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Analytical Modeling of 12/10 stator/rotor pole configuration of Switched Reluctance Motors Considering Magnetic Saturation for Blood Pump Applications

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Abstract— This paper presents a comparative analytical study of torque ripple and radial force characteristics in 3-, 4-, and 6-phase switched reluctance motors (SRMs) with a 12/10 stator/rotor pole configuration, accounting for magnetic saturation effects. While multi-phase SRMs are generally expected to offer improved torque smoothness over conventional 3-phase designs, a systematic comparison of their torque ripple and radial force behavior particularly under saturated conditions has not been thoroughly investigated in the literature. To address this gap, a unified two-dimensional analytical model is developed based on Fourier series, incorporating saturation effects through a modified flux-linkage characteristic. The model predicts electromagnetic torque and radial force as functions of rotor position and phase current for all three phase configurations, and is validated against finite element simulations. The comparative results reveal distinct performance trends across the different phase numbers, highlighting trade-offs between torque ripple suppression, radial force magnitude, and force ripple. A key contribution of this work is linking radial force behavior to bearingless motor technology, where unbalanced radial forces directly affect rotor suspension stability and increase the control effort required for magnetic levitation. This is particularly critical for blood pump applications, where mechanical oscillations and bearing friction impose excessive mechanical stress on blood cells, increasing the risk of hemolysis. By identifying the optimal phase configuration that minimizes both torque ripple and radial force fluctuations, this study provides a foundation for designing quieter, more reliable bearingless drives with reduced mechanical stress and improved hemocompatibility.

Index Terms— Switched Reluctance Motor, Torque Ripple, Radial Force, Magnetic Saturation, Analytical Modeling, Blood Pump

Superscripts

r	Radial direction
θ	Tangential direction
z	Axial direction
i	Slot number (stator and rotor)
n, m, u	Harmonic Order

I. INTRODUCTION

Over the past two decades, blood pumps as shown in Fig.1 have evolved from experimental devices into established therapeutic options for patients with end-stage heart failure. Initially used as a bridge to transplantation, they are now increasingly considered for long-term destination therapy. However, the clinical success of these devices depends on

more than their hydraulic performance. A growing body of evidence indicates that the mechanical interaction between the pump and blood components plays a critical role in long-term patient outcomes. In particular, rotor vibration, torque ripple, and bearing friction can subject blood cells to excessive mechanical stress, leading to hemolysis and thrombosis. These complications not only affect device reliability but also significantly impact patient well-being and daily quality of life. Figure 1 illustrates a simplified schematic of a centrifugal blood pump, consisting of three primary components: the inlet, the impeller, and the outlet. Blood enters the pump through the central inlet and is directed toward the rotating impeller. The impeller, driven by an electric motor, accelerates the fluid radially outward by means of centrifugal force, thereby increasing both pressure and flow rate. The outlet subsequently guides the pressurized blood toward the circulatory system. The geometric design of the impeller blades plays a critical role in minimizing excessive shear stress on red blood cells, which is essential for reducing the risk of hemolysis. Furthermore, smooth motor operation with minimized torque ripple and mechanical vibration is equally important to ensure stable pump performance and long-term device reliability.

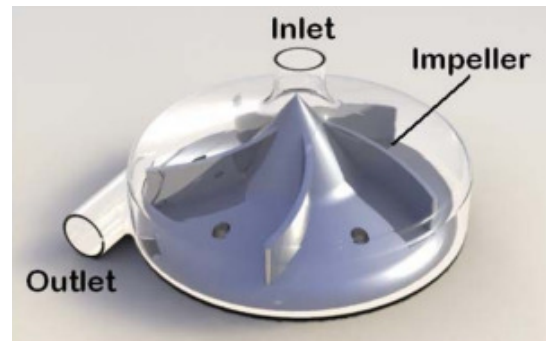


Fig.1. Centrifugal blood pump [1]

Mechanical vibration and torque fluctuations are widely recognized as primary contributors to blood trauma in implantable rotary blood pumps. The choice of motor topology therefore becomes a critical factor in device design. SRMs as shown in Figure 2, present a particularly suitable option for such applications, owing to their inherently robust construction and the absence of permanent magnets or brushes in the rotor. This structural simplicity not only enhances mechanical reliability but also facilitates thermal management, as the primary heat sources the stator windings

are physically separated from the rotating assembly. More importantly, SRMs offer a high degree of controllability over phase currents, enabling effective suppression of torque ripple and radial forces through tailored excitation schemes. This feature is of particular clinical significance, as smoother motor operation directly correlates with reduced mechanical stress on blood cells, thereby lowering the risk of hemolysis and improving long-term device safety and performance which in Figure 3, coupling between SRM shaft and impeller is demonstrated.

