameters, the observed $i_{\alpha\beta f}^{est}$ fundamental currents may differ from the actual fundamental currents $i_{\alpha\beta f}$. To mitigate the errors, their differences, i.e. $\varepsilon i_{\alpha\beta f} = LPF(\varepsilon i_{\alpha\beta f} - i_{\alpha\beta f}^{est})$ are sent to the PI compensators. The outputs of the compensator Δu_{dqf} are added to the voltage references u_{dqf}^* to compensate the errors between the actual load currents and the estimated load currents. The compensator parameters can be the same as the current PIs [BRI00].

(b) HF pulsating voltage injection based sensorless control

For pulsating voltage injection based sensorless control, the same load current compensation method can also be used as shown in Fig. 6.10.

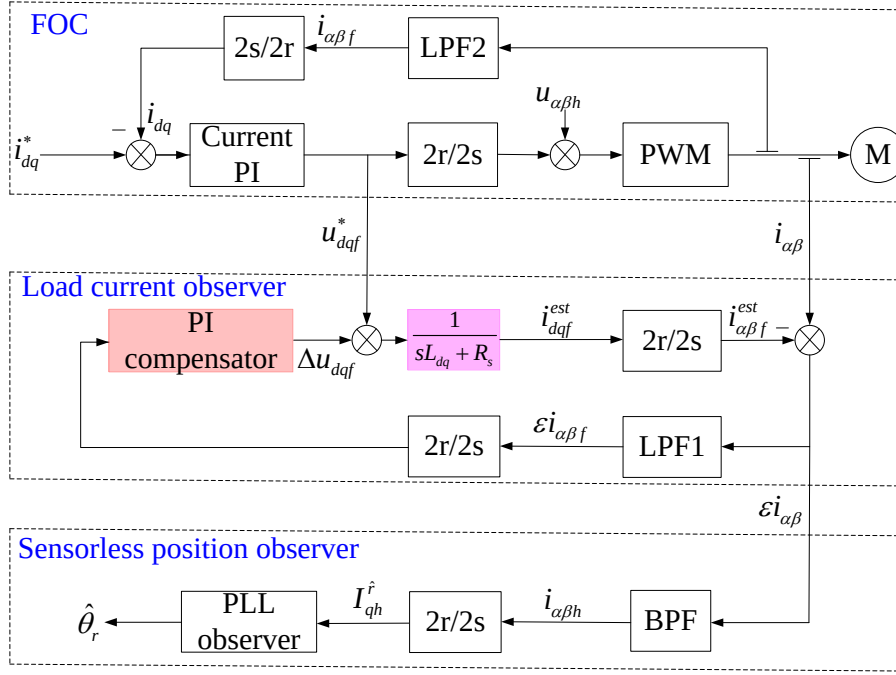


Fig. 6.10. Transient load current compensation in sinusoidal pulsating voltage injection based sensorless control.

6.2.3 Experimental evaluation

Experiments are carried out on the dSPACE platform DS1006 as shown in Fig. 6.11 to evaluate performances of the load compensation strategies in HF injection based sensorless control. A commercial inverter which receives the PWM signals from the dSPACE is used to drive the test machine. The parameters of the machine is shown in Appendix 1. Two external hall-effect sensors are installed on an analog conditioning board for sampling phase currents. The resolution of the ADC is 16bits. An incremental encoder with 1204 pulses is installed to provide rotor position feedback. The PWM and sampling frequency are all set as 10kHz.

The injected HF pulsating voltage is chosen as 15V/200Hz. The position observer parameters are $k_{pobs} = 200, k_{iobs} = 1000$, the current controller parameters are $k_{pcur} = 80, k_{icur} = 1000$, the speed control parameters are $k_{pspd} = 0.4, k_{ispd} = 0.2$. The load current observer parameters are chosen as the same as the current PI, i.e. $k_{pload} = 80, k_{iload} = 1000$, which are the same as those of the current controller.

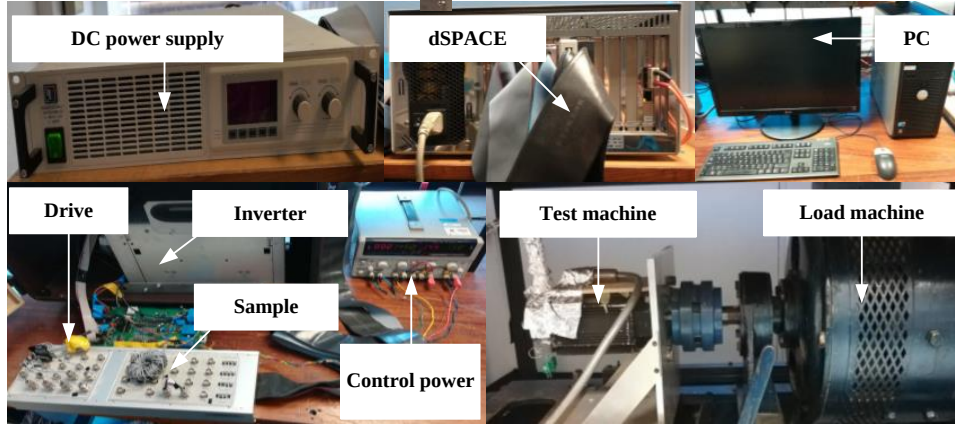


Fig. 6.11. Experimental setup.

(a) Rotating voltage injection based sensorless control without/with load current compensation

In this test, the speed reference is ramped up from zero to 60rpm over 0.3 without/with the load current compensation described in Fig. 6.9. The stator winding resistance is $R_s = 6\Omega$, and the dq-axis apparent inductances are set as $L_d = 40mH, L_q = 60mH$. The initial rotor position before start is fixed as 60 degrees. The injected HF rotating voltage is chosen as 15V/200Hz. The current references are generated according to the maximum torque per ampere law. The speed, dq-axis currents and the position estimation error for the start without/with load current compensation are shown in Figs. 6.12(a) and 6.12(b), respectively.

As can be seen, the maximum position error during start without load current compensation is about 40 degrees, while the maximum error with load current compensation is only 20 degrees. Due to relative small position error, the starting time (from zero to 60r/min) with load current compensation is also less than that without load current compensation. Fig. 6.12(c) shows the observed load currents during starting. The peak values of the differences between the actual and observed currents are less than 1A. By sending these current differences to the band pass filters, the interference between the HF currents and step load currents can be largely reduced. Therefore, with load current compensation technique, the dynamic performance of sensorless control with rotating voltage injection can be improved.

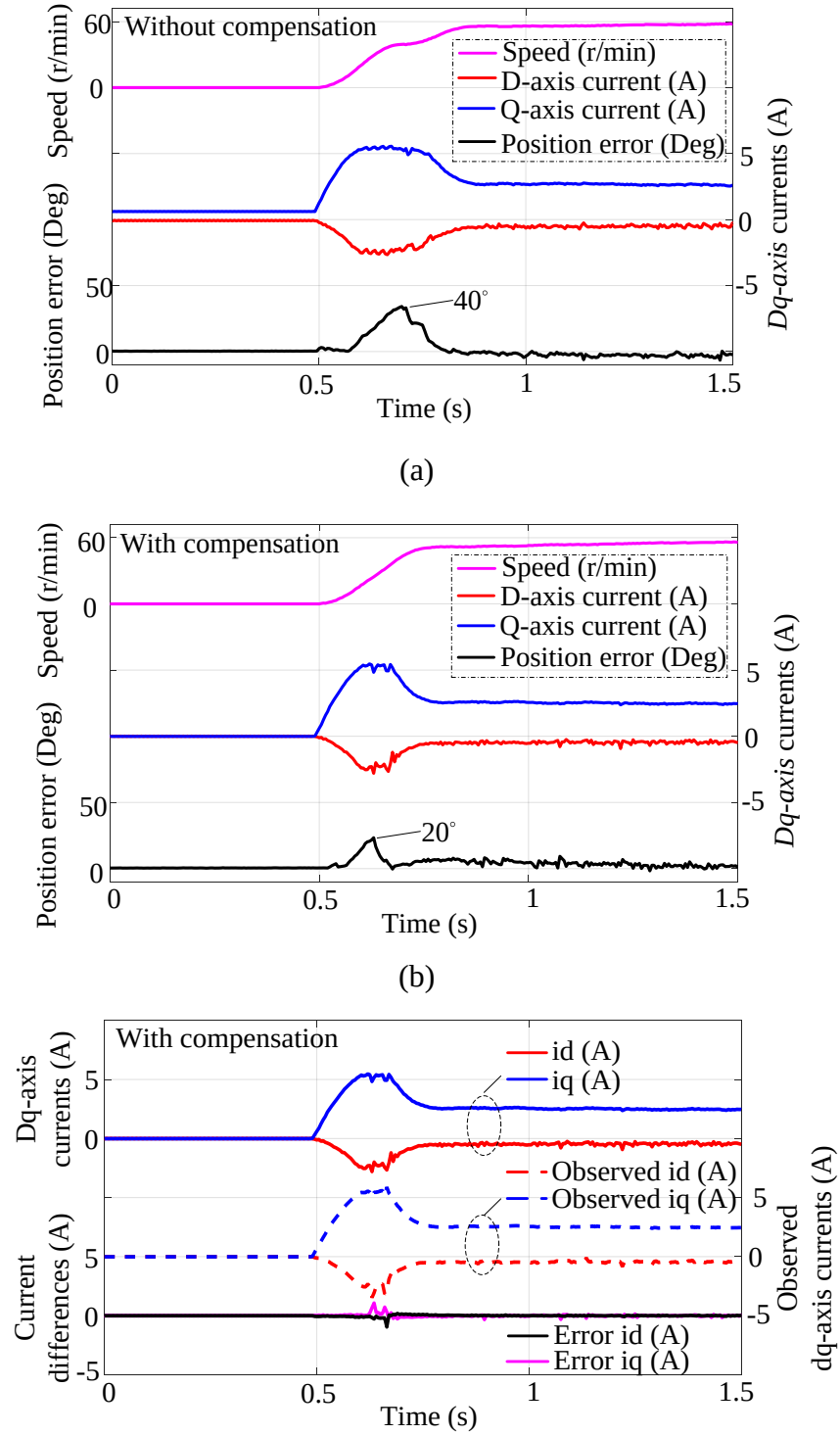


Fig. 6.12. Sensorless start without/with load current compensation based on HF rotating voltage injection. (a) Without compensation. (b) With compensation. (c) Observed currents.

(b) Pulsating voltage injection based sensorless control without/with load current compensation

In this test, the speed reference is ramped up from zero to 60rpm over 0.5s without/with the load current compensation described in Fig. 6.10. The speed, dq -axis currents and the position

estimation error for the start without/with load current compensation are shown in Figs. 6.13(a) and 6.13(b) respectively. The observed load currents are shown in Fig. 6.13(c).

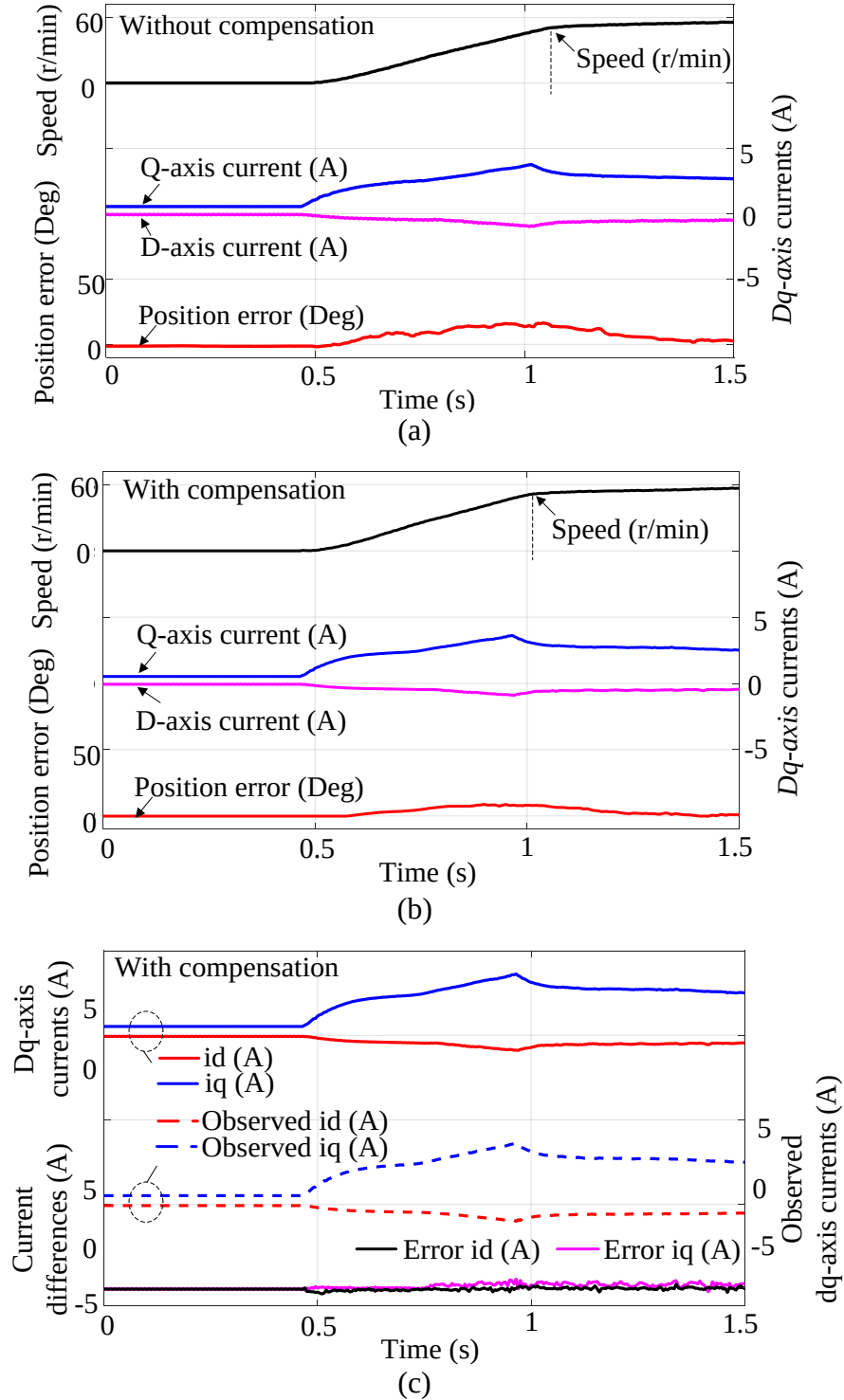


Fig. 6.13. Sensorless start without/with load current compensation based on HF pulsating voltage injection. (a) Without compensation. (b) With compensation. (c) Observed currents.

As can be seen, the maximum position error with load current compensation is decreased. The starting time (from zero to 60r/min) with load current compensation is also less than that

without load current compensation. The observed load currents in Fig. 6.13(c) shows a good agreement with the actual currents. Therefore, with load current compensation technique, the dynamic performance of sensorless control with pulsating voltage injection can be improved.

6.3 Whole Speed Range Sensorless Operation

Whole speed sensorless operation is required in many applications. The saliency detection based sensorless method shows superiority at standstill and low speed, while the fundamental model based sensorless method is effective only at medium and high speeds. The whole speed range cannot be covered by a single sensorless method. Thus, there should be an appropriate strategy to combine these two sensorless algorithms. Many different methods have been proposed and they can be classified into dual-observers solution and single observer solution as described as follows.

6.3.1 Existing methods for smooth switchover between different speed ranges

For dual-observers solution [MA13], [IDE06], [YOU18], the saliency model based position observer and fundamental model based position observer are running in parallel as shown in Fig. 6.14.

