Nine-phase six-terminal pole-amplitude modulated induction motor for electric vehicle applications

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Abstract: In electric drive applications that are based on high-speed induction motors (IMs) with an extended speed range, stator winding pole-changing is a possible technique to avoid oversizing the driving motor. The electronic pole changing employed in multiphase IMs has gained recent interest because it avoids physical winding reconfiguration. The effective number of poles of the air gap flux distribution can be electronically altered by simply changing the applied current sequence to a multiphase stator. The main problem associated with this technique is the significant increase in machine magnetising current with the increase in effective pole number when conventional multiphase distributed windings are employed. This study proposes a new fractional-slot concentrated winding layout with a special stator connection suitable for pole-amplitude modulated IMs that offers a 2:1 pole ratio while maintaining equal magnetising current for both winding pole pairs. Moreover, constant power operation can be achieved for a speed range of over 4 pu. The main concept is discussed and verified through simulations and experimentally. The machine mathematical model and the required vector space decomposition-based controller are also presented.

1 Introduction

Transportation electrification is arguably one of the most promising opportunities to meet the complex environmental challenges facing the modern world. Electric drivetrain architectures based on novel electric machine configurations and power converter technologies have, therefore, been the major focus of many recent research publications. Undoubtedly, machine designs based on permanent magnet (PM) rotors represent the top candidate in this field. The in-wheel motor design has also become an enabling technology in electric vehicles (EVs) as it offers many outstanding operational and constructional merits [1]. However, the past decade has witnessed an extensive use of rare-earth magnets in wind generators, automotive drives, and home appliances, which resulted in the rapid depletion of rare-earth magnet resources. This has driven EV manufacturers to adopt other motor type alternatives with reduced or no PMs [2]. Among the different possible options, induction motors (IMs) with a copper cage rotor have shown promise in some practical EVs such as Tesla [3]. However, the design of IMs with outer rotors for in-wheel direct driven motors is still a challenging research topic [4].

As a matter of fact, electric motors employed in EVs should be capable of providing high torque during starting/acceleration and high power at high speeds for cruising, while the efficiency is maximised. Therefore, designs with high power-to-weight ratios would be crucial to reduce the overall machine weight. Ideally, machines equipped with fractional-slot concentrated windings (FSCWs) are promising candidates in this respect since they offer short end turns, high filling factor, high power density, and high fault tolerant capability [5]. The relatively low number of stator slots, for this winding layout, for a certain pole-pair number highly simplifies the machine construction. This is especially true in direct drive applications when compared with a conventional distributed winding. Unfortunately, the magneto-motive force (MMF) produced by such a winding layout is highly distorted and exhibits numerous sub- and super-harmonic components with non-negligible magnitudes [6]. This characteristic renders FSCWs as a proper choice in PM machines [7]. In recent literature, some

references introduced innovative non-overlapped single tooth winding layouts for both linear and rotary IMs [8–11]. However, most of these winding layouts yield a relatively poor value for the synchronous winding factor, which increases the required copper volume for a given power rating.

Increasing the coil pitch from one, in a single tooth non-overlapped winding, to two, although yields an overlapped winding layout, was found to significantly diminish the effect of slot harmonics, the main undesirable MMF space harmonics [12]. Moreover, it represents an acceptable compromise between a single tooth winding and distributed winding with a slight increase in the overall copper volume in comparison to single tooth windings. This concept has found its way into some recent slot/pole combinations proposed in the recent literature for PM motors, namely 24-slot/10-pole [12], 18-slot/10-pole [13], and 18-slot/8-pole [14]. On the other hand, employing the same concept to three-phase IMs has not still met the same interest due to the effect of sub-harmonic components, which yield opposite torque components and poor machine performance. Fortunately, the multiphase windings have recently shown promise as a viable means to mitigate the space harmonic problem of FSCW since the MMF harmonic spectrum can simply be divided among the available subspaces [15]. By selecting a certain current sequence, some of the parasitic sub- and super-harmonics will completely be eliminated [16, 17].

A rather important requirement of motors employed in EV applications is the ability to run with constant power over a wide speed range (4–5 rated speed) [18]. Although field weakening can easily be applied in IMs, extending the speed across this relatively high range at constant rated power will necessitate an oversized driving motor [18]. This can be explained by referring to Fig. 1, where a family of typical torque–speed curves of an IM operating under constant stator voltage is shown. The machine breaking torque is inversely proportional to the frequency squared. According to NEMA standards, normal IMs are usually designed to have a breakdown torque of 2–2.5 pu [18]. In order to obtain a maximum torque of 0.25 pu at a speed of 4 pu, the employed motor should be able to develop a maximum torque of 4 pu under rated

Fig. 1 Torque-speed curves of IM under constant power operation

Fig. 2 Proposed PAM-9P6T machine

(a) Winding layout, (b) Conventional asymmetrical nine-phase winding, (c) Nine-phase six-terminal connection

frequency, which entails a motor with an approximate rated power of 2 pu. Alternatively, employing pole-changing techniques can be more practical in this regard [18]. In the literature, different electronic pole-changing techniques using multiphase IMs are proposed to avoid physical winding reconfiguration and the extra needed hardware contactors [19]. The pole-phase modulation (PPM) and pole-amplitude modulation (PAM) have shown effectiveness in achieving several speed ratios [18–22]. The effective number of pole pairs can simply be changed by changing the applied stator current sequence order. However, most of these techniques are proposed for IMs with distributed windings with a relatively low pole-pair number. Another critical problem of all of these techniques is the significant reduction in the machine magnetising inductance as the effective number of poles increases, yielding a higher magnetising current and a lower power factor [23].

