DESIGN STRATEGIES USING ROTOR, STATOR AND MAGNET DISPOSITION FOR FLUX WEAKENING IN PM MOTORS WITH WIDE OPERATING SPEED RANGE

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Supervisor
Prof. Antonino Di Gerlando

A thesis submitted in partial fulfillment of the requirements for the degree of
Master of Science(Laura Magistrale)
in Electrical Engineering

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ABSTRACT

Environmental and regulatory requirements has lead to improved efficiency demands in industrial application, IEC 60034 – 30 standard demands very stringent requirements, which induction motors are hard pressed to satisfy. With easier and cheaper access to rare earth (NdFebB or SmCo) magnets and improved Permanent Magnet Synchronous machine (PMSM) design and control technologies, PMSM are being employed for applications field, hitherto occupied by IMs, such as traction, wind turbines, aerospace, etc. The objective of this thesis work is to introduce, analyze and compare different design topologies employed in PM machines for flux weakening operation at wide operating speed range. Different PMSM are described and the design topologies employed in each are analyzed with respect to their effect in issues involved with PMSM operations such as rotor eddy current loss, stator iron loss, partial permanent demagnetization, cogging and torque ripples, and the solutions for them with the help of modification in rotor, stator winding and magnet design has been presented in this thesis.
Requisiti ambientali e normativi hanno portato ad accresciute esigenze di efficienza in applicazioni industriali, con stringenti specifiche normative (cfr IEC 60034-30), che i Motori a Induzione (IM) difficilmente riescono a soddisfare. Grazie alla disponibilità e al minor costo dei magneti permanenti in terre rare (NdFeB o SmCo) e al miglioramento nel progetto e nel controllo dei Motori Sincroni a MP (PMSM), questi ultimi vengono impiegati per applicazioni nelle quali finora erano impiegati motori IM, come ad esempio la trazione, le turbine eoliche, le applicazioni aerospaziali, etc. L'obiettivo di questo lavoro di tesi è la descrizione, l'analisi e il confronto tra diverse topologie di macchina a MP, impiegate in applicazioni ad estesa zona di funzionamento con indebolimento di campo, in un'ampia intervallo di velocità. Sono descritti diversi tipi di PMSM e vengono analizzati i relativi aspetti di progettazione, riguardo a varie problematiche di funzionamento, quali perdite per correnti parassite nel rotore, perdite nel ferro di statore, parziale smagnetizzazione irreversibile dei MP, coppia di cogging e ondulazione di coppia a carico: nella tesi sono analizzate le soluzioni a tali problemi, basati su opportune modifiche di progetto del rotore, dell'avvolgimento statore e dei MP.
ACKNOWLEDGMENT

I would like to thank my advisor professor Antonino Di Gerlando, whose expertise and guidance help made this work possible. He was readily available to answer any of my queries and showed considerable forbearance with me. It was a great pleasure working with him.

I would also like to thank my parents, brother and sister, whose steadfast support has helped me grow as an individual. A special thanks, goes to all my professors and lecturers at Politecnico di Milano, whose invaluable guidance and teaching helped me in my studies.

Lastly, I'd like to convey my thanks to all my friends and family, they helped me silently in background without them life would be a lot less fun.
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1. INTRODUCTION

Electric motors in industrial applications consume between 30% and 40% of the generated electrical energy worldwide. In the European Union (EU), electric motor systems are by far the most important type of load in industry, using about 70% of the consumed electricity. In the tertiary sector (nonresidential buildings), although not so relevant, electric motor systems use about one-third of the electricity consumed. Their wide use makes electric motors particularly attractive for efficiency improvements, to satisfy environmental or legal regulation (binding or recommended), and for cost and loss minimization, etc. Three-phase, squirrel-cage induction motors (IMs) represent, by far, occupy the vast majority of the market of electric motors [Almeida et al. (2011)]. But, growing environmental concerns and high energy costs emphasize the importance of considering the life-cycle costs of nonstandard technologies such as Permanent Magnet (PM) machines. This has lead to the development of higher efficiency electric motors, leading to significant reductions in energy consumption, thereby reducing environmental impact.

1.1 Improving standards

The International Electro-technical Commission (IEC) is a non-profit, non-governmental international standards organization that prepares and publishes International Standards for all electrical, electronic and related technologies. A new international standard, in IEC 60034-30 standard, was formed in November 2008, for single-speed, line-fed, three-phase, squirrel-cage IMs. The objective was to promote a competitive motor market transformation, with a view to harmonize globally, motor energy efficiency classes. In this standard, three efficiency classes were proposed,

- standard efficiency (IE1) [the designation of the energy efficiency class consists of IE (short for International Energy Efficiency Class), directly followed by a numeral representing the classification],
- high efficiency (IE2), equivalent to EPAct
- premium efficiency (IE3), equivalent to National Electrical Manufacturers Association (NEMA) premium.

In addition, in the last proposal of the IEC 60034-31 technical specification standard, a super-premium efficiency (IE4) is also proposed, intended to be informative, since no sufficient market
Tab. 1.1: Motor technologies and their energy efficiency potential

<table>
<thead>
<tr>
<th>Motor type</th>
<th>Line-start</th>
<th>IE1</th>
<th>IE2</th>
<th>IE3</th>
<th>IE4</th>
<th>IE5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Three-phase SCM</td>
<td>Random and form wound windings; IP2x (open motors)</td>
<td>Yes</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
<td>No</td>
</tr>
<tr>
<td></td>
<td>IP4x and above (enclosed motors)</td>
<td>Yes</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
<td>No</td>
</tr>
<tr>
<td></td>
<td>Random wound windings</td>
<td>Yes</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
<td>No</td>
</tr>
<tr>
<td></td>
<td>Form wound windings</td>
<td>Yes</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
<td>No</td>
</tr>
<tr>
<td>3-phase wound-rotor induction motors</td>
<td>Yes</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td>Single-phase SCM</td>
<td>1 capacitor</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
<td>No</td>
<td>No</td>
</tr>
<tr>
<td></td>
<td>2 switchable capacitors</td>
<td>Yes</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
<td>No</td>
</tr>
<tr>
<td></td>
<td>VSD-fed PMSM</td>
<td>No</td>
<td>Yes</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
</tr>
<tr>
<td></td>
<td>Wound-rotor</td>
<td>Some</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
<td>No</td>
</tr>
<tr>
<td></td>
<td>LSPM</td>
<td>Yes*</td>
<td>Yes</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
</tr>
<tr>
<td></td>
<td>Sinusoidal-field reluctance</td>
<td>Some</td>
<td>Yes</td>
<td>Yes</td>
<td>Diff</td>
<td>No</td>
</tr>
</tbody>
</table>

and technological information is available to allow its standardization and more experience with such products is required. All the IE1, IE2, IE3, and IE4 efficiency levels are defined for the 0.75 to 375 kW power range, equivalent to the 1 to 500 hp range.

The IE4 class (proposed) can be applied both to line-fed motors and inverter plus motor units. The goal is to reduce the losses of IE4 by about 15% relative to IE3. For low-power levels (up to 7.5 kW). Full-load efficiency levels for motor-VSD units is shown in figure 1.1

Some European manufacturers see no technical feasibility to reach the proposed IE4 proposed levels with IM technology with the same IEC frame sizes as IE1/IE2-class IMs. However, very high-efficiency motors with permanent magnet (PM) rotor technology are now being introduced, developed and researched in the market, these are not only reaching but overtaking the proposed IE4 levels. Now, the levels of the IE5 efficiency class are being envisaged, to be incorporated into the next edition of the IEC 60034 – 30 standard, although IE5 is not yet defined completely. The goal is to reduce the losses of IE5 by some 20% relative to IE4. Table 1.1 shows the potential of different machines for achieving energy efficiency class.

The second edition of IEC 60034 – 30 standard will expand the covered product range significantly. The output-power range has been expanded, starting at 0.12 kW and ending at 800 kW. The standard is now applicable to both fixed speed and variable-speed (frequency converter supplied) motors, but the energy-efficiency classification for both, as given in it, is related to the losses at sinusoidal power supply only. Fixed-speed motors covered by this standard may be used in variable-speed applications (see IEC/TS 60034 – 25) but, in these cases, the actual efficiency of the motor is lower than the rated/marked efficiency due to increased losses associated with the harmonic content of the voltages produced by the variable-speed drive (VSD) (see IEC 60034 – 2 – 3).
1.2 Viable machines for Field weakening

Separately excited DC motors, if controlled properly, can provide ideal Torque / Power characteristics for wide speed application like traction. However, the presence of commutation device and associated problems is a limiting factor for them. Although, three-phase, squirrel-cage induction motors although do not have ideal Torque / Power characteristics as power associated with motor decreases with increasing speed, whereas FW region in general entails constant power beyond the base speed; still they did give a viable cost effective solution with improved control strategies and usage of VFDs (variable frequency drive). PMSM owing to their small size, high efficiency and new design and control topologies.

Permanent magnet synchronous machine (PMSM) with new and improved design and control strategies are encroaching in the application territories, hitherto occupied by IMs. Their advantages include a higher power factor with cooler operating temperature. Moreover, former disadvantages, such as the higher costs, particularly because of the PMs cost, have now been rendered obsolete with the advent of rare-earth elements such as Neodymium Iron Boron magnets (NdFeB). Therefore, even applications that were exclusively limited to asynchronous motors for cost reasons can now profit from the advantages of PM motors [Almeida et al. (2011)]. PMSMs were significantly more efficient than IMs in constant torque region but they had limited or no flux weakening region. But the new designs and control strategies have made these consideration more or less redundant now, and PM usage across industry is increasingly growing in popularity.

It is clear that based on the new regulations and new efficiency levels PMSM enjoy distinct
advantage for the fixed speed operation, especially in constant torque region, but as will be scene in next section the classical PMSM (permanent magnet synchronous machines) with distributed winding had limited or no flux weakening capabilities, which hampered its usage in a multitude of industries like traction, aerospace, direct-drive generators, automotive, etc. and in applications involving flywheel, turbochargers, and machining spindles, etc. The distinct advantages enjoyed by PM can be enumerated as

- small and compact size.
- high power density, relatively simple construction, and high torque density.
- wide speed range, including low speed crawling and high speed cruising;
- high efficiency over a wide torque and speed range;
- improved fuel economy and reduce emissions in the transportation sector.
- high reliability and robustness, low noise, and reasonable cost.

1.3 Thesis Objectives and Outline

In this thesis work the techniques for improving the design techniques with a focus on flux weakening capabilities are discussed and various machines with varying topology are compared to consider their effects in wide speed machine performances. In case of separately excited DC machines, two sets of windings resulting in "torque producing" and "field producing" currents. They are controlled independently. In the constant power region, the field current is reduced to produce the rated power. In PMSM motors, there is only a stator (armature) winding and field flux is produced by the permanent magnets.

In chapter 2 classification strategies to categorize permanent magnet are discussed along with the key characteristics of these machines. Following which the operating regions are described and their compatibility with wide speed traction application is demonstrated. Further, flux weakening (FW) concept is introduced for permanent magnet machine.

In chapter 3 the first design strategy using concentrated winding in place of traditional distributed winding in stator is discussed for IPM (section 3.2) and SPM (section 3.3 but they are equally effective in other PMSM design. Comparative analysis between IPM with the distributed winding (DW) and concentrated winding(CNW) has been made (section 3.1). Further SPM and IPM with fractional CNW(FSCW) are compared (section 3.7).

In chapter 4 rotor design topology for reduction of torque ripple and cogging in addition to favorable flux weakening are discussed are design that includes magnet segmentation(section 4.1), geometrical alignment of magnets (section 4.2) and Halbach arrays(section 4.4) are discussed.
In chapter 5 design improvements in axial machine for FW in wide speed range, with special consideration to improvements done using mechanical displacements of rotor and stator (section 5.3) has been discussed.

In chapter 6 the design strategies to improve FW in machines with magnets on stator are described with focus on FSPMs and a comparison is made with the traditional PMSM like IPM is made (section 6.2).

In chapter 7 a detailed conclusion of the thesis work is provided.
2. LITERATURE REVIEW

Permanent magnet machines are brushless machines, as they in general do not need a field winding in rotor and hence does not require a commutating device. Based on the shape of the back emf generated in the machine they can be classified into

1. **Brushless DC (BLDC)** machines with trapezoidal back emf: It is desirable to operate this kind of machine in BLDC mode, in order to maximize the torque density while minimizing associated pulsations.

2. **Brushless AC (BLAC)** machines with sinusoidal back emf. These are operated in BLAC mode to maximize the torque density and minimize torque pulsations.

In a BLDC drive, the phase current waveforms are essentially rectangular, while in a BLAC drive the phase current waveforms are essentially sinusoidal as shown in figure 2.1. However, the back-EMF waveforms may depart significantly from the ideal, in practice. Also, irrespective of their back-EMF waveform PM brushless machines may be operated in either BLDC or BLAC mode. This although may lead to compromise in performance, with respect to efficiency, torque ripple and other design goals.

2.1 classification based on magnet position

Based on this topology the basic topologies of PM brushless machine, classified according to the location of the permanent magnets as described in [Zhu and Howe (2007)].

1. **Radial-Field Machines**- Permanent Magnets on Rotor : They may have either an internal rotor or an external rotor, while the PMs may be located either on the surface or interior of the rotor.

   (a) **Surface-Mounted Permanent-Magnet (SPM) Machines**
   This is the most widely used topology for PM brushless machines shown in fig 2.2(a). As the d-axis and q-axis stator winding inductances of such machines are the same, they exhibit no saliency and zero reluctance torque. In general, the armature reaction field is relatively small and the stator windings have low inductance. The magnet with a relative permeability that approximates relative permeability of air, therefore, the effective air
gap can be considered as a sum of the actual air gap length and the radial thickness of the magnets. SPMSMs with classical design with distributed winding, generally have a relatively limited flux-weakening capability. Also, during flux weakening operation, the magnets are exposed directly to the armature reaction field, they therefore are susceptible to partial irreversible demagnetization, which can be a big drawback. However, the flux-weakening capability, can be improved for SPMSMs with a fractional number of slots per pole and a concentrated stator winding, which will be discussed later in 3.

Fig 2.2(a-2) shows a schematic of SPM, here the magnets are inset into the rotor surface. The magnet pole-arc is, less than a full pole-pitch. However, in this topology the q-axis inductance is now greater than the d-axis inductance. For this type of SPM motors as reported in [Zhu and Howe (2007)] a reluctance torque can hence be developed due to the rise in saliency ratio.

(b) Interior Permanent-Magnet (IPM) Machines

In the fig 2.2(b) brushless machines in which the magnets are accommodated within the rotor is shown. The magnets are radially magnetized in fig. 2.2(b-1), while in fig. 2.2(b-2) they are circumferentially magnetized. During flux-weakening (FW) operation, as the magnets are buried inside the rotor iron, they are effectively shielded from the demagnetizing armature reaction field. The magnet Leakage flux in IPM, is significantly greater than SPMSMs. The d-axis inductance of IPM is high compared to an equivalent SPM motor topology. But, IPM d-axis inductance is smaller than q-axis, hence the resulting saliency makes reluctance torque possible. Such machine topologies are eminently appropriate for extended speed, constant power operation in the flux-weakening mode although to a limited value in classical designs with distributed winding as shown in the figure. Indeed, a variant of the topology illustrated in 2.2(b-1) is employed in the Toyota hybrid.
By virtue of the rotor topology shown in Fig. 2.2(b-2), for these IPM machine design, when the pole number is relatively high, flux focusing can be exploited and the air-gap flux density can be made significantly higher than the magnet remanence. Hence, low-cost, low-energy product magnets, such as sintered ferrite, may be employed. Flux-focusing enables an air-gap flux density of 0.6 T to be achieved when sintered ferrite magnets, having a remanence of 0.38 T are employed. Such a machine topology also exhibits a higher d-axis inductance since the armature reaction flux only passes through a single magnet, rather than two magnets as in the other machine topologies, making it appropriately suitable for extended constant power operation.

