CHAPTER 2
LITERATURE REVIEW

2.1 Introduction

Literature review of some of the relevant research in the areas of multi-phase machines and multi-motor drives has been detailed in Jones (M. Phil.-2002), Jones (2005) and Iqbal (2006). The various aspects covered have included advantages of multi-phase machines over their three-phase counterparts, modelling and control of multi-phase machines, three-phase multi-motor drives, etc. This chapter reviews the latest progress made in the area of multi-phase machine drives, highlighting several other relevant aspects and addressing the features of the drives not covered in Jones (2002), Jones (2005) and Iqbal (2006). Although the references are available for drive phase numbers equal to five, six, seven, nine and even fifteen, the literature review focuses on five-phase and five-phase two-motor drives. The reasons for concentrating on these two drives are twofold. Firstly, the thesis examines these two specific drives only. Secondly, these are the two most frequently discussed drives in literature.

The chapter gives an overview of the following aspects relevant for the thesis: application areas of multi-phase machines, the speed sensorless techniques for multi-phase induction machines, and the modelling and control of multi-phase and three-phase inverters. It additionally includes a review of the latest developments in five-phase and six-phase motor drives.

2.2 Application areas of multi-phase machines

Electric traction, ship propulsion, more-electric aircraft, hybrid electric vehicles and battery-powered electric vehicles have rapidly emerged during the last few years as the main potential application areas for multi-phase motor drives. The reasons behind this are primarily twofold. In high power applications (such as ship propulsion) use of multi-phase drives enables reduction of the required power rating per inverter leg (phase). In safety-critical
applications (such as more-electric aircraft) use of multi-phase drives enables greater fault
tolerance, which is of paramount importance. Finally, in electric vehicle and hybrid electric
vehicle applications, utilisation of multi-phase drives for propulsion enables reduction of the
required semiconductor switch current rating (although these drives are not characterised with
high power, low voltage available in vehicles makes current high).

One particular interesting potential application of multi-phase motor drives is for an
integrated starter/alternator in hybrid electric vehicles and ordinary vehicles with combustion
engines. This idea enables replacement of two electric machines with a single machine. Due
to its numerous good features, an induction machine is a candidate for this role. A single
induction machine was proposed as the integrated starter-generator set in Miller et al. (2001a,
2001b) and Miller and Stefanovic (2002). A special method of control, termed pole-phase-
modulation (PPM) speed control, was developed for use in passenger hybrid electric car. An
integrated starter-generator must suffice the very differing needs of both the starter (high
starting torque at low speed) and the generator (wide speed constant-power range with fast
voltage control). For this purpose, a nine-phase inverter was utilised and a nine-phase twelve-
pole induction motor with toroidal winding was reconfigured into three-phase four-pole
machine under vector control conditions by using PPM. By using a multi-phase machine in
conjunction with this discrete speed control method it became possible to meet the
requirements for an integrated starter-generator.

A six-phase permanent magnet (PM) synchronous motor with a specially designed
stator and an outside rotor was investigated by Rattei (2001) for use in parallel hybrid electric
vehicle drive for propulsion purposes in conjunction with internal combustion engine. The
proposed motor nominal power was 8 kW at 2000 rpm and the maximum speed was 6000
rpm. The stator and rotor structures were designed with two poles and ten poles, respectively.
The proposed structure offers several advantages over conventional surface mounted
permanent magnet synchronous machine. Another example of an electric vehicle related
proposal for utilisation of a six-phase machine is the work of Jiang et al (2003), where a novel
PWM technique has been proposed. The method was aimed at six-phase induction motor
drives, with the stator winding consisting of two sets of three-phase windings which were
used to form dual-star six-phase winding with separate neutral points. The idea behind the
concept was to extend the constant-power speed range. The proposed sinusoidal PWM
reduces the phase current and dc link current harmonics and thus can increase the battery life
which forms a considerable fraction of the operating cost of an electric vehicle.
Simoes and Vieira (2002) have proposed a five-phase high-torque low-speed PM brushless machine, which can be used as an in-wheel motor arrangement for electric vehicles. Very much the same proposal is contained in Simoes et al (2001) as well. The machine is in this case of the so-called brushless dc motor design, meaning that the spatial distribution of the MMF in the air-gap is trapezoidal and the machine therefore requires square-wave currents for normal operation.

As already noted, there is considerable ongoing research interest in the utilisation of multi-phase machine in safety-critical applications, such as more-electric aircraft fuel pump drives and flight control surface actuators. The conventional technology is to run the aircraft fuel pump from the engine gearbox. The fuel pumps are typically designed for maximum-output fuel flow rate at low engine speed, which is the take-off condition. At high altitude the engine runs at very high speed but requires less fuel. Consequently, a fuel by-pass system is used to return the excess fuel into the tank. Thus a need is felt to have an independent variable speed electric drive for the fuel pump. One of the most important aspects of such drives is the fault tolerant property to meet the crucial high reliability requirement. A four-phase six-pole, 16 kW, 15000 rpm PM machine was proposed by Mecrow et al (2003) to be used in more-electric aircraft fuel pump drive. The four phases of the machine are supplied independently by four single-phase inverters. The four phases are essentially isolated physically, magnetically and thermally, leading to a fault tolerant high reliability motor structure. Only alternate stator teeth carry a winding in this type of PM machine. A comprehensive design of such modular multi-phase PM brushless machines for use in more-electric aircraft fuel pumps and flight control surface actuators was presented by Ede et al (2002). The design is based on higher value of reactance so as to limit the short circuit current. It was pointed out that the machines have to be overrated if they are to be used as fault tolerant structures. The overrating factors for four, five and six-phase machines are found to be 1.33, 1.25 and 1.2, respectively. Thus a six-phase machine is suggested to be the optimum choice. Green et al (2003) have highlighted the problem of using the position sensor in more-electric aircraft fuel pump fault tolerant drive. The drive utilises a 16 kW, 13000 rpm six-phase permanent magnet motor with six independent single-phase inverters supplying each of the six phases. The authors proposed an alternative sensor-less drive scheme. The proposed technique makes use of flux linkage-current-angle model to estimate the rotor position.

A comprehensive description of possible three-phase fault tolerant motor drive structures was given by Ertugrul et al (2002) for use in safety-critical applications such as aerospace, nuclear power plants, military and medical services. The proposed modular motor
drive (two-motor system with separate controllers and inverters for each machine) system configuration for three-phase permanent magnet ac motor increases the reliability through redundancy. Another proposed scheme is to feed the three windings of the three-phase motor from three single-phase inverters rather than a standard three-phase inverter.

The use of multi-phase electric generator sets for providing power to electric drives in ship propulsion is being investigated as well. Calfo et al (2002) have presented the comparative study regarding the use of conventional turbo synchronous generator and specially designed synchronous generators for such an application. It was shown that by using multi-phase (fifteen-phase) generator system, there is no need for special phase shifting transformers, which reduces the weight of the overall generating system. Further, by using high frequency generators, high pole number construction can be used, which leads to the reduction in the volume of the generator and associated transformer (if required), and vibration (resulting in more robust solution, resistant to mechanical shock). In aircrafts and ships the dc power is normally supplied by ac/dc converters. This requires special filters to eliminate the current ripple. Weiming et al (2002) have proposed an integrated multi-phase generator (with three-phase and twelve-phase windings) design to provide simultaneous ac and dc power generation. The twelve-phase winding has embedded rectifier to generate dc power and three-phase winding provides ac power. Use of multi-phase generator reduces the rectified dc voltage ripple as the ripple frequency is proportional to the number of pulses in the rectifier output, which is twice the number of phases.

2.3 The latest developments in the areas of quasi six-phase (dual three-phase) and five-phase single-motor drives

2.3.1 Quasi six-phase (dual three-phase) motor drives

Among different multi-phase drive solutions, so-called quasi six-phase machines (having dual three-phase windings with 30° spatial phase displacement) are probably considered most often in the literature. This is so since such a machine can be supplied from two three-phase inverters, thus eliminating the need for a design of a customised power electronic supply.

The major drawback of a true six-phase machine (with 60 degrees spatial displacement between any two consecutive phases) is the production of large stator current harmonics, if the machine is fed by two six-step VSIs. These generate additional losses resulting in increase in the size and cost of the machine and the inverter. These negative effects reduce significantly if quasi six-phase machine is used instead of a true six-phase
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machine [Klingshirn (1983)]. These, together with the easiness of obtaining a quasi six-phase machine from an existing three-phase machine, are the main reasons why quasi six-phase machines are predominantly dealt within literature.

The process of dynamic modelling of a quasi-six phase induction machine decomposes the original six-dimensional phase space into three two-dimensional orthogonal subspaces (d-q, x-y and 0+-0-). The neutral points of the two three-phase windings are in vast majority of cases isolated. In such a case the 0+-0- sequence components are eliminated since they cannot appear in any of the two star-connected three-phase windings. The d-q current components are flux and torque producing currents, while x-y stator current components appear due to stator current harmonics and are non flux/torque producing current components. Modelling procedures for quasi six-phase induction and synchronous machines have been discussed in detail in Bojoi et al (2002a), Hadiouche et al (2000), Lyra and Lipo (2002) and Benkhoris et al (2002). The dynamic decoupled model of a double-star synchronous machine with general spatial displacement between two three-phase windings, including decoupling algorithm, was presented by Benkhoris et al (2002). Two decoupling algorithms (called 'complete diagonalisation' and 'partial diagonalisation') were developed, based on the coupling due to mutual inductance between windings and due to rotational terms, respectively. The model contains two pairs of stator d-q axis currents (one pair for d-axis and another pair for q-axis) for use in a vector control scheme. The developed model was verified by simulation studies and by experimental investigation. A novel d-q-0 synchronous reference frame model of a quasi six-phase induction machine, which includes the third harmonic current injection effect, was presented by Lyra and Lipo (2002). Third harmonic current injection makes the stator current waveform and resulting air gap flux almost rectangular leading to an improvement in the flux density and an increase in the output torque of the machine. The model was validated by simulation and by experiments.

A complete model of a dual-stator induction motor, which consists of a single rotor winding and two identical stator windings, was presented by Pienkowski (2002). Two stator windings have in this case no magnetic coupling between them but they are magnetically coupled with the rotor winding and are supplied from the same or independent three-phase supply of the same frequency.

The major advantages of using a quasi six-phase machine, in contrast to a true six-phase machine, are once again shown by Singh et al (2003). The model developed in an arbitrary reference frame takes into account the common mutual inductances between two three-phase winding sets. The detailed comparison of performance of quasi six-phase and true
six-phase induction machine has been presented. The simulation results are given for acceleration transients and steady state behaviour under six-step and PWM inverter supplies. The results show a significant reduction in torque and rotor current pulsation in quasi six-phase configuration. The frequency of torque pulsation in quasi six-phase machine is found to be doubled compared to true six-phase machine case. The reason for the reduction in the torque pulsation is due to the complete elimination of $6k \pm 1 \ (k = 1,3,5,\ldots)$ harmonics from the air-gap mmf. Further investigation under loaded conditions reveals that the torque pulsation is minimum in quasi six-phase machine. However, very little difference in the behaviour of the two configurations is found under PWM supply condition.

Two solutions are in general available as a mean of improving a six-phase motor drive performance. These include modified machine design and application of an appropriate PWM technique. Attempts have been made to exploit both possible solutions, with more effort being put into the second one. Hadiouche et al (2002) have proposed a new winding configuration for dual three-phase (i.e. dual-stator, as the machine is often referred to) induction machine aiming to maximise the stator slot leakage inductance to limit the harmonic current. The basic principle of the winding design consists in the placement of the conductors of stator 1 (the first three-phase winding) and stator 2 (the second three-phase winding) in two alternating slots with $8/9$ pole pitch. This winding arrangement increases the $x$-$y$ leakage inductance and thus limits the stator harmonic current.

A digital PWM technique called double zero-sequence injection modulation technique was proposed by Bojoi et al (2002a) to act on $x$-$y$ components of voltage to limit the harmonic current in vector controlled quasi six-phase induction motor drive with isolated neutral points. Here one six-phase inverter is considered as combination of two identical three-phase inverters sharing a common dc link. Two sets of three-phase reference voltages are obtained from two reference voltage vectors, shifted in phase by $30^\circ$ (electrical). This method was shown to produce satisfactory results with easy implementation on a low cost DSP platform. A double d-q synchronous reference frame current control, which uses four simple PI regulators instead of six, is proposed for inverter current control. The two sets of stator currents are independently controlled to compensate the inherent asymmetries in the two three-phase windings.

A comprehensive comparison of performance of quasi six-phase induction machine drive for six types of digital PWM techniques (space vector, multi-level space vector, vector space decomposition, multi-level vector space decomposition, vector classification and
double zero-sequence injection) based on simulation and experimental study was given by Bojoi et al (2002b). The study reveals that the vector classification and double zero sequence injection PWM techniques offer good results by minimizing the stator harmonic currents and simultaneously reduce the implementation complexity in a low-cost fixed-point DSP controller.

A number of papers, discussed further on, deal with some more specific issues related to six-phase motor drives. These include speed controller design, parameter estimation, dc link current calculation and operation under fault conditions.

A fuzzy logic based speed controller for quasi six-phase induction machine drive was proposed by Kalantari et al (2002). The flexible nature of the fuzzy logic speed controller gives good results for high precision speed control in wide speed range. With simple modification in fuzzy rules the same controller can be used under fault conditions as well.

On-line stator resistance estimation technique was proposed by Jacobina et al (2002) for quasi six-phase induction machine with connected neutral points. In the proposed technique non-torque producing homopolar voltage component (zero-sequence component) is injected along with the symmetrical six-phase voltages in the machine. This distorted voltage is used as the modulating signal for the six-phase pulse width modulator. Three line voltages and currents are measured and the non-torque producing voltage and current are calculated using the developed model. These non-torque producing current components are filtered by low pass filter and their derivatives are calculated. By processing the filtered data through the developed least square algorithm the values of resistances and leakage inductances are determined.

An analytical technique of formulating the dc link RMS current in PWM VSI fed quasi six-phase induction motor drive with isolated neutral points has been presented by Bojoi et al (2002c). Two PWM techniques were considered, sinusoidal PWM and sinusoidal PWM with 16.6% third harmonic addition. Both techniques were found to give similar results.

A double-star synchronous machine with an arbitrary spatial displacement between two three-phase windings with isolated neutral points, fed by two three-phase PWM inverters, was considered by Merabtene and Benkhoris (2002). A model was developed and it was further used to study the machine behaviour under open-circuit fault condition.

2.3.2 Five-phase single-motor drives

Direct torque control (DTC) is one of the powerful methods for high performance control of motor drives, which has become an industrially accepted standard for three-phase
induction machines. The basic operating principle of DTC is based on instantaneous space vector theory and relies on utilisation of the non-ideal inverter nature to achieve good dynamic control. Toliyat and Xu (2000) have recently extended the DTC concept to a five-phase induction motor control and they presented a comparison between the three-phase and five-phase DTC drives. The implementation of the control system was done using 32 bit floating point TMS320C32 DSP. The three-phase inverter has only eight voltage space vectors that can be applied to a motor, while a five-phase inverter has 32 possible voltage space vectors. There is therefore a greater flexibility in controlling a five-phase drive system. The authors achieved high performance in terms of precise and fast flux and torque control and a smaller torque and flux ripple for five-phase induction machine as compared to three-phase induction machine.