In this paper, a new PAM winding with a pole ratio of 2:1 is proposed based on an FSCW with an 18-slot/8-pole combination, or its multiples. The proposed winding layout comprises nine phases that are connected using the so-called nine-phase six-terminal (9P6T) connection, which has recently been introduced in [24] to allow for a high-quality flux distribution using a dual three-phase supply. The motor synchronous speed can simply be altered by changing the value of the current phase shift between the two three-phase current sets. The proposed PAM-9P6T machine can be designed with equal magnetising current under both pole modes, namely, 8-pole and 4-pole distributions. This indeed represents a salient merit over most available winding layouts in the literature. The proposed concept is analysed using finite-element (FE) simulations by designing a practical 50 kW IM, while a small-scale prototype IM is used for experimental validation. The machine vector space decomposition (VSD)-based model, as well as the required controller, is also introduced.

2 Proposed winding layout

2.1 Nine-phase six-terminal connection

The proposed winding layout is simply based on the asymmetrical nine-phase winding. It was proven in the literature that a nine-phase FSCW would correspond to a significant reduction in both sub- and super-harmonics [17]; however, a nine-phase converter

will be needed. To simplify the converter complexity, it has been shown in [24] that the 9P6T connection can be a potential option to feed a nine-phase machine from a six-phase supply. This promising connection will, therefore, be employed with the proposed winding layout in order to simplify the power converter. In the proposed PAM-9P6T winding layout, all coils have a coil pitch of two, which represents a compromise between single tooth and distributed windings, as indicated in the introduction section. The required number of stator slots will be 18, or its multiples for a higher number of stator poles. The 18-slot/8-pole combination [14] represents a valid combination where this winding can be employed, as shown in Fig. 2a. The asymmetrical nine phases can be divided into three three-phase winding sets with a spatial phase shift of 20 electrical degrees between different sets, as shown in Fig. 2b. With the 9P6T connection shown in Fig. 2c, only six stator terminals will be provided.

In a normal asymmetrical nine-phase case, the phase shift between the current groups of the winding sets s_{147} and s_{369} should be 40° . With the 9P6T connection employed, the magnitude of the current group of the winding set s_{258} is given by (1) assuming a phase shift angle of γ between the current groups $\vec{i}_{s_{147}}$ and $\vec{i}_{s_{369}}$ while taking $\vec{i}_{s_{147}}$ as a reference

$$\begin{aligned}\vec{i}_{s_{258}} &= \vec{i}_{s_{147}} + \vec{i}_{s_{369}} = I_m e^{-j0} + I_m e^{-j\gamma} \\ &= 2I_m \cos \frac{\gamma}{2} e^{-j(\gamma/2)}\end{aligned}\quad (1)$$

where I_m is the magnitude of the line current.

For $\gamma = 40^\circ$, which emulates a conventional asymmetrical nine-phase system, the current magnitude of $\vec{i}_{s_{258}}$ obtained from (1) will be 1.88 times the line current with a phase shift of 20° . To ensure the same MMF contributions from different phases, the number of turns of the current group s_{258} should, therefore, be 0.532 times the number of turns of other phases, while the conductor cross-section relationship should be the same as that of the current to maintain the same copper volume of a conventional nine-phase winding [24]. The MMF distribution with a one per unit line current for the 18-slot stator shown in Fig. 2a and a shift angle γ of 40° is shown in Fig. 3. With the given winding design, the MMF produced by all three-phase winding sets will be equal to 1 pu, as is clear from

Fig. 3a. The fundamental component of the total MMF distribution is the fourth harmonic, which indicates an 8-pole flux distribution. The lowest order harmonics are the 14th and 22nd. Therefore, in order to have an acceptable machine performance, the cage rotor design including the number of rotor bars and the corresponding bar skew angle should minimise the effect of these harmonics on the induced rotor harmonic currents and parasitic torque components.

2.2 Effect of current phase shift angle, γ , on the air gap flux distribution

In this subsection, the effect of the current angle, γ , on the order of the dominant torque producing MMF harmonic component will be explored. This is carried out by simply obtaining the phasor summation of the MMF produced by each three-phase winding sets under a general current angle γ . It has been proven in [25] that the flux distribution of a 9P6T connection depends on the current phase shift angle γ . This will be proven mathematically for the proposed winding layout in this subsection.

According to the winding function theory, the MMF harmonic of order h produced by a three-phase winding set, for example, s_{147} , assuming balanced three-phase stator currents with a current magnitude I_m , is given by

$$\bar{F}_{s_{147}}^h = \frac{3N_c I_m K_w^h}{\pi h} e^{jh\theta} \cdot e^{-j\omega t} \quad (2)$$

where N_c is the number of turns per coil, θ is the peripheral angle in mechanical degrees, and ω is the stator angular frequency. Since all coils have a coil span of two slots, the winding factor K_w^h of a harmonic of order h is, therefore, calculated as shown below:

$$K_w^h = \sin \frac{h\pi}{9} \quad (3)$$

The same can be written for the winding set s_{369} , hence

$$\bar{F}_{s_{369}}^h = \frac{3N_c I_m K_w^h}{\pi h} e^{jh(\theta + (4\pi/9))} \cdot e^{-j(\omega t + \gamma)} \quad (4)$$

Since the number of turns per coil for the winding group s_{258} is $0.532N_c$, the MMF harmonic of order h produced by this winding will be written as

$$\bar{F}_{s_{258}}^h = -\frac{3.192N_c I_m K_w^h}{\pi h} \cos \frac{\gamma}{2} \cdot e^{jh(\theta + (2\pi/9))} \cdot e^{-j(\omega t + (\gamma/2))} \quad (5)$$

The negative sign of the MMF phasor $\bar{F}_{s_{258}}^h$ in (5) is used because this winding set is reversed with respect to the other two groups, as shown in Fig. 2a.