2. Radial-Field Machines- Permanent Magnets on Stator When the permanent magnets are located on the stator, the rotor must have a salient pole geometry, similar to that of an SR machine, which is simple and robust, and suitable for high-speed operation, owing to its better mechanical integrity. The stator carries a non-overlapping winding, with each tooth having a concentrated coil. The permanent magnets can be placed on the inner surface of the stator teeth, sandwiched in the stator teeth, or mounted in the stator back-iron. The torque results predominantly from the permanent-magnet excitation torque, irrespective of magnet
location, as the reluctance torque is negligible, although the torque production mechanism
does rely on the rotor saliency. Compared with conventional permanent-magnet brushless
machine topologies, generally, it is easier to limit the temperature rise of the magnets as heat
is dissipated more effectively from the stator, because of better thermal dissipation available
at fixed stator structure.

(a) **Doubly Salient PM Machine** - Permanent Magnets in Stator back-Iron The machine
topology which is shown in Fig. 2.3(a) is referred to as a doubly salient permanent-
magnet machine. For a three-phase machine a magnet is required in the stator back-iron
for every three teeth, while for a four-phase machine a magnet is required for every four
teeth. The variation of the flux-linkage with each coil as the rotor rotates is unipolar,
while the back-EMF waveform tends to be trapezoidal. Thus, this topology is more
suitable for BLDC operation. However, the rotor may be skewed in order to obtain a
more sinusoidal back-EMF waveform to make it more appropriate for BLAC operation.
The air-gap reluctance as seen by the permanent magnets is essentially invariant, with
rotor position, therefore the cogging torque is not significant. However, due to the
unipolar flux-linkage, the torque density is relatively poor compared to that of other PM
brushless machines, a significant disadvantage w.r.t other machine design.

(b) **Flux-Reversal Permanent Magnet Machine** - Permanent Magnets on Surface of
Stator Teeth, topology shown in 2.3(b). A pair of magnets having different polarity are
mounted at each stator tooth surface. When a coil is excited, the field under one magnet
is reduced while that under the other is increased, and the salient rotor pole rotates
towards the stronger magnetic field. The flux-linkage with each coil reverses polarity
as the rotor rotates. Thus, the phase flux-linkage variation is bipolar, while the phase
back-EMF waveform is, again, essentially trapezoidal. Such a machine topology exhibits
a low winding inductance, while the magnets are more vulnerable to partial irreversible
demagnetization. In addition, significant induced eddy-current losses in the magnets
may be induced, the magnets may also experience a significant radial magnetic force.
Further, since the air-gap flux density is limited by the magnet remanence, the torque
density may very well have to be compromised for their design.

(c) **Flux-Switching PM Machine** - Permanent Magnets in Stator Teeth, topology shown
in 2.3(c). The stator consists of U-j-shaped laminated segments between which circumfer-
entially magnetized permanent magnets are sandwiched, the direction of magnetization
being reversed from one magnet to the next. Each stator tooth comprises of two adjacent
laminated segments and a permanent magnet. Thus, flux-focusing may be readily incor-
porated, so that low-cost ferrite magnets can be employed. In addition, in contrast to
conventional PM brushless machines, the influence of the armature reaction field on the
working point of the magnets is minimal. As a consequence, the electric loading of flux-
switching PMSMs can be very high. Therefore, since the phase flux-linkage waveform is
bipolar, the torque capability is significantly higher than that of a doubly salient PMSM
as reported in the article [Hua et al. (2005)]. The back-EMF waveform of flux-switching PMSMs is essentially sinusoidal, which makes them more appropriate for BLAC operation. In addition, since a high per-unit winding inductance can readily be achieved, such machines are eminently suitable for constant power operation over a wide speed range.

3. Other PM Brushless Machine Topologies

(a) **Axial-Field Machines** Axial-field PMSMs have an axially directed air-gap flux and can comprise a single-sided stator and a single rotor, a double-sided stator and a single rotor, or a single stator and a double-sided rotor. In each case, a large axial force exists between the stator and the rotor. As with conventional radial-field PM brushless machines, the stator can be slotted or slotless, although it is more difficult to manufacture a slotted stator for axial-field machines. Thus, slotless designs are more common. However, while this eliminates cogging, it exposes the winding air-gap flux. Hence, a multi-stranded conductor or Litz wire may be required to minimize the eddy-current loss. Further, since the effective air gap is large, the winding inductance is generally relatively small, which may limit the constant power speed range. Flux weakening design strategies for this machine are described in chapter 5.

(b) **Transverse-Flux Machine** Generally, transverse flux machines have a relatively large number of poles, all of which interact with the total ampere-conductors of each phase. This enables very high electric loadings and, hence, high torque densities to be achieved. However, they have a significant leakage flux and a relatively high winding inductance, as well as a poor power factor. This impacts significantly on the associated VA rating of the power electronics converter, which has inhibited its application.

2.2 Operating regions

PMSM have been gaining acceptance and wide spread usage across industry. Every application area require specific operating curves. A typical new application area for PMSM is traction applications which requires an operating curve as shown in figure 2.4(a). The operating curve obtained from IM, SRM and PMSM can be designed to exhibit torque/power-speed characteristics having the form shown in figure 2.4(b), it has three regions

- the constant torque region I, the maximum torque capability is determined by the current rating of the inverter.

- the constant power region II, flux-weakening or commutation phase advance has to be employed due to the inverter voltage and current limits.

- In region III, the torque and power reduce due to the increasing influence of the back-electromotive force (back-EMF).
Fig. 2.3: Alternative radial-field PMSM topologies with magnets on stator. (a) doubly salient PMSM. (b) flux-reversal PMSM. (c) Flux- switching PMSM.

In constant torque region PMSM are extremely efficient and easily satisfy the IE4 energy class and in many cases exceeds it. However as the primary field flux originated from permanent magnets, the wide speed operation in flux weakening region is either not attainable or attainable to a limited extend, in convention PMSM like SPM and IPMSM with distributed and simple classical rotor structure as shown in figure 2.2. The focus of this thesis work is development of design techniques for flux weakening in constant power operation. The basics of flux weakening operation is explained below.

2.2.1 Flux Weakening and Constant Power Operation

Constant torque region ends when back- emf reaches inverter operating voltage. The speed of rotor at this instant is called base speed. Flux weakening essentially as the name implies the reduction in flux linkage associated with the stator winding. This may be done by reducing air gap flux itself, or by changing reluctance path, etc. Various constraints needs to be observed during flux weakening, these constraints are describes here:
2.2.1.1 Voltage constraint

For machine operation beyond rated or base speed, rotor field flux must be reduced to keep the motor terminal voltage constant at its rated value. Flux weakening operation in a PMSM is not as simple as in dc motors, with electromagnetic field excitation, and separate winding for controlling field flux. In typical IPMSM (Interior permanent magnet synchronous motors), armature reaction is used to weaken the rotor magnet flux by means of special control algorithm. The vector control is a typical control strategy employed in PMSM. The three phase rotating field phasor are transformed into two rotating orthogonal phase phasors (dq transformations), with Clarke-Park transforms. The stator terminal voltage put a limit on maximum speed.

A typical d-q axis phasor diagram of IPMSM is shown in figure 2.5. Resistive drop has been neglected which is comparatively low for high speed operations, while drawing the d-q phasor diagram. If all quantities are expressed in per unit and saturation is neglected, the d-q axis voltage equation for any speed can be given as,

\[ V_d = -\omega_{pu}X_qI_q \]
\[ V_q = \omega_{pu}(E_0 + X_dI_d) \]  

(2.1)

Also,

\[ I_d = I_s \sin \gamma \]
\[ I_q = I_s \cos \gamma \]  

(2.2)

where, \( \omega_{pu} \) - per unit speed, \( E_0 \) - open circuit emf, \( I_s \) - stator current, \( \gamma \) - current angle, \( X_d \) - d-axis reactance, \( X_q \) - q-axis reactance.
Therefore, per unit stator terminal voltage can be given as

\[ V_s = \sqrt{V_q^2 + V_d^2} \]

\[ = \omega_{pu} \sqrt{(E_0 - I_d X_d)^2 + (X_q I_q)^2} \]

\[ \therefore V_s = \omega_{pu} \sqrt{(E_0 - X_d I_s \sin \gamma)^2 + (X_q I_s \cos \gamma)^2} \]  \hspace{1cm} (2.3)

From the equation (2.3) it is observed that the terminal voltage is directly proportional to speed for a given stator current and angle. The terminal voltage can not exceed its rated value. Thus, it provides a constraint over maximum speed achievable. The flux weakening operation starts beyond this speed.

2.2.2 Constant power

The flux weakening range is also known as constant power speed range (CPSR) because of constant power nature of this mode. The figure 2.6 shows ideal power-speed and torque-speed characteristic of an electrical machine.

We know that the expression for the average torque for a typical IPMSM is:

\[ T_e = -\frac{m.p}{2} [I_q \psi_m + (L_d - L_q) I_q I_d] \]

\[ = -\frac{m.p}{2} p[I_s \cos \gamma + \frac{1}{2}(L_d - L_q) I_s^2 \sin(2\gamma)] \]  \hspace{1cm} (2.4)
from equation 2.4 the first part is the permanent magnetic torque which is proportional to \( \cos \gamma \), so the maximal value of the permanent magnetic torque is at \( \gamma = 0 \) electric degree, while the maximal value of the second part (the reluctance torque), that is proportional to \( \sin(2\gamma) \) is at \( \gamma = 45^\circ \) electrical degree. So the maximal value of the total torque is at \( \gamma_m \), which is between \( 0^\circ < \gamma_m < 45^\circ \). Figure 2.7a shows the torque versus current control angle profile from the [Wang et al. (2009)].

Neglecting losses in the machine, output power from [Rahman and Dutta (2003)] can be expressed as

\[
P_0 = V_q I_q + V_d I_d = \omega_pu [E_0 I_s \cos \gamma + \frac{(X_d - X_q) I_s^2}{2} \sin(2\gamma)]
\]

(2.5)

As seen from the expression of terminal voltage in equation 2.3, the relationship between speed and terminal voltage is of a linear nature. Hence, for low speed operation, the voltage constraint does not prohibit peak power output or maximum torque per ampere operation; but as speed increases above the base speed, the voltage constraint start to limit the speed so that peak power at rated current occurs at a current angle \( \gamma \) that lies between maximum torque per ampere current angle (say \( \gamma^* \)) and 90°. Below this base speed the motor operates with a voltage less than rated value at optimal current angle referred here as \( \gamma^* \). Above base speed owing to voltage constraint peak power can be obtained for rated terminal voltage only if current angle, i.e. \( \gamma \) is greater than \( \gamma^* \) and approaches 90°. When \( \gamma \) becomes 90°, output power is zero. Thus for each motor design there is a limit over the speed at which output power will ultimately become zero [Schiferl and Lipo (1990)].

Hence as per the aforementioned discussion, the power-speed characteristic of a practical ma-
Machine at flux weakening region can be expected to be as in the figure 2.7b. FW in PMSM discussed here and whose power curve is described in 2.7b have distributed winding in stator with typical traditional PMSM topology. In the following chapters flux weakening capability improvement using design modification in classical PMSM like SPM, IPM is described, which include stator winding modification, rotor motor design modification, also some design topology for axial and FSPM machine has been described.

2.3 Factors determining FW capability

Following factors determine the FW capability associated with a PM machine. These are enumerated below

1. Saliency ratio $\xi$: This is defined as ratio of q-axis and d-axis inductance value. The higher the saliency ratio, More is the reluctance torque associated with the machine. Many authors and books define it as the ratio of d-axis and q-axis inductance, Here following convention is used.

$$\xi = \frac{L_q}{L_d} \quad (2.6)$$

2. Characteristic current $I_{ch}$. This is defined as ratio of flux linked to the d-axis inductance.

$$I_{ch} = -\frac{\Psi_m}{L_d} \text{[Arms]} \quad (2.7)$$

where, $\Psi_m$ is the rms magnet flux linkage and $L_d$ is the d-axis inductance.

It is well known that the condition for optimal flux weakening in an SPM machine occurs when the machine characteristic current equals the rated current. i.e $I_{ch} = I_R$. Unfortunately,
the inductance values of PM machine such as SPM and to some extent IPM machines are typically low with conventional stator winding designs (for SPM, its because the permanent magnets mounted on the rotor surface has permeability as air, hence it can be taken as air gap in the machine’s magnetic circuit. Furthermore, there is limited opportunity to lower the magnet flux linkage $\psi_m$ in SPM machines without degrading the machine’s torque production capability). So for SPM $I_{ch}$ is usually significantly higher than the rated current, hence a limited CPSR(constant-power speed range) during FW operation. Concentrated winding increases the $L_d$ for SPM machines and hence a lower $I_{ch}$ is obtained however for IPM, saliency decreases as $L_q$ increments by a low value so reluctance torque decrements, even though CPSR improves.
Distribution windings have been a de-facto choice for PM machines, but recently the CNW windings have gained popularity. Distributed winding can achieve almost unit winding factor, and hence were a reasonable choices, traditionally, for winding designs. But PMSM with distributed windings have long end windings, large copper loss, and are characterized with difficulties in winding automation, where as, even though the winding factor for concentrated windings is generally smaller than distributed windings, CNW enable easy winding automation, additionally they have shorter end windings, smaller copper loss, and as a consequence small space requirements than DW-PMSM.

3.1 Comparison of distributed and concentrated winding in IPM

For IPMSM, inductances vary with rotor position and current phase angle, unlike the SPM motors. The variation in inductance can have significant effects on motor performances. The results of this analysis hold more or less true for the PMSM. Following machines where analyzed by Soon-O Kwon and et al. in the paper [Kwon et al. (2006)] for a comparative study of the distributed and CNW in field weakening region.

- Distributed winding model with 4 poles and 24 slots and
- concentrated winding model with 4 poles and 6 slots.

The major geometric parameters of two machine under consideration, i.e. DIS(PMSM with distributive winding in stator) and CON(PMSM with CNW in stator) are identical; namely, axial length, air gap length, rotor outer radius, stator outer radius, etc. The only difference is the winding structure as evident from the nomenclature. The effect of winding configuration on the motor performance is observed. Flux density distribution in both concentrated winding (CON) and distributed winding (DIS) is shown in fig 3.1.

Since concentrated winding machines have higher THD (total harmonic distortion) in back emf waveform as compared to distributed windings, a suitable measure has to be taken to remove these distortions, i.e. to reduce the harmonic content. Therefore, CNW are designed to have minimized THD in back-emf waveform by teeth tip and slot open width design, this can be done either by skewing or changing slots per pole(SPP) or both. Also, if flux linked in the air gap is sinudoidal, we still get a sinusoidal back-emf. Due to lower THD of back emf and flux densities, CNWs inact show
lower core loss in the entire speed region. Although cogging, even with skewing is more pronounced in CNWs compared to distributed windings.