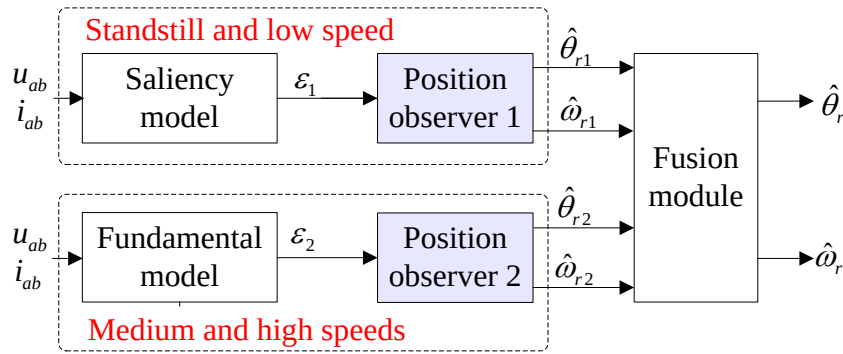


Fig. 6.14. Dual-observers solution for whole speed range sensorless operation.

The saliency model and fundamental model extract the position error signals $\varepsilon_1, \varepsilon_2$ from the terminal voltages and currents respectively. The position observer 1 and observer 2 estimate the rotor positions by controlling $\varepsilon_1 = 0$ and $\varepsilon_2 = 0$. The estimated positions and speeds from different observers are fused in the fusion module to output the estimated position and speed.

The fusion module usually employs a weighting function to combine the estimated positions and speeds as expressed in (6.13)-(6.15). The weighting function can be linear switching or direct switching as shown in Fig. 6.10. The linear switching provide a smooth transition

between low and high speeds, while direct switching is more simple by changing sensorless mode at the switching point. A hysteresis band ω_{hys} is usually inserted to avoid chattering at the switching point.

$$\begin{aligned}\hat{\omega}_e &= k(\hat{\omega}_e)\hat{\omega}_{r1} + [1 - k(\hat{\omega}_e)]\hat{\omega}_{r2} \\ \hat{\theta}_r &= k(\hat{\omega}_e)\hat{\theta}_{r1} + [1 - k(\hat{\omega}_e)]\hat{\theta}_{r2}\end{aligned}\quad \text{fusion function} \quad (6.13)$$

$$k(\hat{\omega}_e) = \begin{cases} 1, & |\hat{\omega}_e| < \omega_L \\ \frac{\omega_H - \hat{\omega}_e}{\omega_H - \omega_L}, & \omega_L \leq |\hat{\omega}_e| \leq \omega_H, \text{ smooth switching} \\ 0, & |\hat{\omega}_e| > \omega_H \end{cases} \quad (6.14)$$

$$k(\hat{\omega}_e) = \begin{cases} 1, & |\hat{\omega}_e| < \omega_{sw} \\ 0, & |\hat{\omega}_e| > \omega_H \end{cases}, \text{ direct switching} \quad (6.15)$$

where ω_L, ω_H are the limits of the transition region for smooth switching, ω_{sw} is the speed point for direct switching.

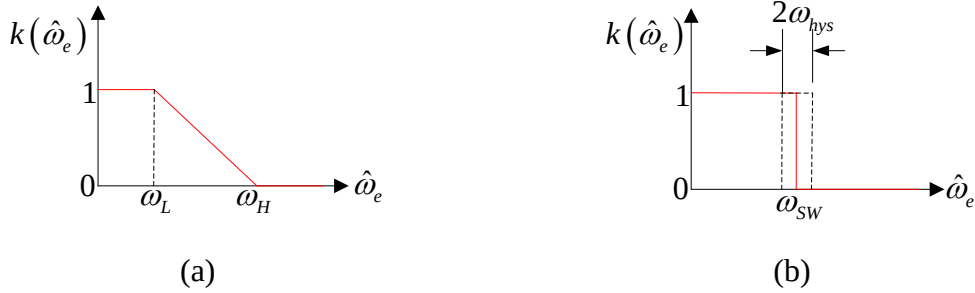


Fig. 6.15. Different switching functions. (a) Linear switching. (b) Direct switching.

With dual-observer structure, the dynamic performance of position estimation at low speed and high speed ranges can be adjusted separately. However, this will increase the computation burden due to two closed-loop observers. To overcome this limit, the single observer structure is used [GUG04], [FLI18], [YAN17], [QIN18].

The single observer structure shown in Fig. 6.16 only requires one closed-loop observer. The position error signals from saliency model and fundamental model are normalized and combined by the error fusion module as shown in (6.9). With normalized error signal, the saliency model and fundamental model can share the same PI parameters with unified dynamic performance.

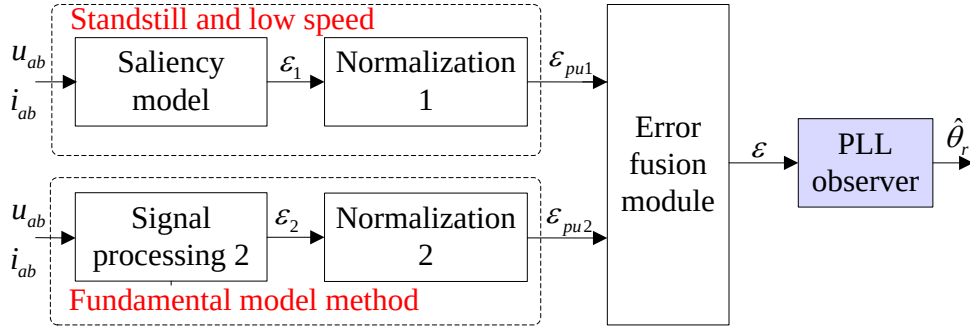


Fig. 6.16. Single-observers solution for whole speed range sensorless operation.

The fused error signal is then sent to the shared position observer. The weighting function $q(\hat{\omega}_e)$ can be linear or direct switching function as that in (6.14) and (6.15).

$$\varepsilon = q(\hat{\omega}_e) \varepsilon_{pu1} + (1 - q(\hat{\omega}_e)) \varepsilon_{pu2} \quad (6.16)$$

In summary, for whole speed sensorless operation, a fusion of the sensorless methods for low speed and high speed ranges is required. Dual observer structure and single observer structure are widely used. The main difference between the two structures is the variable that is fused. For dual observers, the estimated positions and speeds obtained by different sensorless methods are fused, while for single observer, the position errors are fused. The fusion functions can be linear or direct switching types.

Table 6.1 Summary of methods for whole speed sensorless operation

Types	Fusion variables	Fusion function	Reference
Dual observers	Position	Linear switch Direct switch	[MA13]/[YOU18]
	Speed		[IDE06]
Single observer	Position error signals		[FLI18]
			[GUG04], [YAN17], [QIN18]

6.3.2 Proposed method based on single position observer

In this chapter, a whole speed sensorless operating method incorporating HF sinusoidal rotating voltage injection based sensorless control and active-flux method is proposed as shown in Fig. 6.17. The proposed method can obtain smooth switchover between different speed regions with single PLL observer.

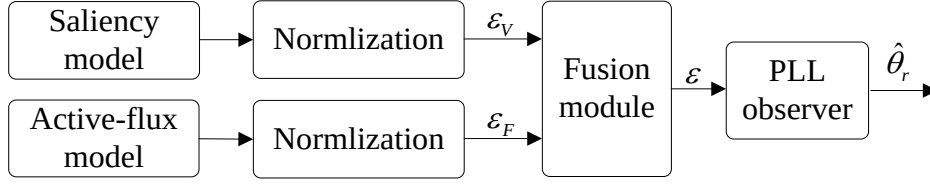


Fig. 6.17. Proposed whole speed range sensorless operation method.

6.3.2.1 Saliency model at standstill and low speed

For whole speed operation, the position at standstill and low speed region is usually estimated with HF voltage injection. In [YAN17], single observer based whole speed sensorless operation is achieved by pulsating square wave voltage injection at standstill and low speed, and the back EMF model at medium and high speeds. For obtaining unity length position error signal from pulsating square wave voltage injection, a scaling factor k_{sal} is required in [YAN17]. This factor varies with the fundamental currents and might change the PLL estimation bandwidth. In this section, rotating sinusoidal voltage $u_{\alpha\beta h}^s$ in are injected to stator stationary frame and the scaling factor is not required for obtaining a unity length error signal.

The resulting HF currents $i_{\alpha\beta h}^s$ at standstill and low speed due to rotating voltage injection are obtained from (6.4) as

$$i_{\alpha h}^s + ji_{\beta h}^s = I_p e^{j(\theta_h - \pi/2)} + I_n e^{-j(\theta_h - 2\theta_r - \pi/2)} \quad (6.17)$$

$$I_p = \frac{V_h (L_{dh} + L_{qh})}{2\omega_h L_{dh} L_{qh}}, I_n = \frac{V_h (L_{qh} - L_{dh})}{2\omega_h L_{dh} L_{qh}} \quad (6.18)$$

The induced HF currents are composed of positive and negative sequence currents which are phase-modulated waveforms. In chapter 2, it has been revealed that the HF currents are also amplitude-modulated waveforms as

$$\begin{bmatrix} i_{ah}^s \\ i_{bh}^s \\ i_{ch}^s \end{bmatrix} = \begin{bmatrix} Amp - i_{ah}^s \sin(\theta_h + \varphi_{ah}) \\ Amp - i_{bh}^s \sin(\theta_h + \varphi_{bh}) \\ Amp - i_{ch}^s \sin(\theta_h + \varphi_{ch}) \end{bmatrix} \quad (6.19)$$

$$\begin{bmatrix} Amp - i_{ah}^s \\ Amp - i_{bh}^s \\ Amp - i_{ch}^s \end{bmatrix} = \begin{bmatrix} \sqrt{I_p^2 + I_n^2 + 2I_p I_n \cos(2\theta_r)} \\ \sqrt{I_p^2 + I_n^2 + 2I_p I_n \cos(2\theta_r + 2\pi/3)} \\ \sqrt{I_p^2 + I_n^2 + 2I_p I_n \cos(2\theta_r - 2\pi/3)} \end{bmatrix} \quad (6.20)$$

It is observed in Fig. 6.18 that the whole electrical period is divided into 12 separate sectors by the crossings of the amplitude traces. The width of each sector is 30° . The initial sector can be detected by the relationship between the three phase HF current amplitudes. The middle angle of the sector can be treated as the initial rotor position. Although the initial position has a maximum error of $\pm 15^\circ$ to the actual initial position, it will only result in less than 5% decrease in the startup torque as analysed in chapter 2. Thus, the machine can be started by comparing the three phase HF current amplitudes.

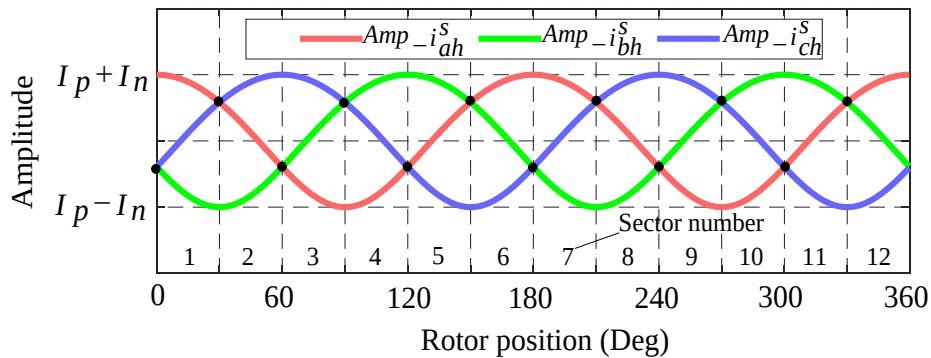


Fig. 6.18. HF current amplitudes during one electrical period.

After the machine is started, the HF current amplitudes are changing with rotor position. The number of sector which is separated by amplitude crossings is changed every 30° . Thus, by detecting the change of sector numbers, the crossings can be identified. The discrete boundary angles $\theta_{discrete}$ are also known and can be used for FOC. To reduce the computation burden, sliding discrete Fourier transform is introduced to calculate the HF current amplitudes.

Based on this phenomenon, the discrete positions $\theta_{discrete}$ can be used for FOC. A sensorless startup experiment discrete boundary positions is carried out. The related waveforms are shown in Fig. 6.19. The machine parameters are attached in Appendix 1. The injection voltage is chosen as 15V/500Hz. The machine is started to 2Hz with half rated load torque.

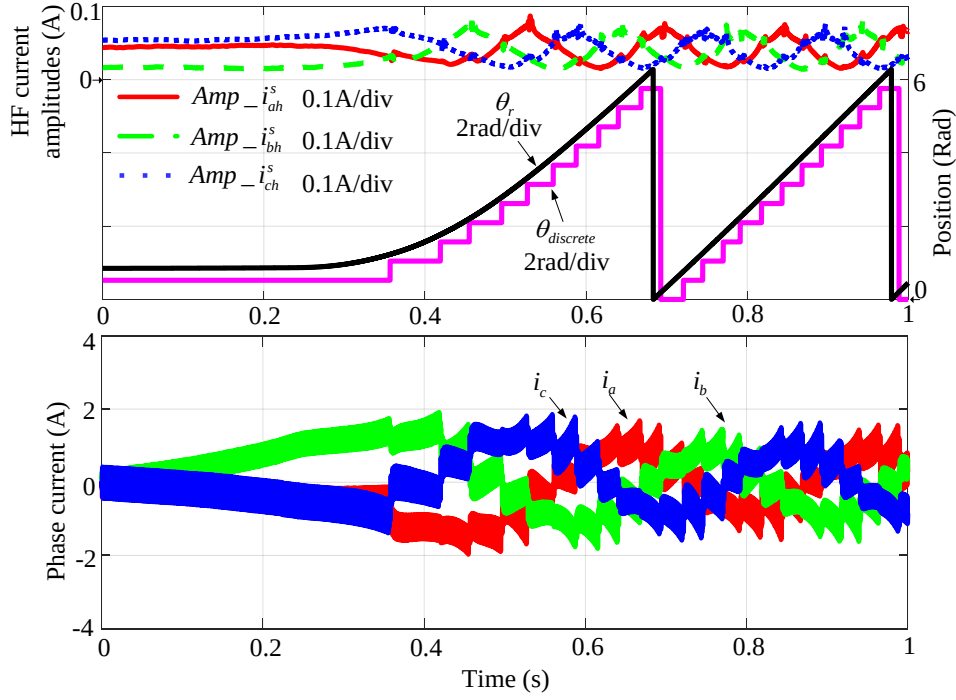


Fig. 6.19. Sensorless startup with discrete boundary positions based on HF rotating voltage injection.

It is observed that the machine is successfully started with discrete positions. However, the phase currents are not continuous at the sector boundaries due to discrete characteristic of the angle $\theta_{discrete}$ used for FOC. Such phase currents contain many harmonics and will result in the following.

- ① Torque ripple and noise.
- ② Deteriorated performance of back EMF based sensorless method when switchover to high speed range is required.

To overcome these problems, continuous position is required from the discrete positions which can be modelled by space vector model as

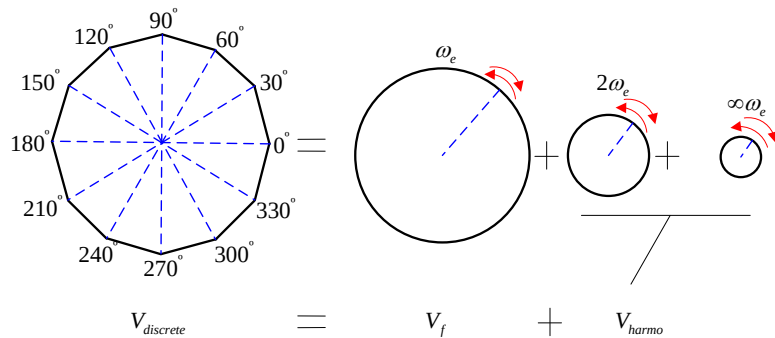


Fig. 6.20. Space vector model for decoupling space harmonics.