Vector control and direct torque control of a five-phase induction motor with concentrated full-pitch winding was also developed and implemented again using 32 bit floating point DSP TMS320C32 in Xu et al (2002a). The proposed vector controller uses fundamental current in conjunction with 15% third harmonic stator current injection to provide quasi-rectangular current, which yields rectangular air gap flux in the concentrated-winding induction motor. It was shown that this approach enhances torque output by 11.2% under dynamic condition and by 10% during steady state operation, compared to the case when only fundamental current is fed to the machine. The DTC provides high performance in terms of smaller current, flux and torque ripples due to large number of space vectors for controlling the machine. Further, zero switching vectors are not needed to implement space vector modulation for five-phase PWM inverter for DTC and thus the wear and tear of motor bearing can be avoided. Shi and Toliyat (2002) have developed the vector control scheme based on space vector PMW for a five-phase synchronous reluctance motor drive. The control system for the proposed drive was again implemented using 32 bit floating point DSP TMS320C32.

A special current control scheme has been developed for a five-phase induction motor drive by Xu et al (2002b), which enables operation under open-circuit fault condition with loss of one or two phases. Concentrated winding induction motor was considered for the study and thus third harmonic current was used in conjunction with the fundamental component to obtain rectangular air-gap flux profile. The amplitudes of the fundamental and the third harmonic current need proper adjustment under fault condition. The speed and load have to be lowered under loss of two phases in order to prevent the stator current from
exceeding the rated value. The whole system was implemented to validate the theoretical findings.

A detailed performance analysis of concentrated winding multi-phase induction motors encompassing multiple of three and non-multiple of three numbers of phases was carried out by Toliyat and Qahtany (2002) using finite element analysis (FEA) method. The study reveals the fact that the torque pulsation decreases with an increase in the number of phases due to smaller step changes in MMF, except in seven-phase case where ripple was high due to cogging phenomena. The efficiency was seen to improve with increasing number of phases because of the reduction in ripple in rotor current. Further the five-phase machine was seen to provide the highest torque to current ratio due to an increase in the amplitude of fundamental MMF. The FEA results also showed a decrease in the stator back iron flux and an increase in stator tooth flux if third harmonic current is injected along with the fundamental in the five-phase induction motor. This suggests a new geometry (pancake shape) for a five-phase machine stator supplied by the third harmonic along with the fundamental. A comparison of performance was also given for different multi-phase machines with respect to the conventional three-phase distributed winding induction motor.

A general modelling approach encompassing dynamic and control issues for a five-phase permanent magnet brushless dc machine was presented by Franceschetti and Simoes (2001). The simulation was done for a five-phase, twelve-pole machine with rated torque of 30 Nm, with concentrated stator winding. The same machine type, which requires square-wave current for its normal operation, has been considered in Simoes et al (2001) as well, where experimental implementation was based on Motorola 56824 DSP. The switching frequency of the inverter was set to 10 kHz. Five-phase trapezoidal back-emf permanent magnet synchronous machine was elaborated by McCleer et al (1991) as well. The motor was supplied from a five-phase VSI in 144° conduction mode with square wave currents. The five-phase machine was shown to have higher torque capacity compared to similar sized three-phase machine and lower peak VA requirement of the switching devices.

The modelling and analysis of a five-phase permanent magnet synchronous machine supplied from a five-phase PWM inverter under normal and fault conditions (one phase open-circuited) was examined by Robert-Dehault et al (2002). The linear permanent magnet machine model was developed in phase variable form for fault condition and also its d-q form was given for normal operating conditions. The machine leakage reactance was considered as 5% with PWM inverter commutating at 2 kHz. The proposed control strategy allows the machine to produce the same torque under fault condition as under normal condition, with
very small torque ripple (6% torque ripple was observed because of the current controller
type). Pereira and Canalli (2002) have presented the design, modelling and performance
analysis of a five-phase permanent magnet synchronous machine operating as a generator
feeding a resistive load through five-phase bridge rectifier. The parameters of the machine
were determined using FEA. The performance in terms of load voltage versus current, output
power, rectified voltage waveform, phase to neutral and phase to phase voltages and phase
currents of the machine was assessed and examined using simulation and actual
measurements.

2.4 Control of three-phase and multi-phase voltage source
inverters

2.4.1 Control of three-phase voltage source inverters

Control of three-phase VSIs is nowadays, except in the highest power range, always
based on PWM schemes. PWM is the basic energy processing technique, used to obtain the
converter output power of required properties. Semiconductors are switched at a high
frequency, ranging from a few kilohertz (motor control) to several megahertz (resonant
converters for power supply) [Kazmierkowski et al (2002)]. A broad classification of PWM
techniques groups the methods into two classes: open-loop PWM and closed-loop PWM.

Among the best known open-loop PWM techniques is the carrier-based sinusoidal
PWM, which is also called ramp-comparison or sine-triangle or sub-oscillation method. In
this technique triangular carrier signals are compared with sinusoidal modulating waves to
generate the switching signals for power switches [Bose (1996)]. With the advent of high
speed and cheap digital signal processors, space vector pulse width modulation (SVPWM)
has become a standard for power inverters as it gives superior performance compared to the
ramp-comparison technique. The reference voltage space vector is generated on average by
imposing two neighbouring active vectors and a zero space vector in the three-phase VSI. The
basic theory and implementation of SVPWM are discussed at length by Neascu (2001). The
major advantages of SVPWM include wide linear modulation range for output line-to-line
voltages and easiness of digital implementation. The digital implementation can be done
using memory look-up table for sinusoidal function within a 60° interval. The alternative
solution is based on interpolation of a minimized look-up table, which can be implemented by
a fuzzy logic controller as well. A detailed comparison of carrier-based PWM methods,
including those with additional zero sequence signal injection, with SVPWM was presented
by Zhou and Wang (2002). The comparison looked at the relationship between modulation
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signals and space vectors, modulation signals and space vector sectors, switching pattern of space vectors and the type of the carrier, distribution of zero vectors and different zero-sequence signals. It was observed that the methods are closely related and can be derived from each other.

Multi-level (especially three-level) inverters are nowadays normally employed in high power applications. There have been many three-level modulation techniques reported in literature. Similar to two-level inverters, the popular choices are again carrier-based PWM and SVPWM. In carrier-based PWM, each phase reference voltage is compared to two identically shaped (but with different offset content) triangular waveforms in order to generate the switching pattern. In the case of SVPWM a higher number of space vectors (sixty four) are available for precise and flexible control of the three-level inverters. A comparative analysis of ramp-comparison PWM and SVPWM for three-level inverters was reported by Wang (2002). The three-level SVPWM equivalent can be realised by ramp-comparison PWM using a proper third harmonic injection, while three-level sine-triangle PWM can be realised through SVPWM by selecting proper dwell times (times spent in a given switching state).

The second group of methods encompasses closed-loop PWM techniques. In this case the inverter PWM pattern is determined by closed-loop current control of inverter output currents. The basic reason of introducing current control is the elimination of stator dynamics [Novotony and Lipo (2000)]. The main task of a current controlled PWM inverter is to force the load current to follow the desired reference current. By comparing the actual and reference phase currents, the current controller generates the switching signals for power devices in order to reduce the current error. The current controller therefore performs two functions, current error compensation and PWM. Various current control techniques have been developed over the years. The examples include hysteresis, ramp-comparison and predictive current control. Application of these methods to three-phase drives were examined by Andriux and Lajoie-Mazenc (1985), Brod and Novotony (1985), and Gio et al (1988). Among these techniques, hysteresis current control is the simplest to implement, it gives fast dynamic response, is insensitive to load parameter variations and it offers inherent current limitation. The major disadvantage of the hysteresis current control is that it gives a varying switching frequency over one cycle of the inverter output. Several other techniques are available, including already mentioned ramp-comparison current control, which give constant switching frequency of inverter. These include adaptive hysteresis controller proposed by Bose (1990), in which the hysteresis band is modulated with load and supply parameters, and dual-band hysteresis current control. In dual-band hysteresis current control also known as space-vector
based hysteresis current control, the switching states and thus the inverter output voltage vectors are chosen such as to minimize the derivative of the current error vector. A fully digital implementation of this scheme, using field programmable gate array (FPGA), was done by Brabandere et al (2002) in three-phase induction motor drive. The ramp-comparison current control is however still the most popular choice because of its simplicity and inherent advantages. Almost all the motor DSP controllers available today have hardware peripherals for implementation of digital modulation based on the ramp-comparison technique.

The main objective in a high performance drive is to control the torque and flux of the machine, which are governed by the fundamental current. A fast current control scheme must be incorporated to achieve the set goal. All the current control methods, utilised in vector controlled drives, essentially belong to one of the two categories: current control in stationary reference frame or current control in rotational reference frame. A comparative study of the methods belonging to these two categories was presented by Sokola et al (1992) in conjunction with a permanent magnet synchronous motor drive. Current control by means of hysteresis controllers and ramp-comparison controllers (in the stationary reference frame) and current control in the rotational reference frame (in conjunction with voltage generation by ramp-comparison method and by space vector modulator) was examined. The current control in the rotational reference frame was found to be superior to the current control in the stationary reference frame.

A space vector approach has been proposed by Bolognani and Zigliotto (2002) for analysis and design of stationary and rotating reference frame current controllers which allows deeper understanding of the behaviour of the controllers. An adaptive high-bandwidth current control algorithm was introduced by Telford et al (2003) for an indirect rotor flux oriented induction motor drive. A comparison between the conventional rotating reference frame PI controller, a conventional dead-beat controller and the proposed adaptive high-bandwidth controller has been conducted by simulation and experimentation. The proposed controller gives similar dynamic performance as the dead-beat controller, with no overshoot. However, it is immune to machine parameter variations, errors in flux estimates and dc bus voltage level fluctuations. Further, it was found to be superior to synchronous rotating reference frame PI controller since it was capable of producing faster dynamic response. However, its steady state error was higher.
2.4.2 Control of multi-phase voltage source inverters

The use of multi-phase inverter was first reported by Ward and Härer (1969) in a variable speed five-phase induction motor drive application. It utilised a forced commutated thyristor based inverter in ten-step operating mode. The torque ripple was decreased to one third compared to the equivalent three-phase case and was at an increased frequency. However, the machine current contained strong third harmonic component, which generated additional losses. To avoid these losses and to obtain fast current control, several PWM techniques for multi-phase VSIs have been developed, such as those reported in Pavitharan et al (1988), Toliyat (1998) and Toliyat et al (2000).

A complete mathematical model of a five-phase VSI, based on space vector representation, was developed by Toliyat et al (1993). The inverter operation in ten-step mode and PWM mode was discussed. The hysteresis type PWM current regulation was used for the drive under rotor flux oriented indirect vector control conditions. A SVPWM was proposed by Gataric (2000) for a five-phase VSI control in conjunction with induction motor drives. The same strategy was employed by Shi and Toliyat (2002) and Toliyat et al (2000) in a five-phase synchronous reluctance motor drive. Gopakumar et al (1993) employed SVPWM technique in split-phase induction motor drive fed by six-phase VSI.

Takami and Matsumoto (1993) have proposed optimum pulse pattern PWM for large capacity nine-phase VSI feeding a nine-phase induction motor (with three sets of three-phase windings on the stator with isolated neutral points and a single three-phase winding on the rotor). The current control loop was eliminated from the inverter control system. A new configuration, which includes nine low-rating (20% of the motor capacity) single-phase reactors, was introduced. The reactor turns ratio was selected equal to 1:2sin(π/18):1 so as to eliminate the lower order harmonics (5th, 7th, 11th and 13th) from the inverter/motor phase voltages and currents and also to balance the fundamental currents in the event of unbalancing. To eliminate even harmonics from the current/voltage waveform the optimal pulse pattern was developed based on Lagrangian multiplier method. The proposed optimal pulse pattern PWM technique was compared with the ramp-comparison PWM and it was shown to reduce the harmonic amplitudes to a significantly lower value. The proposed configuration also reduces the electromagnetic noise of the motor to a great extent.

Kelly et al (2001) have examined general n-phase (leg) inverter control techniques taking nine-phase inverter as a specific example. An n-phase inverter has (n-1)/2 possible load equivalent circuits and each operate in n-step mode to produce a unique step voltage
waveform with different fundamental and harmonic contents. The nine-phase inverter was examined for four different load equivalent configurations in eighteen-step mode and 4-5 configuration was found to be the optimum choice because of the highest switching efficiency (only one switch changes state between conduction intervals), maximum phase current delivery and maximum fundamental content. Four different SVPWM techniques have been developed for the general n-phase inverter. The first technique is a natural extension of the conventional three-phase SVPWM and it resulted in lower dc bus utilisation. The second technique asks for injection of $\sqrt{n}$ order harmonic, resulting in higher dc bus utilisation (maximum attainable fundamental component increases). However, $\sqrt{n}$ separate neutrals have to be used. The third proposed technique utilises smaller number of space vectors (74 instead of 512) but the switching efficiency was found to be poor in this case (in the case of nine-phase inverter, eighteen switches change state between conduction intervals). To improve the switching efficiency the fourth technique was proposed, which does not use zero space vectors (in the case of a nine-phase inverter, only six switches change state between conduction intervals). The first proposed technique (SVPWM) in the above referenced work was compared analytically and by experimentation to the ramp-comparison PWM by Kelly et al (2003) in conjunction with a nine-phase inverter fed nine-phase induction motor drive. The SVPWM was shown to enhance the fundamental by 1.55% compared to the sine-triangle PWM and thus enables better utilisation of the dc bus.

From various publications related to current control of inverter fed multi-phase machines it is evident that the same current control strategies as for a three-phase machine are in principal applicable regardless of the number of phases. Hysteresis current control scheme for a five-phase induction motor drive was reported by Toliyat (1998). In the vector control of a five-phase synchronous reluctance motor, hysteresis current controllers were used by Shi et al (2001). A global current control method for a five phase H-bridge VSI (i.e. an inverter system consisting of five single-phase inverters) was presented by Martin et al (2002). It was based on space vector control method and was aimed at control of a five-phase permanent magnet synchronous machine (PMSM). The independent current control of each phase leads to high current ripple as the dynamics of one phase depend on the states of the other inverter legs and because of the magnetic coupling between the phases. The basic principle consists in the representation of a machine by fictitious machines consisting of two groups, named the main machine and the secondary machines. For an odd number of phases equal to $n$, the equivalent number of fictitious machines are $(n+1)/2$ without mutual magnetic coupling. Out of this number there are $(n-1)/2$ two-phase machines and one single-phase machine. The
torque-producing machine is only the first two-phase machine and the remaining machines just add to the losses. The current dynamics of non-torque producing machines are limited by only small leakage inductance, which leads to high current ripple. Thus the duration of excitation to these non-torques producing fictitious machines are reduced by means of global current control strategy. The current controller produces the voltage reference vector, which is reconstituted over each sampling period by an optimal combination of the voltage vectors. It should be noted that the concept of space vector decomposition was applied in this paper, similar to Gataric (2000). This thesis will use both the concept of matrix transformations [White and Woodson (1959)] and space vector decomposition.