Following observation where made: For concentrated winding, a higher value of current is required compared to the distributed winding model due to lower back emf and lower saliency ratio in constant torque region. The reason for higher input current requirement is the decrement in associated magnetic and reluctance torque. Magnetic torque is reduced because of the decreased winding factor and the decrement in reluctance torque owes mainly to increase of d-axis inductance, $L_d$ based on the paper [o. Kwon et al. (2006)]. Hence decrement of saliency ratio. While in the constant power region, lower current is required for concentrated winding due to lower back emf to weaken and increased d-axis inductance.

Therefore, even though more current is required for concentrated windings in the constant torque region, copper loss is still close to the distributed winding and less current is required in FW region as well as high efficiency is achieved in the FW region due to low back emf and higher d-axis inductance, and the other concludes that concentrated winding is more suitable than distributed winding for field weakening operation at high speed [o. Kwon et al. (2006)]. In nutshell, the reduction of end winding length, simplified manufacturing process and proven ability to achieve a wide CPSR with high efficiency, non overlapping windings (i.e. concentrated windings(CNW)) has gained popularity specially in traction applications.
3.2 IPM with fractional slot concentrated windings (FSCW)

The conventional IPM machine with distributed winding has almost no or small field weakening range. The main focus of this section is the design to achieve wide CPSR and optimal efficiency in the entire speed range. In the article [Chong et al. (2010a)], the design of 1 kW motor with CNW is discussed in detail.

In [EL-Refaie and Jahns (2005)], the design technique with optimal SPP in SPM machines causes over a six times increase in $L_d$ for a 2/7-slots/pole/phase (SPP) is discussed. The authors found that the characteristic current $I_{ch}$ is reduced and optimal flux weakening condition is significantly improved. El-Refaie et al. also showed a 5:1 wide CPSR is achievable for a 2/5 (SPP) in SPM with concentrated windings in [El-Refaie et al. (2005)]. In the section 3.1 the advantages of CNW in IPM has been detailed.

IPM enjoy certain significant advantages, which tilt the scale in their favour for many application, some of them can be enumerated as 1) An increased mechanical durability due to buried magnets instead of surface magnets, 2) Ease of manufacturing and 3) Magnets have a lower chance of demagnetization. single-layer windings results in large MMF harmonic contents, which are not suitable for machines with small air gaps. Hence designing double-layer winding machines with larger air gaps is a viable option.

3.2.1 Design Aspect

The application of concentrated windings in the IPM machine can reduce the axial length to some extent, which is quite important in traction application (hybrid electric vehicle), due to the limited space availability. The influence of armature reaction on magnetic flux can also be improved during flux-weakening. The associated disadvantages with the usage of the CNW are higher cogging torque, low winding factor and reduction of as well as the total torque developed. Using appropriate optimization techniques these disadvantages can although not be negated but minimized if applied to the IPM machine.

In the figure 3.2(a) there is overlap of end windings for distributed winding in stator. The windings in fig 3.2(b) eliminates overlap of end windings resulting in insulation material reductions, hence the axial length of the machine is reduced significantly. Figure 3.2(c) shows the double-layer windings (also known as segmented stator), here each slot contains coils from two different phases. Each phase in a single slot contains only half the amount of conductors compared to single-layered windings, making it useful for applications where the motors are subjected to spatial limitations such as electric/hybrid vehicle. A detailed analysis of the design method has been approached in [Chong et al. (2007)]. Its important to note that torque in an IPM machine consists of two terms -(a) Reluctance torque and (b) Magnet alignment torque. The magnitude of induced electromotive force (emf) in phase windings contributes to the the magnetic alignment torque, whereas the inherent saliency in IPM contributes to reluctance torque. Following is the detailed analysis of selecting suitable parameters for the machine design, so that the drawbacks associated with CNW are
3.2.1.1 Selection of suitable SPP

Here, the need for selecting a suitable SPP, and its importance in the design process has been iterated. The open circuit emf induced per phase in an IPM machine is defined as:

$$E_q = 2\pi \frac{k_w}{\sqrt{2}} N_{ph} \phi_{M1} f \quad [V_{rms}]$$  \hspace{1cm} (3.1)

where,

- $E_q$ - induced per phase (lying on the q-axis)
- $k_w$ - winding factor
- $N_{ph}$ - Number of series turns per phase
- $\phi_{M1}$ - Fundamental component of airgap flux/pole
- $f$ - Input frequency

The effective number of turns per phase in stator windings is reduced if the winding factor is non unity. From eq (3.1), a unity winding factor is only attainable in the distributed windings with SPP = 1. The total winding factor, $k_w$ depends on three sub factors, namely, Pitch factor, distribution factor and skew factor. These are easily calculated in distributed windings with well-known formulas. However, they are hard to visualize in concentrated windings. An effective method to calculate overall winding factor has been enumerated in [Magnussen and Sadarangani (2003)] by using the emf phasors. This method is governed by the following equation:

$$\vec{E}'_x = e^{j\left(\frac{2\pi p}{x}\right)}$$  \hspace{1cm} (3.2)

where,

- $\vec{E}'_x$ - Reference emf phasor element (lp.u. length)
- $p$ - Number of pole pairs
Fig. 3.3: a) Single-layer concentrated winding with associated element numbers, b) Emf phasor consisting of six elements

\[ x = n - 1 \] - number of winding elements

\[ s \] - Total number of stator slots

The winding factor \( k_w \) can be calculated by subtracting \( \vec{E}'_A, \vec{E}'_B \) or \( \vec{E}'_C \) from the vector \( \vec{E}''_A, \vec{E}''_B, \) or \( \vec{E}''_C \), respectively. SPM and IPM machine rotor structure produce similar open-circuit air gap flux distribution, but as the IPM have a smaller magnet pole area than the pole face area. Hence they have an air gap flux density lower than the case when magnets are on stator surface. So, each emf element is reduced by a factor of \( \alpha \) due to reduction of flux density. The resultant open circuit phase emf is shown in figures in IPM(b). Hence we get,

\[
\vec{E}_x = \alpha \cdot e^{j\left(\frac{\pi}{2}x\right)}
\]  

(3.3)

As is the case with SPM machines having concentrated windings discussed in detail in the next section, careful selection of an appropriate SPP value results in a winding factor close to unity as explained in [Cros and Viarouge (2002)], be obtained in IPM too. It is possible because winding factor is affected by the argument of individual phase elements rather than \( \alpha \), the scaling factor. The stator winding structure in figure 3.3 is used as an example to show how to calculate the appropriate value of SPP, the stator has a structure that consists of 6 slots, 4 poles and 3 phases as represented in the figure 3.3.

\[ S_{pp} = \frac{6}{3 \times 4} = \frac{1}{2} \text{[slots/pole/phase]} \]  

(3.4)

This configuration would produce a winding factor of 0.866. The winding factors \( \geq 1/2 \) for various SPP values and these are tabulated in table 3.1. The least common multiple (LCM) for the pole/slot combination is also shown in the table, whose importance is discussed in the next sub section.
3.2.1.2 Cogging torque

The undesired torque ripple known as cogging torque is due to the interaction between rotor magnets and stator teeth, which contributes to the vibrations and noise, causing severe restrictions in machine operation, especially during low speed operations. Several methods are available by which cogging torque can be significantly reduced. However, most of these methods also reduce the back emf and the total developed torque. Owing to higher cogging torque value, reduction process is even more crucial for the design of IPM machines despite a probable drop in emf. Unlike distributed winding, CNW however, should not be skewed, as it would cause a drastic drop in open circuit voltage, and hence not a very successful strategy in IPMSM.

A very useful method for reducing cogging torque is presented in the proceedings [Magnussen and Sadarangani (2003)]. Cogging period per rotor revolution is calculated from LCM of slot and pole numbers. Also, higher cogging frequency means lower cogging amplitude. Hence a design with the highest possible LCM value, will help reduce cogging. The LCM for different slot and pole combinations is indicated in Table 3.1. Hence while considering the $k_w$ for as close as possible sinusoidal waveform for back emf, selection preference should be given to the combination with highest LCM value of slot and poles to lower cogging torque. Consider two winding types, a 12/16 slots/poles and a 12/14 slots/poles, both combinations gives a $k_w$ of 0.966, but the latter (12/14) is preferred as it is able to achieve lower cogging torque, due to higher LCM (48 and 84, respectively).

<table>
<thead>
<tr>
<th>Slots</th>
<th>Poles</th>
<th>$S_m$</th>
<th>$K_w$</th>
<th>LCM</th>
</tr>
</thead>
<tbody>
<tr>
<td>3</td>
<td>1/2</td>
<td>0.966</td>
<td>0.966</td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>1/2</td>
<td>0.966</td>
<td>0.966</td>
<td></td>
</tr>
<tr>
<td>9</td>
<td>1/2</td>
<td>0.966</td>
<td>0.966</td>
<td></td>
</tr>
<tr>
<td>12</td>
<td>1/2</td>
<td>0.966</td>
<td>0.966</td>
<td></td>
</tr>
<tr>
<td>15</td>
<td>1/2</td>
<td>0.966</td>
<td>0.966</td>
<td></td>
</tr>
<tr>
<td>18</td>
<td>1/2</td>
<td>0.966</td>
<td>0.966</td>
<td></td>
</tr>
<tr>
<td>21</td>
<td>1/2</td>
<td>0.966</td>
<td>0.966</td>
<td></td>
</tr>
<tr>
<td>24</td>
<td>1/2</td>
<td>0.966</td>
<td>0.966</td>
<td></td>
</tr>
</tbody>
</table>

**Tab. 3.1: Winding Factor and LCM for Different SPP values**
3.2.1.3 Reducing characteristic current

The Variation of the $I_{ch}$ value will affect the pitch factor of coils, which results in a change of mmf distribution. A scaling factor $\alpha_s$ is introduced, which is the ratio of two pitches i.e. slot pitch $\tau_{sp}$ and one pole pitch $\tau_{pp}$. To simplify and standardize calculations, the start of the pole pitch is taken to be at the inter-coil region with a displacement of $\gamma$. Consider the figure 3.4, here a four pole six slot single-layer concentrated winding with just one phase coil is represented. The modified mmf equation, considering pitch and displacement factor is given as [Chong et al. (2007)].

\[
F_n = \frac{4N_p I}{n:\pi} \cdot \sin \left( \alpha_s \cdot \frac{n:\pi}{2} \right) \cdot \cos(n(\theta_e - \gamma)) \tag{3.5}
\]

where, $\alpha_s$ - Scaling from pitch factor $\gamma$ - Angular displacement.

The shape of emf wave depends on the SPP value and the way the slots are distributed. This shape of emf waveform details input current waveform i.e. sinusoidal or rectangular. Emf waveform will have flat top implying rectangular input currents, while the commonly used slot-pole combination with 2/7 slot/pole has a near sinusoidal emf waveform; therefore, sinusoidal currents can be used for them. Sinusoidal input currents causes the development of forward and backward rotating mmf wave [Magnussen and Sadarangani (2003)]. Each emf components produces harmonics at different frequencies. The resultant mmf produced per phase is given by:

\[
F_n = \frac{4N_p I}{n:\pi} \cdot \sin \left( \alpha_s \cdot \frac{n:\pi}{2} \right) \cdot \cos(n(\theta_e - \gamma)) \cdot \hat{I} \cos(\omega t) \tag{3.6}
\]

31
Equation 3.6 can be broken into constituent forward and backward mmf as:

\[
\mathcal{F}_{n(f)} = \frac{4N_p}{n\pi} \cdot \sin\left(\alpha_s \cdot \frac{n\pi}{2}\right) \cdot \hat{I} \cos(\omega t - n(\theta_e - \gamma)) \cdot \left[\frac{1}{2} + \cos\left(\frac{2\pi(n - 1)}{3} + \delta\right)\right] \tag{3.7}
\]

\[
\mathcal{F}_{n(b)} = \frac{4N_p}{n\pi} \cdot \sin\left(\alpha_s \cdot \frac{n\pi}{2}\right) \cdot \hat{I} \cos(\omega t + n(\theta_e - \gamma)) \cdot \left[\frac{1}{2} + \cos\left(\frac{2\pi(n + 1)}{3} + \delta\right)\right] \tag{3.8}
\]

here, \(\delta\) - Phase displacement from reference phase

For a 3-phase system with six slots, the phase displacement is \(\pm 120\). Both forward and backward wave fluctuates once per mechanical cycle, which is \(n/p\) times the electrical angular displacement.

The resultant 3-phase mmf wave \([\text{Chong et al. (2007)}]\) is given as:

\[
\mathcal{F} = \sum_{n=\frac{-p}{2}}^{\infty} \left[|\mathcal{F}_{n(f)} + \mathcal{F}_{n(b)}|_{\delta=0} + (\mathcal{F}_{n(f)} + \mathcal{F}_{n(b)})|_{\delta=\frac{2\pi}{3}} + (\mathcal{F}_{n(f)} + \mathcal{F}_{n(b)})|_{\delta=\frac{2\pi}{3}}\right] \tag{3.9}
\]

The mmf waveform in (3.9) contains extra harmonic components resulting in additional leakage flux, as compared to the distributed winding. Leakage flux has three components - armature, rotor and winding distribution differential leakage flux \([\text{Chong et al. (2007)}]\). The additional flux created by the extra harmonic components contributes to the armature leakage flux. Ignoring the third (i.e. winding distribution differential) leakage flux component, a simple expression for the d-axis inductance can be obtained as follows:

\[
L_d = L_{le} + \frac{l_c}{T_d} \int_{-\pi}^{\pi} mmf_{d} \frac{B_r U}{2} + \frac{B_r L}{2} d\theta \tag{3.10}
\]

where, \(L_{le}\) - Armature leakage inductance

\(l_c\) - Axial length of stator coil

\(B_r U\) - Upper layer armature flux distribution

\(B_r L\) - Formula Lower layer armature flux distribution.

This condition leads to an increase in armature leakage inductance (because of the increase in leakage flux), leading to a higher \(d\)-axis inductance. From equation 2.7, there is a decrement in the \(L_{ch}\) required, hence improved CPSR, but this also leads to decrement in the saliency ratio, which is not so ideal as large contributions of reluctance torque to the total machine torque production is not available. Hence, the machine must depend on the magnets to deliver the dominant share of the machine’s torque production. So a compromise is garnered. The decrement in saliency ratio is because the change in SPP value leads to an increment in only \(L_d\) value while the \(L_q\) increases marginally. In the paper \([\text{Chong and Rahman (2008)}]\) L. Chong et al. has discussed how to calculate \(L_d\) and \(L_q\) of various winding configurations and to determine the optimum saliency ratio for a 12-slot 14-pole model with FSCW(fractional slot concentrated winding)-IPMSM using finite element method. Also, FSCW-IPMSM with double-layer windings tend to achieve higher magnetic saliency.
and therefore, a better reluctance torque than comparable machines structure with single-layer windings [Tangudu et al. (2010)]. Also double layer winding help in reducing the axial length of the FSCW-IPMSM (fractional slot concentrated winding). In the paper [Chong et al. (2010a)], a highly efficient 1kW machine for field weakening applications has been designed with steps for achieving the widest CPSR and minimal losses.

3.3 SPM with fractional slot concentrated windings (FSCW)

Surface permanent-magnet synchronous motors (SPMSM), have received wide attention in the field of medium and high performance electrical drives. However, for SPMSM, CPSR is limited as it has relatively large air-gap (PM acts as air gap). Recently, some noticeable attempts have been made to improve CPSR with SPMSM for wide speed applications, using concentrated windings. This is particularly promising for the direct drive automotive starter/generator system where a capability of precise torque generation in the flux-weakening region is essential. For example, automotive drive systems require torque regulation specifications less than 5% of the rated value in the overall operating range [Kim and Sul (2007).