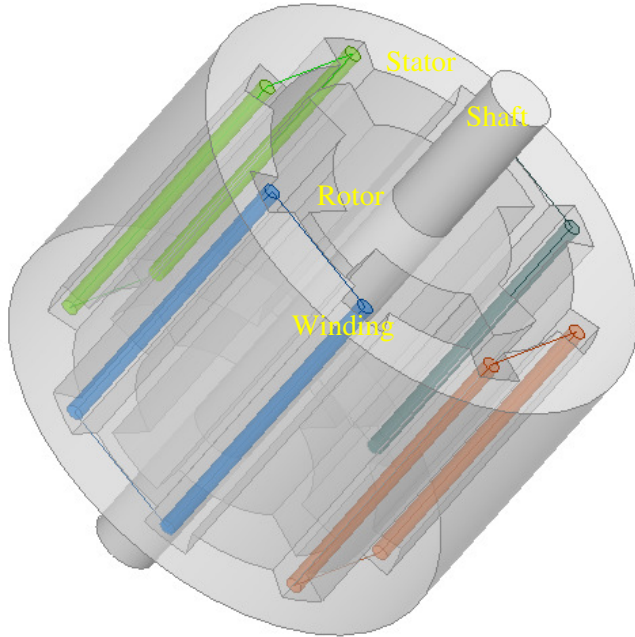


Fig.2. Schematic of SRM

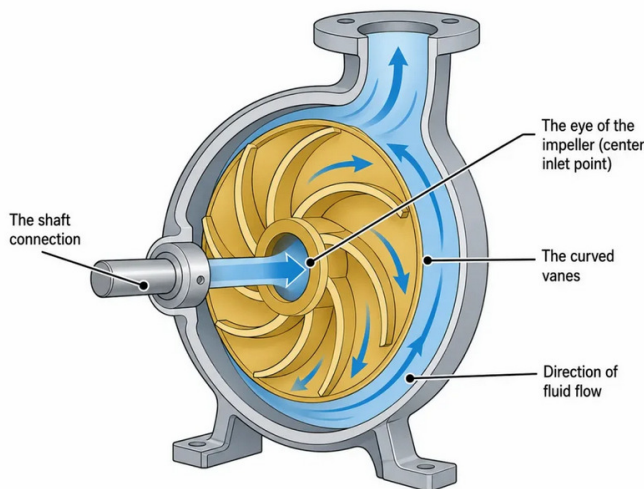


Fig.3. Centrifugal pump coupling with SRM shaft [2]

The literature on SRMs covers a wide range of topics, including analytical modeling, torque ripple reduction, and application-specific designs for automotive, aerospace, and medical sectors. However, a critical review of the existing work reveals that no comprehensive study has yet compared the torque ripple and radial force performance of 3-, 4-, and