The fundamental and harmonic vectors can be estimated by the estimated rotor position $\hat{\theta}_r$ as

$$\hat{V}_f = e^{j(\hat{\theta}_r - \varphi)} \quad (6.21)$$

$$\hat{V}_{harmono} = \sum_{k=1}^{\infty} \left[\frac{-1}{kN-1} e^{-j((kN-1)\hat{\theta}_r - \varphi)} + \frac{1}{kN+1} e^{j((kN+1)\hat{\theta}_r - \varphi)} \right] \quad (6.22)$$

where $N=12$ denotes the sectors in one electrical period, $\varphi=\pm\pi/N$ is the phase delay and of the same sign as rotor speed, ω_e is the fundamental electrical angular frequency. A vector tracking observer is used to obtain smooth position output from the discrete positions as

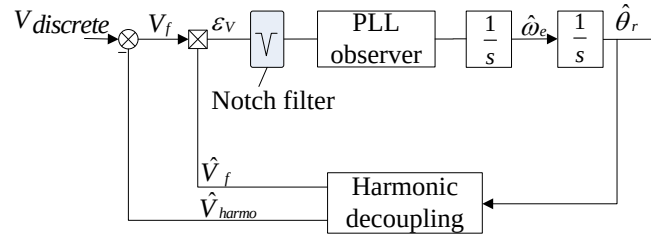


Fig. 6.21. Vector tracking observer for continuous position observer.

The position error ε_v is obtained by cross product as

$$\varepsilon_v = V_f \times \hat{V}_f = \sin(\theta_r - \hat{\theta}_r) \quad (6.23)$$

The estimated position can be obtained by controlling $\varepsilon_v = 0$. The notch filter is used to mitigate the harmonic components caused by position estimation error. It can be seen that the error signal ε_v is inherently unity length. In this chapter, the position error obtained by the vector tracking observer will be combined with the error signal obtained by active-flux model to provide whole speed range sensorless operation.

6.3.2.2 Active flux model at medium and high speeds

The fundamental model is employed for sensorless position estimation at medium and high speeds. For non-salient pole machine ($L_d=L_q=L_s$), the position can be extracted from the flux-linkages as

$$\theta_r = \tan^{-1} \left(\frac{\psi_\beta - L_s i_\beta}{\psi_\alpha - L_s i_\alpha} \right) \quad (6.24)$$

where ψ_α, ψ_β are the two phase stator flux linkages which can be obtained by integrating the back EMF signals.

However, for salient pole machine, the phase inductance is coupled with the rotor position, which makes it difficult to separate the permanent flux linkages. The active-flux $\vec{\psi}_s^a$ as shown in (6.20) was introduced to tackle this problem [BOL08].

$$\vec{\psi}_s^a = \vec{\psi}_s - L_q \vec{I}_s \quad (6.25)$$

where $\vec{\psi}_s$ is the stator winding flux linkage vector, the subscript ‘s’ means the stator winding. As presented in the introduction, the active-flux vector is aligned with the d -axis. Thus, the rotor position can be obtained by observing the position of the active-flux vector.

The $\alpha\beta$ -axis components of the active-flux can be calculated with (6.26). To remove the integration drift, high pass filters (HPFs) are added after the integration as shown in Fig. 6.22.

$$\psi_s^a = |\vec{\psi}_s^a| e^{j\theta_r} = \int (u_{\alpha\beta} - R_s i_{\alpha\beta}) dt - L_q i_{\alpha\beta} \quad (6.26)$$

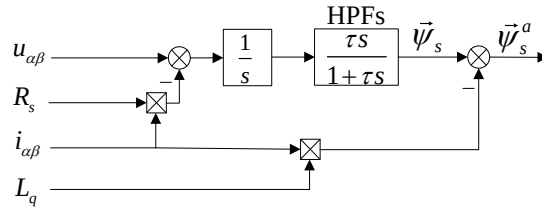


Fig. 6.22. Calculating active-flux.

The rotor position can be then calculated by arctangent function as

$$\begin{aligned} \psi_{s\alpha}^a &= \int (u_\alpha - R_s i_\alpha) dt - L_q i_\alpha = |\psi_s^a| \cos(\theta_r) \\ \psi_{s\beta}^a &= \int (u_\beta - R_s i_\beta) dt - L_q i_\beta = |\psi_s^a| \sin(\theta_r) \\ \theta_r &= \tan^{-1} \left(\frac{\psi_{s\beta}^a}{\psi_{s\alpha}^a} \right) \end{aligned} \quad (6.27)$$

However, direct calculation may introduce much noise to the calculated results. A PLL observer is used to obtain smooth position output as shown in Fig. 6.23. Different from conventional PLL observer, a normalization module is inserted for unity-length error signal ε_F as shown in (6.28). As amplitude of the active-flux $|\vec{\psi}_s^a|$ is proportional to the electrical speed, the normalization is of high accuracy at medium and high speeds.

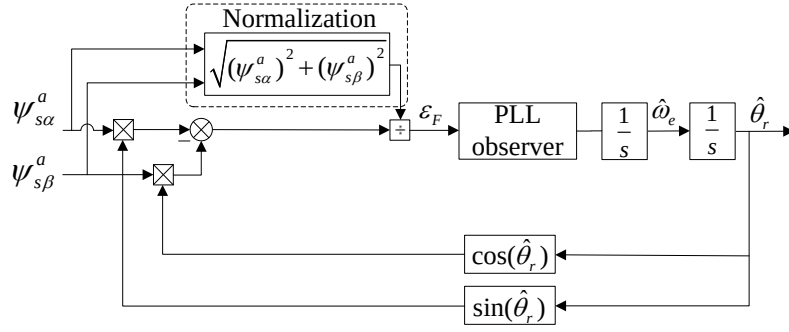


Fig. 6.23. Proposed position observer based on active-flux with unity-length error signal.

$$\varepsilon_F = \frac{\vec{\psi}_s^a \times e^{j\theta_r}}{|\vec{\psi}_s^a|} = \sin(\theta_r - \hat{\theta}_r) \quad (6.28)$$

As can be seen, the error signal from the fundamental model has the same form as that from the saliency model in (6.23). Thus, the two error signals can share the same PLL observer with unified dynamic performance.

6.3.2.3 Proposed method based on rotating voltage injection and active-flux

As previously analysed, the position error signals from the saliency model and the fundamental model are both unity-length. A hybrid position estimation method for whole speed range sensorless operation is proposed as Fig. 6.24.

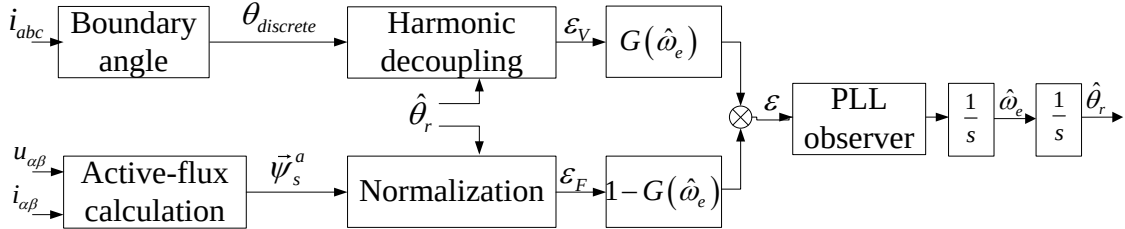


Fig. 6.24. Proposed whole speed range sensorless operation method with single observer.

The position error signal ε_V from the saliency model and ε_F from the flux linkage model are forged by a linear weighting function $G(\hat{\omega}_e)$ which is shown in Fig. 6.25.

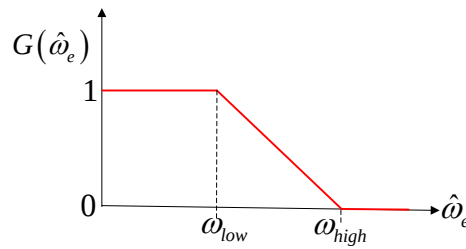


Fig. 6.25. Speed dependent weighting function demonstration.

Determination of the transition band frequencies should take the following guidelines into consideration.

① The value of ω_{low} should be above the low speed limit of back EMF based sensorless methods. According to [HEJ11], under 10% of the rated speed ($\omega_N = 50\text{Hz}$), dynamic stiffness of the back EMF based method is significantly decreased. Thus, $12\%\omega_N$ (6Hz) is chosen for ω_{low} .

② The value of ω_{high} should be less than 10% of the injection frequency for easy separation of the HF currents. In this chapter, the injection frequency is $f_{inj} = 500\text{Hz}$. Thus, $2.4\%f_{inj}$ (12Hz) is chosen for ω_{high} .

With the above parameters, the width of the transition band is 6Hz in which the position estimation is contributed by both the saliency and fundamental models. Various experiments will be carried out to verify the effectiveness of the proposed whole speed range sensorless operation method.

6.3.3 Experimental verification

The proposed whole speed range sensorless operation is verified on a dSPACE platform with a 0.6-kW IPMSM. The parameters of IPMSM are shown in Appendix 1. A direct current generator with resistance load is connected to the machine to produce the load torque. An incremental encoder of 1024 pulses per revolution is used to provide accurate rotor position θ_r and speed ω_r . The switching frequency and sampling frequency are both set as 10kHz. A rotating HF voltage of 15V/500Hz is selected for saliency detection considering estimation bandwidth and accuracy. The rated frequency of the machine is 50Hz. The values of ω_{low} and ω_{high} for the weighting function are determined as $\omega_{low}=37.68\text{rad/s}$ (6Hz/120rpm) and $\omega_{high}=75.36\text{rad/s}$ (12Hz/240rpm). The parameters of the PLL are designed by placing two poles at one point as

$$\frac{\hat{\theta}_r(s)}{\theta_r(s)} = \frac{k_p s + k_i}{s^2 + k_p s + k_i} = \frac{2\rho s + \rho^2}{s^2 + 2\rho s + \rho^2} \quad (6.29)$$

where k_p, k_i are the PI parameters, ρ is the pole value. The PI parameters are set as $k_p = 2\rho, k_i = \rho^2$. The bandwidth under 3dB amplitude attenuation is $\omega_{bw} = 2.48\rho$. In following experiments, ρ is set as 100 for a bandwidth of 248rad/s considering the balance between

dynamic performance and noise filtering. The PI parameters of speed controller are $k_{pspd} = 0.03$, $k_{ispd} = 0.1$ and the PI of current controller are $k_{pcnt} = 80$, $k_{icnt} = 1000$.

(a) Sensorless start to a speed in transition region ($120\text{rpm} < n < 240\text{rpm}$)

In this experiment, the machine is sensorless started to 180rpm which is in the speed transition region, i.e. between 120rpm and 240rpm, and then decelerated to zero. The position offset caused by the cross-coupling effect is compensated by the offline measured errors. Fig. 6.26 shows the waveforms during this process.

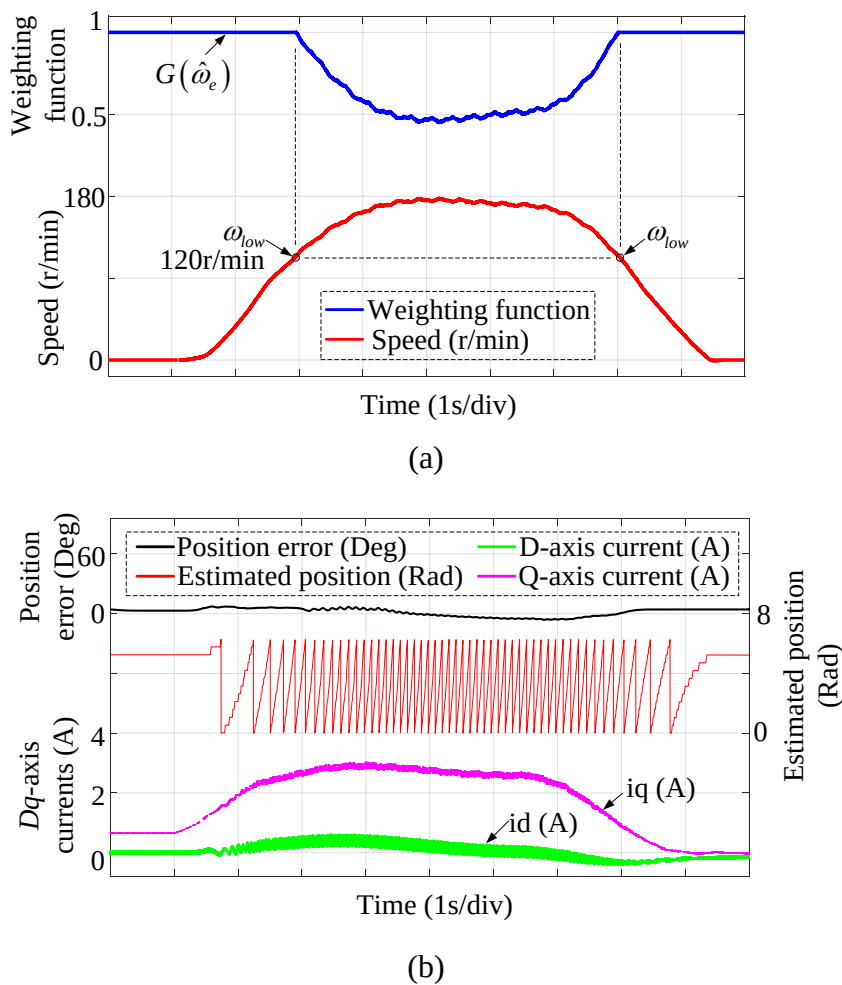


Fig. 6.26. Sensorless start to 180rpm and decelerated back to zero. (a) Speed, weighting function. (b) Position, dq -axis currents.

It can be seen that the speed trace crosses the low speed limit (180rpm) of the transition region during acceleration and deceleration. The weighting coefficient $G(\hat{\omega}_e)$ varies with the speed. Smooth start and stop can be achieved.

(b) Step load and unload in the transition region

The machine is sensorless operated at 180rpm with the proposed hybrid method. Step load torque is applied and then removed after several seconds. The waveforms during the process are shown in Fig. 6.27.