Ramp-comparison current controller was used by Bojoi (2002a) for a six-phase PWM inverter. Several types of current control schemes were proposed by Figueroa et al (2002) for controlling a seven-phase VSI for a brushless dc motor drive. These include the sinusoidal reference current control, synchronous reference frame current control, sampled sinusoidal current control, dc bus current control, square voltage and sampled sinusoidal voltage control. The comparative performance analysis was conducted using simulation and experimentation and it was observed that the last strategy provides an optimum result. It yielded high efficiency, low torque ripple, good torque-speed characteristic and simple control with low cost Hall position sensors.

2.5 Sensorless vector control drive

If a vector control algorithm is applied as a part of a drive with closed loop speed control, information regarding actual speed of rotation becomes necessary.

Speed-sensor-less vector controlled drive can be implemented by using the same speed estimation techniques as for direct torque control. All the speed estimation techniques fall into one of the following three categories [Vas (1998)]:

1. Speed estimation from stator current (or voltage) spectrum;
2. Speed estimation based on induction machine model:
   i. Open-loop estimators that use model equations only;
   ii. Closed-loop estimators that utilise an induction motor model and some additional corrective action (observers, extended Kalman filter and model reference adaptive system);
3. Speed estimation based on artificial intelligence techniques (artificial neural networks, neuro-fuzzy systems, etc).
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The good feature of methods belonging to the first category is that they do not depend on the machine equivalent circuit parameters. However, they require significant amount of signal processing (that leads to limited accuracy in transient operation) and the level of signal from which speed information is extracted can be very low (or indeed insufficient). Methods of the second category are rather easy to implement (except for the extended Kalman filter), but they are all too some extent affected by parameter variation effects in the machine. The artificial intelligence based estimators are expected to be widely used in the future. They are less sensitive to parameter variations and are currently under extensive development [Vas (1998)].

Sensorless variable speed high performance drives offer in general reduction of hardware complexity, improved reliability in hostile environment, increased mechanical robustness, decreased maintenance requirement, and reduced cost [Vas and Rashed (1999)].

Many speed prediction techniques have been suggested for sensorless control of induction motors over the years. The original techniques were designed to extract the information regarding the rotor speed from stator voltage and stator current only in steady state operation. These techniques can be used for low cost induction motor drives for applications that do not require highly dynamic performance [Abbondanti and Brennen (1975)]. More sophisticated speed estimation techniques have been suggested for high performance induction motor drives. This section reviews the speed estimation methods of the first group, using the spatial-saturation stator phase third-harmonic voltage, using saliency (geometrical, saturation) effects, and the estimation methods of the second group, using machine model (including open-loop speed estimators and closed-loop estimators).

The first technique utilises saturation phenomena to obtain the rotor speed of an induction motor and also the position and magnitude of the magnetising flux space vector [Vas (1998)]. The spatial saturation third-harmonic voltage component, which is due to main flux saturation, is obtained by monitoring the sum of the phase stator voltages in a symmetrical three-phase induction motor. The third-harmonic voltage component is then integrated to obtain the third-harmonic flux. The fundamental is further calculated by using a saturation function, which can be determined experimentally [Vas (1998)].

The angle of the magnetising flux space angle is also obtained from the monitored third-harmonic voltage. To obtain this information, a phase stator current of the induction motor has to be monitored as well. Then the displacement angle between this stator current maximum and the maximum of the fundamental magnetising flux is found. From the information about the angle of the fundamental magnetising flux with respect to the stator
current space vector and the angle of the stator current space vector with respect to the direct axis of the stationary reference frame, the angle of the fundamental magnetising flux space vector with respect to the direct axis of the stationary reference frame can be obtained [Vas (1998)].

Once when the fundamental magnetising flux space vector's modulus and phase angle with respect to the direct axis of the stationary reference frame are known they can then be used to determine the stator flux space vector and rotor flux space vector. The rotor speed then can be determined by differentiating the phase angle of stator flux space vector and deducting the angular slip frequency. The obtained magnetising flux space vector, stator flux space vector and rotor flux space vector can also be used in vector control [Vas (1998)].

As already mentioned, spatial saturation (zero sequence) third-harmonic voltage is required to obtain the fundamental of the magnetising flux. The third-harmonic voltage can be obtained in two ways [Consoli et al (2000)]. In the first approach, the summation of stator phase voltages is performed. This is carried out in a three-phase induction motor with wye-connected stator windings without a neutral point by using three wye-connected resistors in parallel to the machine, whose voltages are measured and summed. The resultant voltage of the summation contains a zero-sequence third-harmonic voltage component and the high frequency slot harmonic component. The third harmonic component is dominant, since the fundamental and the other non-triple harmonic components, which could be present in the air-gap magneto-motive force and magnetising flux, will cancel out. The slot harmonic component is then filtered out by using an analogue or digital filter [Vas (1998)]. In an alternative approach, the third harmonic voltage is evaluated directly by measuring the voltage across the midpoint of the DC bus and the neutral point of the machine. A cleaner third-harmonic signal can be obtained in this way because of the reference to a more stable potential point [Consoli et al (2000)]. If the induction motor's stator windings are wye connected without a neutral point, there is no flowing of the zero-sequence components of stator current in the motor. This will eliminate any influence of the third-harmonic component of the air-gap flux on the torque production [Consoli et al (2000)]. Phase shift of the third-harmonic voltage due to load currents is also avoided with the absence of zero-sequence stator currents. This guarantees the orthogonality between the third-harmonic components of stator voltage and air-gap flux [Consoli et al (2000)].

Modern induction machines are designed to operate in the saturated region of the magnetising characteristic for better utilisation [Vas (1998)]. When the machines are saturated, the sinusoidal distribution of air-gap flux density, which is caused by the
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A speed predicting technique can therefore utilise the phenomenon of saturation in the induction motors [Vas (1998)]. The saturation will result in the spatial distribution of the air-gap flux in a form of flattened non-sinusoidal waveform. The harmonic content of the waveform is characterised by the odd components but with the dominance of the third harmonic [Consoli et al (2000)]. The third-harmonic component of the zero-sequence component of stator voltage is always in quadrature with the third-harmonic flux component, without being affected by the load current variations [Consoli et al (2000)].

Speed predicting techniques based on the estimation of air-gap flux position by utilising the spatial saturation third-harmonic voltage are developed to improve the performance of back-EMF based direct field oriented control of induction motor drives [Moreira and Lipo (1990), Kreindler et al (1992)]. Traditional direct field oriented control using back EMF measurement fails at low and zero speed because of low induced voltages, which are practically difficult to measure correctly, and no voltage being induced in the stator winding at zero frequency. The reasons for failure also include motor asymmetries, stator voltage variation and electromagnetic noise, which prevent correct calculation [Consoli et al (2000)]. The harmonic based speed estimation techniques do not require any modification of the motor design. They are easy to implement and insensitive to parameter variations and also allow the low speed sensorless operations. However, the methods still fail at low and zero speed because of insufficient amplitude of the third-harmonic voltage. Detection of flux position at zero speed is impossible with third-harmonic voltage measurement because the third-harmonic component is already too low to be utilised effectively when motor's speed is below 1% of the rated speed [Consoli et al (2000)].

Suggestions for improvement of the technique by injection of a high frequency signal into stator voltage components have been set forth [Consoli (2000), Schroedl (1992)]. The additional voltage components are used to create high frequency current components with amplitudes being related to the spatial air-gap flux distribution inside the motor [Jansen and Lorenz (1995a and 1995b)]. The frequencies of the current and voltage harmonics created by the introduction of a high frequency signal in the stator voltage are related to the frequency of the injected signal and that of the fundamental rotating field [Consoli (2000)]. Demodulation of the high frequency current components will help to detect the angular position of the air-gap flux. The techniques have been verified successfully on salient AC machines [Degner and Lorenz (1998), Consoli et al (1999a)] and are being under development for the application in
non-salient machines, such as induction motors where saturation produces a small saliency effect on the direction of air-gap flux [Consoli et al (1999b), Manjrekar et al (1998)].

Rotor speed in the induction machines can also be estimated by utilising different types of geometrical effects (such as normal slotting, inherent air-gap asymmetry, intentional rotor magnetic saliency caused by spatial modulation of rotor-slot leakage inductance or rotor resistance) and saliency effects as a result of saturation, especially with the squirrel-cage induction motors [Vas (1998)]. The first of several speed predicting techniques using saliency effects discussed here is speed estimating technique utilising rotor slot harmonics. Rotor and stator slots in induction machines produce variations in the air gap permanence, that then interact with the stator winding distribution to result in inductance variations, as reviewed in [Degner and Lorenz (2000)]. This technique, up to now, is not directly used for rotor speed estimation. It is however used for tuning the parameters of speed estimators based on model reference adaptive systems [Vas (1998)]. The advantage of this technique is that its accuracy of estimation is not influenced by the variation of the machine's parameters and it is reliable for all loads. The rotor slot harmonics can be obtained by using either monitored stator voltage or monitored stator current. The latter is preferable because it is anyway necessary to measure stator current in high performance drives, so that the monitoring of stator voltage can be avoided [Vas (1998)].

The slot harmonics detection using stator voltage monitoring utilises the slot harmonic voltages induced in the stator winding of the induction motor. The slot harmonic voltages are induced in the stator winding when the rotor is rotating because of the presence of slot harmonics in the air-gap magneto-motive force. These harmonics are produced due to the variation of the reluctance caused by the stator or rotor slots, and are therefore called stator or rotor slot harmonics, respectively [Vas (1998)]. Both amplitudes and frequencies of slot harmonic voltages depend on the rotor speed. However, only the frequencies are used to estimate the rotor speed because the amplitudes also depend on loading conditions and flux levels [Vas (1998)]. Stronger slot harmonics can be produced in the induction machines without the skewed rotor slots. However, skewed rotor slots are normally created inside the machine to reduce audible noise and eliminate the asynchronous crawling during the line starting [Vas (1998)].

During the monitoring of stator voltages to detect rotor slot harmonics, summation of stator phase voltages is carried out. It has been shown that the summation voltage contains a slot harmonic component and a third-harmonic component, due to main flux saturation. When the stator voltage is supplied by an inverter, the summation voltage also contains extra time
harmonic voltages [Vas (1998)]. It has also been shown that the frequency of the slot harmonic voltage depends on angular stator frequency, slip frequency and the number of rotor slots per pole pair [Vas (1998)]. The slot harmonic voltage in the summation voltage is obtained by removing the third-harmonic voltage and extra harmonic voltages. The removal can be carried out by employing various circuits [Vas (1993)]. The frequency of slot harmonic voltage is then evaluated. With the help of available information regarding the stator angular frequency and the number of slots per pole pair, the rotor speed of the induction motor is finally calculated. However, when the induction motor is running in the low speed region, special considerations are required for this speed estimation technique, because amplitude of the slot harmonic voltage decreases at low speed [Vas (1998)]. Speed estimation can also be done with the machine terminal voltage spectrum obtained from tapped stator windings of an induction machine [Zinger et al (1990)].

Speed estimation technique using monitored stator current is preferable. Harmonic spectral estimation is carried out on the obtained signal of the stator current [Vas (1998)]. The measured signal is scaled and low-pass filtered to eliminate high frequency pulse width modulation harmonics (when the power supplies for the induction motors are the voltage source inverters). Digital Fast Fourier Transform is then performed to find the rotor slot harmonic [Ferrah et al (1992a)]. When the rotor slot harmonic frequency is found and if the information about the angular stator frequency and the number of rotor slots per pole pair is available, rotor speed can be estimated. The accuracy of the speed estimation depends therefore on the accuracy of the rotor slot harmonic frequency and the angular stator frequency. Angular stator frequency is obtained by differentiating the angle of the rotor flux space vector when the rotor flux oriented control is used for the drive. The five main steps for the estimation of rotor slot harmonic frequency are summarised as follows [Vas (1998)]:

1. Identification of angular stator frequency.
2. Determination of the no-load slot harmonic around a specific stator harmonic.
3. Definition of the width of the slot harmonic tracking window. Difference window placement for motoring or generating operation of the motor.
4. Search for the harmonic with the highest amplitude (highest spectrum line), which is a non-triple harmonic of angular stator frequency, in the window.
5. An increase in the accuracy of the rotor slot harmonic frequency.

When an adequate frequency resolution is employed, rotor slot harmonic can be identified from the neighbouring harmonics at no-load and load and over a wide range of
speed. Real value Fast Fourier Transform based on the split-radix algorithm is suitable for
digital implementation of the estimator. Digital Fast Fourier Transform can give a good result
of the estimation when the operating frequency is over a few Hz. However, when the speed of
the motor is too low, noise can affect the accuracy of the speed estimation [Vas (1998)].
Accuracy of the estimating process can be increased by employing appropriate windowing
(such as Henning data window, hamming window or rectangular window) and interpolation
techniques. This means that the signals are analysed and processed for a short length (or
window) of the sampled data by the Fast Fourier Transform analyser. However, when signals
are transformed into frequency domain, false results could be obtained due to the
discontinuities at the window’s edges [Vas (1998)].

Fast Fourier Transform has been used to obtain rotor slot harmonic in a speed
predicting technique proposed by [Ferrah et al (1992b)]. The speed estimation only considers
slotting harmonics and saturation harmonics. This technique gives satisfactory results over a
wide speed range, down to 2 Hz. Attempts of estimating the rotor speed by this technique give
relatively poor dynamic performance due to the limitation of processing power and sampling
rate of the hardware. Estimation of slip speed of the motor based on slot harmonic frequency
identification (instead of estimation of rotor speed) is also suggested by [Beguenane et al
(1995 and 1996)]. Real time adaptive filters with recursive maximum likelihood technique
can also be used for tracking slot harmonics [Cilia et al (1998), Ferrah et al (1996 and 1998)].
They give an error of less than 0.1% for the estimated speed. The parameters of the control
system are updated regularly with recursive algorithm for rotor slot harmonic analysis to give
a better performance.

Insufficiency of the measurement’s bandwidth, the main limitation of this method for
its use as an effective speed feedback signal in the past, has been overcome recently
[Ferrah et al (1996 and 1998)]. Rotor slot harmonic approach can be used to tune the parameters of a
sensorless induction motor drive with speed estimation based on machine models as well
processing techniques to obtain the speed related harmonic frequency signal induced in the
motor by rotor slots and rotor eccentricity. The estimation is then used for tuning the
parameters of the machine model and load model [Turl et al (2001)]. Speed estimator based
on a mechanical model gives speed feedback signal to sensorless control of the motor.
Because the estimated speed signal obtained from rotor slot harmonics analysis is parameter
independent, the sensorless control can give good performance at very low speed [Hurst et al
(1994)].
Speed estimation techniques based on inductance variation due to saturation and geometrical effects have been used for sensorless vector control of permanent magnet synchronous machines and synchronous reluctance machines. The technique can also be applied to sensorless control of induction motors even when the motor is at standstill [Schroedl (1988)]. In permanent magnet synchronous machines, stator inductance is a function of the rotor position because of saturation effects. This can be used for estimation of the rotor position or speed. It has also been shown that stator inductance depends on the angle of the magnet flux, with respect to the stator voltage space vector [Schroedl (1993)]. In synchronous reluctance machines, the stator inductance is a function of rotor position because of rotor saliency. Therefore, rotor position and speed can be estimated with high accuracy by using the inductance variation due to geometrical effects. This method is especially useful for speed estimation at zero and low speed. The speed estimation can be carried out either by monitoring the stator current ripple or by injecting a suitable test stator voltage signal [Vas (1998)].