6-phase SRMs under saturated conditions, nor has such a comparison been linked to the specific mechanical requirements of bearingless blood pump drives. This paper aims to fill that gap. To put this into perspective, the progress made so far in SRM-based blood pump research is worth reviewing, as in [3], an 18/15/6-pole dual-stator bearingless SRM is introduced, in which the torque and levitation flux paths are spatially separated. This architecture allows for independent and high-precision control of each function, a feature that is particularly attractive for implantable applications where stability and reliability are paramount. Similarly, [4] presents a centrifugal blood pump with a disposable pump head that uses a 12/8 bearingless SRM slice motor. In this design, the impeller and rotor are integrated without permanent magnet, which simplifies the mechanical construction and reduces the cost of single-use components. Other studies have focused on linear topologies and modeling strategies. In [5], for instance, a linear-switched reluctance motor is analyzed, modeled, and controlled for use as a left ventricular assist device. Along the same line, [6] develops a reluctance-network-based linear model of the actuator and then validates it through two-dimensional finite element simulations. Moving toward more complex structures, [7] presents an 18/16-pole permanent-magnet-biased bearingless SRM with radial force windings driven by a three-phase full-bridge inverter, while [8] investigates the electromagnetic characteristics of the suspension force in an axial split-phase bearingless SRM to accurately capture the time-varying behavior of the suspension system. To improve fault tolerance and reliability, [9] proposes a dual-channel bearingless SRM that combines the inherent robustness of SRMs with the levitation capability of magnetic bearings. The rotor poles in this design are scalloped and evenly distributed on both the inner and outer walls of the rotor. Meanwhile, [10] takes a more fundamental approach by comparing various rotor pole configurations. The study concludes that the 12/10 structure offers superior performance among the configurations examined, providing a useful reference for choosing the pole combination. On the control side, [11] introduces an enhanced direct instantaneous torque control (DITC) method based on online optimization of turn-on and turn-off angles, aiming to minimize torque ripple and improve efficiency in six-phase SRMs. In a similar spirit, [12] proposes a simulation-based strategy for determining the optimal angular interval of current supply to reduce torque ripple under motor working conditions. Several contributions have also been made in the area of modeling. In [13], a switched lumped reluctance network is developed to represent an SRM under saturation, accounting for position-dependent switching and relative motion between the rotor and stator. A quasi-3-D nonlinear analytical model is proposed in [14] for axial-flux SRMs, where harmonic modeling is combined with an adaptive convergence algorithm to account for material nonlinearity. In [21], an analytical model for inductance and torque is derived based on 2D finite element data from an 8/6 prototype. A few works have specifically addressed vibration and control complexity. In [15], the mechanical vibration behavior of conventional and mutually-coupled SRMs is compared for a 12/8 structure. In [16], a levitation control strategy is

implemented for a 12/8 bearingless SRM, aiming to maximize radial load capacity while improving torque output through real-time adjustment of the instantaneous torque. The control aspect is further explored in [18] and [19], where a 4-phase double-stator SRM with full-pitch winding is analyzed under saturated and unsaturated conditions and a multi-level direct torque control method is developed with discretized duty cycles. Design considerations are also addressed in [17], where high-speed flux-switching permanent magnet machines with ferrite and rare-earth magnets are compared. Finally, [20] introduces a novel bearingless SRM with decoupled torque and suspension stator poles, offering a promising direction for further research. Based on previous research, this paper introduces an analytical model designated as the Two-Component Saturation Correction Model. Unlike conventional saturation models that rely solely on the radial component of flux density, the proposed approach accounts for both radial and tangential flux density components in the saturation process. This comprehensive treatment enables a more accurate prediction of the electromagnetic torque and consequently the torque ripple as well as the radial pressure distribution. The model is applied to three distinct switched reluctance motor configurations with 12/10 stator/rotor pole (due to its superiority of 12/10 stator/rotor pole [10]): 3-phase, 4-phase, and 6-phase machines with concentrated windings, as illustrated in Figure 4.

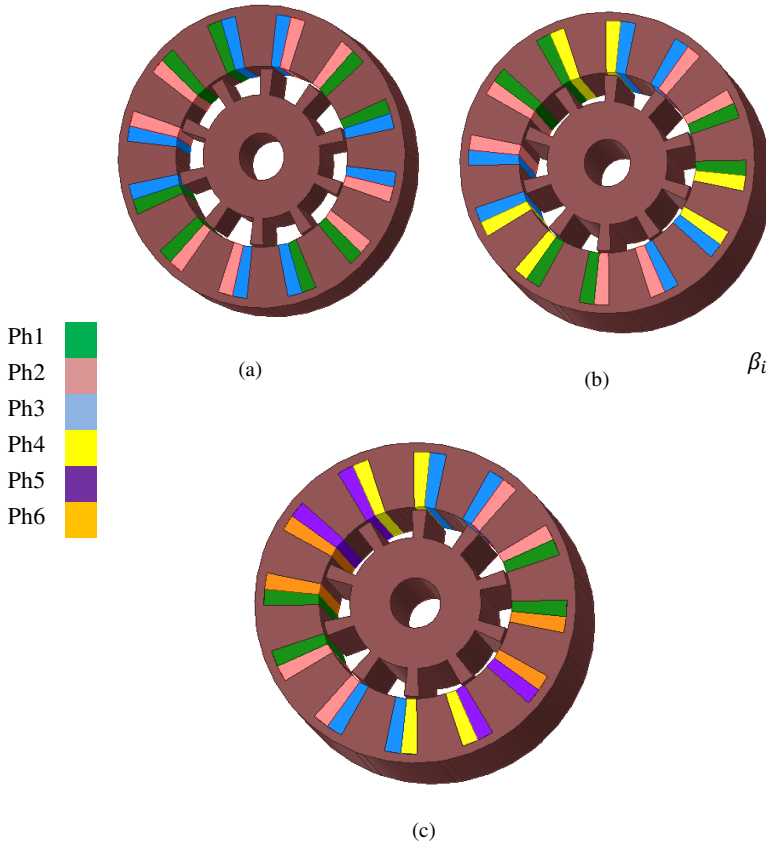


Fig.4. Three-layout of winding for configurations 12/10 stator/rotor pole: a) 3-phase b) 4-phase c) 6-phase