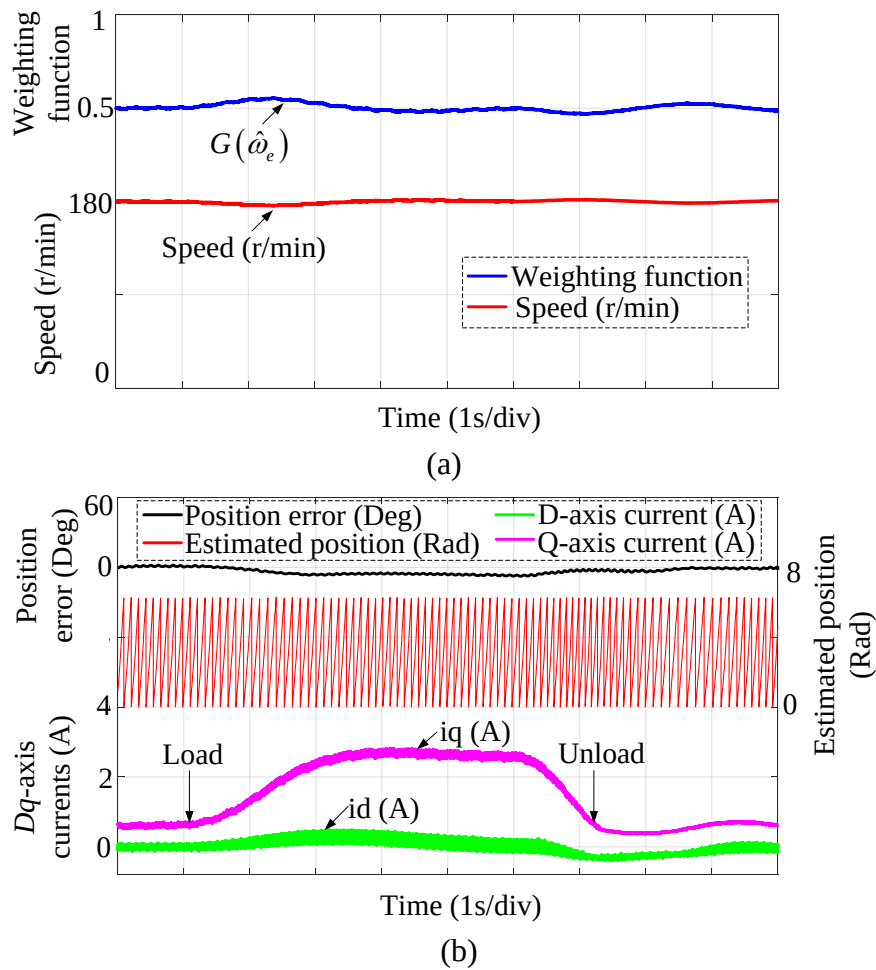
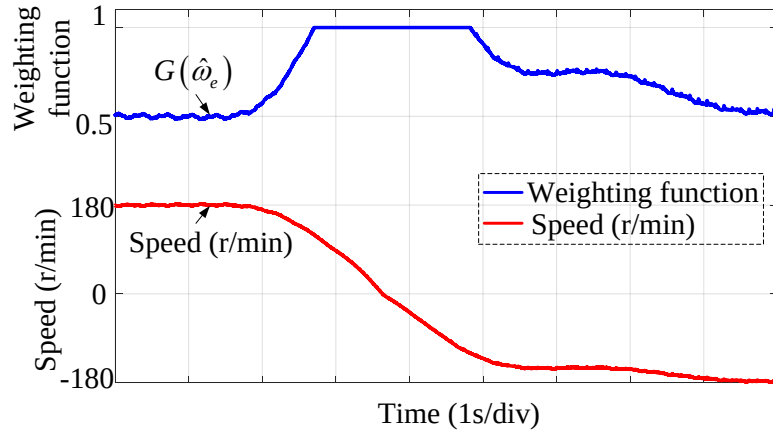


Fig. 6.27. Sensorless operation at 180rpm with step load torque and unload. (a) Speed, weighting function. (b) Position, dq -axis currents.

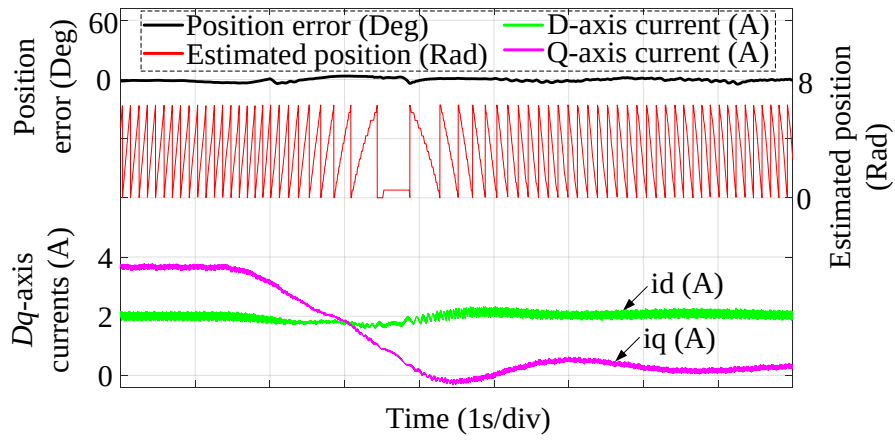
It can be seen that the speed remains stable under step load and unload torque. The weighting function $G(\hat{\omega}_e)$ stays at near 0.5. The position estimation at this speed is contributed by 50% of the fundamental model and 50% of the saliency model. Therefore, the proposed hybrid sensorless method is capable of working in the transition region.

(c) Sensorless speed reversal between 180rpm and -180rpm

The machine is sensorless operated at 180rpm and the speed reference is decreased from 180rpm to -180rpm in 3s. The waveforms during the process are shown in Fig. 6.28.



(a)



(b)

Fig. 6.28. Sensorless speed reversal between 180rpm and -180rpm. (a) Speed, weighting function. (b) Position, dq -axis currents.

As can be seen that the speed can be controlled from 180rpm to -180rpm smoothly. There are some discontinuities in the estimated position at the region near zero speed region. This is due to discontinuous position output of the vector tracking observer at low speed. The weighting function is 0.5 at 180r/min and decreased to zero when the speed is approaching zero. After the speed crosses zero and is accelerated backward to -180r/min, the weighting function is increased to 0.5 again. Therefore, sensorless reversal operation can be achieved with the proposed method.

(d) Sensorless speed variation between 120rpm (ω_{low}) and 240rpm (ω_{high})

Sensorless operation between the low speed and high speed limits is also carried out. The speed reference is increased from 120rpm to 240rpm and decreased back to 120rpm.

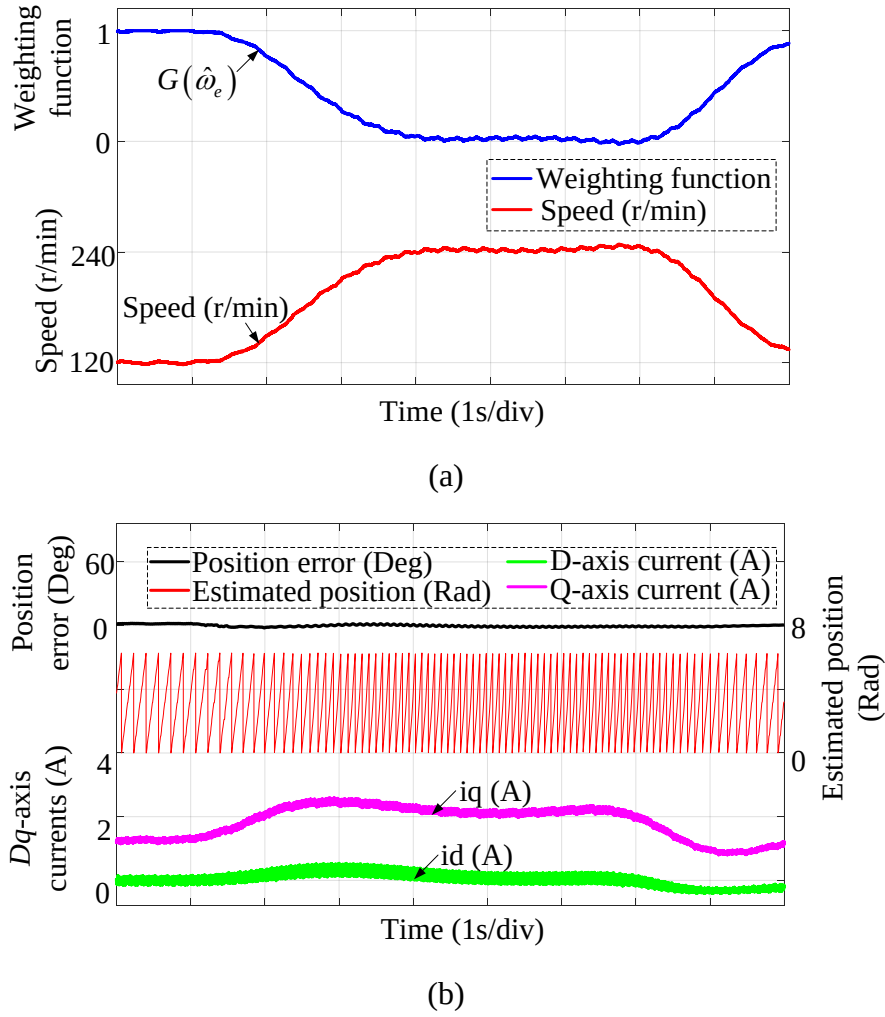


Fig. 6.29. Sensorless variation between low and high speed limits in the transition region. (a) Speed, weighting function. (b) Position, dq -axis currents.

It can be seen that the speed is changed smoothly from 120 rpm to 240 rpm and back to 120rpm. The weighting function is changing from 1 to 0 and back to 1. The position error is above 0.1rad. The proposed method can achieve satisfactory operation between the two boundary speed points in the transition region.

6.4 Conclusion

In this chapter, the impacts of load current transient on sensorless position estimation are analysed. The load current prediction technique is tested for compensating the load current transient. Then, the whole speed sensorless operation method is presented, i.e. the single observer solution and dual-observer solution. Their advantages and disadvantages are analysed. In this chapter, a single observer based whole speed sensorless operation method is proposed based on HF rotating voltage injection at standstill and low speed, active-flux model at medium

and high speeds. Position error signals from saliency model and active flux model can be obtained and fused for position estimation. Experiments are carried out to verify effectiveness of the proposed whole speed sensorless method.

CHAPTER 7

GENERAL CONCLUSIONS

This thesis is focused on sensorless torque control of interior permanent magnet synchronous machines at standstill and low speed. The main structure of this thesis is shown in Fig. 7.1.

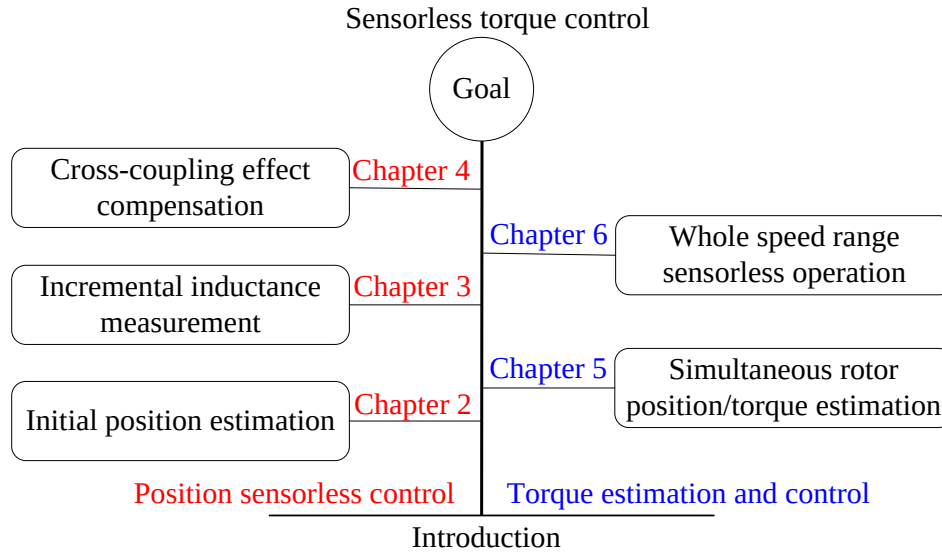


Fig. 7.1. Main structure of this thesis.

Sensorless torque control topic has two important aspects: sensorless rotor position estimation, and torque estimation and control. Hence, the research is carried out along the two sub-topics. The chapters are organized in several chapters focusing on sensorless torque position estimation and several chapters focusing on the torque estimation and actual applications. This chapter on conclusions is organised as follows.

- 1) What is the purpose for doing this research?
- 2) What has the thesis achieved?
- 3) What can be done in the future?

7.1 What is the Purpose for Doing This Research

Rotor position sensorless control methods and torque control techniques for interior permanent magnet synchronous machines have been developed for decades. Some research findings have been successfully applied in domestic and industrial applications, such as wash machine and general-purpose inverters. However, the applications of sensorless control in

electric vehicles and safety critical products are still limited. For example, in electric vehicles, the sensorless control method is only treated as a backup strategy in case of a resolver failure. Besides, the sensorless control should overcome many nonlinear factors, e.g. temperature variation, inverter nonlinearities, and magnetic saturation. Hence, there is still a long way to go before the sensorless methods can achieve comparable performance with hardware based high accuracy position sensors. The improvements on sensorless torque control can be made from following directions.

- 1) Fast and reliable initial position estimation is important for starting.
- 2) Reliable sensorless position estimation and torque control at heavily saturated conditions are required for many applications. In the IPMSMs, the magnetic field is often designed to be working in saturation state to maximize the capability and utilisation of the magnetic materials. For saliency tracking based sensorless methods, high saturation conditions may result in reduced saliency, cross-coupling effect and spatial harmonics. Therefore, for guaranteeing a reliable sensorless position estimation, these should be compensated.
- 3) Low noise sensorless position estimation at standstill and low speed is essential for the products that are close to the consumers. For example, an electric vehicle driver may be very sensitive to the noise caused by the HF voltage injection which is used for estimating the rotor position.
- 4) High accuracy torque estimation considering parameter variations, i.e. temperature, dc bus voltage, and speed ripple suppression, is important for improving the torque control performance.

7.2 What Has This Thesis Achieved

This thesis has investigated three important issues in saliency tracking based sensorless methods, i.e.

- (a) the initial rotor position estimation,
- (b) the incremental inductance measurement and cross-coupling effect compensation,
- (c) the two important issues on torque control, i.e. the torque estimation and whole speed range sensorless operation.

7.2.1 Sensorless rotor position estimation

The sensorless position estimation issues and their significances for sensorless control are summarised in Fig. 7.2. Their implementation and novelties will be discussed in the following sections.

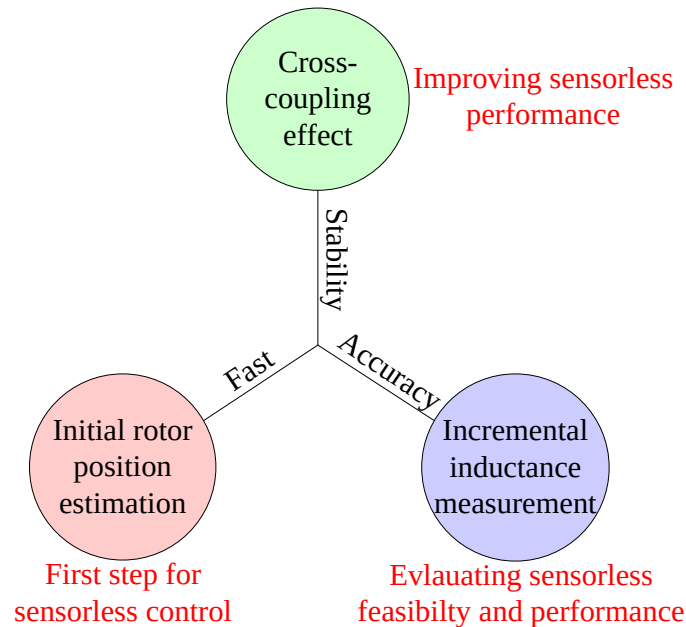


Fig. 7.2. Sensorless position estimation issues and their significances for sensorless control in this thesis.

(a) Initial rotor position estimation

Initial rotor position estimation is the first step for sensorless operation. Conventionally, in sensorless operation, the initial rotor direction is detected with a closed-loop position observer by injecting HF voltage (rotating or pulsating). In the estimation, the observer parameters should be tuned depending on the initial machine parameters which may be not known. Besides, the convergence time of the closed-loop observer is unpredictable and could be very long. In this case, a fast initial position estimation method is proposed without using a position observer.

The HF currents due to HF rotating voltage injection are amplitude-demodulated waveforms as the three phase HF current amplitudes vary with the rotor position. With this phenomenon, the three phase HF current amplitudes can be used for rotor position estimation which is different from conventional negative sequence current based position demodulation method. In the proposed method, by comparing the three phase HF current amplitudes, the rotor position with maximum $\pm 15^\circ$ error can be estimated without using closed-loop position observer. Then, the initial polarity can be judged by employing double voltage pulses injection to the estimated position.

With the proposed initial rotor position estimation method, several advantages can be obtained as

- 1) Faster initial rotor position estimation has been achieved compared to conventional methods.
- 2) Closed-loop position observer has not been used in the estimation.
- 3) Low noise initial rotor position can be achieved when multi-frequency rotating signals are injected as described in the thesis.