The total MMF magnitude of the harmonic component h is then given by

$$\bar{F}_s^h = \bar{F}_{s_{147}}^h + \bar{F}_{s_{369}}^h + \bar{F}_{s_{258}}^h \quad (6)$$

which reduces to (7) as follows:

$$\begin{aligned} \bar{F}_s^h &= \frac{3N_c I_m K_w^h}{\pi h} \cdot e^{jh(\theta - \omega t)} \\ &\times \left(1 + e^{j(h(4\pi/9) - \gamma)} - 1.064 \cos \frac{\gamma}{2} e^{j(h(2\pi/9) - (\gamma/2))} \right) \end{aligned} \quad (7)$$

From (7), and for a current angle $\gamma = -40^\circ$, the harmonic order $h = 4$ dominates (main torque producing component) and can be written as

$$\bar{F}_s^4 = \frac{9N_c I_m}{4\pi} \sin \frac{4\pi}{9} \cdot e^{j4(\theta - \omega t)} \quad (8)$$

This simply represents the 8-pole, or the high-pole, mode of operation, which is represented by the case shown in Fig. 3. This total component is also three times the MMF component of each separate three-phase group that is given by (2).

From (7), the 8-pole MMF harmonic component, $\bar{F}_s^4$, can completely be eliminated for a current angle $\gamma = -206.16^\circ$. In this case, the dominant harmonic will, however, correspond to $h = -2$. This case simply represents the 4-pole, or the low-pole, mode of operation and is represented by (9). The negative sign used with this MMF harmonic order stands for the direction of rotation, which is basically opposite to the 8-pole component

$$\bar{F}_s^2 = \frac{3.12N_c I_m}{\pi} \sin \frac{2\pi}{9} \cdot e^{-j2(\theta - \omega t)} \quad (9)$$

Fig. 4 shows the MMF distribution for this excitation case, which clearly exhibits a 4-pole flux distribution. For this phase shift angle, the current relation between different three-phase current sets is given by

$$\begin{aligned} \bar{I}_{s_{258}} &= \bar{I}_{s_{147}} + \bar{I}_{s_{369}} = I_m \angle 0^\circ + I_m \angle -206.16^\circ \\ &= 0.4526 I_m \angle 76.92^\circ \end{aligned} \quad (10)$$

Under this case, the MMF produced by the winding set s_{258} will significantly be reduced to only 24% of the other winding groups. Hence, the total MMF will be dominated by the other two winding groups, which emulates winding reconfiguration of conventional pole-changing techniques. The lowest order harmonics for this case are the 8th and 10th.

The ratio between the MMF torque producing components in both modes can be found from (11) using (8) and (9) by assuming the same magnetising current

$$\frac{|\bar{F}_s^2|}{|\bar{F}_s^4|} = \frac{3.12}{2.25} \times \frac{\sin 40^\circ}{\sin 80^\circ} = 0.905 \quad (11)$$

Comparing the flux distributions shown in Figs. 3 and 4, the following conclusions can be made:

- For the given 18-slot stator equipped with a 9P6T winding, the equivalent number of poles can be changed from 8-pole to 4-pole by simply changing the current phase shift from 40° to 206.16° . However, the flux waveforms will be travelling in opposite directions.
- The amplitude of the dominant MMF component under the 4-pole mode equals 91% of the 8-pole mode for the same magnetising current magnitude, as in (11).
- The flux distribution under the 4-pole mode is better than the 8-pole mode. However, for the latter case, the dominant harmonics are non-integer multiples of the fundamental torque producing MMF component. Hence, proper rotor design will be imperative for an acceptable operation under both modes.

3 Phase voltage magnitude under different modes

In this section, the voltage magnitude relation between different modes as well as the machine magnetising inductance will be presented. The provided analysis will be vital for the machine design phase. For simplicity, the stator voltage drop will be neglected.

3.1 Terminal phase voltage magnitude

Since the machine originally has nine phases, the terminal phase voltage can be derived from the phase voltages of the original nine-phase winding. The terminal phase voltage can be calculated using the phases s_1 , s_2 , and s_3 (first phase in each group). Since the voltage induced across these phases will be proportional to the number of turns for each specific phase, thus, by taking phase s_1 as

Fig. 3 MMF distribution and spectrum for the 8-pole mode
(a) Phasor diagram, (b) MMF distribution, (c) MMF spectrum

Fig. 4 MMF distribution and spectrum for the 4-pole mode
(a) Phasor diagram, (b) MMF distribution, (c) MMF spectrum

a reference, the induced voltages across these phases can be written as follows:

$$\vec{V}_{s1}^h = V_{ph}^h e^{j0} \quad (12)$$

$$\vec{V}_{s2}^h = -0.532 V_{ph}^h e^{-j(2h\pi/9)} \quad (13)$$

$$\vec{V}_{s3}^h = V_{ph}^h e^{-j(4h\pi/9)} \quad (14)$$

where $h=4$ and -2 , and V_{ph}^h is the induced voltage due to a harmonic h .

Using the first phase of each group, the applied inverter input voltages can simply be found from the following equation:

$$\vec{V}_{a1}^h = \vec{V}_{s1}^h + \vec{V}_{s2}^h \text{ and } \vec{V}_{a2}^h = \vec{V}_{s3}^h + \vec{V}_{s2}^h \quad (15)$$

Substituting from (12)–(14) into (15), the voltage relations for both cases are given by (16) and (17), respectively

$$\vec{V}_{a1}^4 = 1.511 V_{ph}^4 \angle 6.92^\circ, \quad \vec{V}_{a2}^4 = 1.511 V_{ph}^4 \angle 33.08^\circ \quad (16)$$

$$\vec{V}_{a1}^2 = 1.048 V_{ph}^2 \angle -30^\circ, \quad \vec{V}_{a2}^2 = 1.048 V_{ph}^2 \angle -170^\circ \quad (17)$$

It can be concluded from (16) and (17) that the required voltage phase shift between the two three-phase winding sets will be 26.16° and 140° for the 8-pole and 4-pole modes, respectively. While, the terminal phase voltage is 1.511 and 1.048 times the winding phase voltage, respectively.