SPMSM can be designed to achieve optimal flux-weakening conditions by introducing CFSW. In the sections 3.3.1, this objective is achieved by significantly increasing the phase inductance, thereby reducing $I_{ch}$ sufficiently so that it match its rated current. An analytical model is also proposed to make precise conjecture for the machine parameters for SPM with CNW. A brief analysis of the scalability of CNW with machine aspect ratio, number of poles, and machine output power is also proposed.

3.3.1 Design aspect

Surface permanent-magnet (SPM) synchronous machines have generally been considered to be poor candidates for achieving wide ranges of constant-power operation in FW region. The reasoning behind this can be understood by considering the characteristic current of an SPM machine, defined as

$$I_{ch} = \frac{\Psi_m}{L_d} \text{ A (rms)} \quad (3.11)$$

For optimal flux weakening, $I_{ch} = I_R$. SPMSM with conventional stator winding designs have typically low inductance value owing to large air gap, also mounted PM act as air in the magnetic circuit. Also, it is difficult to lower the linked $\Psi_m$ in SPMSM without degrading machine’s torque production capability. These factors contribute to the $I_{ch} > I_R$, forcing severe limitation on the machines CPSR In FW operation. A design approach for achieving optimal FW in SPMSM is by designing the machine’s stator windings using FSCW, it has been proposed in [EL-Refaie and Jahns (2005)].
Fig. 3.5: Comparison of basic stator winding configurations. (a) Distributed with SPP=1. b) Concentrated winding with SPP=0.5

CNW result in elimination of any end-winding overlap with other phase windings. Fig 3.5 shows a comparison between conventional distributed and concentrated windings. The term "fractional slot" refers to stator windings with slot/phase/pole (SPP) values less than one.

This FSCW stator design makes it possible to significantly increase the machine inductance in order to reduce the $I_{ch}$ to the point of establishing equality with the rated current ($I_R$). It was shown in the article [EL-Refaie and Jahns (2005)] that the conditions for optimal flux weakening can be achieved while simultaneously delivering near-sinusoidal back-electromotive-force (EMF) waveforms and low cogging torque by considering proper SPP. Care must be taken in selecting the SPP value for a particular machine as if it is non-optimal, it may lead to unbalanced radial force acting on the rotor as shown in [Magnussen et al. (2004)], here it was shown that SPM machine equipped with FSCW, can achieve a wide CPSR value.

A systematic machine design procedure based on closed-form analytical techniques can is followed to achieve desired performance characteristics in a 6-kW SPMSM in [EL-Refaie and Jahns (2005)]. A flux weakening with a CPSR of (10 : 1) is shown to be achieved in this publication.

As already discussed Concentrated windings offer some significant advantages over distributed windings. These include: 1) significant reduction in the copper volume and copper losses in the end region; 2) significant reduction in the machine total length; 3) reduction in machine manufacturing cost; and 4) compatibility with segmented stator structures that makes it possible to achieve significantly higher slot fill factor values. Most importantly for field weakening performance improvement they provide higher inductance compared to distributed windings.

An important design aspect is the choice of SPP value to achieve the highest machine perfor-
mance. The criteria for choosing the preferred SPP values have been identified by various authors and is summarized [EL-Refaie and Jahns (2005)] here.

- The winding factor for the spatial frequency that matches the rotor magnet fundamental spatial frequency (henceforth referred to as simply the synchronous frequency) should be as high as possible. This leads to high effective numbers of turns and, hence, lower current for the same torque.

- The lowest common multiple (LCM) of the number of stator slots (S) and the number of rotor poles (2P) should also be as high as possible. The harmonic frequency that corresponds to this LCM order value represents the cogging torque frequency. As a result, a high LCM value is selected so that the cogging torque frequency is high but with lower magnitude.

- The greatest common divisor (GCD) of the product of the number of stator slots and the number of rotor poles must be an even number. This GCD value is an indication of the machine’s symmetry. If it is an even number, the net radial force on the machine will be very low.

Two families of stator windings corresponding to SPP values of 2/5 and 2/7 emerge as the best candidates, for our purpose 2/7 has been used, the same consideration can be applied for 2/5 [Shah and EL-Refaie (2009)]. The authors also proved that a 5:1 wide CPSR is achievable for a 2/5 SPP, in SPM with FSCW. In [EL-Refaie and Jahns (2005)], the authors have used 12-slot/14-pole, single-layer design (SPP=2/7) is compared to a distributed 42-slot/14-pole design(SPP=1), both having the same rotor structure, to quantitatively explain why CNWs, have potential for improving the flux-weakening capabilities.

The phase A winding configuration of both designs is shown in Fig 3.6. It can be seen that there is a significant difference between the two designs. Over the 14 rotor poles, phase A of the 12-slot/14-pole design (winding 1) has two coils occupying four slots in contrast to seven coils occupying 14 slots in the case of the 42-slot/14-pole design (Winding 2). The winding function of both the winding configuration is shown in figure 3.8 and fig3.7.

The objective of our design is to decrease the characteristic current 3.11, making it possible to achieve the conditions for optimum FW operation, this is done here in design using fractional winding of Spp = 2/7. The following discussion will prove that it is indeed the case.

3.4 Comparison of FSCW and Distributed Windings in SPM

In order to do a comparative analysis of the winding a constraint of equal flux linkage is imposed. For this N1 turns/coil in 12-slot/14-pole CNW and N2 turns/coil in distributed winding are considered. The corresponding winding function is represented above in figure 3.7 and 3.8. Flux linkage in the phase winding A can be represented by the general equation 3.12. The assumption of sinusoidal flux density and same rotor design has been made. Also winding function only represents magnetizing
Fig. 3.6: Comparison of fractional-slot concentrated and distributed full-pitch windings.

Fig. 3.7: Winding function of the 12-slot/14-pole (SPP=2/7) concentrated fractional-slot winding design.

Fig. 3.8: Winding function of the 42-slot/14-pole (SPP=1) distributed winding design.
and harmonic inductance and the effect of slot and end windings are not included. Never the less, results will more or less comply with the study below even upon there consideration.

\[ \Psi_a = r_g l_{\text{eff}} \int_0^{2\pi} N_a(\theta) B(\theta) d\theta \text{ Wb} - \text{turns} \] (3.12)

where,
- \( \Psi_a \) - phase A magnet flux linkage (Wb);
- \( r_g \) - air-gap radius (m);
- \( l_{\text{eff}} \) - active length of the machine (m);
- \( N_a \) - phase A winding function (turns);
- \( B \) - air-gap magnet flux density (T);
- \( \theta \) - angle along the air-gap periphery (mechanical radians).

Because of the aforementioned assumption flux density in both machine types is given as

\[ B(\theta) = B_{\text{max}} \sin(7\theta) \text{ T} \] (3.13)

The flux linkage in FSCW and distributed winding (DW) machines can be represented by

Winding 1: For CNW 12/14 slot/pole

\[ \Psi_{a1} = r_g l_{\text{eff}} * \frac{4}{p} K_{w1} N_1 B_{\text{max}} \text{ Wb} - \text{turns} \] (3.14)

Winding 2: For CNW 42/14 slot/pole

\[ \Psi_{a2} = r_g l_{\text{eff}} * 2 N_2 B_{\text{max}} \text{ Wb} - \text{turns} \] (3.15)

Since from the design assumption flux linkage is identical for two winding type machine, from equation 3.14 and 3.15 we have the relationship between turns per coil for the CNW and DW windings i.e \( N_1 \) and \( N_2 \)

\[ N_1 = \frac{p}{2K_{w1}} N_2 = 3.623 * N_2. \] (3.16)

Self and mutual inductance of stator winding can be represented as

\[ L_{aa} = \frac{\mu_0 r_g l_{\text{eff}}}{g} \int_0^{2\pi} N_a^2(\theta) d\theta \text{ H} \] (3.17)

\[ L_{ab} = \frac{\mu_0 r_g l_{\text{eff}}}{g} \int_0^{2\pi} N_a(\theta) N_b(\theta) d\theta \text{ H} \] (3.18)

where,
- \( L_{aa} \) - self-inductance of phase "A" (H)
- \( L_{ab} \) - mutual inductance between phases "A1" and "B1" (H)
- \( \mu_0 \) - permeability of air (H/m)
- \( g \) - air-gap thickness (m)
- \( L_d \) - d-axis inductance = \( L_{aa} + L_{ab} \) (H).
Upon evaluation the inductance for winding 1 can be written as

\[ L_{aa1} = \frac{\mu_0 r g_{\text{eff}}}{g} N_2^2 \frac{\pi}{3} \text{ H} \]

\[ L_{ab1} = L_{ac1} = 0 \text{ H} \]

\[ \therefore L_{d-\text{conc}} = \frac{\mu_0 r g_{\text{eff}}}{g} N_1^2 \frac{\pi}{3} \text{ H} \]  

(3.19)

Similarly, upon evaluation inductance values for winding 2 can be found out.

\[ L_{aa2} = \frac{\mu_0 r g_{\text{eff}}}{g} N_2^2 \frac{\pi}{2} \text{ H.} \]

\[ L_{ab2} = \frac{L_{aa2}}{3} \text{ H} \]

\[ \therefore L_{d-\text{dist}} = \frac{\mu_0 r g_{\text{eff}}}{g} N_2^2 \frac{2\pi}{3} \text{ H.} \]  

(3.21)

Dividing equation 3.20 and 3.22 and using the equation 3.16 we have

\[ \frac{L_{d-\text{conc}}}{L_{d-\text{dist}}} = \frac{N_2^2}{2N_2^2} = 6.56. \]  

(3.23)

The above equation 3.23 leads to a characteristic current refer equation 3.11 for a CNW machine (spp= 2/7) 6.56 times less than a distributed winding with spp value of 1. This helps in closing the gap between characteristic current and rated current, bringing in the possibility of optimal condition for field weakening which is \( I_{ch} = I_R \).

### 3.5 Analytical model for the FSCW-SPM

In the above section it has been proven that the characteristic current for the SPM machine can be much lower than in the DW case. The reasoning for this has been explained in the article [EL-Refaie and Jahns (2005)] as the presence of subharmonics in the CNW winding function fourier analysis. This lead to a higher inductance value because of the higher harmonic leakage inductance value. The large spatial harmonic components in the winding function may result in higher harmonic magnetic flux density (B) components. Careful consideration during the machine design process must be put on these additional flux density components in order to minimize their impact on magnetic saturation of the iron core as well as to minimize eddy-current losses. There are Various technique available for reducing these undesirable effects, but this is beyond the scope of this thesis work. The general steps for designing the machine are similar to designing any other ac machine. However for satisfying optimal flux weakening following consideration has to be taken. Special attention at the calculation of the magnet remanence \( B_r \) and the number of series turns \( (N_s) \) in the stator winding.
must be taken, as they are critical for determining the flux-weakening capability of the machine, the means of determining their values must be addressed. The determination of \((B_r)\) and \((N_s)\) focuses on the machine parameter \(\psi_m\) and \(L_d\) as they in turn determine the \(I_{ch}\). Simplified equation for the magnetic flux linkage of the machine is presented in equation (3.24)

\[
\Psi_m = \frac{\sqrt{2}B_1r_s l_{s\text{eff}} N_s K_w \lambda_0}{p} = \frac{k_1 B_r N_s}{p} \text{ Wb – turns}
\]

(3.24)

where, \(B_1\) - Peak fundamental component of the air-gap magnet flux density distribution \(\lambda_0\) - average value of the permeance function to account for the effect of the stator slots, \(p\) - number of pole pairs, and \(k_1\) - machine constant.

The equation (3.24) does not include the impact of harmonics, but it does represent a very good approximation, since \(SPP = 2/7\) leads to flux linkage and back-EMF waveforms that is almost sinusoidal, hence it is a good approximation as harmonic content is low. Therefore, machine is excited with sinusoidal current waveforms in order to achieve very good performance characteristics. Since \(L_d \propto N_s^2\), following proportionality is found upon evaluation

\[
I_{ch} = \frac{\Psi_m}{L_d} = \frac{k_2 B_r N_s}{p N_s^2} = \frac{k_2 B_r}{p N_s} \text{ A rms}
\]

(3.25)

where, \(k_2\) - machine design constant ( dependent upon machine dimensions and configuration.)

From the design constraint of characteristic current value and desired corner-point frequency \(\omega_c\), and using above equation of \(I_{ch} = I_R\), \(B_r\) and \(N_s\) can be uniquely determined if the machine parameters are assumed to be already determine, then

\[
\frac{k_2 B_r}{p N_s} = I_R \text{ A rms}
\]

(3.26)

\[
V_s = \sqrt{(\omega_c \Psi_m)^2 + (\omega_c L_d I_R)^2} \text{ V rms}
\]

(3.27)

where, \(V_s\) - maximum fundamental rms phase voltage available to the machine (Vrms).

Equation (3.27) neglects the stator resistance voltage drop that is anyhow, typically negligible for elevated rotor speeds. Following the design methodology CPSR value as high as 10:1 is achievable as per the author.

### 3.6 Scalability of FSCW-SPMSM

SPMSM are now being considered as serious candidates for applications requiring wide speed ranges of constant-power operation by properly designing the machine with FSCW. In the paper [El-Refaie and Jahns (2005)] a valuable contribution is provided by demonstrating that the key design objective of achieving optimal FW can be achieved over a wide range of application conditions. Following observation and results where obtained considering various machine.
Selection of SPP with lower slot and pole numbers leads to increased machine volume, mass, and magnet remanent flux density $B_r$ values. \[ \therefore \text{If the SPP value is fixed, even though optimal flux weakening can be achieved with a variety of stator slot / rotor pole combinations, but meeting the optimal flux weakening is easier as the slot and pole numbers are increased.} \]

The results of scaling investigation done in the paper has shown that the machine active volume as well as slot current density tend to increase as the diameter to length $(D/L)$ aspect ratio is reduced for the same power level. For wide CPSR in SPM, large-diameter machines that are compatible with high stator slot and pole numbers are most convinient. If the number of stator slots and rotor poles are reduced along with the machine diameter, it will only increase further increase the upward pressure on machine volume and mass.

Design of SPM rated from 1 to 60 kW can show optimal flux weakening over a considerable CPSR. Although, evidently, the machine active volume per kW and/or current density tend to increase as the machine power rating and diameter are reduced.