At the same time, the proposed method also has the following disadvantage:

- 1) The identified initial rotor position with the proposed method has a maximum $\pm 15^\circ$ error although this is already the best resolution compared with any existing initial rotor position estimations. This is an inherent drawback caused by the principle of the proposed method. As presented in the thesis, it will not result in significant decrease in the starting torque (about maximum 5%).

(b) Incremental inductance measurement

Conventionally, the incremental inductances including the cross-coupling effect can be directly measured with sensorless estimation technique by injecting either HF rotating voltage or HF pulsating voltage. However, as evaluated in Chapter 3, these conventional methods are affected by the delay effects and require a closed-loop position observer which is utilized for measuring the angle offset caused by the cross-coupling effect.

In this thesis, a novel incremental inductance measuring method by using both HF rotating voltage injection and pulsating voltage injection is proposed. The incremental inductances are calculated from the measured amplitudes of the positive sequence current and negative sequence current as a result of rotating voltage injection, and the amplitudes of the dq -axis HF currents as a result of pulsating voltage injections in dq -axes.

Compared to the conventional methods, the proposed method has several advantages such as

- 1) Immune to the delay effects.
- 2) No need of closed-loop position observer.

However, as the proposed method needs both rotating and pulsating voltage injections, the disadvantage compared to conventional method is

- 1) The measuring time is increased. However, it is noted that as the injection frequency is relatively high, the additional injection time, e.g. several milliseconds for 200Hz~1kHz injection frequency, is acceptable for the offline measuring process.

(c) Cross-coupling effect compensation

The cross-coupling effect can result in position error in saliency tracking based sensorless control. The relationship between L_{dqh} and L_{qh} is offline measured and used for cross-coupling effect compensation in pulsating voltage injection based sensorless control. However, this method is only used in pulsating voltage injection based sensorless control and its dynamic performance is limited by the PI and integral modules in the position observer. In this thesis, the offline measured ratio L_{dqh}/L_{qh} is extended for use in rotating voltage injection based sensorless control.

The offline measured error caused by the cross-coupling effect can also be used for online cross-coupling effect compensation. In this thesis, a novel method is proposed to offline measure the error by injecting HF pulsating voltages to the dq -axes, respectively. In heavily saturated conditions, due to reduced saliency level and also spatial harmonics, the compensation method by just using the offline measured error may lose stability and even has no convergence point. The stability margin and convergence criterion can be analysed by plotting the observer input signal with rotor position error. The observer input-position error trace should have positive slop zero-crossings at $\Delta\theta_r = 0$ for convergence. Besides, the interval between the positive slop zero-crossing and the adjacent negative slop zero-crossing should be large enough for sufficient stability margin. At heavily saturated conditions, the compensation method by modifying the observer input and the compensation method by adding the offline measured error can be combined to guarantee reliable convergence.

The proposed method has made some contributions to the sensorless control such as

- 1) The offline measured L_{dqh}/L_{qh} is extended for use in the cross-coupling effect compensation in HF rotating voltage injection based sensorless control.
- 2) The rotor position estimation error caused by the cross-coupling effect is offline measured by injecting HF pulsating voltages in the dq -axes.
- 3) It is verified by experiments that the compensation method using offline measured position error has better dynamic performance than that using the offline measured L_{dqh}/L_{qh} before the position observer.

There are still some unsolved issues for cross-coupling effect compensation with the proposed method such as

- 1) As mentioned in the Appendix 4, the offline measured $\frac{L_{qh}-L_{dh}}{L_{qh}+L_{dh}}$ can be used for cross-coupling effect compensation. However, it is only applicable for HF rotating voltage injection based sensorless control.
- 2) Considering cross-coupling effect compensation in heavily saturated conditions, the method for determining the optimal compensation angle and the compensation current, i.e. modifying the position observer input, is not given. The challenge is that the stability margin and convergence criterion are highly dependent on the magnetic characteristics of the machine.

7.2.2 Torque estimation and whole speed range sensorless operation

In position sensorless control, the torque control performance is affected by the sensorless performance, e.g. cross-coupling effect and magnetic saturation. Hence, the research of torque control is closely related to the research on sensorless position control. In this thesis, the torque control issues are shown in Fig. 7.3. Their implementation and novelties are summarized as follows.

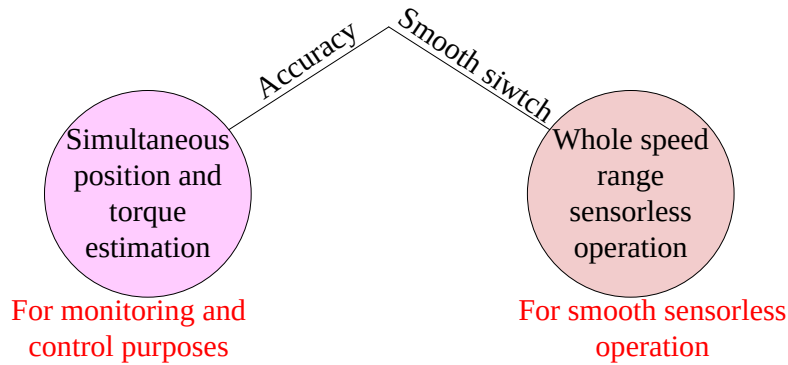


Fig. 7.3. Torque control issues and their significances for torque control in this thesis.

(a) Simultaneous rotor position and torque estimation

Electromagnetic torque is usually estimated by the product of flux linkage and current. At medium and high speeds, the flux linkage is estimated by integrating the back EMF, while at standstill and low speed, the flux linkage is obtained from the current model. For voltage model, the variation of resistance with temperature can reduce accuracy of the observed flux linkage. Hence, different online resistance identification method can be employed for improving the estimation accuracy. Whereas for standstill and low speed with current model based flux linkage observer, the inductances vary with i_d, i_q . The accuracy of the estimated flux linkage may be decreased if the inductance variation is not taken into account in the current model. Furthermore, at standstill and low speed, the fundamental model based inductance

identification method has deteriorated performance as the amplitude of the back EMF signal is very low. Hence, the HF injection technique is employed for identifying the incremental inductances at standstill and low speed which are further used to calculate the flux linkages.

For HF inductance identification with HF voltage injection, the rotor position is usually provided by a position sensor. In this thesis, a simultaneous sensorless rotor position and incremental inductance identification method is developed with the cross-coupling taken into account by injecting square wave voltage to the estimated dq -axes. The identified incremental inductances are then used to calculate the torque by approximating the dq -axis flux linkages.

The proposed method torque estimation method has the following benefits, including:

- 1) Simultaneous sensorless rotor position and torque estimation is achieved.
- 2) The proposed method has high torque estimation accuracy (about $\pm 5\%$) compared to the conventional method without considering the inductance variation (about $\pm 15\%$).

There are still imperfections for the proposed method such as

- 1) The PM flux linkage may vary with temperature. With merely HF voltage injection, it is difficult to identify the PM flux linkage. The recent research on temperature estimation using incremental inductance provides a good direction for solving this problem, which can be carried out as future work.
- 2) The injected HF square wave voltage on q -axis may increase the noise level.

(b) Whole speed range sensorless operation

For whole speed range sensorless operation, to avoid using two separate observers, single-observer solution is used by normalizing the position error signals from saliency model and fundamental model respectively. The normalization process often requires a scaling factor which may vary with the working currents. In this thesis, a whole speed sensorless operation method based on single position observer is proposed by making use of HF rotating voltage injection at standstill and low speed, and active flux model at medium and high speeds. The discrete rotor position signal at standstill and low speed is obtained by comparing the three phase HF current amplitudes. Then, the position error signal with unity-length is obtained from a vector tracking observer and the discrete positions. For active flux model, the unity position error signal is obtained by dividing the position error signal with the amplitude of the active flux which is relatively high at medium and high speeds. The unity error signals from saliency model and active flux model are then fused with a weighting function and sent to the PLL observer for position estimation.

The proposed whole speed sensorless operation method has the following benefits:

- 1) Single position observer based whole speed range sensorless operation is achieved.
- 2) Unity-length position error signals can be obtained without using a scaling factor.

However, the proposed method is not perfect since

- 1) A transition region is still required for smooth switching between the saliency model and the fundamental model.

7.3 What Can Be Done in the Future

This thesis has investigated some important areas of the sensorless torque control and achieved some results. Further research work for improving the sensorless torque control performance can be carried out as shown in Fig. 7.4.

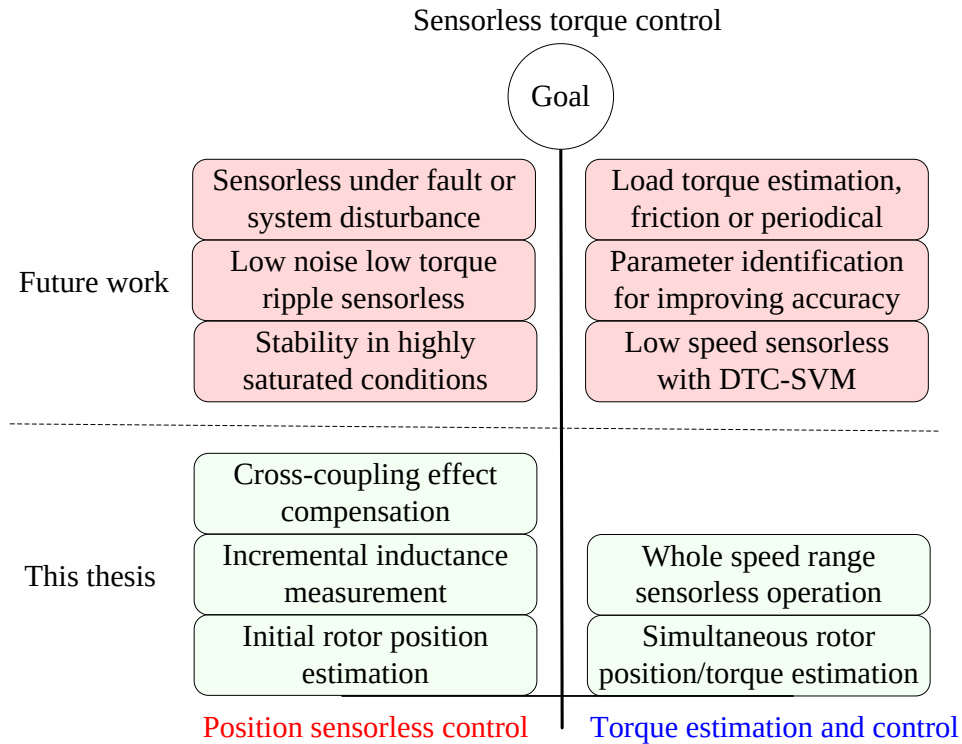


Fig. 7.4. Future work for sensorless torque control.

(a) Sensorless position estimation

Sensorless position estimation at highly saturated conditions still needs to be investigated. As presented before, although by angle compensation and observer input compensation, the convergence can be achieved, the compensation is highly dependent on the characteristics of the machine and may be disturbed by inverter nonlinearities and spatial inductance harmonics. The torque ripple and noise caused by HF voltage injection are unwanted in many applications.

Hence, the sensorless method by making uses of the PWM waveforms should be investigated due to its low noise emission. Besides, the robustness of the sensorless position estimation under fault conditions or voltage variations is also important for guaranteeing reliable operation.

(b) Torque control

Firstly, sensorless torque control with DTC strategy can be investigated. Comparison can be made between sensorless torque control with DTC and FOC, especially at standstill and low speed. For DTC, the HF injection based sensorless control can be combined with DTC-SVM. Then, the research on parameter identification, i.e. resistance, inductance and permanent magnet flux linkage, should be continued for improving the torque estimation accuracy. Finally, the research on load torque estimation, including friction, periodical load torque disturbance, can be carried out for effectively improving the speed control performance.

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APPENDICES

Appendix 1 Specification of the prototype machine

Table A1.1 Specification and dimensions of the IPMSM [CHE99]

Rated voltage (peak)	158V
Rated current (peak)	4.0A
Rated power	0.6kW
Rated torque	4.0 Nm
Number of pole pairs	6
Slot number	18
Number of series connected conductors/pole/phase	152 turns
Wire diameter (2 wires stranded)	0.36mm+0.51mm
Winding coil pitch	3 slots
Winding resistance per phase	6.0Ω (20°C)
Skew	1 slot-pitch
Stator outer radius	53.3 mm
Stator inner radius	31.00 mm
Stator outer radius	30.25 mm
Core length (stator)	32.00 mm
Core length (rotor)	30.00 mm
Shaft diameter	25.00 mm
Air gap length	0.75 mm
Magnet	UGIMAX 35HC1 ($B_r=1.17T$, $\mu_r=1.07$)
Stator steel	Transil 310-50
Rotor Steel	Transil 300-35