The speed predicting techniques based on variation of inductance due to saliency or geometrical effects can be used for sensorless control of induction motors because the stator inductance in induction motors is dependent on the level of saturation as well as the position of main flux. This dependency is caused by the saturation of stator and rotor teeth. In a saturated induction machine at standstill, the rate of change of stator currents is dependent on the stator voltage space vector and stator transient inductance. The magnitude and angle of this inductance also depend on the magnetic operating point and the direction of the magnetising flux space vector. Application of appropriate test voltage signal would give the rate of change of stator currents at test signal frequency. The angle of magnetising flux space vector is then found, which will be used to extract the information about rotor speed, because the minimum of the ellipse locus of the modulus of complex transient inductance is in the direction of the magnetising flux space vector [Vas (1998)]. Estimation of the flux space vector's position can also be carried out by tracking the high frequency magnetic saliency caused by magnetic saturation at zero or low speed, with the injection of high frequency voltages (or currents) into the stator [Jansen and Lorenz (1995a), Briz et al (2001)].

High frequency magnetic saliency caused by magnetic saturation (main flux or leakage flux saturation) can also be used for speed estimation at zero and low speed. A high frequency signal is injected into the stator winding to create high frequency current with speed related harmonics [Jansen and Lorenz (1996), Degner and Lorenz (1997)]. The technique of combining rotor slot harmonic method and carrier frequency signal injection to
estimate rotor speed gives robust response over a wide range of speeds, even at low or zero speed. Multiple saliencies, although not desirable, could also be used for speed estimation, by developing an induction machine model with multiple spatial saliencies [Degner and Lorenz (1998)].

A comparative study carried out by [Hurst and Habetler (1997)] has pointed out a significant limitation of all the speed estimation techniques with spectrum analysis discussed so far. It is the lack of robust performance with respect to machine specific structural characteristics. For example, rotor slot harmonic analysis depends on the number of rotor slots, which is a typically unknown parameter. The control schemes then have to be programmed again for each different induction machine. In addition to that, the estimation techniques are intensive computationally; therefore they are only suitable for speed estimation in steady state operation. Robust and fast hardware is required for speed estimation in transient state, and only a limited accuracy can be obtained.

Estimation of rotor speed of induction machines, especially at zero and low speed can be achieved by utilisation of saliency effects intentionally created by the special design of the rotor, which gives rise to spatial modulation of rotor leakage inductance or rotor resistance [Vas (1998)]. Rotor eccentricity, which is always present in induction motors because of the compromise between cost and performance, can also be used for speed estimation [Holliday et al (1995)]. Specially designed rotor eccentricity will give position signal more robust than those obtained from dynamic eccentricity [Vas (1998)].

The special designs of the rotor may include variation of the rotor slot openings or variation of the depths of rotor slot openings [Jansen and Lorenz (1995a)]. Specially designed asymmetry of rotor resistance can also be used for speed estimation. The asymmetry is created to make rotor position harmonics dependent on the machine flux and load conditions [Cilia et al (1997)]. The speed estimation techniques discussed previously can again be used here, the only difference is that the saliency is created by physically existing asymmetry, not by saturation. Similarly to those techniques, symmetrical three phase high frequency voltages are injected into stator to track the saliency. It has been proven that when the special designs of the rotor are employed, the stator transient inductance is position-dependent because of physical asymmetry. This dependency will make stator currents position-dependent when the high frequency stator voltages are applied to the machines. Therefore, information about rotor position and speed can be obtained by monitoring and analysing the currents [Vas (1998)]. An attempt of rotor speed estimation using spatial modulation of the rotor leakage inductance, created by periodic variation of rotor slot opening width, has been made by [Jansen and
Lorenz (1995b). Addition of specific irregularities to the rotor to create small air gap variations in the machine for the purpose of speed estimation has also been suggested. The irregularities are detected by injection of high frequency signal into the stator currents. These irregularities produce amplitude oscillations in the stator currents generated by the high frequency carrier, and the oscillations are proportional to the rotor position [Dixon and Rivarola (1996)].

However, the variation of rotor slot opening width may cause adverse effects on the magnetising inductance and this could result in torque pulsation. Variation of the depths of rotor slot openings to spatially modulate the rotor leakage inductance will almost have no effect on the magnetising inductance, but the rotor lamination asymmetry may not be desirable by the manufacturers. Variation in rotor slot opening fill can help to obtain leakage inductance modulation. However, it could also have some other adverse effects [Vas (1998)]. Iron saturation in induction machines also degrades the performance of speed estimation technique using rotor asymmetry (such as variation of rotor resistance, variation of rotor leakage inductance or rotor eccentricity) because it causes some degree of saliency in addition to the saliency used to estimate the speed [Staines et al (1998)]. Periodic high frequency voltage burst is injected into the machine instead of continuous high frequency voltage to avoid the saturation.

An entirely different approach to induction machine speed estimation is the one that relies on the induction machine model. An open loop speed estimator that utilises only the induction machine equations and has no corrective action will be explained in more detail in section 6.2. The other model based speed estimators reviewed further on are of closed-loop structure, where apart from an induction motor model equations some corrective action is introduced to improve the accuracy of the speed estimate. Closed-loop speed estimators based on machine models can give satisfactory performance in sensorless induction machine drives. Speed estimators using model reference adaptive system (MRAS) are one category of the closed-loop machine-model based estimators, which can give reliable estimation and good performance for speed sensorless induction motor drives [Griva et al (2001)]. Measured stator voltages and currents are used in these estimators for extracting the information about speed from two machine models. In the first model, certain state or other variable of the induction machine is calculated independently of the rotor speed, and the model is called reference model. In the second model, the same variable is calculated from a different machine model that is dependent on the rotor speed. The estimated rotor speed is used in the second model as feedback signal. The second model is called adaptive model. The difference between the
variables from the two models is used to find the estimated value of the rotor speed through an adaptation mechanism. The adaptation mechanism tunes the adaptive model to obtain a satisfactory estimation of speed [Vas (1998)].

The adaptation mechanism can be derived by using Popov’s criterion of hyperstability, which results in a stable and fast response system [Vas (1998)]. The difference between the state variables mentioned above is formulated into an error signal. This signal is used as input into a PI controller, whose output is the estimated speed. The variables of an induction machine that have been used so far for MRAS speed estimators are rotor flux, back EMF, reactive power and air gap power. Rotor flux based MRAS speed estimator is, however, the most frequently used MRAS speed estimator [Schauder (1992), Tajima and Hori (1993)]. Each of these variables will give different performance characteristics of the speed estimator.

It has been shown that rotor flux based MRAS speed estimator gives good estimation in medium performance sensorless drives over a wide range of speed excluding zero and near-zero speed [Conroy et al (1995)]. Back EMF based estimator also produces good estimation over a wide range of speed but the performance at low speed is affected by the inaccuracy of stator resistance [Conroy et al (1995)]. Reactive power based estimator has the advantage of eliminating the dependency on stator resistance [Peng and Fukao (1994)]. It is, however, characterised with difficulty in obtaining stable control under all operating conditions and it still depends on stator transient inductance [Conroy et al (1995)]. Air gap power based speed estimator is independent of the stator transient inductance. Nonetheless, stator resistance variation affects the performance at low speed and an on-line identification scheme for stator resistance is needed to have accurate estimated speed [Conroy et al (1995), Zhen and Xu (1995)].

The transient response of MRAS based sensorless vector control is similar to the response of vector control with speed measurement [Wang (1999)]. Rotor flux based MRAS speed estimator offers advantages in terms of hardware implementation requirements, when compared to other MRAS based schemes [Marwali and Keyhani (1997)].

Artificial-intelligence-based MRAS speed estimators, which do not require mathematical model and whose adaptation mechanism is a part of the tuning system of an appropriate artificial intelligence network, can also offer good performance, especially at low and very low speed [Vas (1998)]. The problems experienced with the MRAS speed estimators in the low speed region can be therefore avoided.

A closed loop speed estimator can be defined as an observer [Vas (1998)]. A closed loop estimator contains a correction term, which involves the estimation error. This correction
term is used to adjust the response of the observer. The observer is more robust than an open loop estimator in terms of immunisation against deviations of parameters and noise, both in steady state and transient states. Observers can be classified as deterministic observers or stochastic observers according to the representation of the systems, for which they are used to estimate variables or parameters [Vas (1998)]. Luenberger observers and Kalman filters are the most commonly used closed loop estimators [Du et al (1994)]. Luenberger observers are of deterministic type and Kalman filters are of stochastic type. The basic forms of both types of observers can only be used with a linear system. For non-linear systems, the extended forms of these observers have to be used [Vas (1998)].

Many kinds of observers can be used for speed sensorless high performance induction motor drive, such as full order adaptive state observer (which is built from the equations of an induction machine in a stationary reference frame and additional error compensators [Vas (1998)]), extended Kalman filters (EKF) and extended Luenberger observers (ELO). While in a full order adaptive state observer rotor speed is a parameter, it is a state variable in EKF and ELO. When an appropriate speed observer is used in speed sensorless high performance induction motor drive, stable operation can be achieved over a wide speed range, even at very low speed. The observers have been used widely in industrial applications of induction motor drives and will have an increasing role in the future [Vas (1998)].

Full order (fourth order) adaptive state observer can be obtained by modification of a rotor flux estimator to give a speed signal. The stability of the system can be assured by deriving the adaptation mechanism of the system from state error dynamic equations of the induction machine and Lyapunov's stability theorem [Vas (1998)]. Adaptive state observers can be used for estimation of states and/or parameters of a non-linear system in real time. During the calculation, the states are estimated based on a mathematical model. The predicted states are continuously adjusted by a feedback correction scheme. Extended Kalman filters, when used in conjunction with speed sensorless induction motor drives, can give high accuracy of estimated speed over a wide range of speeds and joint estimation of other state variables and parameters can also be incorporated in the control scheme. However, the intensive computational task is the most important drawback of the method. It requires powerful hardware to provide fast response and it is not preferable in terms of economy. EKF is a recursive stochastic state estimator, which can be used for non-linear dynamic systems by utilising the monitored signals disturbed by random noises [Vas (1998)]. It is also assumed that the measurement noise and disturbance noise are not correlated. Inaccuracy in modelling and measurement is also accounted for through the noise.
Extended Luenberger observers are used for the control schemes with joint estimation of flux and speed. They are applicable to non-linear time-varying deterministic systems. ELO is used for replacement of EKF for the joint estimation in a high performance induction motor drive. Extended Kalman filters expose inherent problems created by the stochastic approach [Vas (1998)]. An EKF has problems when the noise contents of the system and the measurements are too low. Unlike standard linear Kalman filter, extended Kalman filters are not optimal. There may be a problem of biasing when the assumed characteristics of stochastic noise do not match those of the real noises. Extended Luenberger observers also give better performance at steady state with the higher accuracy of the estimated speed [Vas (1998), Cirrincione (2005)].

2.6 Summary

A comprehensive literature review, related to the relevant aspects of the research in this project, is presented in this chapter. The potential application areas for multi-phase motor drives, such as electric traction, more-electric aircraft, ship propulsion and electric and hybrid vehicles are surveyed at first. It is concluded that multi-phase machine drives in such applications offer many advantages when compared to their three-phase counterparts, primarily due to the increased reliability, flexible control features and potential for reduction of the switch power rating. The latest developments in the area of multi-phase motor drives are examined next, with an emphasis on five-phase and six-phase (true and quasi) drives.

The papers related to the modelling and control of three-phase and multi-phase inverters, encompassing both the open-loop and closed-loop PWM schemes, is finally studied and a review is presented. The choice of an appropriate PWM technique for a multi-phase inverter is an important issue, as the use of multi-phase machines introduces severe distortion of stator current, leading to large winding losses and increased stress on inverter. It is concluded that, in principle, the already available current control methods for three-phase inverters are applicable to multi-phase inverters as well.

The last section of this chapter discusses numerous sensorless speed estimation techniques used in vector controlled drives. It is concluded that in principle, the sensorless techniques used for three-phase induction machine may be equally applied to multi-phase machine because d-q modelling of multi-phase machine is same as one for three-phase machine and also the vector control principle used for three-phase machine may be used for multi-phase machine.
CHAPTER 3

FIVE-PHASE AND SERIES CONNECTED FIVE-PHASE TWO MOTOR DRIVE SYSTEM

3.1 Introduction

The first part of this chapter details at first the modelling and control of a five-phase drive system including inverter, controllers and motor. The five-phase machine model is developed next in phase domain and then it is transformed into a system of decoupled equations in orthogonal reference frames (section 3.3). The $d$-$q$ axis reference frame currents contribute towards torque and flux production, whereas the remaining $x$-$y$ components plus the zero-sequence components do not. This allows a simple extension of the rotor flux oriented control (RFOC) principle to a five-phase machine, as elaborated in section 3.4.

The inverter current control techniques, used in vector control, are further reviewed in section 3.5. Current control in stationary reference frame is elaborated using hysteresis method. Tuning of the speed controller and the current controller is performed for the given drive parameters and the procedure is described in detail. A simulation study is finally performed for speed mode of operation, for a number of transients, and the results are presented in section 3.6.


The next part of this chapter deals with a two-motor five-phase drive system in which stator windings of two five-phase induction machines are connected in series. Introduction of an appropriate phase transposition in the series connection leads to a complete decoupling of the flux/torque producing current of one machine from the flux/torque producing current of the second machine as shown in Levi et al (2003d). Thus an independent control of two five-phase induction machines, fed by a single PWM voltage source inverter becomes possible by means of vector control. The drive structure is reviewed first in section 3.7.

3.2 Modelling of a five-phase voltage source inverter: A Review

3.2.1 Power circuit and switch control signals

Power circuit topology of a five-phase VSI, which was used probably for the first time by Ward and Härer (1969), is shown in Fig. 3.1. Each switch in the circuit consists of two power semiconductor devices, connected in anti-parallel. One of these is a fully controllable semiconductor, such as a bipolar transistor or IGBT, while the second one is a diode. The input of the inverter is a dc voltage, which is regarded further on as being constant. The inverter outputs are denoted in Fig. 3.1 with lower case symbols (a,b,c,d,e), while the points of connection of the outputs to inverter legs have symbols in capital letters (A,B,C,D,E). The basic operating principles of the five-phase VSI are developed in what follows assuming the ideal commutation and zero forward voltage drops.

Figure 3.1. Five-phase voltage source inverter power circuit.