It is now required to determine the relationship between the magnitudes of the phase voltages in both cases that ensure the same no-load magnetising current. This will help determine the required winding voltage rating. Starting from the winding phase voltage, V_{ph}^h , one can write

$$\frac{V_{ph}^4}{V_{ph}^2} = \frac{N_{ph}^4}{N_{ph}^2} \cdot \frac{\phi^4}{\phi^2} \cdot \frac{K_w^4}{K_w^2} = \frac{N_{ph}^4}{N_{ph}^2} \cdot \frac{B_g^4}{B_g^2} \cdot \frac{p^2}{p^4} \cdot \frac{K_w^4}{K_w^2} \quad (18)$$

where N_{ph}^h is the number of turns per phase, ϕ^h is the average flux per pole, B_g^h is flux density per pole, and p^h is the number of pole pairs.

The number of turns per phase will be the same for the two modes, while the winding factor, K_w^h , in both cases is calculated from (3) as 0.9848 and 0.6427 for the 8-pole and 4-pole modes, respectively. By neglecting the effect of saturation, the air gap flux density ratio can be approximated by the ratio between the torque producing MMF components in both cases given by (11) as shown in the following equation:

$$\frac{V_{ph}^4}{V_{ph}^2} = \frac{1}{0.905} \cdot \frac{2}{4} \cdot \frac{0.9848}{0.6427} = 0.847 \quad (19)$$

Based on (16) and (17), the relation between the supply phase voltages, V_{supply}^k , will therefore be that shown in the following equation:

$$\frac{V_{supply}^4}{V_{supply}^2} = 0.847 \cdot \frac{1.511}{1.048} = 1.22 \quad (20)$$

This ratio, of course, represents the ratio between the induced emf voltages and neglects the effect of the stator voltage drop; however, it can be used as a general guide in the design phase. For the 8-pole mode, the stator leakage inductance is expected to be higher due to the presence of the non-negligible 14th-order harmonic. Therefore, the actual voltage ratio is expected to be higher than the calculated value in (20).

3.2 Magnetising inductance under different modes

It is also valuable to show the effect of the selected mode on the machine equivalent magnetising inductance to highlight the advantage of this winding over conventional PPM, and PAM modulated multiphase IMs with conventional distributed windings. The ratio between the magnetising inductances under the two modes can simply be calculated from the ratio between the induced back emf and the line magnetising current. Since the voltage ratio given by (20) is calculated assuming the same magnetising current, the same ratio will represent the ratio between the magnetising inductances of both modes, hence

$$L_m^4 = 1.22 L_m^2 \quad (21)$$

Interestingly enough, for the proposed winding, the inductance of the 8-pole mode is 22% higher than the 4-pole mode. In contrast to this, the magnetising inductance of conventional multiphase

Fig. 5 Flux density distributions

(a) 8-pole mode, (b) 4-pole mode, (c) Rotor-teeth flux density – 8-pole mode, (d) Rotor-teeth flux density – 4-pole mode

Table 1 Machine dimensions of the 50 kW design example

number of stator slots	18
number of rotor bars	40
outer stator diameter	290 mm
inner stator diameter	182 mm
outer rotor diameter	180.5 mm
air gap length	0.75 mm
axial core length	145 mm
stator back iron width	25 mm
rotor back iron width	25 mm
stator tooth width	15.7 mm
rotor tooth width	8 mm

machines with distributed winding is inversely proportional to the square of the harmonic order [26].

4 50 kW design example

To demonstrate the proposed concept using a practical design example, similar design data for the driving motor used in the 2004 Toyota Prius hybrid EV drive will be employed. Although this car model employs an 8-pole interior PM motor, it has been shown in [27] that a copper cage IM with the same stator design will be a competitive option. A 50 kW IM is, therefore, designed as an 8-pole stator having 18 stator slots, as shown in Fig. 5a. The machine's base speed is 1500 rpm (constant torque range), and a constant power range with field weakening is employed to extend the speed range from 1500 to 6000 rpm. The design steps for conventional induction machine given in [23] are followed, and then, the design is confirmed using JMAG Express software. The machine is first designed assuming an 8-pole nine-phase stator with a phase voltage of 173 V and a magnetising current of 21 A. According to (16), after configuring the stator winding as a 9P6T connection, the input phase voltage will be 262 V.

In 4-pole mode and using the same rated magnetising current of 21 A, the input phase voltage will ideally be reduced to 215 V, by neglecting the stator voltage drop as in (20). According to the FE simulation, the same magnetising current is obtained for a phase voltage of only 200 V. This is because the stator voltage drop of the 4-pole mode is relatively lower than the 8-pole mode. This stems from the fact that the 8-pole mode corresponds to higher low-order

space harmonics, and hence, a higher leakage inductance will be expected. For the same air gap flux magnitude under both modes, the stator and rotor back iron fluxes will be higher under the 4-pole (low-pole) mode. Hence, the stator, as well as the rotor back iron thicknesses, should be designed based on the 4-pole flux to avoid excessive core saturation [18]. The air gap space harmonics will also give rise to undesirable torque ripple components. A rotor bar skew angle of 1 rotor slot is found suitable in this design to relatively reduce the induced torque ripple magnitude.

The final machine dimensions and specifications are given in Tables 1 and 2, respectively. The transient model developed with JMAG Studio is used to simulate the machine torque–speed curves under both modes. The machine flux distribution, as well as the rotor teeth flux density distributions for both modes at rated conditions, are shown in Fig. 5. Clearly, the peak flux density is approximately the same in both modes, and the core flux distributions do not exceed the recommended standard steel saturation level.

For the same rated output power, the machine efficiency values will be closed under both modes. This conclusion neglects the effect of core loss, which is ignored in this simulation study. The machine power factor is relatively low under the 8-pole mode, which is logical for this high pole order mode [23]. However, under the 4-pole mode, the power factor is significantly improved.