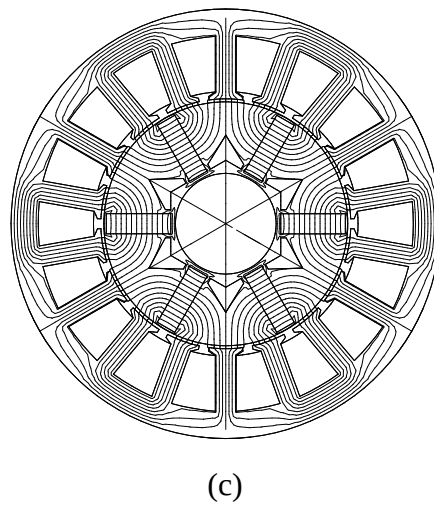
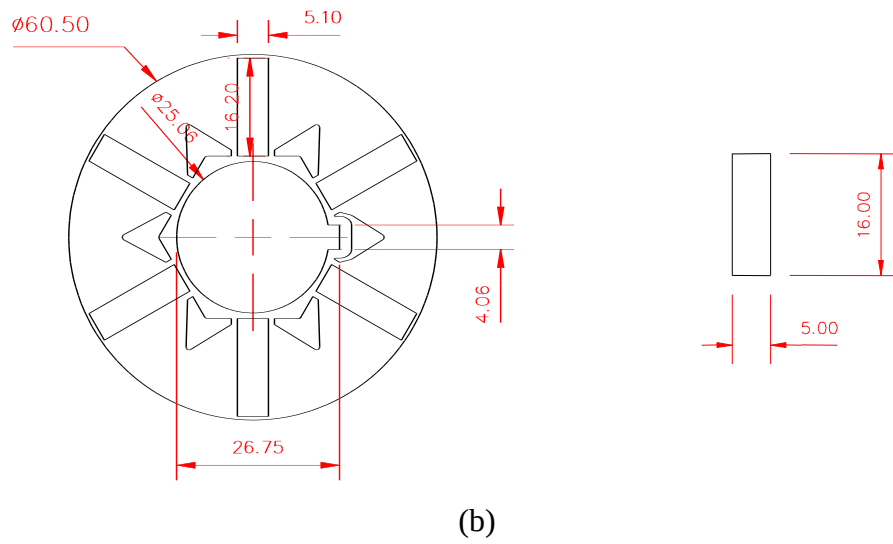
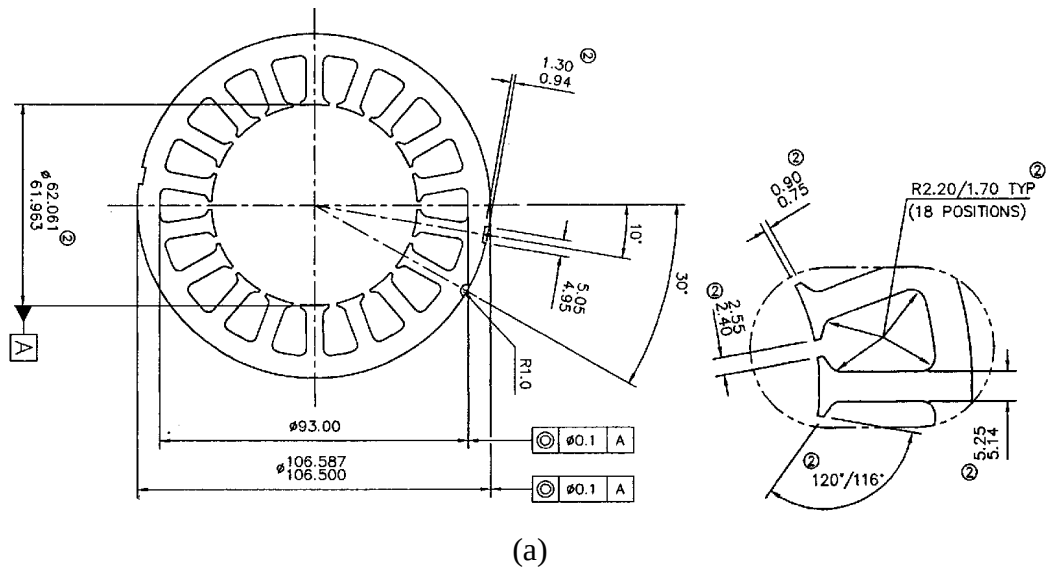
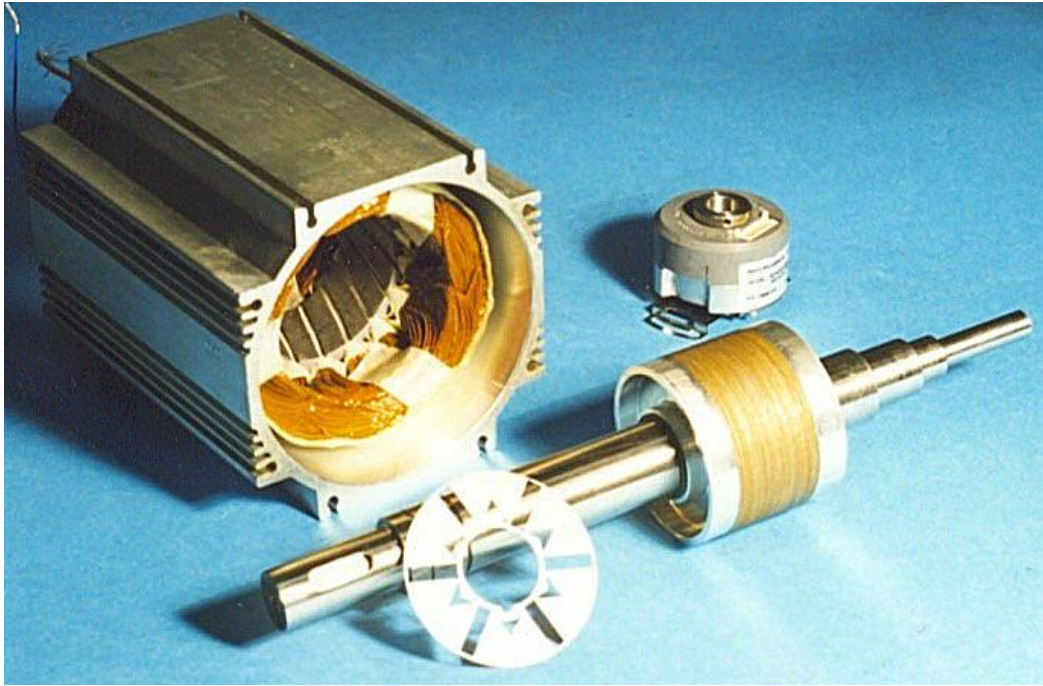
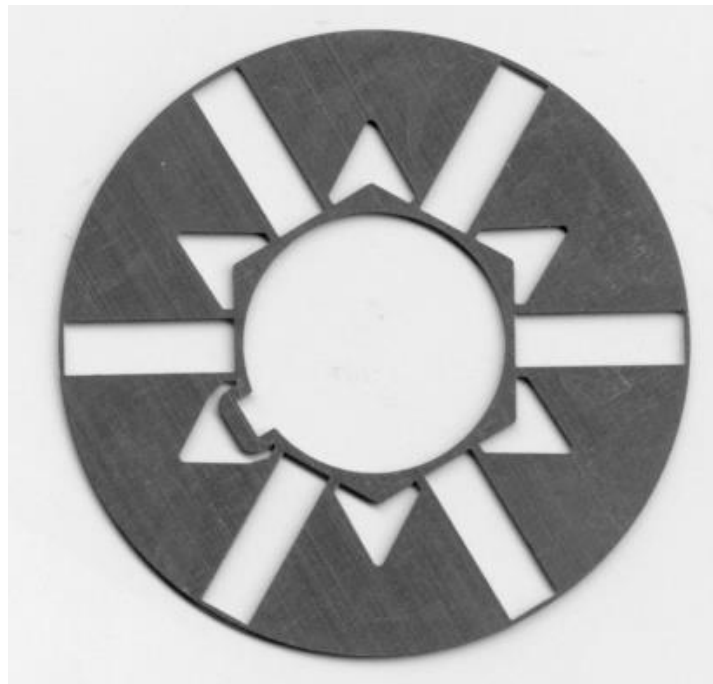


Fig. A1.1. Cross-section of prototype machine [CHE99]. (a) Stator. (b) Rotor. (c) Flux line distribution.



(a)



(b)

Fig. A1.2. Prototype machine [CHE99]. (a) Stator and rotor. (b) Rotor lamination.

Appendix 2 Finite element predicted incremental inductances

Table A2.1 FEA predicted d-axis incremental inductances

$L_{dh}(\text{mH})$	$i_d=-3\text{A}$	$i_d=-2\text{A}$	$i_d=-1\text{A}$	$i_d=0\text{A}$	$i_d=1\text{A}$	$i_d=2\text{A}$	$i_d=3\text{A}$
$i_q=-8\text{A}$	33.23	33.92	34.07	33.38	31.55	28.61	24.80
$i_q=-7\text{A}$	35.73	35.98	35.50	34.06	31.67	27.71	23.77
$i_q=-6\text{A}$	37.88	37.87	36.87	34.77	31.31	26.93	23.04
$i_q=-5\text{A}$	39.20	39.60	38.39	35.34	30.86	26.22	22.29
$i_q=-4\text{A}$	41.30	41.60	40.88	37.38	31.54	26.36	22.21
$i_q=-3\text{A}$	42.13	42.36	42.06	38.15	31.48	26.17	21.74
$i_q=-2\text{A}$	42.93	43.14	43.00	39.21	31.43	25.74	21.26
$i_q=-1\text{A}$	43.73	43.86	43.56	39.89	31.87	25.73	20.81
$i_q=0\text{A}$	44.27	44.41	44.09	40.56	32.19	25.67	20.43
$i_q=1\text{A}$	43.72	43.86	43.55	39.89	31.87	25.72	20.80
$i_q=2\text{A}$	42.92	43.12	43.00	39.20	31.44	25.74	21.25
$i_q=3\text{A}$	42.11	42.35	42.04	38.15	31.47	26.17	21.74
$i_q=4\text{A}$	41.28	41.58	40.86	37.38	31.55	26.37	22.22
$i_q=5\text{A}$	39.19	39.57	38.37	35.35	30.86	26.22	22.29
$i_q=6\text{A}$	37.88	37.87	36.85	34.76	31.31	26.94	23.04
$i_q=7\text{A}$	35.75	35.98	35.49	34.03	31.67	27.71	23.77
$i_q=8\text{A}$	33.25	33.92	34.06	33.36	31.51	28.61	24.80

Table A2.2 FEA predicted q-axis incremental inductances

$L_{qh}(\text{mH})$	$i_d=-3\text{A}$	$i_d=-2\text{A}$	$i_d=-1\text{A}$	$i_d=0\text{A}$	$i_d=1\text{A}$	$i_d=2\text{A}$	$i_d=3\text{A}$
$i_q=-8\text{A}$	17.56	17.61	17.38	17.23	17.05	16.71	16.21
$i_q=-7\text{A}$	22.27	21.47	20.62	20.07	19.60	18.94	18.34
$i_q=-6\text{A}$	28.16	26.45	24.71	23.66	22.72	21.83	21.13
$i_q=-5\text{A}$	33.48	32.18	30.07	28.02	26.60	25.37	24.43
$i_q=-4\text{A}$	38.72	37.45	35.52	33.42	31.42	29.76	28.5
$i_q=-3\text{A}$	43.56	42.86	41.57	39.54	37.5	35.62	33.8
$i_q=-2\text{A}$	50.48	50.03	49.46	47.92	45.34	43.18	41.38
$i_q=-1\text{A}$	58.18	57.38	56.3	55.36	53.8	52.03	50.12
$i_q=0\text{A}$	61.3	61.2	61.05	60.7	58.55	55.87	53.34
$i_q=1\text{A}$	57.5	56.6	55.67	54.45	53.1	51.18	49.45
$i_q=2\text{A}$	49.84	49.45	48.72	47.01	44.41	42.42	40.51
$i_q=3\text{A}$	43.06	42.26	40.9	38.86	36.84	34.93	33.08
$i_q=4\text{A}$	38.27	36.92	34.94	32.88	30.84	29.29	28.01
$i_q=5\text{A}$	33.01	31.57	29.53	27.61	26.13	24.99	24.06
$i_q=6\text{A}$	27.57	25.85	24.23	23.31	22.41	21.52	20.85
$i_q=7\text{A}$	21.70	21.03	20.28	19.77	19.35	18.71	18.12
$i_q=8\text{A}$	17.18	17.21	17.09	17.00	16.82	16.56	16.06

Table A2.3 FEA predicted mutual incremental inductances

$L_{dqh}(\text{mH})$	$i_d=-3\text{A}$	$i_d=-2\text{A}$	$i_d=-1\text{A}$	$i_d=0\text{A}$	$i_d=1\text{A}$	$i_d=2\text{A}$	$i_d=3\text{A}$
$i_q=-8\text{A}$	7.93	10.00	11.58	12.54	12.79	12.44	11.49
$i_q=-7\text{A}$	8.12	10.11	11.67	12.52	12.83	12.17	11.32
$i_q=-6\text{A}$	7.67	9.51	11.26	12.28	12.39	11.86	11.14
$i_q=-5\text{A}$	7.10	8.38	10.12	11.41	11.65	11.26	10.73
$i_q=-4\text{A}$	6.36	7.18	8.41	9.9	10.44	10.27	9.85
$i_q=-3\text{A}$	5.2	5.83	6.64	8.07	8.54	8.59	8.42
$i_q=-2\text{A}$	3.68	4.04	4.41	5.31	6.15	6.25	6.15
$i_q=-1\text{A}$	1.52	1.92	2.33	2.68	3.27	3.47	3.45
$i_q=0\text{A}$	-0.02	-0.03	-0.07	-0.12	-0.16	-0.12	-0.14
$i_q=1\text{A}$	-1.67	-2.11	-2.49	-2.91	-3.45	-3.67	-3.63
$i_q=2\text{A}$	-3.79	-4.14	-4.54	-5.51	-6.35	-6.39	-6.3
$i_q=3\text{A}$	-5.25	-5.92	-6.73	-8.17	-8.64	-8.69	-8.51
$i_q=4\text{A}$	-6.39	-7.23	-8.47	-9.95	-10.49	-10.31	-9.9
$i_q=5\text{A}$	-7.09	-8.38	-10.15	-11.44	-11.67	-11.27	-10.74
$i_q=6\text{A}$	-7.65	-9.53	-11.24	-12.28	-12.39	-11.86	-11.14
$i_q=7\text{A}$	-8.10	-10.08	-11.64	-12.48	-12.81	-12.17	-11.31
$i_q=8\text{A}$	-7.87	-9.94	-11.53	-12.50	-12.74	-12.44	-11.49

Appendix 3 Measured incremental inductances

Table A3.1 Measured d-axis incremental inductances

$L_{dh}(\text{mH})$	$i_q=-4\text{A}$	$i_q=-3\text{A}$	$i_q=-2\text{A}$	$i_q=-1\text{A}$	$i_q=0\text{A}$	$i_q=1\text{A}$	$i_q=2\text{A}$	$i_q=3\text{A}$	$i_q=4\text{A}$
$i_d=-3\text{A}$	41.00	42.02	43.15	43.93	44.33	43.58	43.05	42.30	40.80
$i_d=-2\text{A}$	41.10	42.50	43.00	43.68	44.00	43.40	42.99	42.10	41.20
$i_d=-1\text{A}$	39.70	42.02	42.90	43.78	43.99	43.40	42.80	41.80	40.03
$i_d=0\text{A}$	36.80	38.00	39.40	39.80	40.70	40.20	38.70	37.90	36.90
$i_d=1\text{A}$	31.10	31.20	31.00	31.90	32.00	31.80	31.20	31.25	30.80
$i_d=2\text{A}$	25.90	26.30	25.90	25.45	25.60	25.88	25.30	25.30	26.00
$i_d=3\text{A}$	21.50	21.10	20.90	19.68	19.88	19.80	20.50	20.50	21.50

Table A3.2 Measured q-axis incremental inductances

$L_{qh}(\text{mH})$	$i_q=-4\text{A}$	$i_q=-3\text{A}$	$i_q=-2\text{A}$	$i_q=-1\text{A}$	$i_q=0\text{A}$	$i_q=1\text{A}$	$i_q=2\text{A}$	$i_q=3\text{A}$	$i_q=4\text{A}$
$i_d=-3\text{A}$	39.40	43.20	50.30	58.00	61.89	57.55	51.00	43.50	38.60
$i_d=-2\text{A}$	38.40	42.11	50.40	57.01	60.20	57.80	49.90	42.60	38.20
$i_d=-1\text{A}$	36.00	41.90	49.50	56.20	60.20	56.70	49.01	41.10	35.65
$i_d=0\text{A}$	34.40	39.99	47.50	55.72	60.70	55.00	47.90	39.99	35.20
$i_d=1\text{A}$	31.50	37.68	45.00	52.90	57.77	52.52	45.90	36.80	32.20
$i_d=2\text{A}$	30.70	35.90	43.52	51.20	55.12	51.85	43.00	35.75	30.25
$i_d=3\text{A}$	28.60	34.01	40.60	50.50	53.16	50.25	42.20	35.20	27.90

Table A3.3 Measured mutual incremental inductances

$L_{dqh}(\text{mH})$	$i_q=-4\text{A}$	$i_q=-3\text{A}$	$i_q=-2\text{A}$	$i_q=-1\text{A}$	$i_q=0\text{A}$	$i_q=1\text{A}$	$i_q=2\text{A}$	$i_q=3\text{A}$	$i_q=4\text{A}$
$i_d=-3\text{A}$	6.70	5.70	3.80	1.90	0.20	-1.80	-4.20	-5.80	-6.80
$i_d=-2\text{A}$	7.30	6.10	4.30	1.60	0.30	-2.50	-4.30	-6.30	-7.00
$i_d=-1\text{A}$	8.10	7.00	4.60	2.60	0.10	-2.20	-6.50	-7.60	-9.30
$i_d=0\text{A}$	9.20	8.80	7.60	3.20	0.90	-4.90	-6.90	-7.40	-9.60
$i_d=1\text{A}$	10.90	9.80	7.80	4.60	0.50	-5.20	-7.50	-7.60	-9.70
$i_d=2\text{A}$	10.30	9.30	7.90	5.00	0.60	-5.20	-6.70	-7.50	-9.50
$i_d=3\text{A}$	10.00	8.90	8.20	5.20	0.20	-5.00	-6.70	-7.20	-9.60

Appendix 4 Applying offline measured γ for cross-coupling effect compensation

As can be seen, in HF rotating voltage injection based sensorless control, (4.37) for calculating $\Delta\theta_m$ with $\lambda = \frac{L_{dqh}}{L_{qh}}$ is quite complicated and requires many multiplication operations, which increase the calculation burden. Therefore, an alternative method is proposed by introducing a new factor γ , which is the ratio between the difference and sum values of the d -axis and q -axis HF inductances, i.e.

$$\gamma = \frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}} \quad (\text{A4.1})$$

In rotating voltage injection based sensorless control, the error $\Delta\theta_m$ caused by the cross-coupling effect can be calculated by γ and the amplitudes I_p, I_n as

$$\Delta\theta_m = 0.5 \tan^{-1} \left(\frac{2L_{dqh}}{L_{qh} - L_{dh}} \right) = \pm 0.5 \tan^{-1} \left(\sqrt{\frac{I_n^2}{I_p^2 \gamma^2} - 1} \right) \quad (\text{A4.2})$$

where the sign of $\Delta\theta_m$ is the same to that of L_{dqh} as L_{qh} is usually bigger than L_{dh} for IPMSM. The sign of L_{dqh} can be identified by the finite element calculation or experiment. Compared with (4.37) which uses λ for calculating the error $\Delta\theta_m$, (A4.2) is simpler and requires less multiplication operations.