The main contribution of this work lies in the development of a saturation model that simultaneously incorporates the effects of radial and tangential flux density components. While most existing analytical models either neglect tangential flux or treat saturation in a simplified, single-component manner, the proposed Two-Component Saturation Correction Model offers a more physically consistent representation of the magnetic field distribution. This is particularly important for switched reluctance motors, where the interaction between radial and tangential fields significantly influences torque production and radial force behavior. By capturing these coupled effects, the model provides a more reliable foundation for predicting torque ripple and vibration two critical factors in sensitive applications such as blood pump drives. Furthermore, the comparative analysis across 3-, 4-, and 6-phase topologies reveals how the phase number affects the saturation behavior and the resulting electromagnetic performance, offering practical insights for motor selection in low-vibration medical devices.

II. NON-LINEAR MODEL

The proposed nonlinear model is developed on SRM illustrated in Figure 5 in two main stages that in Figure 5 whole SRM geometry and domains are shown. First, a linear field solution is obtained using the subdomain method. Second, an iterative saturation compensation technique is applied to correct the linear solution by accounting for the nonlinear behavior of the magnetic materials, considering both radial and tangential flux density components.

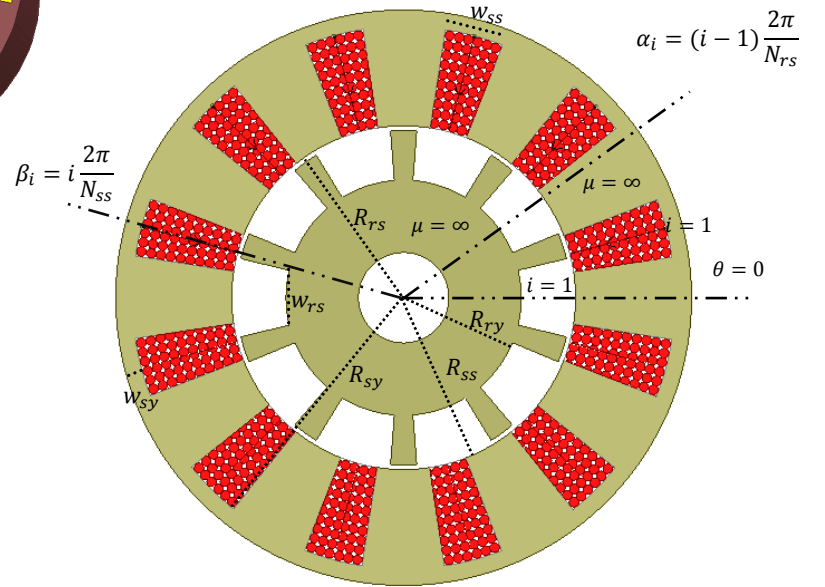


Fig.5. Parameters of Geometry

Stage 1: Linear Field Solution via Subdomain Method

In the first stage, the machine geometry is divided into distinct subdomains (e.g., stator slots, airgap, rotor slots). Under the assumption of infinite permeability for the iron core as shown in Figure 6, the magnetic vector potential A is calculated for each subdomain by solving Maxwell's equations as shown in relations (1). In regions without energy sources (no current density J or magnetization M), the equation reduces to Laplace's equation. In winding regions, the current density J serves as the source term, while in permanent magnet regions, the magnetization vector M (with its radial and tangential components) acts as the source. The general solution for each subdomain is expressed in terms of Fourier series, with unknown coefficients determined by applying boundary conditions at the interfaces between adjacent regions. Once the magnetic vector potential A is known, the radial and tangential components of flux density B and the magnetic field intensity H by (2) and (3) are obtained based on the permeability coefficients of the materials, where μ_0 represents the permeability of air and μ_r represents the relative permeability of each material.