The machine equivalent parameters under both modes are estimated based on the simulated characteristic curves and are given in Table 3. These estimated parameters at a rated frequency of 100 Hz are used to compare the calculated machine

Table 2 Machine rated values of the 50 kW design example

Mode	8-pole	4-pole
phase voltage, V	262	200
rated current, A	52	52
rated speed, rpm	1458	2974
efficiency, %	93.7	92.7
power factor	0.66	0.87

Table 3 Estimated machine parameters under different modes

8-pole	$r_s^4 = 0.119 \Omega$, $l_s^4 = 2.41 \text{ mH}$, $L_m^4 = 18.77 \text{ mH}$, $r_r^4 = 0.127 \Omega$, $l_r^4 = 2.12 \text{ mH}$
4-pole	$r_s^2 = 0.083 \Omega$, $l_s^2 = 0.467 \text{ mH}$, $L_m^2 = 15.3 \text{ mH}$, $r_r^2 = 0.037 \Omega$, $l_r^2 = 0.467 \text{ mH}$

Fig. 6 Comparison between calculated characteristic curves and FE simulation results
(a) 8-pole mode, (b) 4-pole mode

characteristic curves with the FE simulation results. This comparison is shown in Fig. 6. From Table 3, it can be concluded that the rotor impedance in the 4-pole mode is much lower (approximately one-fourth) than the 8-pole rotor impedance. This is due to the relative difference in the equivalent winding factor between both modes. Of course, this ratio will also depend on the rotor cage design, including the end ring dimensions.

Based on the estimated machine parameters, the machine characteristic curves under different frequencies are given in Fig. 7. Clearly, the machine can successfully supply rated power output over the full speed range of 4 pu without the need for oversizing the propulsion motor, as clarified in the introduction section. For speeds below rated speed ($f < \text{base frequency}, f_b$), the 8-pole mode is employed with constant V/f control. To increase the speed above rated speed, field weakening is then applied. The switching between the two modes will be based on the frequency at which the corresponding torque-speed curve will be asymptotic to the constant power locus. Based on the estimated machine parameters, this switching frequency value was approximately found 1.5 pu. For higher frequencies, the 4-pole mode is applied with a constant V/f control up to rated frequency, f_b , which corresponds to a synchronous speed of 2 pu. Field weakening is then applied till the maximum machine torque asymptotes the constant power loci. With the designed machine, the speed can clearly be extended above 4 pu speed since maximum torque is not reached. The maximum speed of the constant power region corresponds to the frequency which yields rated output power at the maximum torque point of the torque-speed curve corresponding to this frequency. This value is found to be 7 pu for the designed machine. However,

many other aspects will of course limit this maximum reachable speed. Ideally, the maximum torque for both modes will preferably be equal, and a switching speed point of 2 pu between the two modes is favoured, which entails an optimal machine design case study since these values will highly depend on the cage rotor design. Improving the air gap flux distribution will also reduce the stator leakage inductance yielding a higher maximum torque under the 8-pole mode. This design case study will be postponed to a future study.

The rated torque profiles under both modes are also simulated and shown in Fig. 8. For the selected skew angle, the torque ripple is $< 8\%$. This relatively large value represents a notable challenge and needs further investigation to come up with a 9P6T FSCW design with better quality flux distribution, which is also postponed to a future study. Obviously, the frequency of the torque ripple component is different in both modes, which is dependent on the harmonic order of the low-order space harmonics under each case.

5 Machine VSD modelling

To properly control the proposed machine, the proposed PAM-9P6 T winding is considered as an equivalent six-phase machine with two main subspaces, namely, $\alpha - \beta$ and $x - y$ subspaces, as described in [25]. These two subspaces will correspond to the 8-pole and 4-pole modes of operation, as proven in Section 2. The excited subspace will mainly depend on the current angle γ . By exciting the $\alpha - \beta$ subspace with two balanced current components, an 8-pole flux distribution will be obtained, while the 4-pole mode can simply be enabled through exciting the $x - y$ subspace. Since the machine is originally nine-phase, the first step is to derive the

Fig. 7 Characteristic curves over full speed range
(a) Torque–speed curves, (b) Speed–RMS line current curves

Fig. 8 Rated torque waveform under both modes

required VSD transformation that decomposes the six terminal currents/voltages into their equivalent sequence components. The equivalent six-phase VSD-based model is then derived. This model assumes the following assumptions:

- Linear magnetic circuit is assumed.
- The core losses are neglected.
- Dominant harmonic for each subspace is only considered, while higher order space harmonics are neglected.

Since the relationship between currents of different winding sets is governed by the connection of different phases irrespective of the winding layout itself, the same current connection matrix, $[C_i]$, derived in [25] for a distributed winding relating the phase and line currents $[i_{ph}^s] = [C_i][i_l^s]$ and given by (22), can simply be used here

$$[C_i] = \begin{bmatrix} 1 & n & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & n & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 1 & n & 0 \\ 0 & n & 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & n & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 & 0 & n & 1 \end{bmatrix} \quad (22)$$

where $n = 0.532$, $[i_{ph}^s] = [i_{s1} \ i'_{s2} \ i_{s3} \ i_{s4} \ i'_{s5} \ i_{s6} \ i_{s7} \ i'_{s8} \ i_{s9}]^t$ and $[i_l^s] = [i_{a1} \ i_{b1} \ i_{c1} \ i_{a2} \ i_{b2} \ i_{c2}]^t$ are the phase and line current vectors referred to the supply side, respectively. The superscript ‘ $'$ ’ indicates a variable reduced to the supply side.

The terminal phase voltage vector can also be obtained from the phase voltage vector $[v_{ph}^s] = [C_v][v_{ph}^s]$ using the voltage connection matrix $[C_v] = [C_i]^t$, as detailed in [25], where $[v_{ph}^s] = [v_{s1} \ v'_{s2} \ v_{s3} \ v_{s4} \ v'_{s5} \ v_{s6} \ v_{s7} \ v'_{s8} \ v_{s9}]^t$ and $[v_{inv}^s] = [v_{a1} \ v_{b1} \ v_{c1} \ v_{a2} \ v_{b2} \ v_{c2}]^t$.