3.2.2 Space vector representation of a five-phase voltage source inverter

In order to introduce space vector representation of the five-phase inverter output voltages, an ideal sinusoidal five-phase supply source is considered first. Let the phase voltages of a five-phase pure balanced sinusoidal supply be given with
Chapter 3: Five-phase and series connected two five-phase induction motor drive system

\[ v_a = \sqrt{2} V \cos(\omega t) \]
\[ v_b = \sqrt{2} V \cos(\omega t - 2\pi / 5) \]
\[ v_c = \sqrt{2} V \cos(\omega t - 4\pi / 5) \]
\[ v_d = \sqrt{2} V \cos(\omega t + 4\pi / 5) \]
\[ v_e = \sqrt{2} V \cos(\omega t + 2\pi / 5) \]

(3.1)

Space vector of phase voltages is defined, using power invariant transformation, as:

\[ \mathbf{V} = \sqrt{\frac{2}{5}} (v_a + a^2 v_b + a^4 v_c + a^6 v_d + a^8 v_e) \]

(3.2)

where \( a = \exp(j2\pi/5), \quad a^2 = \exp(j4\pi/5), \quad a^* = \exp(-j2\pi/5), \quad a^{*2} = \exp(-j4\pi/5) \) and * stands for a complex conjugate. The space vector is a complex quantity, which represents the five-phase balanced supply with a single complex variable. Substitution of (3.1) into (3.2) yields for an ideal sinusoidal source the space vector

\[ \mathbf{V} = \sqrt{5} V \exp(j\omega t) \]

(3.3)

However, the voltages are not sinusoidal any more with the inverter supply. They are in general of quasi-square waveform. Leg voltages (i.e. voltages between points A, B, C, D, E and the negative rail of the dc bus N in Fig. 3.1) are considered first.

Phase-to-neutral voltages of the star connected load are most easily found by defining a voltage difference between the star point \( n \) of the load and the negative rail of the dc bus \( N \). The following correlation then holds true:

\[ v_A = v_a + v_{an} \]
\[ v_B = v_b + v_{bn} \]
\[ v_C = v_c + v_{cn} \]
\[ v_D = v_d + v_{dn} \]
\[ v_E = v_e + v_{en} \]

(3.4)

Since the phase voltages in a star connected load sum to zero, summation of the equations (3.4) yields

\[ v_{an} = (\frac{1}{5}) (v_A + v_B + v_C + v_D + v_E) \]

(3.5)

Substitution of (3.5) into (3.4) yields phase-to-neutral voltages of the load in the following form:

\[ v_a = (\frac{4}{5}) v_A - (\frac{1}{5}) (v_B + v_C + v_D + v_E) \]
\[ v_b = (\frac{4}{5}) v_B - (\frac{1}{5}) (v_A + v_C + v_D + v_E) \]
\[ v_c = (\frac{4}{5}) v_C - (\frac{1}{5}) (v_A + v_B + v_D + v_E) \]
\[ v_d = (\frac{4}{5}) v_D - (\frac{1}{5}) (v_A + v_B + v_C + v_E) \]
\[ v_e = (\frac{4}{5}) v_E - (\frac{1}{5}) (v_A + v_B + v_C + v_D) \]

(3.6)
3.3 Modelling of a five-phase induction motor: A Review

A model of a five-phase induction motor is developed initially in phase variable form. In order to simplify the model by removing the time variation of inductance terms, a transformation is applied and so-called d-q-x-y-0 model of the machine is constructed. It is assumed that the spatial distribution of all the magneto-motive forces (fields) in the machine is sinusoidal, since only torque production due to the first harmonic of the field is of relevance in this project. All the other standard assumptions of the general theory of electrical machines apply. The model derivation is summarised in the following sub-sections. A more detailed discussion of the modelling procedure is available in Jones (2005) and Atif (2006).

![Figure 3.2. Phase-to-neutral voltage space vectors.](image)

3.3.1 Phase variable model

A five-phase induction machine is constructed using ten phase belts, each of 36 degrees, along the circumference of the stator. The spatial displacement between phases is therefore 72 degrees. The rotor winding is treated as an equivalent five-phase winding, of the
same properties as the stator winding. It is assumed that the rotor winding has already been referred to stator winding, using winding transformation ratio. A five-phase induction machine can then be described with the following voltage equilibrium and flux linkage equations in matrix form (underlined symbols):

\[
\begin{align*}
\dot{v}^s_{\text{abcde}} &= R_s^s i^s_{\text{abcde}} + \frac{d\psi^s_{\text{abcde}}}{dt} \\
\psi^s_{\text{abcde}} &= L_s^s i^s_{\text{abcde}} + L_{sr}^s i^r_{\text{abcde}} \\
\dot{v}^r_{\text{abcde}} &= R_r^r i^r_{\text{abcde}} + \frac{d\psi^r_{\text{abcde}}}{dt} \\
\psi^r_{\text{abcde}} &= L_r^r i^r_{\text{abcde}} + L_{sr}^r i^s_{\text{abcde}}
\end{align*}
\] (3.7) (3.8)

The following definition of phase voltages, currents and flux linkages applies to (3.7)-(3.8):

\[
\begin{align*}
\dot{v}^s_{\text{abcde}} &= [v_a^s \ v_b^s \ v_c^s \ v_d^s \ v_e^s]^T \\
\dot{i}^s_{\text{abcde}} &= [i_a^s \ i_b^s \ i_c^s \ i_d^s \ i_e^s]^T \\
\psi^s_{\text{abcde}} &= [\psi_a^s \ \psi_b^s \ \psi_c^s \ \psi_d^s \ \psi_e^s]^T \\
\dot{v}^r_{\text{abcde}} &= [v_a^r \ v_b^r \ v_c^r \ v_d^r \ v_e^r]^T \\
\dot{i}^r_{\text{abcde}} &= [i_a^r \ i_b^r \ i_c^r \ i_d^r \ i_e^r]^T \\
\psi^r_{\text{abcde}} &= [\psi_a^r \ \psi_b^r \ \psi_c^r \ \psi_d^r \ \psi_e^r]^T
\end{align*}
\] (3.9) (3.10)

The matrices of stator and rotor inductances are given with \(\alpha = 2\pi/5\):

\[
L_s^s = \begin{bmatrix}
L_{aaa} & L_{aab} & L_{aac} & L_{aad} & L_{aee} \\
L_{aba} & L_{bab} & L_{bac} & L_{bad} & L_{bee} \\
L_{aca} & L_{aca} & L_{cas} & L_{cad} & L_{ace} \\
L_{ada} & L_{bda} & L_{bdc} & L_{bde} & L_{bce} \\
L_{aee} & L_{bde} & L_{cde} & L_{ede} & L_{ece}
\end{bmatrix}
\]

\[
L_r^s = \begin{bmatrix}
L_{ia} + M & M \cos \alpha & M \cos 2\alpha & M \cos \alpha & M \cos \alpha \\
M \cos \alpha & L_{ia} + M & M \cos \alpha & M \cos 2\alpha & M \cos 2\alpha \\
M \cos 2\alpha & M \cos \alpha & L_{ia} + M & M \cos \alpha & M \cos 2\alpha \\
M \cos \alpha & M \cos 2\alpha & M \cos \alpha & L_{ia} + M & M \cos \alpha \\
M \cos \alpha & M \cos 2\alpha & M \cos \alpha & M \cos \alpha & L_{ia} + M
\end{bmatrix}
\]

\[
L_r^r = \begin{bmatrix}
L_{ia} & L_{iab} & L_{iac} & L_{iad} & L_{iae} \\
L_{ira} & L_{ira} & L_{ira} & L_{ira} & L_{ira} \\
L_{ida} & L_{ida} & L_{ida} & L_{ida} & L_{ida} \\
L_{iia} & L_{iia} & L_{iia} & L_{iia} & L_{iia}
\end{bmatrix}
\] (3.11)
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$$L_e = \begin{bmatrix} L_a + M & M \cos \alpha & M \cos 2\alpha & M \cos 2\alpha & M \cos \alpha \\ M \cos \alpha & L_a + M & M \cos \alpha & M \cos 2\alpha & M \cos 2\alpha \\ M \cos 2\alpha & M \cos \alpha & L_a + M & M \cos \alpha & M \cos 2\alpha \\ M \cos \alpha & M \cos 2\alpha & M \cos \alpha & L_a + M & M \cos \alpha \\ M \cos \alpha & M \cos 2\alpha & M \cos \alpha & M \cos 2\alpha & L_a + M \end{bmatrix}$$ (3.12)

Mutual inductances between stator and rotor windings are given with:

$$L_{sr} = M \begin{bmatrix} \cos \theta & \cos(\theta + \alpha) & \cos(2\alpha) & \cos(\alpha - \theta) \\ \cos(\theta - \alpha) & \cos \theta & \cos(\alpha + 2\alpha) & \cos(\alpha - \theta) \\ \cos(\theta + 2\alpha) & \cos(\theta - \alpha) & \cos \theta & \cos(\alpha + \theta) \\ \cos(\theta - \alpha) & \cos(\theta + 2\alpha) & \cos(\alpha - \theta) & \cos \theta \\ \cos(\theta + \alpha) & \cos(\theta + 2\alpha) & \cos(\theta - \alpha) & \cos \theta \end{bmatrix}$$ (3.13)

The angle $\theta$ denotes the instantaneous position of the magnetic axis of the rotor phase 'a' with respect to the stationary stator phase 'a' magnetic axis (i.e. the instantaneous position of the rotor with respect to stator). Stator and rotor resistance matrices are 5x5 diagonal matrices,

$$R_s = \text{diag}(R_s, R_s, R_s, R_s, R_s)$$
$$R_r = \text{diag}(R_r, R_r, R_r, R_r, R_r)$$ (3.14)

Motor torque can be expressed in terms of phase variables as

$$T_e = \frac{P}{2} \int \frac{dL_{abcd}}{d\theta} i + \frac{P}{2} \int \frac{dL_{abcde}}{d\theta} \left[ \int L_{abcde} \right]$$ (3.15a)

$$T_e = P_s \int \frac{dL_{abcde}}{d\theta} i_{abcde}$$ (3.15b)

Substitution of stator and rotor currents from (3.9)-(3.10) and (3.13) into (3.15b) yields the torque equation in developed form:

$$T_e = -PM \left( (i_{ab}i_{ar} + i_{ac}i_{as} + i_{ad}i_{ar} + i_{ae}i_{as})\sin \theta + (i_{ab}i_{ar} + i_{ac}i_{as} + i_{ad}i_{ar} + i_{ae}i_{as})\sin(\theta + \alpha) + (i_{ac}i_{ar} + i_{ad}i_{ar} + i_{ae}i_{as})\sin(\theta + 2\alpha) + (i_{ad}i_{ar} + i_{ae}i_{as} + i_{ac}i_{ar} + i_{ad}i_{ar} + i_{ae}i_{as})\sin(\theta - \alpha) \right)$$ (3.16)

3.3.2 Model transformation

In order to simplify the model, it is necessary to apply a co-ordinate transformation that will remove the time varying inductances. The co-ordinate transformation is utilised in the power invariant form. The following transformation matrix is therefore applied to the stator five-phase winding:

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\[
A_s = \sqrt{\frac{2}{5}} \begin{bmatrix}
\cos \theta_s & \cos(\theta_s - \alpha) & \cos(\theta_s - 2\alpha) & \cos(\theta_s + 2\alpha) & \cos(\theta_s + \alpha) \\
-\sin \theta_s & -\sin(\theta_s - \alpha) & -\sin(\theta_s - 2\alpha) & -\sin(\theta_s + 2\alpha) & -\sin(\theta_s + \alpha) \\
1 & \cos(2\alpha) & \cos(4\alpha) & \cos(4\alpha) & \cos(2\alpha) \\
0 & \sin(2\alpha) & \sin(4\alpha) & -\sin(4\alpha) & -\sin(2\alpha) \\
\frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}}
\end{bmatrix}
\]

(3.17)

Transformation of the rotor variables is performed using the same transformation expression, except that \( \theta_s \) is replaced with \( \beta \), where \( \beta = \theta_s - \theta \). Here \( \theta_s \) is the instantaneous angular position of the d-axis of the common reference frame with respect to the phase ‘a’ magnetic axis of the stator, while \( \beta \) is the instantaneous angular position of the d-axis of the common reference frame with respect to the phase ‘a’ magnetic axis of the rotor. Hence the transformation matrix for rotor is:

\[
A_r = \sqrt{\frac{2}{5}} \begin{bmatrix}
\cos \beta & \cos(\beta - \alpha) & \cos(\beta - 2\alpha) & \cos(\beta + 2\alpha) & \cos(\beta + \alpha) \\
-\sin \beta & -\sin(\beta - \alpha) & -\sin(\beta - 2\alpha) & -\sin(\beta + 2\alpha) & -\sin(\beta + \alpha) \\
1 & \cos(2\alpha) & \cos(4\alpha) & \cos(4\alpha) & \cos(2\alpha) \\
0 & \sin(2\alpha) & \sin(4\alpha) & -\sin(4\alpha) & -\sin(2\alpha) \\
\frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}}
\end{bmatrix}
\]

(3.18)

The angles of transformation for stator quantities and for rotor quantities are related to the arbitrary speed of the selected common reference frame through:

\[
\begin{align*}
\theta_s &= \int \omega_d \, dt \\
\beta &= \theta_s - \theta = \int (\omega_d - \omega) \, dt
\end{align*}
\]

(3.19)

where \( \omega \) is the instantaneous electrical angular speed of rotation of the rotor.

3.3.3 Machine model in an arbitrary common reference frame

Correlation between original phase variables and new variables in the transformed domain is governed with the following transformation expressions:

\[
\begin{align*}
\psi'_{dq} &= A_s \psi_{abcde} \\
\psi'_{dq} &= A_r \psi_{abcde} \\
\psi'_{dq} &= A_r \psi_{abcde} \\
\psi'_{dq} &= A_r \psi_{abcde}
\end{align*}
\]

(3.20)

Substitution of (3.7)-(3.8) into (3.17) and application of (3.17)-(3.18) yields the machine’s voltage equations in the common reference frame:

\[
\begin{align*}
v_x &= R_i l_x - \omega_x \psi_x + P \psi_x \\
v_y &= R_i l_y + \omega_x \psi_x + P \psi_x \\
v_z &= R_i l_z + P \psi_x \\
v_x &= R_i l_x + P \psi_x \\
v_y &= R_i l_y + P \psi_x \\
v_z &= R_i l_z + P \psi_x
\end{align*}
\]

(3.21)
Transformation of flux linkage equations of (3.7)-(3.8) results in:

\[ \psi_\alpha = (L_\alpha + 2.5M)I_\alpha + 2.5Ml_\phi \]
\[ \psi_\beta = (L_\beta + 2.5M)I_\beta + 2.5Ml_\phi \]
\[ \psi_\alpha = L_\alpha I_\alpha \]
\[ \psi_\beta = L_\beta I_\beta \]
\[ \psi_\alpha = L_\alpha I_\alpha \]
\[ \psi_\beta = L_\beta I_\beta \]

(3.22)

Introduction of the magnetising inductance \( L_m = 2.5M \) enables writing of (3.22) in the following form:

\[ \psi_\alpha = (L_\alpha + L_m)I_\alpha + L_m l_\phi \]
\[ \psi_\beta = (L_\beta + L_m)I_\beta + L_m l_\phi \]
\[ \psi_\alpha = L_\alpha I_\alpha + L_m l_\alpha \]
\[ \psi_\beta = L_\beta I_\beta + L_m l_\beta \]
\[ \psi_\alpha = L_\alpha I_\alpha + L_m l_\alpha \]
\[ \psi_\beta = L_\beta I_\beta + L_m l_\beta \]

(3.23)

Finally, transformation of the original torque equation (3.15b) yields

\[ T_r = \frac{5P}{2} M \left[ q_\phi l_\phi - d_\phi l_q \right] \]
\[ T_r = PL_m \left[ q_\phi l_\phi - d_\phi l_q \right] \]

(3.24)

Mechanical equation of rotor motion is invariant under the transformation and is

\[ T_r - T_L = \frac{J}{P} \frac{d\omega}{dt} \]

(3.25)

3.4 Rotor flux oriented control of a five-phase induction motor

3.4.1 Indirect vector controller

The basis of vector control is the selection of the speed of the common reference frame. In rotor flux oriented control scheme the speed of the reference frame is selected as equal to the speed of the rotor flux space vector. The rotor flux space vector is kept aligned at all times with the real axis (d-axis) of the common reference frame, while q-axis is perpendicular to it. As the rotor flux space vector is aligned with the real axis its imaginary component always remains equal to zero.