$$-\nabla^2 A_z^i = \mu_0 \mu_r J_z^i + \frac{\mu_0}{r} \left(\frac{\partial M_r}{\partial \theta} - r \frac{\partial M_\theta}{\partial r} \right) \quad (1)$$

$$B = \nabla \times A \quad (2)$$

$$H = \frac{B}{\mu_0 \mu_r} - \frac{M}{\mu_r} \quad (3)$$

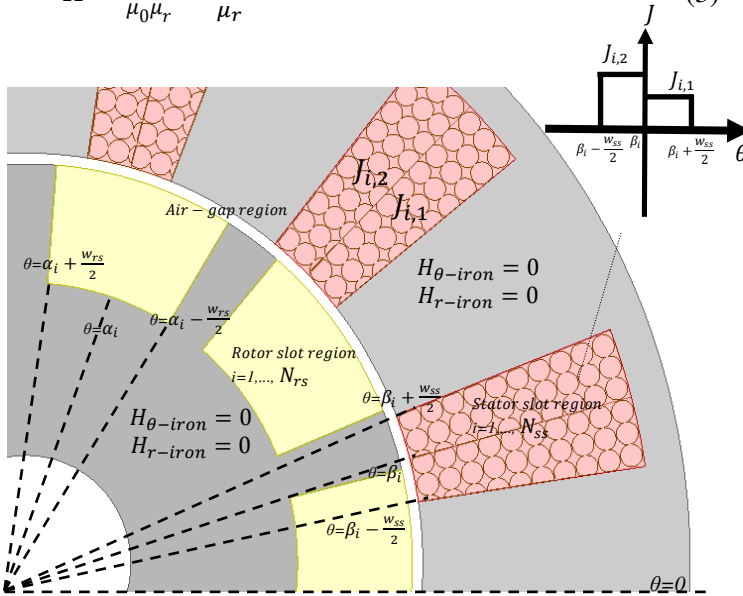


Fig.5. Sub-domain for solution of linear analytical model

Stage 2: Two-Component Saturation Correction Model

While the linear solution provides a fast initial estimate, it fails to capture the saturation effects that occur in the iron teeth and yoke under high-load conditions. To address this limitation, a Two-Component Saturation Correction Model is developed, which accounts for both radial and tangential flux density components in the saturation process. The compensation procedure is performed iteratively. First, the flux passing through each stator tooth, rotor tooth, and yoke is calculated by integrating the radial flux density obtained from B . The average flux density in each section is then determined, and the corresponding magnetic field intensity H is obtained from the actual B-H curve of the lamination steel. Local saturation factors are defined for each tooth and yoke section, representing the reduction in effective airgap flux due to iron saturation. The linear flux density is then corrected locally by dividing both radial and tangential components by the product of the corresponding saturation factors. This process is repeated until the relative change in flux density falls below a specified tolerance. The summarized algorithm of the solution path is brought in Figure 7.

START

INPUT: Motor geometry, B-H curve, currents, rotor position.

STAGE 1 – Linear Solution:

- Solve field using subdomain method → get A , then B , then H .

STAGE 2 – Saturation Correction (Iterative):

- Calculate flux in teeth and yoke.
- Get H from B-H curve.
- Define local saturation factors.
- Correct both B_r and B_θ .
- Repeat until change < tolerance.

STAGE 3 – Performance:

- Compute torque, torque ripple, and radial pressure.

OUTPUT: Corrected field, torque, ripple, pressure, saturation factors.

APPLY TO: 3-phase, 4-phase, 6-phase SRMs.

END

Fig. 7. Summary of solution algorithm

A. ROTOR SLOTS EQUATIONS IN LINEAR MODEL

In these regions, there is no current or magnetization, so Laplace's equation applies:

$$-\nabla^2 A_z^i = 0 \quad (4)$$

The general solution is a cosine Fourier series (due to symmetry and the boundary condition $\mathbf{H} = 0$ at the iron boundaries):

$$A_z^i(r, \theta) = a_{i,0} + \sum_{m=1}^{\infty} \left(a_{i,m} f(m, r) \cos \left(\frac{m\pi}{W_{rs}} \left(\theta - \alpha_i + \frac{W_{rs}}{2} \right) \right) \right) \quad (5)$$

$$f(n, r) = r^{\frac{-m\pi}{W_{rs}}} + (R_{ry}^{\frac{-m\pi}{W_{rs}}} r^{\frac{m\pi}{W_{rs}}} / R_{ry}^{\frac{m\pi}{W_{rs}}}) \quad (6)$$