### 3.7 Comparison of FSCW- SPM and IPM machines

The objective of this section is to develop a comparative analysis of fractional slot concentrated winding (FSCW) in SPM and IPM. The majority of comparison results are taken from the paper [Reddy et al. (2012)]. Here the author uses the Freedom-Car 2020 requirements to test two motors through a set of specifications as summarized in Table 3.2 and Fig. 3.9. Some of the key set of specifications are maximum speed of 14 000 r/min, continuous power of 30 kW over 20–100% speed range, peak power of 55 kW for 18 s at 20% speed, minimum efficiency of 95% over 10 – 100% speed range, 105°C cooling inlet temperature, fixed nominal 325-V dc source.
Tab. 3.2: FREEDOMCAR 2020 ADVANCED MOTOR PERFORMANCE REQUIREMENTS

<table>
<thead>
<tr>
<th>Requirement</th>
<th>Target</th>
<th>Condition</th>
</tr>
</thead>
<tbody>
<tr>
<td>Minimum top speed</td>
<td>14,000 rpm</td>
<td>at 20% speed and nominal voltage</td>
</tr>
<tr>
<td>Peak output power</td>
<td>55 kW for 18 sec</td>
<td>at 100% speed and nominal voltage</td>
</tr>
<tr>
<td>Continuous output power</td>
<td>30 kW</td>
<td></td>
</tr>
<tr>
<td>Weight</td>
<td>≤ 35 kg</td>
<td>in quantities of 100,000</td>
</tr>
<tr>
<td>Volume</td>
<td>≤ 9.7 L</td>
<td></td>
</tr>
<tr>
<td>Unit cost</td>
<td>≤ $275</td>
<td></td>
</tr>
<tr>
<td>Operating DC bus voltage</td>
<td>200–450V, 325V nominal</td>
<td></td>
</tr>
<tr>
<td>Maximum phase current</td>
<td>400Ams</td>
<td></td>
</tr>
<tr>
<td>Characteristic current</td>
<td>&lt; Max. current</td>
<td>at 10–100% speed</td>
</tr>
<tr>
<td>Efficiency</td>
<td>&gt; 95%</td>
<td>at 50% rated torque</td>
</tr>
<tr>
<td>Line-to-line Back-EMF</td>
<td>&lt; 600V peak</td>
<td>at 100% speed</td>
</tr>
<tr>
<td>Torque pulsation</td>
<td>&lt; 5% peak torque</td>
<td>at any speed</td>
</tr>
<tr>
<td>Ambient operating temp.</td>
<td>-40–140°C</td>
<td>outside housing</td>
</tr>
<tr>
<td>Coolant inlet temperature</td>
<td>105°C</td>
<td></td>
</tr>
<tr>
<td>Max. coolant flow rate</td>
<td>10 liters/min</td>
<td>terminals to ground</td>
</tr>
<tr>
<td>Max. coolant pressure drop</td>
<td>2 psi</td>
<td></td>
</tr>
<tr>
<td>Max. coolant inlet pressure</td>
<td>20 psi</td>
<td></td>
</tr>
<tr>
<td>Min. isolation impedance</td>
<td>1MΩ</td>
<td></td>
</tr>
</tbody>
</table>

The 12-slot IPM machine using high-strength sintered NdFeB magnets is built with 12 individually wound tooth structures, making it a double-layer winding pattern is used. The FSCW-SPM machine has 10 poles and 12 slots and the stator structure is segmented. Also Carbon fiber banding is tightly wound around the rotor outer periphery to provide the required structural containment at high rotor speeds for SPM. The author conducted experimental test and compared the results with the predicted FEM models of these two machine, and following observation where made.

For a rated load of 30 kW, the stator and rotor core losses are the dominant components and form the majority of total losses in IPM. Both the losses increase with speed. The high space harmonic contents in the FSCW contributes to higher core losses in both the rotor and stator for IPM refer fig. 3.10a on . The losses for FSCW-SPM machine are shown in fig 3.10b. In SPM, majority of the losses occur in the magnets and in the stator core, while significantly lower losses are present in the rotor core and copper losses. The magnet losses are unavoidable and are challenging to dissipate because of the lack of good convection paths aggravated by the presence of the carbon fiber retaining ring around the magnets. The back emf of SPM machine would be quite sinusoidal due the absence of higher order space frequency which are very much prevalent in IPM.

Despite these differences the efficiency of two design are quite similar, even though the loss components are clearly different in the two machines. Figure 3.11a shows the predicted efficiency pattern of the two machine under observation and figure 3.11b shows experimental as well as predicted efficiency. Please note that due to the different experiment set up the SPM was tested by the author till 5000 rpm for the article [Reddy et al. (2012)]. Also it can be observed that the
(a) Predicted losses over the speed range under rated condition of 30 kW in FSCW-IPM machine.

(b) Predicted losses over the speed range under rated condition of 30 kW in FSCW-SPM machine.

(a) Predicted efficiencies for FSCW-IPMSM and SPMSM under rated-load

(b) Predicted and experimental efficiencies for FSCW-IPMSM and SPMSM under rated-load

Experimental efficiency with the SPM machine clearly falls lower than the prediction by at least a couple of percent of power.

Both machines are comparable in terms of mass, volume, and, hence, power density. However, FSCW-IPM machine requires lower magnet mass due to the higher reluctance torque component. But, the FSCW-SPM machine exhibits better overload capability, as it can produce the 55-kW peak power at 320 A(rms) versus 400 A(rms) in the FSCW-IPMSM which can mainly be attributed to lower magnetic saturation effects in the FSCW machine due to its larger effective air gap.
4. MODIFICATION OF ROTOR DESIGN

As has already been discussed in previous chapters and sections, that the PM such as IPMSM enjoy advantages of high energy efficiency, power-to-volume ratio, etc. They have a wide constant-power speed range (CPSR) making it suitable for many applications such as traction, wind energy generation, etc. The potential of flux weakening operation in an IPM machine was first introduced in [Jahns (1987)]. The flux weakening operation in the IPM machine can be achieved through appropriate stator current that produces flux which opposes or weakens permanent magnet flux. However, achieving a wide CPSR in radially laminated IPM machines with single magnet layer per pole and standard sinusoidal distributed stator is difficult. In such IPM machines, the CPSR is extended by enhancing the saliency ratio to exploit the reluctance torque during flux weakening, but due to q-axis saturation, a constant ratio of $L_q$ to $L_d$ cannot be maintained, leading to reduction of CPSR. It has been seen that flux weakening range of typical IPM machine is very small. Moreover, the SPM and IPM machines with concentrated winding have torque pulsation at low speed and unwanted space harmonics.

Many attempts have been made to improve the range. Among these, modification of rotor design takes a considerable part. An investigation has been made in this section about rotor designs suitable for wide speed range. In the following section a design strategy has been adopted, while considering IPM, where, flux weakening is achieved, while removing the torque ripples, reducing cogging torque, by altering the air gap flux path (instead of weakening it) with the help of armature reaction. A finite element model has been constructed and variation of magnetic field with armature reaction was done in [Dutta and Rahman (2008)].

4.1 Magnet Segmentation

As seen in chapter 3 although distributed windings were the preferred winding type due to its ability of producing MMF waveforms that are very close to sinusoidal. However, due to key attributes of reduced machine axial length and simplified manufacturing, along with relative fault free machine, concentrated windings have gained popularity now especially in traction applications. A 5:1 CPSR can be obtained with 6kW surface permanent magnet (SPM) machine with FSCW as shown in the article [El-Refaie et al. (2005)]. FSCW produce increased MMF harmonic content therefore affecting the field weakening performance of machines with small air gaps like the interior permanent magnet (IPM) machine as compared to distributed windings as explained in the article [Bianchi et al. (2006)]. In the section ?? the effect of high harmonic content was negated to a large degree because of the
use of optimal SPP value. The method to determine the optimal arrangement of phase coils for different SPP value was derived [Cros and Viarouge (2002)]. Reluctance torque is lowered because of the concentrated winding. Also a higher torque ripple is inflicted on the machine.

Torque ripple is defined as

\[ T_{\text{rip}} = \frac{(T_{\text{max}} - T_{\text{min}})}{T_{\text{ave}}} \]  \hspace{1cm} (4.1)

where, \( T_{\text{max}} \) is maximum torque of one period, \( T_{\text{min}} \) is minimum torque, and \( T_{\text{ave}} \) is the average torque.

Segmenting has been identified as a key method to reduce the torque ripple. In a paper [Stumberger et al. (2001)], segmentation of rotor magnets was identified as a suitable way for wide speed range flux weakening. According to authors, with this design flux weakening for a speed range of up to 5:1 was achieved. Although flux-weakening was achieved with stator current, armature reaction was used to alter the magnet flux path instead of weakening the flux produced by the PM, here. During segmentation of rotor magnets the magnet is segmented and a small iron bridge is placed between two segments.

With segmentation, the magnet flux which passes through the air gap is reduced as when the stator current is applied, a part of the permanent magnet flux is canalized in to the iron sections between the magnet segments. In the article [Rahman and Dutta (2003)], a comprehensive study of this new design using FE model for a 11kW, 415V, 4 pole machine was presented by R. Dutta et al. Its important to note that the air gap flux density of the magnet depends greatly on the width of the iron bridges, and in following paragraph their impact on flux weakening will be further enumerated.

In the article [Rahman and Dutta (2003)], the author theorizes that, normally the gaps will be saturated with magnet leakage flux. The negative d-axis armature reaction however changes the saturation level of these gaps allowing canalization of more magnet flux in these gaps, which in turn reduces the flux passing through the air gap, in other words, this effectively weakens air gap flux, hence amount of flux passing through the air gap can be controlled, achieving desired flux reduction during flux weakening operation, this has an added advantage of having less danger of permanent demagnetization in magnets. The flux distribution in conventional winding and segmented IPM is as shown in figure 4.1 from [Wang et al. (2009)].

Segmenting the magnet is one of the better ways to reduce the ripple torque. The number of segmentation required for optimal ripple reduction without effecting the torque profile of the machine is important. The authors in [Wang et al. (2009)], used two segmented FEA models with two and four segment magnets in each pole in allusion to the conventional one, with same stator parameters (but different rotor design), to analyze the optimal segmentation. Figure 4.2 shows segmented magnet rotors a) with 2 segments, b) with 4 segments.

The author used the machine with design parameters (having distributed winding) as tabulated in [4.1].

Figure 4.3 shows the result of torque ripple comparison between three rotor type IPM machine with respect to rotor current angle. The figure 4.3 makes it clear that the two segment provides
Fig. 4.1: Flux distribution in conventional and Segmented models

Fig. 4.2: Segmented magnet models

<table>
<thead>
<tr>
<th>Design Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated output power (kW)</td>
<td>7.5</td>
</tr>
<tr>
<td>Rated voltage (V)</td>
<td>115</td>
</tr>
<tr>
<td>Rated torque (N.m)</td>
<td>22.4</td>
</tr>
<tr>
<td>Number of poles</td>
<td>8</td>
</tr>
<tr>
<td>Number of stator slots</td>
<td>48</td>
</tr>
<tr>
<td>Rated speed (r/min)</td>
<td>3195</td>
</tr>
<tr>
<td>Outer diameter of stator (mm)</td>
<td>175</td>
</tr>
<tr>
<td>Inner diameter of stator (mm)</td>
<td>120</td>
</tr>
<tr>
<td>Axial length of stator core (mm)</td>
<td>75</td>
</tr>
<tr>
<td>Inner diameter of rotor (mm)</td>
<td>38</td>
</tr>
<tr>
<td>Air gap (mm)</td>
<td>0.5</td>
</tr>
<tr>
<td>Thickness of magnet (mm)</td>
<td>5.5</td>
</tr>
<tr>
<td>Magnet material</td>
<td>NdFeB</td>
</tr>
<tr>
<td>Rated current (Arms)</td>
<td>46</td>
</tr>
</tbody>
</table>

Tab. 4.1: Machine parameters
comparable torque to conventional machine while also reducing the ripples, however the rotor with magnet segmentation of four even though reduces the ripple, also reduces the effective torque available from the machine. Hence two segmented rotors can provide suitable ripple dampening while improving the CPSR too.

The torque and power characteristic for the two machine are charted with respect to speed in the figure 4.4. The improvement shown in the figure is for IPM with distributed winding having segmented rotor magnet design.

The shape of segmented rotor magnet have interesting effect in flux distribution through the rotor iron and in the air gap, a study of the same has been done in [Chong et al. (2010b)]. Following section show the impact of optimal geometrical placement of segmented rotor magnets.

4.2 Magnet geometries

4.2.1 Comparing SPMSM,IPMSM with rectangular magnets, and IPMSM with v-shaped magnets

In the paper [Chong et al. (2010b)], authors used three different rotor types, namely, a) SPM, b) IPM with rectangular flat-shaped magnets, and c) IPM with V-shaped magnets, which shown in figure 4.5. The grade and volume of magnet are kept constant in all three and the same double-layer FSCW stator is used.

The parameters used by the author for rotor geometry comparison are tabulated in table 4.2.

The power versus speed characteristics of the three models are compared in the figure 4.6. The flat-shaped magnets can be observed having weakest performance amongst the three designs. It produces the lowest torque and power magnitudes at all speeds and has the poorest field weakening performance among the three machine considered for comparison. The V-shaped model exhibits a better field weakening performance with a CPSR of up to 7:1. Although, the SPM produces the
Fig. 4.4: Maximum torque/power-speed characteristics with speed

Fig. 4.5: a) SPM, b) IPM with flat magnets, c) IPM with V-shaped magnets
<p>| | |</p>
<table>
<thead>
<tr>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Rated stator current (Arms )</td>
<td>15</td>
</tr>
<tr>
<td>Total number of series turns per phase (mm)</td>
<td>1</td>
</tr>
<tr>
<td>Air gap length (SPM)( mm )</td>
<td>1 + Magnet depth</td>
</tr>
<tr>
<td>Number of Poles</td>
<td>14 poles</td>
</tr>
<tr>
<td>Stack length (mm)</td>
<td>65</td>
</tr>
<tr>
<td>Magnet volume per pole(mm³)</td>
<td>2870</td>
</tr>
<tr>
<td>Magnet remanent flux density(T)</td>
<td>1.3</td>
</tr>
<tr>
<td>Base speed (rpm)</td>
<td>428.571 (50Hz)</td>
</tr>
</tbody>
</table>

Tab. 4.2: Key Machine Parameters for Rotor Geometry Comparison

highest torque and power density up to 2100rpm, its CPSR is comparatively smaller.

As is already known, that IPM have considerably small air gap width and hence are quite prone to the effects of harmonics produced because of the FSCW in stator, the increase in air gap width improves this and also increases CPSR. But the associated power drops down with increasing air gap as the leakage flux of PM end is negligible in IPM. Thus a trade off has to be made in this case between the air gap and desired CPSR value. The optimization of magnet angle in v shaped rotor magnets can improve the CPSR performance while also reducing the torque ripple, this is discussed in the following section.

4.3 Magnet angle variation

The use of V -shaped magnets increases magnet having surface area \(A_m\), leads to an increase in air gap flux density \(B_{air}\). An effective way of increasing magnet surface area is by the variation of the V-angle. A flux density plot comparison in Figure 4.7 shows the values of magnet surface area and air gap flux as the V-angle is changed.

Figure 4.8 shows the FW performance for several angle variations, namely 120°,90°,60° and 30°. Power is normalized for a clear CPSR comparison.Following results where obtained by the authors, as the V -angle decreases, air gap flux increases in proportion to the magnet surface area. An
increase of CPSR was also seen when V-angle decreased from 120° to 60°. A maximum CPSR of 10:1 was achieved for the model with a V-angle of 60°, after which the CPSR began to fall despite an increase in torque generated at 30°. The maximum possible CPSR of 10:1 is obtained without considering the loss components at 60° V-shaped segmented IPM, upon considering the loss components the CPSR is bound to come down a little. In the machine considered by authors for analysis in [Chong et al. (2010b)], when loss was accounted for the CPSR was reduced to 9.2:1.

4.4 Halbach Array Magnets

Segmentation of magnets is primarily used for reducing torque ripples in IPM machine. Another approach to reduce the effect of harmonics in back emf and hence torque is the production of sinusoidal flux in the air gap using the specially shaped rotor magnets or utilizing Halbach array. The performance of the motor can be obviously be thereby, improved. An example of effect of Halbach array on magnetic field has been shown in figure 4.9. As is evident from the figure one side of magnet has high flux concentration while the other side had minimal concentration due to the special alignment of magnets.