Since the 9P6T connection emulates a six-phase winding with a single neutral connection, there will be two additional zero sequence components. These extra components will add to zero during healthy operation.

It has been shown in [25] that the required current and voltage VSD transformations that decompose the terminal variables of a 9P6T connection into their equivalent six sequence components are given by (23) and (24), respectively. Unlike the conventional multiphase Clarke's transformation, the voltage and current transformations for this connection are clearly dissimilar

$$[T_{6i}] = \begin{bmatrix} 0.333 & -0.202 & -0.132 & 0.281 & -0.299 & 0.018 \\ 0.040 & 0.269 & -0.309 & 0.183 & 0.152 & -0.335 \\ 0.333 & -0.312 & -0.202 & -0.281 & 0.299 & -0.018 \\ 0.040 & -0.309 & 0.269 & 0.183 & 0.152 & -0.335 \\ 0.281 & 0.281 & 0.281 & -0.052 & -0.052 & -0.052 \\ 0.177 & 0.177 & 0.177 & 0.511 & 0.511 & 0.511 \end{bmatrix} \quad (23)$$

$$[T_{6v}] = \begin{bmatrix} 0.336 & -0.168 & -0.168 & 0.257 & -0.316 & 0.058 \\ 0 & 0.291 & -0.291 & 0.216 & 0.115 & -0.331 \\ 0.336 & -0.168 & -0.168 & -0.301 & 0.280 & 0.023 \\ 0 & -0.291 & 0.291 & 0.148 & 0.187 & -0.335 \\ 0.374 & 0.374 & 0.374 & -0.130 & -0.130 & -0.130 \\ 0.038 & 0.038 & 0.038 & 0.206 & 0.206 & 0.206 \end{bmatrix} \quad (24)$$

From the inverse of (23), balanced $\alpha - \beta$ current components will produce two three-phase current sets (abc_1 and abc_2) with a phase shift angle of -40° . The phase shift angle will, however, be -206.16° for an $x - y$ excitation, which complies with the discussion made in Section 2. It can also be shown that the inverse

Fig. 9 Proposed VSD-based controller

of (24) will give two balanced three-phase voltage groups (abc_1 and abc_2) with voltage phase shift angles of 26.16° and -140° under balanced $\alpha - \beta$ and $x - y$ excitations, respectively.

The machine voltage equations of the $\alpha - \beta$ subspace can be written as

$$\begin{aligned} v_{s\alpha} &= r_s^4 i_{s\alpha} + \frac{d}{dt} \lambda_{s\alpha} \\ v_{s\beta} &= r_s^4 i_{s\beta} + \frac{d}{dt} \lambda_{s\beta} \\ 0 &= r_r^4 i_{ra} + \frac{d}{dt} \lambda_{ra} - p^4 \omega_m \lambda_{r\beta} \\ 0 &= r_r^4 i_{r\beta} + \frac{d}{dt} \lambda_{r\beta} + p^4 \omega_m \lambda_{ra} \end{aligned} \quad (25)$$

and the flux linkage equations are given by

$$\begin{aligned} \lambda_{s\alpha} &= (L_s^4 + L_m^4) i_{s\alpha} + L_m^4 i_{ra} \\ \lambda_{s\beta} &= (L_s^4 + L_m^4) i_{s\beta} + L_m^4 i_{r\beta} \\ \lambda_{ra} &= (L_r^4 + L_m^4) i_{ra} + L_m^4 i_{s\alpha} \\ \lambda_{r\beta} &= (L_r^4 + L_m^4) i_{r\beta} + L_m^4 i_{s\beta} \end{aligned} \quad (26)$$

The voltage equation for the $x - y$ subspace is shown in the following equation:

$$\begin{aligned} v_{sx} &= r_s^2 i_{sx} + \frac{d}{dt} \lambda_{sx} \\ v_{sy} &= r_s^2 i_{sy} + \frac{d}{dt} \lambda_{sy} \\ 0 &= r_r^2 i_{rx} + \frac{d}{dt} \lambda_{rx} - p^2 \omega_m \lambda_{ry} \\ 0 &= r_r^2 i_{ry} + \frac{d}{dt} \lambda_{ry} + p^2 \omega_m \lambda_{rx} \end{aligned} \quad (27)$$

and the flux linkage equations are written as

$$\begin{aligned} \lambda_{sx} &= (L_s^2 + L_m^2) i_{sx} + L_m^2 i_{rx} \\ \lambda_{sy} &= (L_s^2 + L_m^2) i_{sy} + L_m^2 i_{ry} \\ \lambda_{rx} &= (L_r^2 + L_m^2) i_{rx} + L_m^2 i_{sx} \\ \lambda_{ry} &= (L_r^2 + L_m^2) i_{ry} + L_m^2 i_{sy} \end{aligned} \quad (28)$$

Unlike conventional multiphase machines [25], both the $\alpha - \beta$ and $x - y$ subspaces will be complementarily contributing to the torque production. The developed torques of both subspaces are given by (29) and (30), respectively

$$T_{\alpha\beta} = 3p^4 L_m^4 (i_{s\beta} i_{ra} - i_{s\alpha} i_{r\beta}) \quad (29)$$

$$T_{xy} = 3p^2 L_m^2 (i_{sy} i_{rx} - i_{sx} i_{ry}) \quad (30)$$

6 System control strategy

As clarified in Section 5, the phase variables of the proposed PAM-9P6T IM can be decomposed into two decoupled subspaces, where each subspace will correspond to one of the operational modes based on the desired reference speed value. In this section, the required controller to run the machine over the full speed range is introduced. The proposed controller is basically extracted from the VSD-based controller initially proposed for the 9P6T IM [25], but with some introduced modifications to allow proper pole-switching based on the desired reference speed.