B. AIR-GAP EQUATION IN LINEAR MODEL

In these regions, Laplace's equation applies, but the solution includes both sine and cosine terms (due to the full circumferential opening):

$$A_z^i(r, \theta) = \sum_{n=1}^{\infty} ((a_{1n}(r)^n + a_{2n}(r)^{-n}) \cos(n\theta) + (b_{1n}(r)^n + b_{2n}(r)^{-n}) \sin(n\theta)) \quad (7)$$

C. STATOR SLOTS EQUATION IN LINEAR MODEL

In these regions, Poisson's equation applies with current density J as the source:

$$-\nabla^2 A_z^i = \mu_0 \mu_r J_z^i \quad (8)$$

In the winding regions (stator slots), the magnetic field has a source of the winding current. Therefore, the governing equation is non-homogeneous (Poisson's equation), not homogeneous (Laplace's equation).

To solve the non-homogeneous equation, the complete solution is built from two parts:

1. **General solution:**
The solution to the homogeneous equation (Laplace), which depends only on the geometry and boundary conditions.
2. **Particular solution:**
A specific solution that directly accounts for the source (winding current), which takes the form $-\frac{1}{4} \mu_0 J_i r^2$.

$$A_z^i(r, \theta) = a_{i,0} + b_{i,0} \ln(r) - \frac{1}{4} \mu_0 J_i r^2 + \sum_{u=1}^{\infty} \left((a_{i,u} r^{\frac{-u\pi}{W_{ss}}} + b_{i,u} r^{\frac{u\pi}{W_{ss}}}) \cos \left(\frac{u\pi}{W_{ss}} \left(\theta - \beta_i + \frac{W_{ss}}{2} \right) \right) \right) \quad (9)$$

$$J_i(\theta) = J_{i,0} + \sum_{u=1}^{\infty} \left(J_{i,u} \cos \left(\frac{u\pi}{W_{ss}} \left(\theta - \theta_i + \frac{W_{ss}}{2} \right) \right) \right) \quad (10)$$

Where $J_{i,0} = (J_{i,1} d + J_{i,2} d) / w_{ss}$,
 $J_{i,u} = \frac{2}{u\pi} \sin \left(\frac{u\pi d}{W_{ss}} \right) (J_{i,1} + J_{i,2}) (-1)^u$,

$J_{i,1}, J_{i,2}$ are determined from the current density values on the right and left sides of the slots.

B. BOUNDARY CONDITIONS

At each interface between two adjacent subdomains, two fundamental conditions must be satisfied based on Figure 8:

1. Continuity of the magnetic vector potential (A_z):
This ensures that the magnetic flux passes smoothly from one region to the next, with no abrupt changes or discontinuities.
2. Continuity of the tangential component of the magnetic field intensity (H_θ):
This ensures that there are no surface currents at the interface, which is a physical requirement in the absence of any current sheet.

These conditions are applied at every interface and, after projection onto the Fourier basis; they yield a system of linear equations that determines the unknown coefficients.

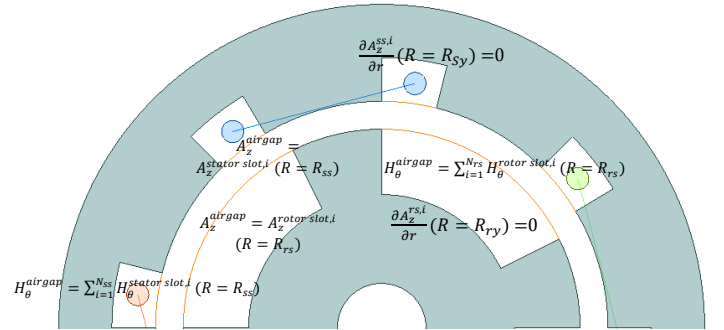


Fig. 8. Boundary condition illustrative

1. Continuity of A_z : Rotor Slots and Airgap (at the interface radius $R=R_{rs}$)

For each rotor slot i with angular width w_{rs} and center angle α_i , the magnetic vector potential must be continuous over the slot span.