Halbach array widens greatly the application fields of PM motor, some of the principle applications area for Halbach motors are flywheel energy storage systems, artificial heart blood supply system and servo system, and in aviations. These motors are used in the aforementioned applica-
tion because they offer some advantages over conventional systems, that are enumerated below.

1. The larger air gap flux can increase the motor power density.
2. The inherently sinusoidal air gap field distribution can be obtained without recourse to skewing or distributed winding of the stator.
3. Due to good self-shielding effect, rotor back-iron is not essential so that the mass and inertia can be reduced and the dynamic performance can be improved.
4. The slotless winding structure can be used to weaken the effect of cogging torque.

4.4.1 Types of Halbach arrays

According to operation principles of different motors, Halbach array magnet can be classified as four types. Though the Halbach array magnet structure is different, Halbach array follows the same rule that the angle between any two adjacent magnetization direction $\Delta \theta_M^\circ$ is a constant. Although the detail explanation of each topology is out of the scope of this thesis work, the four types are enumerate below:

1. Linear magnet array which can be applied on linear motor as in figure 4.9. They are characterized such that the centers of all magnet segments are in a line. The linear magnet can have either a brick shape or a tubular shape according to each magnet segment shape. If $n$ is the number of magnet segments per pole, magnetization direction $\Delta \theta_M^\circ$ is given by

$$\Delta \theta_M^\circ = \pm 180^\circ/n$$

(4.2)

Here, "+" and "−" shows that the main magnetic field is located above and below the magnet, respectively for brick shaped magnet array. For the tubular shaped magnet, "+" and "−" show that the main magnetic field is located inside and outside the magnet, respectively.
2. Planar magnet array are applied on planar motor. In these motors centers of all magnet segments are in a planar. The planar magnet disposition is represented in figure 4.11.

3. The third kind is cylindrical magnet which can make motor rotate. The centers of all magnet segments are in a ring. According to magnetization pattern, the cylindrical magnet can be magnetized in the radial direction and axial direction. The cylindrical magnet disposition is represented in figure 4.10. For Halbach array cylindrical magnet, \( \Delta \theta_M^\circ \) can be expressed that

\[
\Delta \theta_M^\circ = 180^\circ \left( 1 \pm \frac{p}{n} \right)
\]  

For radially magnetized magnet, in equation 4.3 "+" and "−" show that the main magnetic field is located inside and outside magnets, respectively. Similarly, for axially magnetized magnet, "+" and "−" shows that the main magnetic field is located above and below the magnets, respectively.

4. Spherical magnets are applied on spherical motors to realize three-degree of freedom motion. According to magnetization range, the spherical magnet includes one-layer magnetic pole and multi-layer magnetic pole. Though the Halbach array magnet structure is different, Halbach array follows the same rule that the angle between any two adjacent magnetization direction \( \Delta \theta_M^\circ \) is a constant.
Halbach magnetized brushless machines exhibit an essentially sinusoidal air gap field distribution and a sinusoidal back-emf waveform, and therefore are eminently suitable for brushless ac operation. Although they are also applied in brushless dc machines too. In the article [Zhu et al. (2003)] AC operation of anisotropic, bonded, NdFeB Halbach magnetized ring magnet is employed and the steady-state performance of a Halbach magnetized brushless ac machine when operating in the constant torque and flux-weakening modes has been investigated both theoretically and experimentally. A Halbach magnetized machine is a form of surface-mounted magnet machine. The prototype motor is supplied from a dc link voltage of 21 V, its rated peak phase current being 45A. The stator has 18 slots, which accommodate a three-phase non-overlapping winding, while the rotor has no back-iron and is equipped with a 12-pole anisotropic bonded NdFeB Halbach magnetized magnet.

Although the Halbach magnetized machine is also a surface-mounted magnet motor topology, it has been designed to have 1.0 p.u. winding inductance by appropriately designing the stator teeth and slot openings so that the majority of the d-axis armature reaction flux does not pass through the magnets but instead passes across the stator slot openings. Hence, a high level of flux-weakening and demagnetization withstand capability can be achieved [Zhu et al. (2003)]. Also a significantly high value of CPSR is observed even with FSCW.

The author observed that the disadvantage of this method is that during the measured torque in the constant torque mode (i.e., below the base-speed of $\sim 860r/min$) for the above machine is significantly lower than the predicted value (using FEM) when saturation is neglected. This is due to the effect of cross-coupling saturation between the magnetic field produced by the Halbach magnetized magnet, and magnetic field produced by the q-axis armature reaction. This causes $\psi_m$, and consequently the torque, to reduce significantly as the q-axis current increases as evident from the equation 4.4. It has also been shown that the influence of the cross coupling on $\psi_m$ is much more significant than that on the d- and q-axis inductances in terms of compromising the achievable torque. Here, flux linkage is given by

$$\psi_m = \frac{T}{(3/2)pI_q} \quad (4.4)$$

Apart from the application mentioned in the earlier section motor with Halbach rotor magnets find application recently in aerospace, high-precision process control, marine navigation, military, etc. The strong interests owes to the high torque and power density servo-motors with very fast dynamic response capability of these machines.
5. FLUX WEAKENING IN AXIAL FLUX MACHINE

The significant research efforts for past 30 years have resulted in what can now be considered a mature technology for AXIAL-FLUX permanent-magnet (AFPM) operation and usage. A wide variety of applications, from renewable energy systems to transportation, wherever extreme axial compactness coupled with high torque density and high efficiency are necessary. Some of the application areas of AFPM are hybrid electric vehicles, washing machines, aircraft propulsion, combined heat and power, wind energy generation, portable gensets, and starter/alternators. A general classification of AFPM machine topologies is presented, using a tree diagram with five hierarchical levels, as shown in figure 5.1.

Calculation of various design parameters has been described in the paper [Huang et al. (1999)] by Surong Huang et al. using the basic sizing equation which are presented in equation 5.1 and 5.3.

\[
P_{mec} = C_{mec} f D_0^3 L_c.
\]

(5.1)

where C is defined as,

\[
C_{mec} = \frac{1}{1 + K_{\phi}} m K_e K_i K_L K_p \eta B_g A_{avg} \frac{1}{p} (1 - \lambda^2) \frac{1 + \lambda}{2}.
\]

(5.2)

\[
T_{em} = \frac{\pi}{4} B_{ave} A_{in} K_d \lambda(1 - \lambda^2) D_0^3.
\]

(5.3)

Fig. 5.1: AFPM machine topologies
Standard AFPM machine topologies are generally machines with poor CPSR ($\leq 2 : 1$) owing to the low synchronous inductance values. New AFPM topologies and advancement in alternative ways for achieving flux weakening, have demonstrated that CPSR as high as $10 : 1$ is achievable. There are various design topologies to attain flux weakening in axial permanent magnet machines. Some of them are:

5.1 Field Controlled Axial flux Permanent Magnet (FCAFPM) machine

An exploded view of Structure of FCAFPM (Field Controlled Axial flux Permanent Magnet) machine is shown in the figure [5.2] In the papers $[\text{Kosaka et al. (2005)}]$ and $[\text{Aydin et al. (2003)}]$, the author has shown how to design torus FCAFPM machine. They have also used 3-D FEA analysis/optimization, and compared the predicted results with the ones obtained from experimental testing.

Figure [5.2] shows an exploded view of the FCAFPM machine, it is an axial-flux topology that allows flux control and weakening without the use of a negative $d$-axis current component. As shown in the figure [5.2] it has two slotted concentric stators and two rotors. The stators are with two sets of three phase windings, one for each air-gap (not shown in the figure), along with a dc winding placed between the two stators. Each rotor pole consists of NdFeB magnets with an arc-shaped iron pieces. Each rotor has two concentric rings, formed alternatively by PMs and iron pieces. The machine can be classified as a particular version of the slotted NS Torus topology and rotor PMs are arranged with opposite magnetization. Aydin et al. demonstrated that by commanding a positive or negative dc current in the stator dc field winding, the flux linked with the stator windings can be intensified or the flux can be weakened, respectively. Advantage of this particular topology is that the flux density component due to dc excitation passes through the iron pieces and without crossing through the PM, thereby minimizing the risk of PM demagnetization during heavy flux weakening.
To this effect appropriate distances between PMs and the iron pieces, as well as, between dc winding slot width and the concentric stators is needed. The authors, Ayedin et al. designed, simulated and built a 1.2 HP, eight pole, 1200 r/min machine. Using 3-D FEA analysis the variation in air-gap flux density was kept between 0.75 T and 0.2T, and a resultant CPSR of up to 3.75 : 1 is predicted using FEA.

5.2 Negative saliency machine (NSAFPM)

Tapia et al. in the papers [Gonzalez-Lopez et al. (2008)] and [Moncada et al. (2010)] introduced the design, and testing of negative saliency machine (NSAFPM) which allows flux weakening with a reduced amount of d-axis current. A typical NSAFM machine is shown in figure 5.3, it has a single, slotted stator and two rotors with NN magnetization. The stator has two standard overlapping three phase ring windings, one for each air-gap. Each rotor has PMs and iron poles laid out as concentric rings, in a similar fashion to the previous topology (FCAFPM), with the difference that the d-axis has a lower reluctance compared to the q-axis, i.e., $L_q < L_d$, and hence the "negative saliency" similar to the ACFPM topology, here the demagnetizing flux created by d-axis current, passes through the iron pieces avoiding the magnets, again minimizing the possibility of PM demagnetization. But this topology also resulted in a maximum CPSR at a mediocre level of 3:1.

Lately, the idea of mechanical field weakening has been growing popularity as it allows CPSR value as high as 10:1 and two of the design topologies for the same are discussed in the next sections.
5.3 Mechanical flux weakening

The flux can be weakened by increasing the negative d-axis current and decreasing the air gap flux density or by reducing the flux linkage to the winding through other methods like by the rotor and stator displacements.

Some electromagnetic-structure modifications can be introduced in the original AFPM arrangement, such as the use of a slotted winding in conjunction with soft-material slot-closing bars, these modifications increase machine inductance. This allows the stator reaction flux to be profitably used for offsetting the magnets flux, although over a limited extension of speed range. However, the wide speed range over which a direct-drive starter/alternator is expected to deliver its output power, and which is also the objective of this thesis, presents a significant design challenge for improving the flux-weakening capabilities of such a slotted winding AFPM. In addition, PM starter/alternator machines are desired to fulfill the regulation, that the machine output voltage should not simply rely on the control of stator reaction flux, as the machine emf increases linearly with speed and any accidental loss of the flux-weakening current control would expose humans to high machine terminal voltages during high-speed operation.

To overcome the aforementioned limitation, the PM machines relies on the use of a speed dependent mechanical device to provide control of the flux linkage and thereby of the induced emf. A variation of the above method would be the usage of a torque dependent mechanical device for rotor shifting. Both these topological aspects are discussed in the succeeding subsections.

5.3.1 Rotor shifting

Caricchi, Del Ferraro et al. presented and developed the idea of mechanical flux weakening in articles [Ferraro et al. (2004)] with a TDD (torque-dependent device), in article [Caricchi et al. (2001)] with a SDD (speed-dependent device) for relative displacement of rotors. The two topologies with different devices for the rotor displacement are compared in [Ferraro et al. (2006b)]. The authors used machine with following parameters and design requirements: A 4 kW, 42 V, 8 pole pair, slot-less Torus starter/alternator with Litz-wire in the stator winding due to high speed operations for design and analysis of machines suitable for field weakening. The machine is characterized by two operating modes: In starter mode, the speed ranges from 0 to 250 r/min with a maximum torque of 150 Nm; while in generating mode, the speed ranges from 600 to 6000 r/min with a maximum torque rating of 32 Nm. The base speed is set at 1200 r/min, and constant power operation is held from 1200 to 6000 r/min, i.e. a CPSR of 5 : 1

Figure 5.4 shows the rotor displacement of two rotors with respect to one another causing the modification of flux linkage associated with the stator coils (reduced here), without having to vary the air-gap flux density. It can be shown that the displacement angle is a function of speed.

The concept is very simple: AFPM machines having two rotor discs has stator linked flux, which is the vector sum of the flux generated by the magnets on the two rotors. The two rotor discs are generally so aligned, that the vector sum resolves in to an algebraic sum. However, displacing the
two rotors relative to each other by a specified mechanical angle, say \( \theta \) w.r.t each other, such that linked flux which is the resultant of two flux vectors shifted by an electrical angle \( p\theta \), for a \( p \) pole pair machine. The vector diagram is shown in figure 5.5. Its to be noted that this rotor shifting doesn’t change the total magnetic energy in the system, hence linked flux weakening can be ideally achieved with no energy expense. Since such a rotor shifting doesn’t change the total magnetic energy of the system.

As can be seen from the figure 5.4, the magnets drive flux across the annular air gaps into the stator core; the flux then travels circumferentially along the toroidal core, back across the air gaps, and then through the back iron of the rotor discs as shown in figure 5.4A. It clearly appears from 5.4A) that the stator coil actually links only half of the flux being exiting from each rotor disc pole, so that the total flux linkage actually is the result of equal flux contributions given by both two facing magnets. The introduction of a phase displacement between the two rotor discs results in a rotor displacement and the new flux path is modified as shown in figure 5.4B.

If \( \omega \) is the speed of the machine and \( \omega_b \) is the base value. When the two rotor discs rotate at a given base speed \( \omega_b \) and the two rotors are positioned to have no mechanical phase displacement between them, so resultant emf \( E_l \) will be algebraic sum of the two constant vector emfs generated due to the flux linkage in stator because of the two rotors. A displacement angle of \( p\theta \) will hence be included in the emf vector addition. Since the back emf can not be increased beyond the terminal voltage from the inverter circuit, required deviation \( p\theta \) is obtained as below

\[
E(\omega) = 2k\omega \cos(p\theta/2) = 2k\omega_b = \text{const}
\]

(5.4)
and hence, solving for $p\theta$

$$p\theta = 2\cos^{-1}(\omega_b/\omega)$$ (5.5)

The equation (5.5) shows that the electrical phase angle $p\theta$ needed to achieve constant emf amplitude at any machine speed beyond the base speed for flux weakening. The desired angle is therefore proved to be a function of the speed. In order to achieve a constant-power mode of operation, torque must decrease inversely with the speed. With reference to the nominal torque, at any given speed $\omega$ torque value $T$ must then be:

$$\frac{T}{T_b} = \frac{\omega_b}{\omega} = \cos\left(\frac{p\theta}{2}\right)$$ (5.6)

As is clear from the above equations (5.5) and (5.6), the relative displacement between the two rotors can be achieved using either torque-dependent device (TDD) or speed dependent device (SDD). A brief description of these two mechanisms is as follows:

- A torque-dependent device is designed and built by using a linear spring: when torque is applied to the machine, the spring reacts with a deformation and shifts the rotors. Such a device has been developed and tested along with a steady state and transient analysis in [Ferraro et al. (2004)]. The desired mechanical phase displacement as a function of the electromechanical torque can be accomplished by a simple spring-mass system. Figure 5.6 shows the schematic representation of such a proposed device.