Rotor flux oriented reference frame is defined with

\[ \theta_v = \phi_v \]
\[ \theta_r = \phi_r - \theta \]
\[ \omega_v = \omega_r \]
\[ \omega_r = \frac{d\phi_r}{dt} \]

(3.26)
where angle $\phi_r$ denotes instantaneous rotor flux space vector position. Rotor flux space vector becomes a pure real variable in this special frame of reference,

$$\psi_r = \psi_d + j \psi_q = \psi_r$$  \hspace{1cm} (3.27a)

i.e., it follows that

$$\psi_d = \psi_r, \quad \psi_q = 0 \quad \frac{d\psi_q}{dt} = 0$$  \hspace{1cm} (3.27b)

Machine model obtained by transforming only d-q windings with the rotational transformation is considered further on. Assuming that the machine is current fed, one can omit the stator voltage equations from further consideration. From the model of the machine in an arbitrary reference frame (3.21), (3.23), (3.24), with stator equations omitted,

$$0 = R_i l_d - (a_d - \omega)\psi_d + pl \psi_d$$
$$0 = R_i l_q + (a_q - \omega)\psi_q + pl \psi_q$$
$$0 = R_i l_r + pl \psi_r$$
$$0 = R_i l_w + pl \psi_w$$
$$0 = R_i l_0 + pl \psi_0$$

$$\psi_d = (L_d + L_m) l_d + L_m l_w$$
$$\psi_q = (L_q + L_m) l_q + L_m l_w$$
$$\psi_r = L_0 l_r$$
$$\psi_w = L_0 l_w$$
$$\psi_0 = L_0 l_0$$

$$T_e = p L_m (l_d l_q - l_d l_q)$$  \hspace{1cm} (3.29)

and observing that x-y-0 rotor current and flux components are identically equal to zero, one further obtains by substitution of (3.27) into (3.28)-(3.30)

$$\psi_r + T_e \frac{d\psi_r}{dt} = L_m l_d$$  \hspace{1cm} (3.31)

$$(a_r - \omega)\psi_r T_e = L_m l_q$$
$$\omega_m = \frac{L_m l_q}{T_e \psi_r}$$  \hspace{1cm} (3.32)
$$T_e = \frac{p L_m}{l_r} \psi_q l_q$$  \hspace{1cm} (3.33)

where $T_e = L_r / R_r$. It can be seen from (3.31)-(3.33) that the flux and torque producing currents in five-phase machines are only d-q components, thus the vector control scheme for a current fed five-phase machine is identical to the scheme for a current fed three-phase machine. The
only difference is that the co-ordinate transformation now generates five phase current references instead of three. The configuration of the indirect vector controller for operation in the base speed region is illustrated in Fig. 3.3 for the five-phase induction machine. Constants in Fig. 3.3 are determined with the following expressions (which are in essence identical to those for a three-phase induction machine with indirect rotor flux oriented control):

\[
i_{qs}^* = K_1 T_e^* \Rightarrow K_1 = i_{qs}^*/T_e^* = \frac{1}{P} \frac{L_r}{L_m} \frac{1}{\psi_r} = \frac{1}{P} \frac{L_r}{L_m} \frac{1}{i_{ds}}
\]

\[
\omega_{st}^* = K_2 i_{qs}^* \Rightarrow K_2 = \omega_{st}^*/i_{qs}^* = \frac{L_m}{T_r \psi_r} = \frac{1}{T_r i_{ds}}
\]

(3.34)

Determination of these constants is discussed next, in the following subsection. The same applies to the design of the PI speed controller that will be needed in simulations related to the speed mode of operation of the drive. Induction motor data are given in the Appendix A.

Figure 3.3. Indirect vector control of a five-phase induction machine in the base speed region.

### 3.4.2 Design of the indirect vector controller

In order to design an indirect vector controller one needs to determine constants $K_1$ and $K_2$ in Fig. 3.3 and speed PI controller parameters.

In steady state operation under rated operating conditions (index $n$ stands for rated values) one has

\[
T_{en} = \frac{P L_m}{L_r} \psi_r i_{qs} \\
\psi_m = L_m i_{ds} \\
\omega_{sn} = \frac{L_m i_{qn}}{T_r \psi_m}
\]

(3.35)

\[
\psi_r^* = \psi_m \\
T_L = T_{en} \\
\omega_{sl} = \omega_{sn}
\]

(3.36)
The stator current RMS value will equal the rated value. Since power invariant transformation is used, this means that the magnitude of the stator current space vector will be $\sqrt{5}$ greater than the RMS value (which is 2.0 A). Hence

$$i_m = \sqrt{5}I_m = 4.4721\, \text{A}$$

$$i_m = \sqrt{i_{dm}^2 + i_{qn}^2}$$

Taking into account that the rated RMS rotor flux is 0.57 Wb and the rated per-phase torque is 1.667 Nm, one has

$$\psi_{rm} = 0.57 \times \sqrt{5} = 1.274\, \text{Wb}$$

$$T_{em} = 8.0\, \text{Nm}$$

The rated torque is determined with (3.35) where the pole pair number equals two and the magnetising and the rotor inductances are 0.41 H and 0.45 H, respectively. By solving (3.37) and the torque equation of (3.35) one gets the rated stator d-q axis current components

$$i_{dm} = 3.015\, \text{A}$$

$$i_{qn} = 3.5511\, \text{A}$$

The two constants defined in (3.34) and required in the indirect vector control scheme of Fig. 3.10 are finally

$$K_1 = \frac{1}{P} L_r \frac{1}{\psi_r^*} = \frac{1}{P} L_m \frac{1}{i_{dm}} = 0.420$$

$$K_2 = \frac{\omega^*}{i_{qr}} = \frac{L_m}{T_r \psi_r^*} = \frac{1}{T_r i_{dm}^*} = 4.412$$

3.5 Current control techniques and speed controller design

3.5.1 General considerations

Current controlled PWM inverter is the most frequent choice in high performance ac drives as decoupled flux and torque control by instantaneous stator current space vector amplitude and position control is achieved relatively easily.

All the current control techniques for VSIs essentially belong to one of the two major groups. The first group encompasses all the current control methods that operate in the stationary reference frame while the second group includes current control techniques with current controllers operating in the rotational frame of reference. If the current control of an induction machine is performed in rotational reference frame, decoupling of stator voltage equations substitutes local current feedback loops in stationary reference frame, which suppress influence of stator voltage equations.

Current control in stationary reference frame is usually implemented in an analogue fashion. Three types of current control techniques are met in the literature:
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1. Hysteresis control
2. Ramp-comparison control
3. Predictive or adaptive control

Hysteresis controllers utilise a hysteresis band in comparing the actual current with the reference current. The ramp-comparison controllers compare the current error to a triangular carrier waveform to generate the switching signals for the inverter. In predictive control schemes the inverter voltage is calculated to force the current to follow the reference current.

The most pronounced shortcoming of the hysteresis current control is the variable inverter switching frequency over a period of output voltage. Current control by ramp-comparison controllers overcomes this problem. Here current error serves as modulating signal, which is compared to the triangular carrier wave. Deviation of amplitude and phase of phase currents with respect to commanded values unfortunately takes place and some compensation has to be introduced. Another difficulty arises from a possibility that multiple crossing of the carrier may occur if the frequency of the current error becomes greater than the carrier frequency. This can be overcome by adding hysteresis to the controller. The advantage of the ramp-comparison current control with respect to hysteresis current control is the fixed and constant inverter switching frequency. The predictive control is characterised by a fixed switching frequency as well. However, time required for the calculations is significant. Moreover, an efficient prediction requires the load knowledge as well.

Ramp-comparison current control in stationary reference frame requires that current PI controllers process alternating signals, that can be of a large frequency range. Furthermore, controller characteristics in steady state depend on the operating frequency and the machine impedance. These shortcomings can be partially but not completely eliminated by different modifications of the basic current control principles. At low operating speeds the induced rotational electromotive force in the machine is small and current control enables very good tracking between reference and actual currents, with respect to both amplitude and phase. However at higher speeds, due to limited voltage capability of the inverter and finite inverter switching frequency, tracking worsens and an error is met in both amplitude and phase of actual currents compared to reference currents. This feature becomes very pronounced in the field-weakening region where the inverter operates very close to the voltage limit. The problem may be solved by moving current controllers from the stationary to the rotating reference frame. The outputs of the current controllers then become voltage references in rotational reference frame. If the inverter switching frequency is high enough, decoupling circuit for stator dynamics is usually omitted.
Current control in rotational reference frame is well suited to fully digital realisation. The main advantage of this method of current control is that current controllers (most frequently of PI type) process dc signals. As the current control is performed in rotational reference frame, measured currents have to be transformed from stationary to rotational reference frame. When current control in rotational reference frame is applied, different PWM methods may be utilised for creation of the desired voltages at machine terminals. For example, sinusoidal PWM may be selected or voltage space vector modulation may be chosen.

3.5.2 Hysteresis Current controllers

The structure of an induction motor drive with current control in the stationary reference frame is depicted in Fig. 3.4. Five independent controllers in five phases \( a, b, c, d, e \) reference frame are used. The current controllers in Fig. 3.4 can be of either hysteresis type or ramp-comparison type.

The idea of hysteresis current control is in essence very simple and well suited for analogue realisation. Actual currents are allowed to deviate from their reference values for a fixed value termed as hysteresis band (Fig. 3.5). The discrepancy between actual and reference currents will vary in time and will be either positive or negative. The values of the hysteresis band are the same for both positive and negative variation. The state of the appropriate leg of the inverter bridge changes once when the difference between actual and reference current exceeds hysteresis band. For example, suppose that the upper switch in phase ‘a’ leg is closed, while the lower switch is open. This state will be preserved as long as the current error in phase ‘a’ is within hysteresis band. However, when the actual current in phase ‘a’ becomes greater than the reference value plus hysteresis band, the upper switch will be opened and the lower switch will be closed. Thus the actual current will be forced to reduce and fall once more within the hysteresis band. As the actual current change in time is function of the drive dynamics and operating state, the instants of inverter semiconductor switching cannot be predicted and will vary. Furthermore the switching frequency of the inverter varies and is not constant even over one cycle of the output frequency. The principle of hysteresis current control is illustrated in Fig. 3.6, where the inverter leg voltage is shown as well. One clearly observes in Fig. 3.6 how the switching frequency of the inverter varies from one cycle of operation to the other cycle. Periods of inverter switching are denoted as \( T_1, T_2, T_3 \) and \( T_4 \) and one easily observes that \( T_1 \) is the largest out of the four, while \( T_4 \) is the smallest.
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In the system shown in Fig. 3.4 only speed controller is present (its input is the speed error, not shown in Fig. 3.4), whose output becomes after appropriate scaling q-axis current command. The d-axis reference current is obtained by dividing the reference rotor flux by magnetising inductance.

Hysteresis band for hysteresis current control is selected in all simulations (for both a single-motor drive in this chapter and multi-motor drives in subsequent chapter) as ±2.5% of the motor rated (peak) current value, i.e. as equal to ±0.07 A (motor rated RMS current is 2.0 A). Speed controller parameter setting is discussed in detail in section 3.5.4.

![Figure 3.4. Induction motor drive structure with current control in the stationary reference frame.](image)

![Figure 3.5. Hysteresis current controller.](image)

### 3.5.3 Ramp-comparison current control

The structure of the drive system with ramp-comparison controller is again the one shown in Fig. 3.4. Current error is formed in the stationary reference frame and is further passed through PI current controllers. The outputs of the current controllers are phase voltage
references which are compared to the triangular carrier wave of the fixed frequency. As the triangular carrier wave is of fixed frequency, while frequency of the current error varies, the ratio of these two frequencies is not an integer and so-called asynchronous PWM results. Asynchronous PWM in general leads to generation of unwanted sub-harmonics in the output voltage waveform. However, if the triangular carrier wave frequency is high enough this effect can be neglected as it will not have any serious impact on the drive behaviour. In general, five carrier waves are needed, one per phase. However, if the triangular carrier frequency is high enough, one carrier wave may be used for all the five phases. Generation of the inverter voltages using ramp-comparison control is illustrated in Fig. 3.7.

![Diagram of hysteresis current control](image)

Figure 3.6. Principle of hysteresis current control (currents and a leg voltage).

The tuning of PI current controller is discussed next. The frequency of the triangular carrier wave for ramp-comparison control is fixed at all times to 5 kHz and its amplitude is ±1. The outputs of the PI current controllers, which are implemented as discrete PI controllers, are limited to ±1 in order to ensure operation in the full PWM mode at all times. Tuning of the current controller parameters is performed first, with speed control loop kept open. Phase current references are generated through the co-ordinate transformation, from imposed d-q axis current references. Stator d-q axis current references are given as pulsed current waveforms, with a period of 0.015 s and the pulse duration of 0.0045 s. The amplitude of the pulse in d-q axis current references is set to 2.45 A. The integral gain is kept at a sufficiently low value and the proportional gain is gradually increased in order to achieve
good tracking between the pulsed phase current reference and the actual current response. Once the current response becomes of acceptable nature, the proportional gain is fixed and the integral gain is gradually altered to obtain an acceptable overshoot in the current response. The procedure is illustrated in Fig. 3.8, where the reference pulsed phase ‘a’ current and the actual current response are shown for four different pairs of proportional and integral gains. The current response obtained with the proportional gain equal to 0.65 and the integral gain equal to 0.75 was deemed to be acceptable and is selected for further work. It should be noted that, due to the discrete form of the PI controller, what is called here integral gain is, strictly speaking, not the integral gain. The value of 0.75 mentioned above is in essence a product of the sampling time (20 μs) and the integral gain of the continuous PI controller equivalent.

![Figure 3.7. Ramp-comparison current control.](image)

The same tuning procedure was repeated for the two-motor drive systems (five-phase). The impedance seen by current controllers changes due to the series connection of two machines. Figures 3.9 and 3.10 illustrate the phase ‘a’ current reference (which is the same as for one five-phase machine case) and the actual current response for the five-phase two-motor drive and the six-phase two-motor drive system, respectively. As can be seen from these two figures, the current response obtained with the proportional gain of 0.65 and the integral gain of 0.75 is very similar as for the single-motor five-phase drive. These current controller parameters are therefore used in all simulations related to ramp-comparison current control, reported in this and the subsequent chapter.
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3.5.4 Speed controller design

PI speed controller is considered next. Two different speed controllers are designed, a continuous one and a discrete one. Both speed controllers are used in simulations further on. The type of the speed controller used in conjunction with any specific simulation will be indicated in the corresponding section with simulation results. The design of continuous speed
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controller is presented first. For this purpose, and having in mind that the inverter current control will be performed in the stationary reference frame using hysteresis or ramp-comparison technique, the whole current control loop is approximated with unity gain and zero time delay. The structure of the speed control loop is then as shown in Fig. 3.10.