$$\begin{aligned} \text{DC component } (m=0) \quad & \int_{\alpha_i+w_{rs}/2}^{\alpha_i+w_{rs}/2} A_z^{\text{airgap}}(R_{rs}, \theta) d\theta \\ & = \end{aligned} \quad (11)$$

$$\int_{\alpha_i+w_{rs}/2}^{\alpha_i+w_{rs}/2} A_z^{\text{rotor slot}}(R_{rs}, \theta) d\theta$$

$$\begin{aligned} \text{For each harmonic } (m=1, 2, \dots, M) \quad & \int_{\alpha_i-w_{rs}/2}^{\alpha_i+w_{rs}/2} A_z^{\text{airgap}}(R_{rs}, \theta) \cos\left(\frac{m\pi}{w_{rs}}\left(\theta - \alpha_i\right)\right) d\theta \\ & = \int_{\alpha_i-w_{rs}/2}^{\alpha_i+w_{rs}/2} A_z^{\text{rotor slot}}(R_{rs}, \theta) \cos\left(\frac{m\pi}{w_{rs}}\left(\theta - \alpha_i\right)\right) d\theta \end{aligned} \quad (12)$$

2. Continuity of A_z : Stator Slots and Airgap (at the interface radius $R=R_{ss}$)

For each rotor slot i with angular width w_{ss} and center angle β_i , the magnetic vector potential must be continuous over the slot span.

$$\begin{aligned} \text{DC component } (u=0) \quad & \int_{\beta_i+w_{ss}/2}^{\beta_i+w_{ss}/2} A_z^{\text{airgap}}(R_{ss}, \theta) d\theta \\ & = \end{aligned} \quad (13)$$

$$\int_{\beta_i+w_{ss}/2}^{\beta_i+w_{ss}/2} A_z^{\text{stator slot}}(R_{ss}, \theta) d\theta$$

$$\begin{aligned} \text{For each harmonic } (u=1, 2, \dots, U) \quad & \int_{\beta_i-w_{ss}/2}^{\beta_i+w_{ss}/2} A_z^{\text{airgap}}(R_{ss}, \theta) \cos\left(\frac{u\pi}{w_{ss}}\left(\theta - \beta_i\right)\right) d\theta \\ & = \int_{\beta_i-w_{ss}/2}^{\beta_i+w_{ss}/2} A_z^{\text{stator slot}}(R_{ss}, \theta) \cos\left(\frac{u\pi}{w_{ss}}\left(\theta - \beta_i\right)\right) d\theta \end{aligned} \quad (14)$$

3. Continuity of H_θ Rotor Slots and Airgap (at the interface radius $R=R_{rs}$)

Cosine projection (for each harmonic $n=1, 2, \dots, N$):

$$\begin{aligned} & \int_0^{2\pi} \sum_{i=1}^{N_{rs}} H_\theta^{\text{rotor slot}, i}(R_{rs}, \theta) \cos(n\theta) d\theta \\ & = \int_0^{2\pi} H_\theta^{\text{airgap}}(R_{rs}, \theta) \cos(n\theta) d\theta \end{aligned} \quad (15)$$

Sine projection (for each harmonic $n=1, 2, \dots, N$):

$$\begin{aligned} & \int_0^{2\pi} \sum_{i=1}^{N_{rs}} H_\theta^{\text{rotor slot}, i}(R_{rs}, \theta) \sin(n\theta) d\theta \\ & = \int_0^{2\pi} H_\theta^{\text{airgap}}(R_{rs}, \theta) \sin(n\theta) d\theta \end{aligned} \quad (16)$$

4. Continuity of H_θ Stator Slots and Airgap (at the interface radius $R=R_{ss}$)

Cosine projection (for each harmonic $n=1, 2, \dots, N$):

$$\begin{aligned} & \int_0^{2\pi} \sum_{i=1}^{N_{ss}} H_\theta^{\text{stator slot}, i}(R_{ss}, \theta) \cos(n\theta) d\theta \\ & = \int_0^{2\pi} H_\theta^{\text{airgap}}(R_{ss}, \theta) \cos(n\theta) d\theta \end{aligned} \quad (17)$$

Sine projection (for each harmonic $n=1, 2, \dots, N$):

$$\begin{aligned} & \int_0^{2\pi} \sum_{i=1}^{N_{ss}} H_\theta^{\text{stator slot}, i}(R_{ss}, \theta) \sin(n\theta) d\theta \\ & = \int_0^{2\pi} H_\theta^{\text{airgap}}(R_{ss}, \theta) \sin(n\theta) d\theta \end{aligned} \quad (18)$$

C. MATRIX DIMENSIONS FOR A 12/10 SRM WITH HARMONIC ORDER 100

The prototype SRM is of 12/10 stator/rotor pole configuration that for a harmonic order ($n=m=u$) order of 100, the total number of unknowns is $N_{total} = 3834$ due to table 1, resulting in a matrix of size 3834×3834 . The matrix has a 3×3 block structure formulated in relation (19) and the dimension of each block is observable in Table 2 with zero off-diagonal blocks between the rotor and stator subdomains, as they are not directly adjacent.