- A speed-dependent device is built based on a cam-spring governor system acting on a treadle lever: the relative shift of the rotor discs is based on the dynamic equilibrium between the centrifugal force acting on the cam and the reaction force of the spring as in the article [Caricchi et al. (2001)]. Figure 5.7 represents the schematic representation of a cam-spring
Fig. 5.6: a) Schematic representation of the mechanical spring device, of S/A and engine b) Mechanical displacement by $\theta$ between the rotors after a spring extension

governor acting on a treadle level being regulating the phase displacement between the rotor discs: Figure 5.7A shows standstill position (no phase displacement between the rotor discs). Figure 5.7B shows operating condition i.e. the spring-contrasted angular rotation of the cam imposes an angular displacement to one of the rotor discs.

The main difference between the two devices is that the SDD behaves ideally as a constant voltage source while the TDD behaves as a constant current source. The proposed devices introduce innovative concept of mechanical flux weakening for AFPM power regulation in a wide speed range. In the article [Ferraro et al. (2006b)] it has been demonstrated that the introduction of these devices doesn’t modify machine behavior in starting mode, so that the particular features of AFPM machines in terms of high torque density and overload capability remain unaffected.

5.3.2 Stator shifting

The dual of rotor shifting is stator shifting, and it was explored in [Ferraro et al. (2006a)] and [Capponi et al. (2009)] by F. G. Capponi et al. Experimental tests on a 2-kW prototype validate the capabilities of the proposed solution, both during steady-state and transient operating conditions were done by the authors for a proof of concept machine. A 3D model is shown in the figure for the actuator assembly for stator winding shifting in figure 5.8. The machine was designed to supply loads with a constant voltage between 600 and 6000 r/min, i.e., the CPSR is 10 : 1. The stator winding is split into two separate and adjacent Litz-wire windings, each one encapsulated in epoxy resin and both connected in series; the two windings are encased in a fixed frame and separated by low-friction Teflon rings.

Flux-weakening is achieved by rotating the windings with respect to each other by means of a purposely designed actuators. Its important to notice the highest rated machine speed in the
Fig. 5.7: Mechanical weakening of flux linkage via angular phase displacement between the two PM rotor discs: A) aligned rotors, B) displaced rotors

Fig. 5.8: Three-dimensional view of the mechanical device for stators (a) aligned and (b) shifted.
previous section was also 6000 r/min. But because of considering different base speed the CPSR here is 10:1 while in the previous case it was 5:1 and hence it was not inferior to this approach. However, a drawback of the devices mentioned in the previous section was that they were not controlled actively, hence control on machine output voltage was limited, especially during transients. In [Ferraro et al. (2006a)], an active control by means of an electromechanical device is introduced. In order to simplify the mechanical layout of the system, displacement is not obtained on the rotors but on the stator which is, for this purpose, split into two separate adjacent windings connected in series.

In the figure 5.9, a cross-section of two pole-pitches of an iron-less AFPM machine, with two stator windings is shown. In figure 5.9A, the main flux path during \( \dot{\text{fullflux}} \) operating mode can be seen, here the two stator windings are aligned, and thus link the same flux: as the windings connected in series. While, two stators are shifted of a mechanical angle \( \theta \). The machine works in flux weakening mode in this configuration. Also, the two windings link displaced emfs and hence they are summed as phasors, \( \therefore \) Emf is given as

\[
E(\omega) = 2k\omega \cos \left( \frac{p\theta}{2} \right)
\]  

(5.7)

The energy in the airgap doesn’t change, just like in the previous section, hence no torque is required to the actuator for displacement, except for achieving the dynamic performances requested to the machine. Therefore the efficiency doesn’t deteriorate by employing this topology. Also the combination of ironless stator and Litz-wire windings allows for light weight and high efficiency. Experimental results prove proper operation of the system up to 5000 r/min, i.e., a CPSR equal to
Due to mechanical limitations of the test bench, so it can be surmised that the efficiency will not deteriorate significantly and a speed of 6000 rpm can be achieved with flux weakening. Figure 5.10 shows the predicted efficiency for the machine with stator winding actuator for flux weakening.

5.4 Comparison with other PM machines

Following general points can easily be considered from above discussion i.e. the AFPMSM have higher power densities, given magnet material and air-gap flux density are same for machines under consideration, while, radial flux designs have higher moment of inertia for rotor; the active weight of the AFPM machines are smaller; copper losses in the slotless AFPM machines are higher than their slotted counterparts [Capponi et al. (2012)].

In [Cavagnino et al. (2002)], the authors performed a detailed design comparison between RFPM (Radial flux permanent magnet) machines and a Kaman machine, designed for direct drive applications. Both machines had sinusoidal back-emf and slotted stators with following constraints: equal air-gap, teeth and yoke flux densities, equal volume, equal loses/heat exchange surface, including equal rated speed (1000 r/min). The authors concluded that the AFPM topology is preferable for machines having pole pairs (> 5) and short axial length, i.e., the ratio between axial length and outer diameter < 0.3. Fig. 5.11 shows the supporting plots of torque density as a function of number of poles and of the aspect ratio $L_a/D_0$, for both AFPMSM and RFPMSM.

In the article Parviainen et al. (2005), the author has perform a similar comparison between 55 kW radial and double-stator, single-rotor, axial-flux machines, with 150 r/min, 300 r/min and 600 r/min rated speeds; mechanical constraints where included for a faithful comparison. The comparison criteria were: efficiency, total volume occupied, and cost incurred. Following conclusion are drawn by the author: the efficiency is higher for radial flux flux machines, mainly owing to the long end windings of the chosen axial-flux topology; AFPM machines with one or two pole pairs
Fig. 5.11: Torque density comparison as a function of number of poles and aspect ratio $\lambda$: (a) RFPM machine and (b) AFPM machine.

have generally shown extremely poor performance; a larger selection of pole pairs may be used for radial flux machines, but having more than eight pole pairs, the cost of the radial flux machines will be higher than that of the axial-flux machines.

It can therefore, imperative to highlight that dependent on the application axial and radial machines may be favored over one another. So while designing the drive mechanism, it is important to consider the design requirements like space, cost, number of poles, efficiency etc. The answer for selecting suitable motor will vary accordingly.
6. MACHINE WITH MAGNET ON STATOR

Due to the poor thermal capability of traditional PM with magnets on rotor, partial demagnetization problem can occur over the life cycle of machine, placing magnets on the stator allows possibility of better thermal capability. In addition it lends robustness and better mechanical integrity during high speed operations. Also, safety-critical applications need improved fault-tolerance capability, as the PM machine cannot be de-excited, these machine (like FSPM) has an advantage in this regards. These machines have gained popularity recently, and considerable research is being done in their applicability. One such machine, Flux switching permanent magnet (FSPM) machine, has been discussed in the following section. Their design aspect with respect to improved flux weakening capability is discussed and a comparison has been made with a IPMSM (Prius).

6.1 Flux switching permanent magnet (FSPM) machines

6.1.1 Introduction

Permanent magnets (PMs) of conventional permanent magnet machines are mounted with rotor. The magnets usually need to be protected from the centrifugal force by employing a retaining sleeve, which is made of either stainless steel or nonmetallic fiber. Due to its poor thermal dissipation, rotor temperature is raised, which may cause irreversible demagnetization of magnets and ultimately limits the power density of the machine. Its rotor is similar to SRM rotor which has no permanent magnets and windings, higher mechanical strength, better flux-weakening capability and fitter for high-speed operation. Detailed analysis and sizing equation for FSPM can be obtained in the article [Hua et al. (2006)].

6.1.2 Advantages of FSPM

The general research has revealed some advantages of FSPM, such as high power density, high efficiency and robust rotor structure. The PM flux linkage and armature linkage of FSPM are parallel in a period, because of its structure. So the magnetic resistance is small and the work point of permanent magnet is stable. Not only the output torque is high, but also d-axis inductance is large [Yang et al. (2010)]. Moreover, the most attractive performance is its special ability of flux-weakening. FSPM has an infinite speed range in theory with constant power.

In the paper [Yang et al. (2010)], J. Yang et al. chose a typical 3-phase FSPM with 12-stator-tooth and 10-rotor-pole as an example. The authors used finite element analysis techniques to de-
termine the influences of permanent magnet width, air-gap length and rotor-tooth width on flux-weakening capability, and flux-weakening factor is analyzed. The novel topology of PM motor used by the authors has flux-switching permanent magnet motor (FSPM), whose cross sectional view is shown in figure 6.1 with a slot/pole ratio of 12/10. It has a rotor and stator which are salient and permanent magnets are placed in the stator. The FSPM machine shown in figure 6.1, has a stator that contains 12 segments of U-shape magnetic cores and 12 pieces of PMs. Each phase has four coils, here, A1-A4 windings belong to phase A, B1-B4 windings belong to phase B, C1-C4 windings belong to phase C. In addition, the concentrated windings are employed, which lead to lower copper consumption and lower copper losses due to shorter end-connections. The parameter of the machine used are tabulated in table 6.1.

6.1.3 Flux weakening capability

FSPM machine like most PM machine drive configuration is supplied with inverter. In constant torque region while working under the current limit of the inverter, sufficient voltage below the limiting voltage $U$ is supplied. In contrast, as the back-EMF rises, and machine is operated above the base speed, the inverter voltage becomes lower than back-EMF, thus energy cannot flow into the machine, namely entering constant power region. Then, the machine has to be operated under flux-weakening control to reach higher speed. The speed and maximum speed, from [Wang et al. (2008)], are expressed by 6.1 and 6.2 respectively.

$$\Omega = \frac{U}{p\sqrt{(\psi_m + L_d i_d)^2 + (L_q i_q)^2}} \quad (6.1)$$
<table>
<thead>
<tr>
<th>Parameters</th>
<th>Number</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of phases</td>
<td>3</td>
<td></td>
</tr>
<tr>
<td>Rated speed</td>
<td>1500</td>
<td>rpm</td>
</tr>
<tr>
<td>Number of rotor teeth</td>
<td>10</td>
<td></td>
</tr>
<tr>
<td>Number of stator teeth</td>
<td>12</td>
<td></td>
</tr>
<tr>
<td>Stator outer diameter</td>
<td>172</td>
<td>mm</td>
</tr>
<tr>
<td>Tooth width</td>
<td>8.67</td>
<td>mm</td>
</tr>
<tr>
<td>Magnet thickness</td>
<td>6.19</td>
<td>mm</td>
</tr>
<tr>
<td>Air-gap length</td>
<td>0.35</td>
<td>mm</td>
</tr>
<tr>
<td>Rotor iron length</td>
<td>104</td>
<td>mm</td>
</tr>
</tbody>
</table>

Tab. 6.1: FSPM parameters

where, $\Omega$, $U$, $P$, respectively are mechanical angular speed, terminal voltage and number of poles.

In order to rise the motor speed, the armature current should be adjusted. If $i_d = I_{\text{max}}$ and $i_q = 0$, then the armature current is completely used to counteract the PM flux-linkage. Therefore, the maximum operation speed $\Omega_{\text{max}}$ can be obtained as in equation 6.2

$$\Omega_{\text{max}} = \frac{U_m}{p(\psi_m - L_d i_m)} \quad (6.2)$$

If $I_{\text{max}}$ is maximum limiting current and the limiting voltage is $U_m$, with $U_m$ and $I_{\text{max}}$ constant, the maximum speed depends on $\psi_m$ and $L_d$.

If $\psi_m \leq L_d I_{\text{max}}$ is satisfied, the speed can be infinite in theory. However, in most PM motors, this condition can not be satisfied. Thus, a factor $K_{fw}$ is defined to reflect the flux weakening capability of the machine designed, if i.e.

$$K_{fw} = \frac{L_d I_{\text{max}}}{\psi_m}$$

$$= \frac{[(\Lambda_d N^2)(\sqrt{2} J_s A_h s K_p)/N)]/(N \phi_m)}{\sqrt{2} \Lambda_d J_s A_h s K_p/\phi_m} \quad (6.3)$$

Obviously, larger the $K_{fw}$, better will be the flux weakening capability. However, the results from FE analysis, it can be noted that the PM flux $\phi_m$ is significantly high due to the flux focusing effect employed in FSPM, resulting in excellent torque capability. But this also makes the d-axis inductance small due to the heavy magnetic saturation. So reluctance torque contribution in total torque is small. Hence, the flux-weakening capability of the FSPM machine with NdFeB is inferior to the case of adopting ferrite under the current density of say, $5 A/mm^2$, in which it can reach infinite speed [Hua et al. (2006)]. here, $\psi_m = N\phi_m$, $L_d = \Lambda_d N^2$ and $I_{\text{max}} = \sqrt{2} J_s A_h s K_p/N$ as obtained in [Hua et al. (2006)].

Where, $\Lambda_d$, $N$, $J_s$, $A_h$, $K_p$, $\phi_m$ are d-axis permeance per turn winding turns per coil, current density, area of half slot, slot packing factor, and PM flux, respectively. Equation 6.3 reveals that
the flux-weakening capability can be improved by increasing $L_d$ if current is invariant which is a most popular method. steady-state $L_d$, when rotor d-axis coincides with A phase winding, let q-axis current to be zero then from the Park transformation, equation 6.4 and 6.5 can be obtained,

$$i_d = \frac{2}{3}(i_a - \frac{1}{2}i_b - \frac{1}{2}i_c)$$

(6.4)

$$\psi_d = \frac{2}{3}(\psi_a - \frac{1}{2}\psi_b - \frac{1}{2}\psi_c)$$

(6.5)

During flux weakening machine is supplied with currents, $i_a = -2i_b = -2i_c$. Thus, $i_d = i_a$ and steady-state inductance $L_d$ is

$$L_d = [\psi_d - \psi_m]/i_d = [\psi_d - \psi_m]/i_a$$

(6.6)

It needs to be emphasized that the winding axis of stator coincides with the stator tooth axis. The PM flux-linkage is zero if the rotor-tooth aligns with the stator tooth in FSPM. When the rotor staggars $9^\circ$ (mechanical angle), the flux-linkage of windings reaches maximum. Referring to the situation of permanent magnet synchronous motor, when a winding axis coincides with the rotor d-axis, the winding flux-linkage reaches a maximum. However, the flux linkage is minimum if the winding axis coincides with q-axis. In figure 6.2 rotor d-axis coincide with the phase A winding axis implying a zero rotor angle. The author in Yang et al. (2010) used FEM analysis to obtain the flux linked with the three windings. Using equation 6.5, flux linked $\psi_d$ in d axis can be calculated. After further analysis following parameteric values are obtained by the author: $L_d = 13.96$mH, $\psi_m=0.2294$Wb, $i_d=13.37$A a. From equation 6.3, $k_{fw} = 0.814$ is obtained. Also speed extension is by a ratio i.e. CPSR is 5.36 by the author.
6.2 Comparison of FSPM and IPM

The article [Cao et al. (2012)] an IPM machine with FSPM is analytically and experimentally compared using FEA models. PM motors in industrial, residential, and automotive applications are still dominated by interior permanent-magnet motors (IPM) because the claimed advantages of stator PM motors have not been fully investigated and validated. Hence, this paper will perform a comparative study between a stator-PM motor, namely, a flux switching PM motor (FSPM), and an IPM which has been used in the 2004 Prius hybrid electric vehicle (HEV). For a fair comparison, the two motors are designed at the same phase current, current density, and dimensions including the stator outer diameter and stack length. First, the Prius-IPM is investigated by means of finite-element method (FEM). The FEM results are then verified by experimental results to confirm the validity of the methods used in this study. Second, the FSPM design is optimized and investigated based on the same method used for the Prius-IPM. Third, the electromagnetic performance and the material mass of the two motors are compared.