As explained in [18], a machine with a 2:1 pole ratio can ideally be used to obtain a 4:1 field weakening range. Rated torque output can be maintained up to base speed (1 pu) under the 8-pole (high-pole) mode. For this speed range, the voltage increases linearly with speed until reaching its rated value (1 pu) at rated speed (1 pu) allowing for a constant air gap flux magnitude. Above rated speed, the machine operates under constant voltage and power by decreasing the reference air gap flux level to limit the machine voltage at its maximum allowable magnitude. The 8-pole mode will ideally be maintained while field weakening is applied up to double the base speed (2 pu). Above double the base speed, the 4-pole mode is then employed. Under this mode, field weakening will be maintained by reducing the direct current component in conjunction with the rotor speed to limit the voltage within its maximum value. For the same direct current component, the phase voltage magnitude will approximately drop by 18% at the same frequency, as indicated by (20), when switching from the 8-pole mode to the 4-pole mode. This additional voltage reserve will allow for an extended speed range operation above 4 pu. This may be considered as another notable difference over the topology introduced in [18].

Fig. 9 shows the proposed controller, which employs standard indirect rotor field-oriented control to generate the reference flux/torque producing ($\delta\gamma$) current components. Based on the operational mode, these current components are applied to the corresponding subspace, while the reference current components for the other subspace are set to zero. The mode selection block in Fig. 9 is used to determine the intended operational mode based on the rotor speed, namely, the 8-pole mode is enabled for speed range up to 2 pu, while the 4-pole mode is employed for higher rotor speeds.

The reference direct current component, i_d^* , is set to the rated magnetising current component for speeds up to rated speed (1 pu). For speed >1 pu, field weakening is applied and the reference direct current component, i_d^* , is, therefore, calculated based on the operational mode as well as the rotor speed. The mode selection

Table 4 Prototype machine ratings

16-pole mode		8-pole mode	
phase voltage	42 V	phase voltage	30 V
RMS line current	3.5 A	RMS line current	3.5 A
rated speed	450	rated speed	900

Fig. 10 Experimental setup **Fig. 11** Simulated torque–speed characteristic for the prototype machine in per unit values

block also calculates the rotor time constant $\tau_r = L_r^h / r_r^h$, which is used to calculate the rotor slip frequency, ω_{slip} . Each subspace is controlled using a pair of proportional–integral (PI) controllers. The 6-line currents are decomposed into their sequence components using (23). The four PI controllers are then used to regulate these four current components and generate the corresponding stator voltage components. The inverse voltage transformation given by (24) is finally applied to calculate the applied terminal voltages.

7 Experimental results

7.1 Experimental setup

In this paper, an existing 1 Hp, 60 Hz, 4-pole three-phase IM with 36 stator slots and 28 rotor bars is rewound using the proposed winding layout and used to validate the proposed concept. After rewinding the existing stator, the number of turns per coil for the winding groups s_{147} and s_{369} was 36 turns. The number of turns of the winding group s_{258} was selected to be $0.532 \times 36 = 19$ turns, as depicted by (1). With the existing stator, two slot/pole combinations will be obtained, namely, 36-slot/16-pole and 36-slot/8-pole for current shift angles of 40° and 206.16° , respectively. In the following, these different modes are called 16-pole and 8-pole modes of operation.

The original machine was a 4-pole motor. After stator rewinding, it became a 16-pole machine with a rated synchronous speed of 450 rpm at 60 Hz. This results in a significant reduction in machine output power for the same copper volume. With the machine designed to carry same rated current, approximate rated

Table 5 Estimated prototype machine parameters

16-pole	$r_s^4 = 2.64 \Omega$, $l_s^4 = 8.5 \text{ mH}$, $L_m^4 = 15.4 \text{ mH}$, $r_r^4 = 1.53 \Omega$, $l_r^4 = 8.5 \text{ mH}$
8-pole	$r_s^2 = 1.84 \Omega$, $l_s^2 = 4.3 \text{ mH}$, $L_m^2 = 13.4 \text{ mH}$, $r_r^2 = 0.97 \Omega$, $l_r^2 = 4.3 \text{ mH}$

torque can still be achieved, which may be enough to prove the proposed concept. The new rated voltages under both modes, which yield the same rated current of 3.2 A and same output power at rated frequency, are experimentally determined and given in Table 4. It is clear from Table 4 that the relation between the rated phase voltages for both modes is ~ 1.4 , which is higher than the value given by (20) due to the effect of the stator voltage drop, as explained before. Since an existing machine frame is used to verify the presented concept, machine optimisation was not possible at this stage.

The machine is fed from two three-phase pulse width modulated inverters connected to the same dc link. The switching frequency is 5 kHz. The IM is mechanically loaded using a coupled PM generator connected to load bank resistors. The controller is implemented using a digital signal processor board eZdsp™ based on Texas Instruments F28335 DSP while a CAN bus is used for online measurements. A pulley with relatively high inertia is mounted on the rotor to emulate the vehicle inertial load. The whole machine setup is shown in Fig. 10.

The machine parameters of both subspaces are simply estimated by employing standard blocked rotor and no-load tests for each subspace separately, while the current components of the other subspace are controlled to zero [28]. The estimated parameters are given in Table 5.

The machine free-running starting is simulated for both operational modes using FE simulations to estimate the corresponding torque–speed curves under rated and double rated frequency. The simulated torque–speed curves in per unit are shown in Fig. 11. Obviously, the machine can maintain rated output power for a speed range above 4 pu with a pole-switching speed point at 2 pu. The maximum achievable speed will be experimentally determined in the following section.

7.2 Experimental results

The proposed system is investigated under different cases using speed control mode. First, each mode is investigated separately by disabling the mode selection block in Fig. 9, while the required mode is set manually. Under speed control mode, the reference quadrature current component, i_q^* , is derived from the speed error using a PI controller. The reference direct current component, i_d^* , is set to the rated magnetising current value.