γ at different working points can be obtained from pulsating voltage injections in d -axis and q -axis respectively. When HF pulsating voltage is injected to the actual d -axis, the ratio between L_{dqh}, L_{qh} is measured as

$$\lambda_1 = -\frac{i_{qh1}}{i_{dh1}} = \frac{L_{dqh}}{L_{qh}} \quad (\text{A4.3})$$

When HF pulsating voltage is injected to the actual q -axis, the ratio between L_{dqh}, L_{dh} is measured as

$$\lambda_2 = -\frac{i_{dh2}}{i_{qh1}} = \frac{L_{dqh}}{L_{dh}} \quad (\text{A4.4})$$

γ can be obtained from λ_1 and λ_2 as

$$\gamma = \frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}} = \frac{\lambda_2 - \lambda_1}{\lambda_2 + \lambda_1} \quad (\text{A4.5})$$

It can be seen that L_{dqh} is of the same sign as λ_1 or λ_2 . Therefore, the sign of $\Delta\theta_m$ in (A4.2) can be easily identified from the sign of λ_1 or λ_2 . The predicted and measured γ are shown in Fig. A4.1.

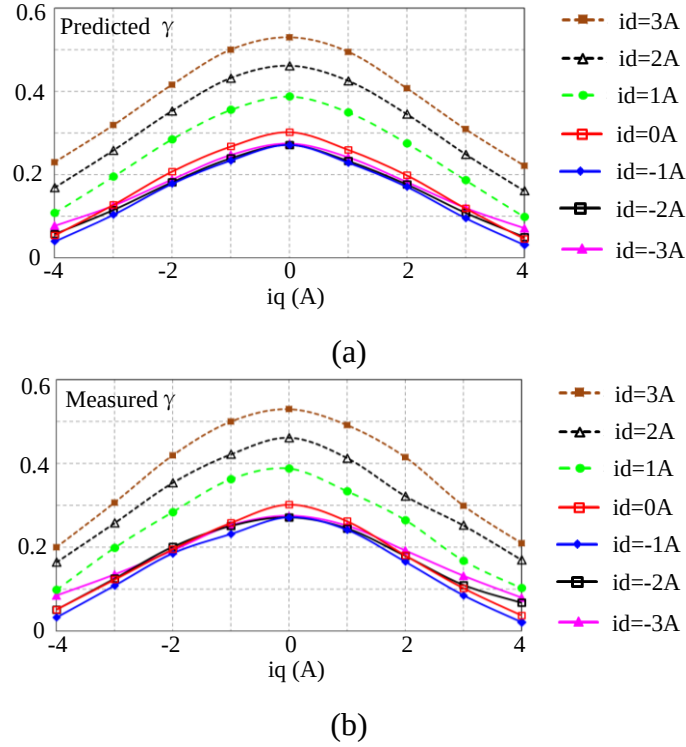


Fig. A4.1. FEA predicted and measured γ at different i_d, i_q . (a) Predicted. (b) Measured.

The cross-coupling effect compensation method with offline measured γ has simple calculation as shown in (A4.2). However, this method is not applicable for use in pulsating voltage injection based sensorless control. Therefore, its universality is poor compared to that of $\lambda = L_{dqh}/L_{qh}$. Table A4.1 compares the cross-coupling effect compensation methods with offline measured λ, γ and $\Delta\theta_m$ respectively.

Table A4.1

COMPARISON OF CROSS-COUPLING EFFECT COMPENSATION METHODS WITH λ, γ AND $\Delta\theta_m$.

	Measuring method	Universality/Dynamic performance		
		Rotating	Pulsating	Other
$\lambda = L_{dqh}/L_{qh}$	d-axis HF voltage injection $\lambda = -i_{qh1}/i_{dh1}$	✓/(High)	✓/(Low)	×
$\gamma = \frac{L_{qh} - L_{dh}}{L_{qh} + L_{dh}}$	d- and q-axis HF voltage injection $\lambda_1 = -i_{qh1}/i_{dh1}, \lambda_2 = -i_{dh2}/i_{qh2}$ $\gamma = \frac{\lambda_2 - \lambda_1}{\lambda_2 + \lambda_1}$	✓/(High)	×	×
$\Delta\theta_m$	d- and q-axis HF voltage injection $\Delta\theta_m = 0.5 \tan^{-1}(\frac{2\lambda_1\lambda_2}{\lambda_2 - \lambda_1})$	✓/(High)	✓/(High)	✓/(High)

Appendix 5 Measuring initial dq -axis HF inductances with LCR meter

The HF voltage equations for the machine in the two phase stationary reference frame are given by

$$\begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} = \begin{bmatrix} L_{aa} & M_{ab} & M_{ac} \\ M_{ba} & L_{bb} & M_{bc} \\ M_{ca} & M_{cb} & L_{cc} \end{bmatrix} p \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \quad (\text{A5.1})$$

where $L_{aa}, L_{bb}, L_{cc}, M_{ab}(M_{ba}), M_{ac}(M_{ca}), M_{bc}(M_{cb})$ are the incremental self and mutual inductances respectively as

$$\begin{aligned} L_{aa} &= L_1 - L_2 \cos(2\theta_r) \\ L_{bb} &= L_1 - L_2 \cos(2\theta_r + 2\pi/3) \\ L_{cc} &= L_1 - L_2 \cos(2\theta_r - 2\pi/3) \\ M_{ab} &= M_{ba} = M_1 - M_2 \cos(2\theta_r - 2\pi/3) \\ M_{bc} &= M_{cb} = M_1 - M_2 \cos(2\theta_r) \\ M_{ca} &= M_{ac} = M_1 - M_2 \cos(2\theta_r + 2\pi/3) \end{aligned} \quad (\text{A5.2})$$

where L_1, L_2, M_1, M_2 are the amplitudes of dc and second harmonics of HF self and mutual inductances.

The dq -axis HF inductances can be derived from (A5.1) by coordinate transformation as

$$\begin{aligned} L_{dh} &= L_1 - M_1 - \frac{L_2}{2} - M_2 \\ L_{qh} &= L_1 - M_1 + \frac{L_2}{2} + M_2 \end{aligned} \quad (\text{A5.3})$$

The dq -axis inductances can be obtained by measuring the line inductance between phases a and phase b at different rotor positions. The line voltage equation between ab is

$$u_a - u_b = L_{aa} p i_a + M_{ab} p i_b + M_{ac} p i_c - M_{ba} p i_a - L_{bb} p i_b - M_{bc} p i_c \quad (\text{A5.4})$$

When line voltage ab is applied, the line voltage equation is obtained from (A5.4) as

$$\begin{aligned} u_{ab} &= u_a - u_b = L_{aa} p i_a - M_{ab} p i_a - M_{ba} p i_a + L_{bb} p i_a \\ &= (L_{aa} - M_{ab} - M_{ba} + L_{bb}) p i_a \\ &= (2L_1 - 2M_1 + (L_2 + 2M_2) \cos(2\theta_r - 2\pi/3)) p i_a \end{aligned} \quad (\text{A5.5})$$

The maximum and minimum values of the measured line inductance can be expressed as

$$\begin{aligned} L_{\min} &= 2L_1 - 2M_1 - L_2 - 2M_2 = 2L_{dh} \\ L_{\max} &= 2L_1 - 2M_1 + L_2 + 2M_2 = 2L_{qh} \end{aligned} \quad (\text{A5.6})$$

As can be seen, the maximum and minimum values of the measured line inductance are two times of the dq -axis HF inductances respectively. The frequency and current of the LCR meter are 100Hz and 10mA respectively. The measured line inductances with position are show in Fig. A5.1.

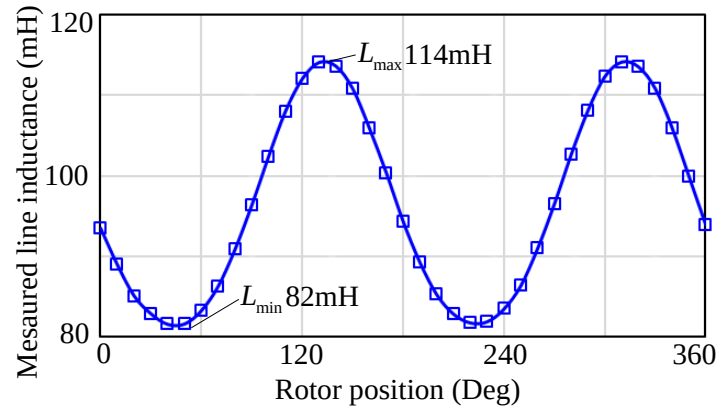


Fig. A5.1. Measured line inductance at different standstill rotor positions with LCR meter.

As can be seen, the dq -axis HF inductances at standstill can be obtained as $L_{dh} = 0.5L_{min} = 41mH$, $L_{qh} = 0.5L_{max} = 57mH$. These values are very close to the values measured with the proposed square wave voltage injection method as shown in Fig. 5.8.

Appendix 6 Simultaneous sensorless rotor position and torque estimation with HF rotating voltage injection

The HF model is the stationary reference frame without considering the cross-coupling effect (will be considered later) is given by

$$\begin{bmatrix} i_{\alpha h}^s \\ i_{\beta h}^s \end{bmatrix} = \frac{1}{L_{dh}L_{qh}} \begin{bmatrix} L_{\Sigma} + L_{\Delta} \cos(2\theta_r) & L_{\Delta} \sin(2\theta_r) \\ L_{\Delta} \sin(2\theta_r) & L_{\Sigma} - L_{\Delta} \cos(2\theta_r) \end{bmatrix} \int \begin{bmatrix} u_{\alpha h}^s \\ u_{\beta h}^s \end{bmatrix} dt \quad (\text{A6.1})$$

For HF rotating voltage injection based sensorless control, sinusoidal voltages $u_{\alpha h}^s, u_{\beta h}^s$ are injected to the stationary reference frame as

$$u_{\alpha h}^s + ju_{\beta h}^s = V_h e^{j\omega_h t}, \theta_h = \omega_h t \quad (\text{A6.2})$$

where V_h, ω_h, θ_h are the amplitude, angular frequency and phase angle of the injected rotating voltage. Substituting (A6.1) into (A6.2) yields

$$i_{\alpha h}^s + ji_{\beta h}^s = I_p e^{j(\theta_h - \pi/2)} + I_n e^{-j(\theta_h - 2\theta_r - \pi/2)} \quad (\text{A6.3})$$

$$I_p = \frac{V_h (L_{dh} + L_{qh})}{2\omega_h L_{dh} L_{qh}}, I_n = \frac{V_h (L_{qh} - L_{dh})}{2\omega_h L_{dh} L_{qh}} \quad (\text{A6.4})$$

where I_p, I_n are the amplitudes of the positive sequence current and negative sequence current.

The rotor position can be estimated by heterodyning demodulation as shown in Fig. A6.1.

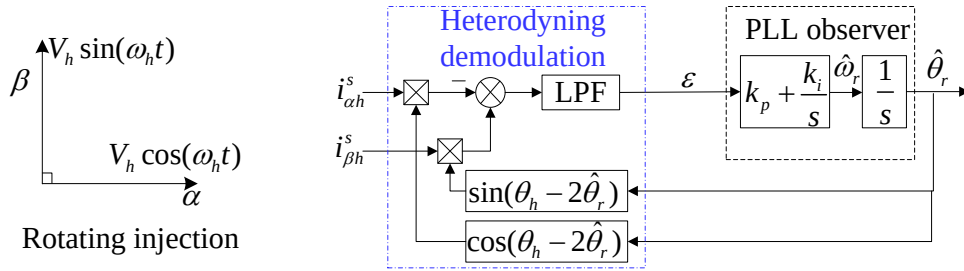


Fig. A6.1. Sensorless position estimation with HF sinusoidal rotating voltage injection.

For sinusoidal rotating voltage injection, the HF inductances can be obtained simultaneously from (A6.4) as

$$L_{dh} = \frac{V_h}{\omega_h (I_p + I_n)}, L_{qh} = \frac{V_h}{\omega_h (I_p - I_n)} \quad (\text{A6.5})$$

As can be seen from (A6.5), the sinusoidal rotating voltage injection based sensorless method is capable of simultaneously position estimation and online HF inductance identification. In this section, the identified HF inductances will be used for torque estimation at low speed. With the increase of load current and saturation level of the magnetic material, the cross-coupling effect which has non-negligible impacts on both position and torque estimation should be accounted for.

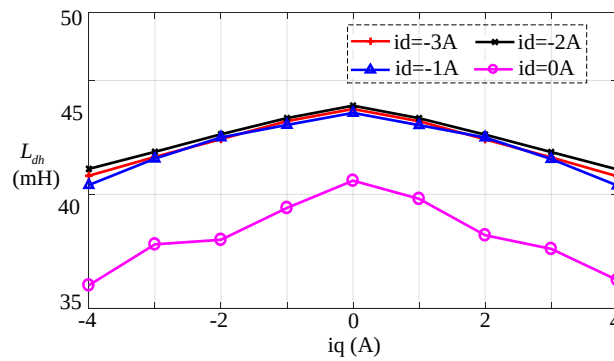
The cross-coupling effect compensation method by adding a compensation angle $-0.16i_q$ is also applicable for rotating voltage injection based sensorless control as the compensation angle is independent of the injection signals [WAN19]. The HF inductances considering the cross-coupling effect can be calculated from I_p, I_n, k_c, i_q as [XU14]

$$L_{dh} = \frac{V_h \left(I_p - I_n / \sqrt{1 + \tan^2(2k_c i_q)} \right)}{\omega_h (I_p^2 - I_n^2)} \quad (\text{A6.6})$$

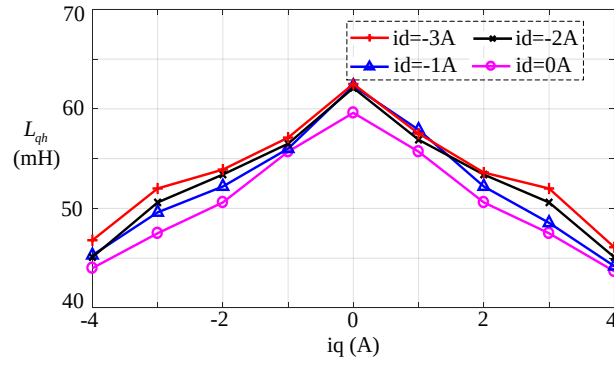
$$L_{qh} = \frac{V_h \left(I_p + I_n / \sqrt{1 + \tan^2(2k_c i_q)} \right)}{\omega_h (I_p^2 - I_n^2)} \quad (\text{A6.7})$$

$$L_{dqh} = \frac{V_h I_n \tan(2k_c i_q)}{\omega_h (I_p^2 - I_n^2) \sqrt{1 + \tan^2(2k_c i_q)}} \quad (\text{A6.8})$$

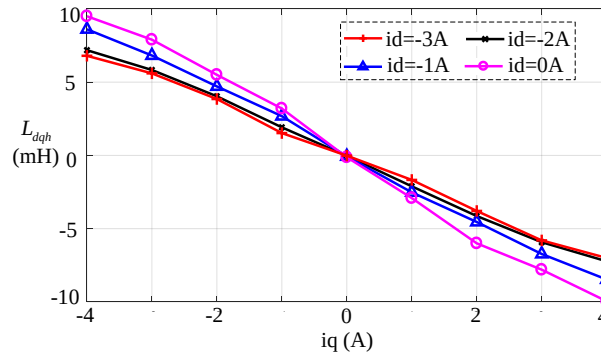
Sensorless torque estimation with sinusoidal rotating voltage injection based sensorless control is also carried out. The injected HF rotating voltage is chosen as 10V/250Hz. The sensorless identified HF inductances at different i_d, i_q are shown in Fig. A6.2.