From the table figure 6.3, where the electromagnetic performance and cost of both motors are summarized and tabulated. Following advantages were surmised by the authors:

<table>
<thead>
<tr>
<th>Items</th>
<th>Prius-IPM</th>
<th>FSPM</th>
</tr>
</thead>
<tbody>
<tr>
<td>$N_{pu}$ (Turns)</td>
<td>8</td>
<td>14</td>
</tr>
<tr>
<td>$k_{io}$</td>
<td>0.601</td>
<td>0.728</td>
</tr>
<tr>
<td>$k_{s rm}$ (mm$^2$)</td>
<td>0.7969</td>
<td>0.7969</td>
</tr>
<tr>
<td>EMF (RMS)</td>
<td>71.5</td>
<td>104.1</td>
</tr>
<tr>
<td>Mass of PM (kg)</td>
<td>1.239</td>
<td>2.837</td>
</tr>
<tr>
<td>Mass of stator iron (kg)</td>
<td>19.05</td>
<td>11.18</td>
</tr>
<tr>
<td>Mass of rotor and shaft (kg)</td>
<td>11.58</td>
<td>15.16</td>
</tr>
<tr>
<td>Mass of copper mass (kg)</td>
<td>6.8</td>
<td>4.76</td>
</tr>
<tr>
<td>Total mass</td>
<td>38.67</td>
<td>33.93</td>
</tr>
<tr>
<td>Inner power angle (deg)</td>
<td>48</td>
<td>5</td>
</tr>
<tr>
<td>Input peak current, $I_{\text{max}}$ (A)</td>
<td>250</td>
<td>250</td>
</tr>
<tr>
<td>Speed (rpm)</td>
<td>1200</td>
<td>1200</td>
</tr>
<tr>
<td>$T_{\text{avg}}$ (Nm)</td>
<td>383.4</td>
<td>347</td>
</tr>
<tr>
<td>$T_{\text{ripple}}$ (Nm)</td>
<td>79.52</td>
<td>21.15</td>
</tr>
<tr>
<td>$T_{\text{ripple}}$ (%)</td>
<td>20.7</td>
<td>5.9</td>
</tr>
<tr>
<td>$T_{\text{cogging}}$ (Nm)</td>
<td>3.7</td>
<td>5.14</td>
</tr>
<tr>
<td>$T_{\text{avg/PM}}$ (Nm/kg)</td>
<td>309.45</td>
<td>122.325</td>
</tr>
<tr>
<td>$R_0$ ($\Omega$)</td>
<td>0.07</td>
<td>0.049</td>
</tr>
<tr>
<td>copper loss (w)</td>
<td>6562.5</td>
<td>4593.75</td>
</tr>
<tr>
<td>Iron loss (w)</td>
<td>150</td>
<td>225</td>
</tr>
<tr>
<td>Output power (w)</td>
<td>48155</td>
<td>43583.3</td>
</tr>
<tr>
<td>Efficiency (%)</td>
<td>86</td>
<td>89</td>
</tr>
<tr>
<td>Theoretical $k_{io}$</td>
<td>Fully weakened</td>
<td>3.73</td>
</tr>
</tbody>
</table>

Fig. 6.3: FSPM vs IPM comparison
• The FSPM has sinusoidal back-EMF, thus offering smaller torque ripple when operating as BLAC.

• Similar to a switching reluctance motor, it has no PMs, armature windings, or brushes on the rotor, thus offering good mechanical integrity and high reliability for high speed operation.

• Different from rotor PM motors, the PMs are located in the stator, which can help achieve better thermal dissipation.

• With concentrated armature winding, the copper loss and copper cost of the FSPM is less.

The FSPM also has some disadvantages in comparison to more traditional IPM machine such as

• The FSPM evidently have smaller PM utilization ratio. As can be seen from Table 6.3, the total mass of the PM material in the FSPM is about 2.29 times that in the Prius-IPM. The ratio output torque per kg PM for Prius-IPM is about 2.53 times that of the FSPM, which the author obtained by dividing average torque to weight of PM used.

• Also, the reluctance torque for FSPM is smaller, if the output power of both (FSP and IPM) motors are designed to be same at equal applied phase current and rotor speed.

• The open circuit back-EMF for FSPM is larger, compared to Prius-IPM.
7. CONCLUSION

The maximum flux-weakening capability, defined as the ratio of the maximum speed to the base-speed, under supply inverter voltage and current limitations, can be achieved when a PM brushless machine is designed to have 1.0 per-unit d-axis inductance such that

\[ L_d = \frac{\psi_m}{I_r} \quad \text{or} \quad \frac{L_d I_r}{\psi_m} = 1 \]  

(7.1)

where \( \psi_m \) is the stator flux-linkage due to the magnets, \( L_d \) is the d-axis inductance, and \( I_r \) is the rated current.

Although it is possible to design a PM brushless machine which satisfies the foregoing requirement, generally, for most PM machines \( L_d I_r/\psi_m < 1 \), since the d-axis inductance is relatively low as a consequence of the recoil permeability of the magnets being approximately equal to 1.0. Nevertheless, the higher the ratio of \( L_d I_r/\psi_m \), the higher will be the flux-weakening capability, which, theoretically, is "infinite" when the ratio is 1.0. However, the higher the flux-linkage, \( \psi_m \) to achieve a high low-speed torque capability, the more difficult it is to realize wide-speed operation.

In [Soong and Ertugrul (2002)], it was shown that it was possible to design any PM brushless machine to achieve "infinite" flux-weakening capability. Clearly, however, if the rated current is high (e.g., the machine is liquid cooled), it is much easier to satisfy equation [7.1] even for surface-mounted magnet machines, which have a high \( \psi_m \) and a relatively low \( L_d \). For example, in [Zhu et al. (2003)] "infinite" FW capability was achieved with an SPM machine equipped with a self-shielding, sinusoidal magnetized rotor having no back-iron. However, in general, it is much easier to achieve a wide operating speed range with machines equipped with an interior permanent magnet rotor, since generally \( \psi_m \) is lower, while \( L_d \) is higher.

The operational characteristics, design features in various permanent-magnet brushless machine types for flux weakening with emphasis on their low-speed torque and high-speed power capability has been considered in this thesis. Various PM brushless machine topologies have been highlighted, and their relative merits have been briefly described. Special emphasis has been given to the comparative analysis for design topologies and their effect in various machine parameters like cogging, torque ripple, magnet partial demagnetization, stator and rotor losses has been given. In general, however, all machine technologies can meet the performance requirements for wide speed operation such as in traction drives, and each machine technology has merits.

SPM brushless machines which have a fractional number of slots per pole and a concentrated winding have been the subject of recent research. They have an inherently low cogging torque,
short end-windings and, hence, a low copper loss, a high efficiency, and a high power density, as well as excellent flux-weakening performance. The stator coils may be wound either on all the teeth or only on alternate teeth. In the latter case, the phase windings are effectively isolated, both magnetically and physically, and a high per-unit self-inductance can readily be achieved to limit the prospective short-circuit current, by utilizing the relatively high air-gap inductance and the leakage flux at the slot openings. Due to the physical separation of the coils and the negligible mutual inductance between phases, the possibility of a phase-to-phase fault is minimized. Therefore, the fault tolerance and flux-weakening capability of such machines can be significantly higher than for more conventional machine designs.

However, since the torque is developed by the interaction of a stator space-harmonic MMF with the permanent magnets, a relatively high rotor eddy-current loss can result from the fundamental and low-order space-harmonic MMFs which rotate relative to the rotor. However, since the torque is developed by the interaction of a stator space-harmonic MMF with the permanent magnets, a relatively high rotor eddy-current loss can result from the fundamental and low-order space-harmonic MMFs which rotate relative to the rotor.

In addition to the SPMSM and IPMSM, which are also compared for different topologies in this thesis work. Beside them, special attention was given to axial flux machine design consideration for FW mode operation, which is specially attractive for traction and wind plant operation. A comparative study in section 5.4 with other PMSM topologies was also done. The advantages and technical constraint were discussed in this thesis work in chapter 5. In addition PMSM with magnets on rotor are described here in chapter 6, one machine type of this kind is discussed in the chapter, namely FSPM, whose design though a little complicated, exhibits vast scope of usage in wide speed application involving operation in FW region. They are compared with IPMSM in the section 6.2.

Apart from the cogging and torque ripple problems associated with flux weakening issues discussed above, following issues are to be handled in each design topologies. They are discussed below.

### 7.1 Demagnetization withstand Capability

Operation in the FW mode is a necessary requirement for traction applications. NdFeB is the most commonly employed permanent-magnet material for PM brushless machines. Adequate demagnetization withstand capability at the maximum operating temperature, when magnets are most vulnerable to partial irreversible demagnetization is an important design parameter for PMSM. Also, effective thermal management is a key mean for improving demagnetization withstand capability, low reluctance path for the demagnetizing d-axis armature reaction flux can be provided such that it effectively does not cross through magnets. it can be done by either having a narrower stator slot openings and thick tooth-tips, or thick rotor slot bridges in IPM machine. However, such features will also have an influence on $\psi_m$ and $L_d$. In general, it is easier to realize a high demagnetization withstand capability for IPM machines. By careful design, the magnet working point for
Fig. 7.1: Variation of iron loss in SPM and IPM machines when their open-circuit stator iron loss are designed to be the same.

SPM machine can be made to stay reasonably high up the magnet demagnetization characteristic, even when the machine has "infinite" FW capability [Zhu and Howe (2007)]. Also machines with magnets in stator show better demagnetization capability like in FSPM machine.

7.2 Stator Loss

The iron loss which results during FW operating mode depends on the machine topology, as illustrated in Fig. 7.1. In general, SPM machines have the lowest full-load iron loss, IPM machines have significantly higher full-load iron loss, since the armature reaction field has a higher harmonic content due to the small effective air gap. Also since magnets are simply inset into the rotor surface the harmonic content in the armature reaction field increases further, and results in the high full-load iron losses. Suitably designed rotor with magnet disposition such that the flux in air gap is sinusoidal results in lower stator loss, also a suitable selection of appropriate SPP value helps in their reduction, even if the stator has CNW [Zhu and Howe (2007)].

7.3 Rotor Eddy-Current Loss

PMSM of both BLAC and BLDC type machines are usually considered to have negligible rotor loss. However, they may be important in SPM machines, in terms of the resulting temperature rise. Eddy currents may be induced in the permanent magnets, the rotor back-iron, and any conducting sleeve which may be employed to retain the magnets, by time and space harmonics in the air-gap field. Eddy current arise because of the following factors (a) stator slotting; (b) stator MMF harmonics; and (c) non sinusoidal phase current waveforms, which result from six-step commutation and PWM [Zhu and Howe (2007)].

In general the rotor eddy-current loss are relatively small compared with the stator copper and iron losses. Still, it may cause significant heating of the magnets, due to the relatively poor heat dissipation capability associated with rotor structure. Partial irreversible demagnetization may happen. Rotor eddy-current loss are important in machines with a) a high fundamental frequency,
i.e high-speed and/or high-pole number; b) windings span of fractional pole-pitch and which have nearly equal pole and slot numbers. Possible solution include magnet segmentation, either axially and/or circumferentially. Also the use of Halbach arrangement of magnets allows to reduce rotor back iron and incidentally, the eddy losses associated with rotor.

From the above discussion it can be concluded that the new design topologies available with PMSM accounts for their increasing popularity in industry. They are finding application in all kinds of fields from traction to aerospace to power generation as in wind turbines to hazardous operating environments. They are leading the race in satisfying higher standards in terms of size and efficiency, and are increasingly finding new roles even in higher power application.
8. ABBREVIATIONS
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Abbreviated for</th>
</tr>
</thead>
<tbody>
<tr>
<td>$E_q$</td>
<td>induced per phase (lying on the q-axis)</td>
</tr>
<tr>
<td>$k_w$</td>
<td>winding factor</td>
</tr>
<tr>
<td>$N_{ph}$</td>
<td>Number of series turns per phase</td>
</tr>
<tr>
<td>$\phi_{M1}$</td>
<td>Fundamental component of airgap flux/pole</td>
</tr>
<tr>
<td>$f$</td>
<td>Input frequency</td>
</tr>
<tr>
<td>$\vec{E}_x$</td>
<td>Reference emf phasor element (lp.u. length)</td>
</tr>
<tr>
<td>$p$</td>
<td>Number of pole pairs</td>
</tr>
<tr>
<td>$x = n - 1$</td>
<td>Number of winding elements</td>
</tr>
<tr>
<td>$s$</td>
<td>Total number of stator slots</td>
</tr>
<tr>
<td>$\alpha_s$</td>
<td>Scaling from pitch factor</td>
</tr>
<tr>
<td>$\gamma$</td>
<td>Angular displacement</td>
</tr>
<tr>
<td>$\delta$</td>
<td>Phase displacement from reference phase</td>
</tr>
<tr>
<td>$L_{le}$</td>
<td>Armature leakage inductance</td>
</tr>
<tr>
<td>$l_c$</td>
<td>Axial length of stator coil</td>
</tr>
<tr>
<td>$B_{rU}$</td>
<td>Upper layer armature flux distribution</td>
</tr>
<tr>
<td>$B_{rL}$</td>
<td>Formula Lower layer armature flux distribution</td>
</tr>
<tr>
<td>FSCW</td>
<td>Fractional slot concentrated winding (SPMSM)</td>
</tr>
<tr>
<td>$\Psi_a$</td>
<td>Surface permanent-magnet synchronous motors</td>
</tr>
<tr>
<td>$r_g$</td>
<td>Air-gap radius (m)</td>
</tr>
<tr>
<td>$l_{e\text{ff}}$</td>
<td>Active length of the machine (m)</td>
</tr>
<tr>
<td>$N_a$</td>
<td>Phase &quot;A&quot; winding function (turns)</td>
</tr>
<tr>
<td>$B$</td>
<td>Air-gap magnet flux density (T)</td>
</tr>
<tr>
<td>$\theta$</td>
<td>Angle along the air-gap periphery (mechanical radians)</td>
</tr>
<tr>
<td>$B_1$</td>
<td>Peak fundamental component of the air-gap magnet flux density distribution</td>
</tr>
<tr>
<td>$\lambda_0$</td>
<td>Average value of the permeance function to account for the effect of the stator slots</td>
</tr>
<tr>
<td>$k_2$</td>
<td>Machine design constant (for FSCW)</td>
</tr>
<tr>
<td>FCAPFM</td>
<td>Field Controlled Axial flux Permanent Magnet</td>
</tr>
<tr>
<td>AFPM</td>
<td>Axial flux permanent magnet</td>
</tr>
<tr>
<td>$\theta$</td>
<td>Rotor phase displacement in axial machines</td>
</tr>
<tr>
<td>$p\theta$</td>
<td>Electrical flux displacement between rotor fluxes</td>
</tr>
<tr>
<td>SDD</td>
<td>Speed dependent device</td>
</tr>
<tr>
<td>TDD</td>
<td>Torque dependent device</td>
</tr>
<tr>
<td>$\Lambda_d$</td>
<td>d-axis permeance per turn winding turns per coil</td>
</tr>
<tr>
<td>$J_s$</td>
<td>Current density</td>
</tr>
<tr>
<td>$A_{hs}$</td>
<td>Area of half slot</td>
</tr>
<tr>
<td>$K_p$</td>
<td>Slot packing factor</td>
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BIBLIOGRAPHY