The 16-pole (high-pole) mode is first enabled, and the reference dq current components are used to control the $\alpha - \beta$ subspace, while the reference current components of the $x - y$ subspace are set to zero. The reference speed is initially set to 200 rpm, then it is step increased to 450 rpm (rated synchronous speed at 60 Hz for the 16-pole mode). The speed is then brought back to its initial value of 200 rpm. This speed profile is repeated multiple times to investigate the machine dynamic performance. Fig. 12 shows the rotor speed, the reference and actual dq current components, and the applied phase voltage magnitude in per unit.

The same previous test is repeated under the 8-pole (low-pole) mode using the same reference direct current component. However, the reference speed is increased from 200 to 900 rpm (the synchronous speed at 60 Hz for the 8-pole mode). The machine dynamic performance is shown in Fig. 13. By comparing the two cases, the phase voltage magnitude under the 8-pole mode is ~ 0.8 times the voltage magnitude under the 16-pole mode for the same

Fig. 12 Machine dynamic response under the 16-pole mode
(a) Speed profile, (b) Reference and actual current components, (c) RMS phase voltage profile in per unit

Fig. 13 Machine dynamic response under the 8-pole mode
(a) Speed profile, (b) Reference and actual current components, (c) RMS phase voltage profile in per unit

Fig. 14 Transition dynamics between both modes
(a) Speed, (b) dq current components, (c) RMS phase voltage magnitude in per unit

direct current component and stator frequency, which complies with (20).

Upon pole switching, the air-gap field comprises two flux components. The first component is the decaying flux component of the outgoing pole number sustained by the cage rotor currents. While the second component is the flux component of the incoming pole number that is excited by the stator currents. The transient period will, therefore, experience a transient loss of the controller current regulation, which yields uncontrolled torque transients. This problem has been explained in [18]. To investigate this effect, the controller's dynamic pole-switching response is shown in Fig. 14 while operating at a constant speed of 450 rpm. First, the selected mode is set to the 8-pole (low-pole) mode. The 16-pole (high-pole) mode is then enabled for ~ 3 s. Then, the operating mode is brought back to the 8-pole mode, while maintaining the same direct current component during the whole period. As explained in [18], although the torque significantly drops due to current sequence reversal, this has a minimal effect on

the speed due to the relatively high motor and load equivalent inertia in EV applications. The most notable change appears in the RMS voltage magnitude, where the voltage is increased for the same speed and reference current components under the 16-pole mode.

The steady-state current waveforms of phases a_1 and a_2 are shown in Figs. 15a and b for the cases given in Figs. 12 and 13, respectively. Clearly, both cases correspond to approximately the same frequency although the speed is doubled, while the current phase shift angle between these two currents are changed from 40° to 206.16° . For the same reference speed of 450 rpm, the effect of the switching transition between the 16-pole mode to the 8-pole mode on the phase current is shown in Fig. 15c. For this latter case, the frequency is halved with pole number reduction.

To investigate machine operation under the whole speed range, the controller is tested under torque mode by applying a constant quadrature current component, $i_q^* = 2$ A, while the mode selection block is enabled. The direct current component, i_d^* , is, therefore,

Fig. 15 Phase current waveform

(a) 16-pole mode, (b) 8-pole mode, (c) Transition between 16-pole to 8-pole mode at the same reference speed of 450 rpm

Fig. 16 Machine acceleration under torque mode

(a) Speed profile, (b) dq current components

defined based on the rotor speed. The quadrature current component is reduced to zero when the phase voltage magnitude reaches rated voltage (1 pu). Fig. 16 shows the machine acceleration and current components profiles under the described torque mode. Starting from zero speed, the 16-pole mode is initially enabled with a constant direct current component until a rated speed of 450 rpm (1 pu) is reached. Field weakening is then enabled to dictate the required direct current component for a speed range from 450 to 900 rpm (2 pu). When the rotor speed reaches 900 rpm, the 8-pole mode is alternatively enabled after ~ 7 s. Since this speed threshold value represents the base speed for the 8-pole mode, the controller suddenly changes the reference direct current component to the rated magnetising current to maximise the torque production. This yields a sudden dip in the quadrature current profile due to this sudden pole switching. For speeds higher than 900 rpm, the controller continues to run the machine under field weakening till the phase voltage magnitude reaches its ceiling value. The maximum achievable speed is found to be 2340 rpm (5.2 pu). This complies with the conclusion made before based on (20). The quadrature current component is then brought to zero, and the machine acceleration is stopped. The rotor, therefore, continues to run at this final speed value thanks to the high inertial load connected to the rotor.

8 Conclusion

This paper proposed a new winding layout for pole-amplitude modulated six-phase IMs, which offers a 2:1 speed ratio while maintaining the same magnetising current under both pole pairs. The pole switching is simply done by changing the current phase shift angle between the two three-phase supply currents. The required applied inverter voltages to ensure a desired current phase shift angle were calculated and the corresponding voltage magnitude ratio was decided. The theoretical concept was presented and supported using FE simulation results of a designed 50 kW machine. Although the simulated torque–speed curves

under the full speed range proved that the proposed winding layout can attain a continuous wide speed range without oversizing the driving motor, an optimisation design case study is still be needed. Machine modelling based on VSD was also presented and used to control the machine over the full speed range. The experimental investigation using a small-scale prototype machine with the proposed VSD-based controller showed that the proposed PAM-9P6T machine can achieve constant power operation up to an approximate maximum speed of 5 pu. Although the study has been done using a normal inner rotor configuration, the proposed winding layout can also be extended to outer rotor IMs, which is preferred in direct drive applications.

The main technical challenge which entails another further future investigation is the high harmonic content of the proposed PAM-9P6T based on FSCW. Optimal rotor design is, therefore, crucial to optimise the machine torque production. Future work is intended to propose improved winding layouts based on the 9P6T connection with better air gap flux distribution.

9 References

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