The transfer function of the PI speed controller is

\[ G_{pr}(s) = K_p \left( 1 + \frac{1}{sT_i} \right) = K_p + K_i \frac{1}{s} \]  

(3.41)

![Figure 3.10. Structure of the speed control loop.](image)

The characteristic equation of the above speed control loop is solved for \( 1 + G_{pr}(s)H(s) = 0 \), where according to Fig. 3.10 \( H(s) = 1/(sJ/P) \). The parameters are \( J = 0.025 \text{kgm}^2 \), \( P = 2 \). Hence the characteristic equation is

\[ s^2 + 66.67K_p s + 66.67K_i = 0 \]  

(3.42)

The coefficients in the above equation are equated with those in the following equation, which defines the desired closed loop dynamics in terms of the damping ratio \( \xi \) and natural frequency \( \omega_o \):

\[ s^2 + 2\xi\omega_o s + \omega_o^2 = 0 \]  

(3.43)

Damping ratio is selected as 0.712. The natural frequency for the speed control loop is dependent on the bandwidth of the inner current control loop. Maximum practical value of the current control loop bandwidth in the case of ramp-comparison control with 5 kHz switching frequency is 1 kHz. For the purpose of the speed controller design, current loop bandwidth is taken as one tenth of the maximum value (i.e. as 100 Hz). Taking the speed control bandwidth as one tenth of this value (10 Hz) and approximating the natural frequency with the bandwidth, one has \( \omega_o = 2\pi 10 = 62.8318 \text{rad/s} \). Substitution of the damping ratio and natural frequency values into (3.43) and comparison with (3.42) yields the following values for the speed controller parameters:

\[ K_p = 1.313, \quad K_i = 59.116, \quad T_i = 0.246 \]  

(3.44)
In the design of a discrete PI speed controller the same procedure is adopted as the one used for the PI current controller tuning. A pulsed speed reference (of 200 rad/s amplitude) is applied as the input of the PI speed controller and the speed of the machine is observed. The reference speed pulsed waveform is applied at 0.2 s. The period of the reference speed pulse is equal to 0.6 s. and the pulse duration is 0.3 s. Tuning is again performed using the single-machine five-phase drive system. The responses obtained with four different pairs of proportional and integral gain values are shown in Fig. 3.11. The speed response obtained with the proportional gain equal to 2.2 and the integral gain equal to 0.1214 is deemed as being satisfactory. The integral gain is rounded to 0.12.

These discrete speed controller parameters \( K_p = 2.2, K_i = 0.12 \) are used further on in simulations of a single five-phase induction machine described in this chapter, in conjunction with both hysteresis current control and ramp-comparison current control. For the speed controller there is no need to perform additional tuning trials for the two-motor drive systems, since addition of the second machine in series with the first one affects only the current control loop, while the speed control loop remains unaffected.

Figure 3.11. Tuning of the speed PI controller for a single five-phase induction motor drive, using a pulsed speed reference. The values of the proportional and the integral gain are, respectively: a. 0.2 and 0.002; b. 2.2 and 0.002; c. 1.5 and 0.002; d. 2.2 and 0.1214.
3.6 Simulation of a single five-phase induction motor drive

A simulation program is written using MATLAB/SIMULINK software for an indirect rotor flux oriented five-phase induction motor drive. The motor is simulated using developed d-q model in the stationary reference frame. The machine is fed by a PWM voltage source inverter and hysteresis current control is exercised upon motor phase currents. The drive is operated in closed loop speed control mode with discrete anti-windup PI speed controller. The anti-windup feature restricts the saturation of the integral part of the controller while working in the limiting region. The torque is limited to twice the rated value (16.0 Nm). The drive is simulated for acceleration, disturbance rejection and speed reversal transients, at operating speed 1200 rpm.

Forced excitation is initiated first. Rotor flux reference (i.e. stator d-axis current reference) is ramped from \( t = 0 \) to \( t = 0.01 \) s to twice the rated value. It is further reduced from twice the rated value to the rated value in a linear fashion from \( t = 0.05 \) to \( t = 0.06 \) s and it is then kept constant for the rest of the simulation period. Once the rotor flux has reached steady state, a speed command of 1200 rpm is applied at \( t = 0.3 \) s in ramp wise manner from \( t = 0.3 \) to \( t = 0.35 \) s. The inverter dc link voltage is set to \( 420\sqrt{2} = 593.1 \) V. A step load torque, equal to the motor rated torque (8.0 Nm), is applied at \( t = 1 \) s and the machine is allowed to run for sufficient time so as to reach the steady state condition. A speed reversal is then initiated in the ramp-wise manner (ramp duration from \( t = 1.2 \) to \( t = 1.25 \) s).

Simulation results for the 1200 rpm speed command are shown in Figs. 3.12-3.14. In particular, Fig. 3.12 illustrates rotor flux and rotor flux reference for the complete duration of the transient, as well as motor speed response, torque response and reference and actual current during the acceleration transient. Rotor flux settles to the reference value after initial transient and then it remains constant throughout the simulation period (2 seconds), indicating that full decoupling between rotor flux and torque control has been achieved. During acceleration motor torque and speed follow the commanded value. Acceleration takes place with the maximum allowed value of the motor torque. Actual motor phase current tracks the reference very well. Consequently, torque response closely follows torque reference. There is sufficient voltage reserve to enable the complete acceleration transient to take place in the torque limit.

Disturbance rejection properties of the drive are studied by applying rated load torque to the machine and the resulting responses are shown in Fig. 3.13. Application of the load torque causes an inevitable speed dip. Motor torque quickly follows the torque reference and
enables rapid compensation of the speed dip. The motor torque settles at the value equal to the load torque in less than 100 ms and the motor current becomes rated at the end of the transient. Fig. 3.13 also depicts the stator phase voltage, which is typical for a PWM inverter fed motor drive.

Speed reversal transient study is also simulated and the resulting responses are shown in Fig. 3.14. Once more it is observed that the actual torque closely follows the reference, leading to the rapid speed reversal, with torque in the limit, in the shortest possible time interval (approximately 350 ms). The change of phase sequence in stator current because of the change in the rotational direction is clearly observed from the plot of the stator current.

![Figure 3.12. Excitation and acceleration transients (1200 rpm speed command) using hysteresis current control: a) actual and reference rotor flux, b) actual and reference torque, and speed, c) actual and reference stator phase 'a' currents.](image)

![Figure 3.13. Disturbance rejection transient at 1200 rpm using hysteresis current control: a) actual and reference torque, and speed, b) actual and reference stator phase 'a' currents, c) stator phase 'a' voltage.](image)

![Figure 3.14. Reversing transient (1200 rpm) using hysteresis current control: a. actual and reference torque, and speed, b. actual and reference stator phase 'a' currents, c. stator phase 'a' voltage.](image)
3.7 Five-phase series-connected two-motor drive structure: A Review

The concept and detailed treatment of five-phase series connected two motor drive is presented in Iqbal (Ph.D. Thesis, 2005). Vector control enables independent control of flux and torque of an ac machine by means of only two stator current components (one component pair: d-q). This leaves one pair of components as additional degrees of freedom in case of five-phase machines as shown in Levi et al (2003d). Hence, if it is possible to connect stator windings of two five-phase machines in such a way that what one machine sees as the d-q axis stator current components the other machine sees as x-y current components, and vice-versa, it would become possible to completely independently control the speed (position, torque) of these two machines while supplying them from a single current controlled PWM voltage source inverter. In simple terms, it is possible to independently realise vector control of two five-phase machines using a single VSI, provided that the stator windings of the two machines are connected in series and that an appropriate phase transposition is introduced so that the set of five-phase currents that produce rotating mmf in the first machine, does not produce rotating mmf in the second machine and vice-versa. This explanation constitutes the basis of the two-motor five-phase drive system. The proof and exhaustive explanation of the concept for general n-phase machine series connections is given in Jones (2002), Jones (2005) and Iqbal (2006).

![Diagram of Five-phase two-motor drive with series connection of phase windings and an appropriate phase transposition.](image)

Figure 3.15. Five-phase two-motor drive with series connection of phase windings and an appropriate phase transposition.

The required phase transposition is analysed in detail in Jones (2002), Jones (2005) and Levi et al (2003d) and Iqbal (2006). On the basis of considerations in these references it
is possible to construct a connection table, given in Table 3.1. The corresponding connection diagram is given in Fig. 3.15 (with the change from machine phase order symbols 1,2,3,4,5 to a,b,c,d,e).

Table 3.1 Connectivity matrix for five-phase two-motor drive.

<table>
<thead>
<tr>
<th>Machine number</th>
<th>A</th>
<th>B</th>
<th>C</th>
<th>D</th>
<th>E</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>2</td>
<td>3</td>
<td>4</td>
<td>5</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>3</td>
<td>5</td>
<td>2</td>
<td>4</td>
</tr>
</tbody>
</table>

Phase variable model of two five-phase induction machines connected in series according to Fig. 3.15 is developed in state space form (Iqbal 2005). Each of the two five-phase induction machines can be represented with the model developed in first part of this chapter and described with equations (3.7) to (3.14) and (3.16).

Due to the series connection of two stator windings according to Fig. 4.1 the following holds true:

$$
\begin{align*}
\bar{v}_A &= v_{a1} + v_{a2} \\
v_b &= v_{b1} + v_{b2} \\
v_c &= v_{c1} + v_{c2} \\
v_d &= v_{d1} + v_{d2} \\
v_e &= v_{e1} + v_{e2} \\
i_A &= i_{a1} + i_{a2} \\
i_B &= i_{b1} + i_{b2} \\
i_C &= i_{c1} + i_{c2} \\
i_D &= i_{d1} + i_{d2} \\
i_E &= i_{e1} + i_{e2}
\end{align*}
$$

(3.45) (3.46)

Capital letters in indices stand for inverter phase-to-neutral voltages and inverter phase currents in equations (3.45)-(3.46).

By omitting the x-y and zero-sequence equations for rotor windings and the zero-sequence equation of the inverter, the complete model in stationary reference frame for the two five-phase series-connected machines can be written in developed form as:

$$
\begin{align*}
\bar{v}_a^\text{inv} &= R_1 i_a^\text{inv} + (L_1 + L_m) \frac{di_a^\text{inv}}{dt} + L_m i_a^\text{inv} + R_{s1} i_a^\text{inv} + L_{s1} \frac{di_a^\text{inv}}{dt} \\
v_b^\text{inv} &= R_1 i_b^\text{inv} + (L_1 + L_m) \frac{di_b^\text{inv}}{dt} + L_m i_b^\text{inv} + R_{s1} i_b^\text{inv} + L_{s1} \frac{di_b^\text{inv}}{dt} \\
v_c^\text{inv} &= R_1 i_c^\text{inv} + (L_1 + L_m) \frac{di_c^\text{inv}}{dt} + L_m i_c^\text{inv} + R_{s1} i_c^\text{inv} + L_{s1} \frac{di_c^\text{inv}}{dt} \\
v_d^\text{inv} &= R_1 i_d^\text{inv} + (L_1 + L_m) \frac{di_d^\text{inv}}{dt} + L_m i_d^\text{inv} + R_{s1} i_d^\text{inv} + L_{s1} \frac{di_d^\text{inv}}{dt} \\
v_e^\text{inv} &= R_1 i_e^\text{inv} + (L_1 + L_m) \frac{di_e^\text{inv}}{dt} + L_m i_e^\text{inv} + R_{s1} i_e^\text{inv} + L_{s1} \frac{di_e^\text{inv}}{dt}
\end{align*}
$$

(3.47)
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\[ 0 = R_1 \frac{di_{\text{aq}}}{dt} + L_{\text{m1}} \frac{di_{\text{aq}}}{dt} + \left( L_{\text{a1}} + L_{\text{m1}} \right) \frac{di_{\text{aq}}}{dt} + a_1 \left( L_{\text{m1}}i_{\text{aq}} + \left( L_{\text{a1}} + L_{\text{m1}} \right) i_{\text{aq}} \right) \]  
\[ \text{(3.48)} \]

\[ 0 = R_1 \frac{di_{\text{qo}}}{dt} + L_{\text{m1}} \frac{di_{\text{qo}}}{dt} + \left( L_{\text{q1}} + L_{\text{m1}} \right) \frac{di_{\text{qo}}}{dt} - a_1 \left( L_{\text{m1}}i_{\text{qo}} + \left( L_{\text{q1}} + L_{\text{m1}} \right) i_{\text{qo}} \right) \]  
\[ \text{(3.49)} \]

In addition to these equations that describe the two machines connected in series with phase transposition, one needs correlation between inverter phase and d-q voltages and currents.

\[ \begin{bmatrix} v_{\text{aq}} \\ v_{\text{qo}} \\ v_{\text{ao}} \\ v_{\text{op}} \end{bmatrix} = C \begin{bmatrix} v_x \\ v_y \\ v_0 \\ v_P \end{bmatrix} = C \begin{bmatrix} i_{\text{aq}} \\ i_{\text{qo}} \\ i_{\text{ao}} \\ i_{\text{op}} \end{bmatrix} \]  
\[ \text{(3.50)} \]

Correlation between inverter current components and stator current components of the two machines is, using (3.46) and the second of (3.50), given with:

\[ i_{\text{aq}} = i_{\text{am}} = i_{\text{aq}} \]
\[ i_{\text{qo}} = i_{\text{pm}} = -i_{\text{qo}} \]
\[ i_{\text{ao}} = i_{\text{am}} = i_{\text{ao}} \]
\[ i_{\text{op}} = i_{\text{pm}} = i_{\text{op}} \]
\[ \text{(3.51)} \]

The equation (3.51) implies that the \( a \)-axis and \( q \)-axis inverter current components are equal to the flux/torque producing stator current components of machine 1 and simultaneously \( x \)-axis and \( y \)-axis (in reverse direction) current components of machine 2. Similarly, \( x \)-axis and \( y \)-axis inverter current components are equal to the \( x \)-axis and \( y \)-axis stator current components of machine 1 and to flux/torque producing current components of machine 2.