(a)



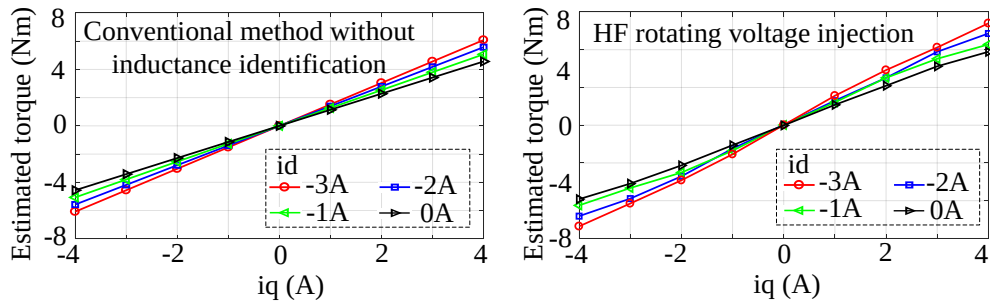
(b)



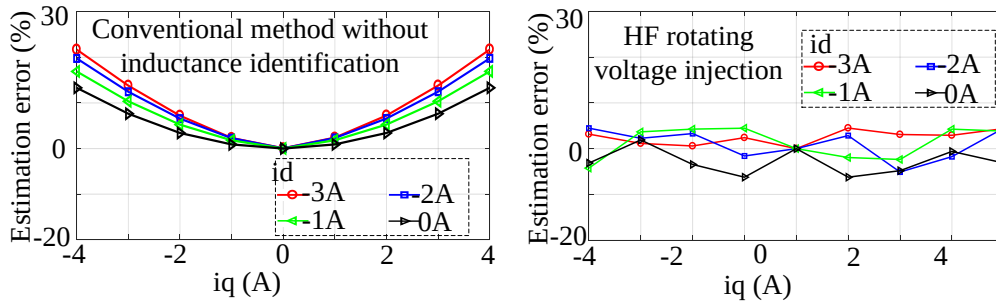
(c)

Fig. A6.2. Sensorless online identified HF inductances with high frequency rotating voltage injection at different i_d, i_q . (a) L_{dh} . (b) L_{qh} . (c) L_{dqh} .

The estimated torque and torque estimation errors with the sensorless identified HF inductances from HF rotating voltage injection are shown in Fig. A6.3.



(a)



(b)

Fig. A6.3. Comparison of the torque estimation performances without/with identified HF inductances from HF rotating voltage injection. (a) Estimated torque. (b) Estimated errors.

As can be seen from Figs. A6.3(a) and (b), the rotating voltage injection based sensorless methods can estimate the torque with errors less than 5%, while the estimated error without using the identified HF inductances can reach up to 20%. Therefore, the HF rotating voltage injection method is capable of simultaneous position and torque estimation with satisfactory performance. However, as the injection frequency of sinusoidal voltage is generally lower than that of square wave voltage injection due to limited sampling points during one injection cycle, the dynamic position estimation performance is generally lower than that of square wave voltage injection with higher injection frequency. Furthermore, with square wave voltage injection, the LPFs for fundamental currents feedback can also be removed due to higher injection frequency compared to the cut-off frequency of the current control loop [YOO11].

Experiments are carried out to compare the dynamic performance of rotating voltage injection and square wave voltage injection based sensorless torque and position estimation performances. The machine is sensorless started from standstill to 60rpm under rated load torque with the proposed method in 2 seconds.

The current control bandwidths for the rotating voltage injection and the proposed method (square wave voltage injection) are all set as 200Hz. The speed control PIs are all set as $k_p = 0.1, k_i = 0.2$, and the position observer PIs are all set as $k_p = 500, k_i = 1000$. For rotating voltage injection, low pass filters of cut-off frequency 100Hz are used in current feedback for reducing interfere between the HF currents (250Hz) and the current control loop. For the proposed method, the low pass filters can be eliminated due to higher injection frequency of the square wave. The speed, phase current, position error with the HF rotating voltage injection based sensorless method and the proposed method are shown in Fig. A6.4(a). The estimated torque and torque estimation errors for the rotating voltage injection based method and the proposed method are shown in Fig. A6.4(b).

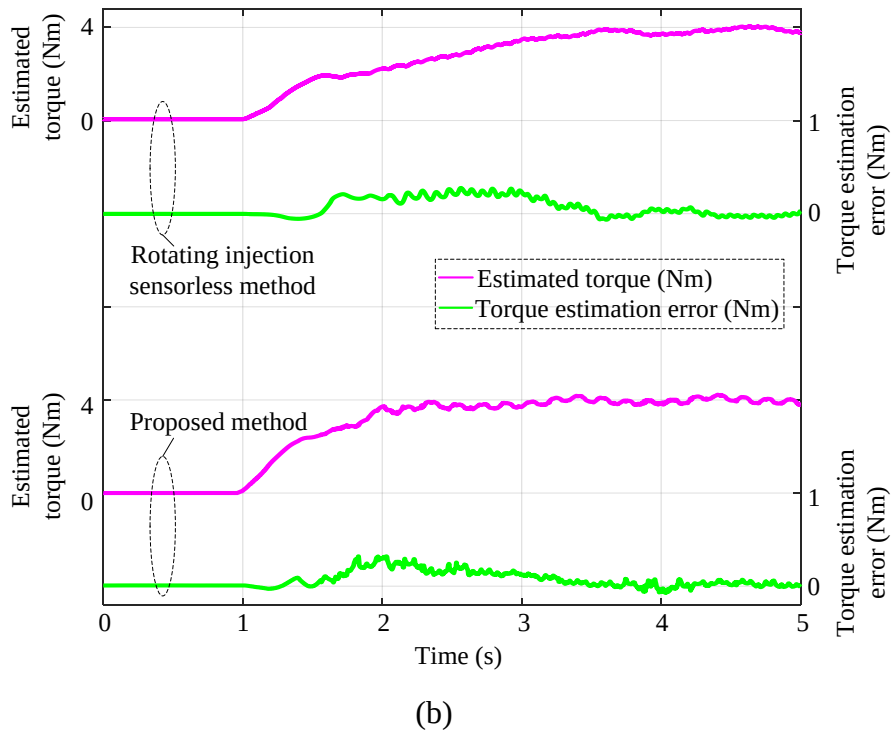
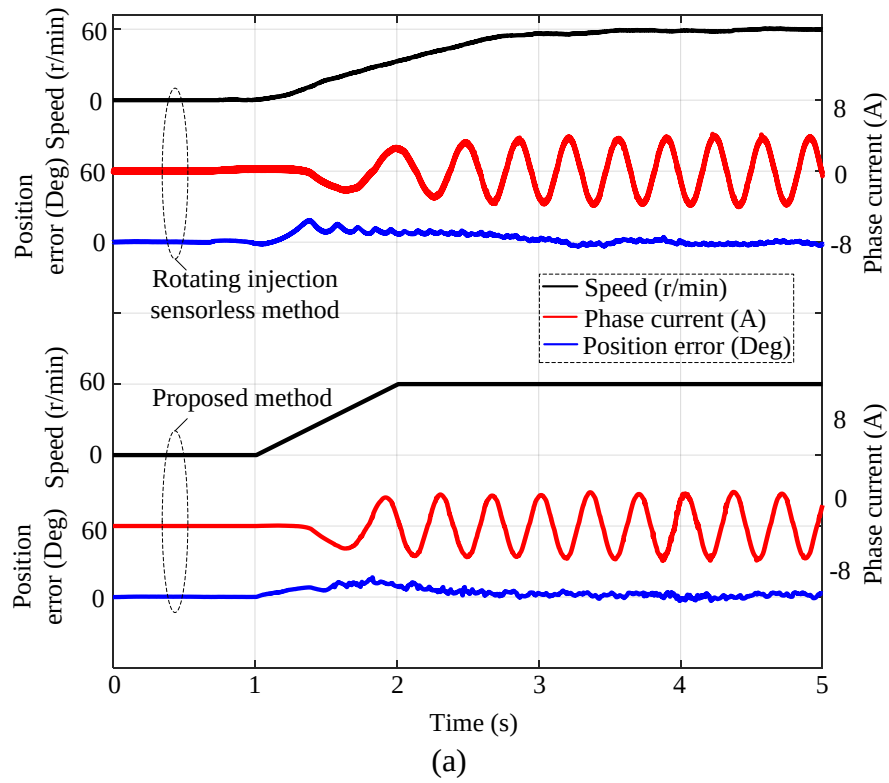


Fig. A6.4. Sensorless position and torque estimation performance during startup for the method with HF rotating voltage injection and the proposed method. (a) Speed, phase current, position estimation error. (b) Estimated torque and torque estimation error.

As can be seen from Fig. A6.4(a), the maximum transient position errors for rotating voltage injection and the proposed methods are both below 15 degrees. However, the position error

with HF rotating voltage injection based sensorless control method takes longer time to return back to zero. The startup time, about 2s, is also longer than that of the proposed method, about 1s. The better dynamic performance with the proposed method is due to higher injection frequency and elimination of the low pass filters compared to rotating voltage injection.

As can be seen from Fig. A6.4(b), the actual load torque is increased with the rotor speed as the output voltage of the dc generator is proportional to the rotation speed. For the proposed method, the estimated torque is increased faster than the method with rotating voltage injection due to its better dynamic performance in current control. The torque estimation errors for both the rotating voltage method and the proposed method are less than 0.2Nm as shown by the torque estimation errors. However, the torque estimation error for the proposed method returns to nearly zero more quickly than the method with rotating voltage injection based method. This means the proposed method with higher injection frequency has better dynamic performance compared to HF rotating voltage injection.

Appendix 7 Experimental test rig specifications

The test rig is composed of three parts: the dSPACE system, the measuring and sampling system, the power module and the load system. They are described as follows.

1) dSPACE system

The dSPACE platform comprises highly flexible and modular systems for hardware-in-the-loop (HIL) and rapid control prototyping applications. The test dSPACE in this thesis is composed of a processor board DS6001, a digital waveform output board DS5010, an incremental encoder interface board DS3001, and a high speed A/D board DS2004.

The Matlab/simulink program is compiled and downloaded to the DS6001. The DS6001 board equipped with an Intel Core i7-6820EQ quad-core processor features maximum 2.8GHz frequency and excellent real-time performance, with minimum jitter and latency. It is the perfect choice for use cases with high requirements regarding computation power and data bandwidth. For motor control application which is usually executed with a frequency of 10kHz, the execution time in DS6001 is negligible.

The DS5010 module generates the digital PWM drive signals according to the voltage references. The digital pulses are then sent to the drive board which transforms the low level PWM signals to high level gate drive signals through optical couplers. The PWM is generated with space vector pulse width modulation (SVPWM) strategy. The dq-axis voltage references are transformed to the two phase stationary reference frame. The SVPWM strategy generate ratios of each phase according to seven sectors distribution, i.e. $U_0, U_a, U_b, U_7, U_b, U_a, U_0$. The voltage vectors U_a, U_b are the two active vectors and U_0, U_7 are the two zero voltage vectors.

2) Measuring and sampling system

The phase currents are sampled by two hall-sensor effect transducers (LEM LA55-P). The primary nominal current is 50A, and the secondary nominal current is 50mA. A resistance is connected to the secondary winding and the voltage drop is measured to obtain the secondary current. To improve the sampling accuracy, there are three turns of the primary winding. The current sampling circuit is shown in Fig. A7.1.

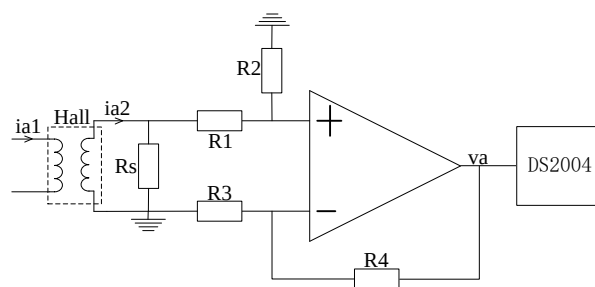


Fig. A7.1. Current sampling circuit.

The output v_a is sent to the DS2004 module which features $\pm 5V$ sampling range and 16bits resolution. The relationship between the primary current and v_a is

$$v_a = 5/8i_a \quad (A7.1)$$

The maximum sampling range is $[-8A, 8A]$. Resolution of the analog to digital converter is 16 bits. The minimum current sampling resolution is $\frac{16A}{2^{16}} = 0.244mA$. If a higher current range is required, the amplification factor should be adjusted by using a higher ratio $R4/R3$. The dc-bus voltage is sampled by a hall-effect sensor LEM LV 25-P. The primary current is 10mA, and the secondary current is 25mA. A resistance is connected to the dc-bus voltage to transform dc-bus voltage signal to current which is sensed by the secondary current. The same sampling circuit in Fig. A7.1 can be used. The relationship between the sampled voltage and the dc-bus voltage is

$$v_{dcsample} = 0.025v_{dc} \quad (A7.2)$$

The rotor position is measured by an incremental encoder, the out signals of which are sent to the DS3001 module. In the Simulink environment, the output of the DS3001 is an index number indicating the pulses that have been detected. As the incremental encoder only provides position increment, a position calibration block is required in the program to obtain the absolute position. This is done by subtracting the position offset when the rotor is locked at the zero position.

3) Power module

The inverter used in the thesis is a prototype driver with IGBT module SKS 35F B6CI 20 V12. The power rating of the IGBTs used in this inverter is 35A/1200V. The auxiliary power, including $\pm 15V$ for sampling, 15V for gate drive is provided by a separate dc source.



Fig. A7.2. Power module of the driver.

4) Load

The load of the test machine is provided by a dc generator which has two sets of windings, i.e. excitation winding and load winding. The excitation winding is connected to an external dc voltage source and the load winding is connected to an external variable resistor. The load torque is adjusted by varying the resistor. There is only one test machine and two load machines for the experiment in this thesis. For load machine one (used in Chapters 2, 3, 4 and 6), torque sensor cannot be installed due to mechanical constraint, while for load machine two (used in Chapter 5), the shaft torque can be measured by the torque sensor which is installed between the test machine and the load machine. The nameplate of torque sensor is shown in Fig. A7.3.



Fig. A7.3. Nameplate of the torque sensor.

Appendix 8 Papers published during PhD study

Journal papers

1. B. Shuang and Z. Q. Zhu, "A Novel Sensorless Initial Position Estimation and Startup Method," *IEEE Trans. Ind. Electron.* vol. 68, no. 4, pp. 2964-2975, Apr. 2021.
2. B. Shuang and Z. Q. Zhu, "Improved cross-coupling effect compensation method for sensorless control of IPMSM with high frequency voltage injection," *IEEE Trans. Energy Convers.*, early access, Jun. 29, 2021, doi: 10.1109/TEC.2021.3093361.
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