Further, in order to calculate stator phase voltages for individual machines, one observes from (3.50) and (3.45) that

\[ v_{\text{aq}} = v_{\text{na1}} + v_{\text{na2}} \]
\[ v_{\text{qo}} = v_{\text{na1}} - v_{\text{na2}} \]
\[ v_{\text{ao}} = v_{\text{na1}} + v_{\text{na2}} \]
\[ v_{\text{op}} = v_{\text{na1}} + v_{\text{na2}} \]
\[ \text{(3.52)} \]

This implies that the \( a \)-axis and \( q \)-axis circuits of machine 1 are connected in series with \( x \)-axis and \( y \)-axis circuit of machine 2, respectively, and vice-versa. Hence
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\[
\begin{align*}
\dot{v}_{a1} &= R_{s1}^{PV} + (L_{a1} + L_{m1}) \frac{di_{a1}^{PV}}{dt} + L_{m1} \frac{di_{p1}}{dt} \\
\dot{v}_{\phi1} &= R_{s1}^{PV} + (L_{a1} + L_{m1}) \frac{di_{\phi1}^{PV}}{dt} + L_{m1} \frac{di_{p1}}{dt} \\
\dot{v}_{a1} &= R_{s1}^{PV} + L_{m1} \frac{di_{a1}^{PV}}{dt} \\
\dot{v}_{m1} &= R_{s1}^{PV} + L_{m1} \frac{di_{m1}^{PV}}{dt} \\
\dot{v}_{a2} &= R_{s2}^{PV} + (L_{a2} + L_{m2}) \frac{di_{a2}^{PV}}{dt} + L_{m2} \frac{di_{p2}}{dt} \\
\dot{v}_{\phi2} &= R_{s2}^{PV} + (L_{a2} + L_{m2}) \frac{di_{\phi2}^{PV}}{dt} + L_{m2} \frac{di_{p2}}{dt} \\
\dot{v}_{a2} &= R_{s2}^{PV} + L_{m2} \frac{di_{a2}^{PV}}{dt} \\
\dot{v}_{m2} &= R_{s2}^{PV} + L_{m2} \frac{di_{m2}^{PV}}{dt} \\
\end{align*}
\]

(3.53)

Individual machine phase voltages are then determined with

\[
\begin{bmatrix}
\dot{v}_{a1} \\
\dot{v}_{\phi1} \\
\dot{v}_{a1} \\
\dot{v}_{m1} \\
\dot{v}_{a2} \\
\dot{v}_{\phi2} \\
\dot{v}_{a2} \\
\dot{v}_{m2}
\end{bmatrix} = \begin{bmatrix}
-\dot{v}_{m1} \\
-\dot{v}_{a1} \\
\dot{v}_{a1} \\
\dot{v}_{m1} \\
-\dot{v}_{m2} \\
-\dot{v}_{a2} \\
\dot{v}_{a2} \\
\dot{v}_{m2}
\end{bmatrix} = C^T \begin{bmatrix}
\dot{v}_{a1} \\
\dot{v}_{\phi1} \\
\dot{v}_{a1} \\
\dot{v}_{m1} \\
\dot{v}_{a2} \\
\dot{v}_{\phi2} \\
\dot{v}_{a2} \\
\dot{v}_{m2}
\end{bmatrix}
\]

(3.54)

3.8 Vector control of the two-motor series-connected five-phase drive using current control in the stationary reference frame

The indirect rotor flux oriented vector control scheme is utilised to independently control the two series-connected five-phase induction machines. The vector controller used for individual machines is the same as the one used for controlling a single five-phase machine (Fig. 3.3). The basic drive structure with vector control scheme is shown in Fig. 3.16. It should be noted that the same vector control structure has been used in Jones (2002), Jones (2005) and Levi et al (2003d) for this two-motor drive system. However, the vector control scheme was developed in these references using steady state analysis and physical reasoning.

The stator windings of the two machines in Fig. 3.16 are connected as per the transposition rule of Table 3.1. The actual speeds of both machines are sensed and used as feedback signal for the speed and vector controllers. The two vector controllers generate appropriate phase current references which are summed according to the transposition rule to form the inverter phase current references. These reference currents are then compared with measured (actual) inverter phase currents to generate the phase current errors. An appropriate
inverter current control algorithm (hysteresis (section 3.6.1) is used to provide necessary switching signals to the power switches of the VSI to eliminate the phase current error.

![Figure 3.16. Vector control scheme for five-phase two-motor drive system (Iqbal 2006).](image)

The constants of Fig. 3.3 for each machine are determined as

\[
\begin{align*}
    \omega_{\alpha 1}^* &= K_{\alpha 1}(\alpha_{\alpha 1}) \rightarrow K_{\alpha 0} = \frac{\omega_{\alpha 1}}{\omega_{\alpha 1}} \Rightarrow \frac{1}{T_{\omega_{\alpha 1} \omega_{\alpha 1}}} = \frac{1}{T_{\omega_{\alpha 1} \omega_{\alpha 1}}} \\
    \omega_{\alpha 2}^* &= K_{\alpha 2}(\alpha_{\alpha 2}) \rightarrow K_{\alpha 0} = \frac{\omega_{\alpha 2}}{\omega_{\alpha 2}} \Rightarrow \frac{1}{T_{\omega_{\alpha 2} \omega_{\alpha 2}}} = \frac{1}{T_{\omega_{\alpha 2} \omega_{\alpha 2}}} \\
    \omega_{\alpha 3}^* &= K_{\alpha 3}(\alpha_{\alpha 3}) \rightarrow K_{\alpha 0} = \frac{\omega_{\alpha 3}}{\omega_{\alpha 3}} \Rightarrow \frac{1}{T_{\omega_{\alpha 3} \omega_{\alpha 3}}} = \frac{1}{T_{\omega_{\alpha 3} \omega_{\alpha 3}}} \\
    \omega_{\alpha 4}^* &= K_{\alpha 4}(\alpha_{\alpha 4}) \rightarrow K_{\alpha 0} = \frac{\omega_{\alpha 4}}{\omega_{\alpha 4}} \Rightarrow \frac{1}{T_{\omega_{\alpha 4} \omega_{\alpha 4}}} = \frac{1}{T_{\omega_{\alpha 4} \omega_{\alpha 4}}}
\end{align*}
\] (3.55)

\[
\begin{align*}
    \omega_{\alpha 1}^* &= K_{\alpha 1}(\alpha_{\alpha 1}) \rightarrow K_{\alpha 0} = \frac{\omega_{\alpha 1}}{\omega_{\alpha 1}} \Rightarrow \frac{1}{T_{\omega_{\alpha 1} \omega_{\alpha 1}}} = \frac{1}{T_{\omega_{\alpha 1} \omega_{\alpha 1}}} \\
    \omega_{\alpha 2}^* &= K_{\alpha 2}(\alpha_{\alpha 2}) \rightarrow K_{\alpha 0} = \frac{\omega_{\alpha 2}}{\omega_{\alpha 2}} \Rightarrow \frac{1}{T_{\omega_{\alpha 2} \omega_{\alpha 2}}} = \frac{1}{T_{\omega_{\alpha 2} \omega_{\alpha 2}}} \\
    \omega_{\alpha 3}^* &= K_{\alpha 3}(\alpha_{\alpha 3}) \rightarrow K_{\alpha 0} = \frac{\omega_{\alpha 3}}{\omega_{\alpha 3}} \Rightarrow \frac{1}{T_{\omega_{\alpha 3} \omega_{\alpha 3}}} = \frac{1}{T_{\omega_{\alpha 3} \omega_{\alpha 3}}} \quad (3.56)
\end{align*}
\]

For simulation purpose the two five-phase induction machines are assumed to be identical. Thus the constants in (3.55)-(3.56) are the same as in (3.37).

The phase current references are formed separately for the two machines, as follows:
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\[ \mathbf{C} = \mathbf{y}_j \left( \mathbf{4}_1 \cos(\mathbf{\phi}_1) - \mathbf{I}_n \sin(\mathbf{\phi}_1) \right) \]

\[ \mathbf{C}_1 = \mathbf{y}_j \left( \mathbf{4}_{i1} \cos(\mathbf{\phi}_{i1}) - \mathbf{I}_{i1} \sin(\mathbf{\phi}_{i1}) \right) \]

\[ \mathbf{C}_2 = \mathbf{y}_j \left( \mathbf{4}_{i2} \cos(\mathbf{\phi}_{i2}) - \mathbf{I}_{i2} \sin(\mathbf{\phi}_{i2}) \right) \]

\[ \mathbf{C}_3 = \mathbf{y}_j \left( \mathbf{4}_{i3} \cos(\mathbf{\phi}_{i3}) - \mathbf{I}_{i3} \sin(\mathbf{\phi}_{i3}) \right) \]

\[ \mathbf{C}_4 = \mathbf{y}_j \left( \mathbf{4}_{i4} \cos(\mathbf{\phi}_{i4}) - \mathbf{I}_{i4} \sin(\mathbf{\phi}_{i4}) \right) \]

\[ \mathbf{C}_5 = \mathbf{y}_j \left( \mathbf{4}_{i5} \cos(\mathbf{\phi}_{i5}) - \mathbf{I}_{i5} \sin(\mathbf{\phi}_{i5}) \right) \]

Overall inverter current references are given with:

\[ \mathbf{I}_s = \frac{2}{5} \left( \mathbf{I}_{i1} \cos(\mathbf{\phi}_{i1}) - \mathbf{I}_{i1} \sin(\mathbf{\phi}_{i1}) \right) \]

\[ \mathbf{I}_s = \frac{2}{5} \left( \mathbf{I}_{i2} \cos(\mathbf{\phi}_{i2}) - \mathbf{I}_{i2} \sin(\mathbf{\phi}_{i2}) \right) \]

\[ \mathbf{I}_s = \frac{2}{5} \left( \mathbf{I}_{i3} \cos(\mathbf{\phi}_{i3}) - \mathbf{I}_{i3} \sin(\mathbf{\phi}_{i3}) \right) \]

\[ \mathbf{I}_s = \frac{2}{5} \left( \mathbf{I}_{i4} \cos(\mathbf{\phi}_{i4}) - \mathbf{I}_{i4} \sin(\mathbf{\phi}_{i4}) \right) \]

\[ \mathbf{I}_s = \frac{2}{5} \left( \mathbf{I}_{i5} \cos(\mathbf{\phi}_{i5}) - \mathbf{I}_{i5} \sin(\mathbf{\phi}_{i5}) \right) \]

(3.57)

(3.58)

3.9 Dynamic behaviour of the five-phase two-motor vector controlled drive

A simulation program is written using Matlab/Simulink software for an indirect rotor flux oriented five-phase two-motor drive. The series-connected induction machines are simulated using the developed model in the stationary reference frame, given with (4.53)-(4.56). The machines are fed by a PWM voltage source inverter and hysteresis current control (section 3.6.1) is exercised upon the total inverter phase currents. The drive is operated in closed-loop speed control mode using anti-windup PI speed controllers. The design of the speed controller has been described in section 3.5.4 and proportional and integral gains correspond to those of (3.53). The torque of both machines is limited to twice the rated value (16.67 Nm). The acceleration, disturbance rejection and speed reversal transients are simulated, with two machines running at two different speeds (the first machine at 1500 rpm and the second at 750 rpm).

Forced excitation is initiated first in both machines at the same time. Rotor flux reference (i.e. stator d-axis current reference) is ramped from \( t = 0 \) to \( t = 0.015 \) s to twice the rated value. It is further reduced from twice the rated value to the rated value in a linear fashion from \( t = 0.055 \) to \( t = 0.065 \) s and it is then kept constant for the rest of the simulation period. Once the rotor flux has reached steady state a speed command of 1500 rpm is applied at \( t = 0.35 \) s in ramp-wise manner from \( t = 0.35 \) to \( t = 0.4 \) s to machine 1 (IM1). The speed command of 750 rpm is applied to machine 2 (IM2) at \( t = 0.4 \) s in ramp-wise manner (from \( t = 0.4 \) to \( t = 0.46 \)).
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= 0.4 s to t = 0.45 s). The inverter dc link voltage is arbitrarily set to $2x(\sqrt{2}x420) = 1187.9$ V (the inverter input dc voltage is doubled since it feeds two machines). A step load torque, equal to the motor rated torque (8.0 Nm), is applied to IM1 at $t = 1$ s and load torque of one half the rated value (4.0 Nm) is applied to IM2 at $t = 1.1$ s. A speed reversal is initiated from these steady states (from $t = 1.26$ s to $t = 1.32$ s for IM1 and from $t = 1.31$ s to $t = 1.36$ s for IM2). The responses obtained by simulation are shown in Figs. 3.17-3.19.

It is seen from Fig. 3.17a that the rotor flux builds up independently in the two machines and, after attaining rated value in less than 0.1 second, remains further on unchanged regardless of what happens to any of the two machines (note that rotor flux plots in Fig. 3.17a apply to all three transients, i.e. to the time interval 0 to 2 s). This indicates that rotor flux control in any of the two machines is completely decoupled from torque control of both machines. Such a situation is confirmed in Fig. 3.17b and Fig. 3.17c, where torque and speed responses of the two machines are shown for the initial acceleration transients (up to 0.9 s). Torque variation in one machine does not affect torque of the other machine and vice-versa, so that both machines accelerate with the torque in the limit. One notices appearance of the torque ripple in IM2 prior to the application of the speed command to it (time interval 0.31 to 0.41 s). This is a consequence of the inverter current higher harmonics and is an expected consequence of the non-ideal nature of the inverter.

Stator phase ‘a’ current references of two machines, as well as the inverter phase ‘a’ current reference and actual current, are shown in Fig. 3.17d and Fig. 3.17e respectively for the same time interval. While the individual phase current references have the familiar waveform and are sinusoidal functions in final steady state, the inverter current reference (and hence the actual current as well) is highly distorted due to the summation described with expression (4.77). In final steady state of Fig. 3.17e inverter current references are sums of two sinusoidal functions, of 50 Hz and 25 Hz frequency, respectively.

Disturbance rejection properties of the drive are illustrated in Fig. 3.18. Since the rotor flux remains undisturbed (Fig. 3.17a), torque responses are the quickest possible, leading to a rapid compensation of the speed dip, caused by the load torque application. The control of the two machines is again completely independent.

Reversing transient is illustrated in Fig. 3.19. Fully decoupled flux and torque control, as well as a fully independent control of the two machines, is again evident from the results for this transient. The change of phase sequence in stator current because of the change in the rotational direction is evident from the current waveforms.

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Figure 3.17. Excitation and acceleration transients (a) actual and reference rotor flux of IM1 and IM2 (b) actual and reference torque, and speed of IM1 (c) actual and reference torque, and speed of IM2 (d) stator phase 'a' current references of IM1 and IM2 (e) actual and reference inverter phase 'a' currents.

Figure 3.18. Disturbance rejection transient: (a) actual and reference torque and speed of IM1 (b) actual and reference torque and speed of IM2 (c) stator phase 'a' current references of IM1 and IM2 (d) actual and reference inverter phase 'a' currents.
3.10 Summary

The first part of this chapter is devoted to review of the components of a single five-phase vector controlled induction motor drive. Principle of operation of a five-phase VSI, d-q modelling of a five-phase induction machine and principle of rotor flux oriented control were discussed. Current control in stationary reference frame was further reviewed and current and speed controllers were designed for the given machine and inverter data.

Performance of a vector controlled single five-phase induction machine drive, obtainable with hysteresis current control method, is further evaluated and illustrated for a number of operating conditions on the basis of simulation results. Full decoupling of rotor flux control and torque control was realised by hysteresis current control technique under the condition of a sufficient voltage reserve. Dynamics, achievable with a five-phase vector controlled induction machine, are shown to be essentially identical to those obtainable with a three-phase induction machine.

The second part of this chapter is devoted to the modelling and simulation of a five-phase two-motor drive with two induction machines connected in series with an appropriate phase transposition. The system is fed by a single five-phase VSI. The phase variable model of the drive system and then transformed into orthogonal set of equations using decoupling transformation and then into rotational transformation in the stationary common reference

Vector control principle is further discussed for the five-phase two-motor drive system. The concept and the models are verified by simulation of the whole system. Hysteresis current control in stationary reference frame is examined as means for the inverter current control. Dynamic performance with this current control method is as obtained. The results show that the two machines can be controlled independently using vector control principle without affecting each other. The torque producing current component of one machine becomes the non-torque producing component of the other and vice-versa. The same applies to the flux producing component.