Table I. Number of Unknowns per Subdomain

Subdomain	Count per Harmonic	Total Unknowns
Rotor Slots (10 slots)	$1+100=101$ per slot	$10 \times 101 = 1010$
Airgap	$4 \times 100 = 400$	400
Stator Slots (12 slots)	$2+2 \times 100 = 202$ per slot	$12 \times 202 = 2424$

$$\begin{pmatrix} M_{11} & M_{12} & M_{13} \\ M_{21} & M_{22} & M_{23} \\ M_{31} & M_{32} & M_{33} \end{pmatrix} \begin{pmatrix} X_{rotor} \\ X_{airgap} \\ X_{stator} \end{pmatrix} = \begin{pmatrix} 0 \\ 0 \\ \text{Winding Source} \end{pmatrix} \quad (19)$$

Table II. Block Sizes

Block	Size	Description
M11	1010 × 1010	Rotor → Rotor self-coupling
M12	1010 × 400	Rotor → Airgap coupling
M13	1010 × 2424	Zero (no direct rotor-stator coupling)
M21	400 × 1010	Airgap → Rotor coupling
M22	400 × 400	Airgap → Airgap self-coupling
M23	400 × 2424	Airgap → Stator coupling
M31	2424 × 1010	Zero (no direct stator-rotor coupling)
M32	2424 × 400	Stator → Airgap coupling
M33	2424 × 2424	Stator → Stator self-coupling

D. TWO-COMPONENT SATURATION CORRECTION MODEL

The linear solution from the subdomain method assumes infinite permeability in the iron. That works for a rough estimate, but not for real conditions. When flux density increases, the iron saturates and the linear B - H relationship no longer holds. In SRMs, where flux runs entirely through iron, saturation is significant. To overcome this, two-component saturation correction model is presented that unlike most approaches that only correct the radial component, the model corrects both radial and tangential components.

Steps are implemented as follow:

Step 1: Calculate the Flux in Each Tooth

For each stator tooth and each rotor tooth as shown in Figure 10, the total flux passing through it is calculated. Since flux is defined as the integral of the normal component of $\mathbf{B}$ over a surface:

$$\Phi_{tooth,i} = R L_{stack} \int_{\theta_{lower}}^{\theta_{upper}} B_{r,linear} d\theta \quad (20)$$

The yoke flux is then obtained by summing the fluxes from the teeth that are connected to it.

$$\Phi_{yoke,\theta} = \sum_{i=1}^{N_{teeth}} \Phi_{tooth,i,\theta} \quad (21)$$

Step 2: Compute Average Flux Density in Each Part

By dividing $\Phi_{tooth,i}$ by the cross-sectional area of the corresponding part the average flux density is obtained:

$$B_{r,tooth,i} = \frac{\Phi_{tooth,i}}{L_{stack} w_{teeth} R} \quad (22)$$

Where L_{stack} motor depth and R is teeth radius of stator. For the yoke, the flux density is obtained by dividing the yoke flux by its cross-sectional area:

$$B_{yoke} = \frac{\Phi_{yoke}}{L_{stack} h_{yoke}} \quad (23)$$

The tangential component also affects saturation. So, total flux density magnitude at each point:

$$|B_{total}| = \sqrt{B_{r,tooth,i}^2 + B_{\theta,tooth,i}^2} \quad (24)$$

$$B_{\theta,tooth,i} = \frac{B_{\theta,airgap,i}}{\mu_r} \quad (25)$$

Step 3: Find the Corresponding H from the B - H Curve

Once the total flux density at each point is determined, the next step is to obtain the corresponding magnetic field intensity. This is where the model transitions from a linear approximation to a realistic nonlinear description. Instead of assuming a constant permeability, the actual B - H curve of the lamination steel is used the same curve that represents the true magnetic behaviour of the material under operating conditions (Figure 9).

By the actual B - H curve of the lamination steel, magnetic field intensity is obtained:

$$f_{B-H}(|B_{total,i}|) = |H_{total,i}| \quad (26)$$

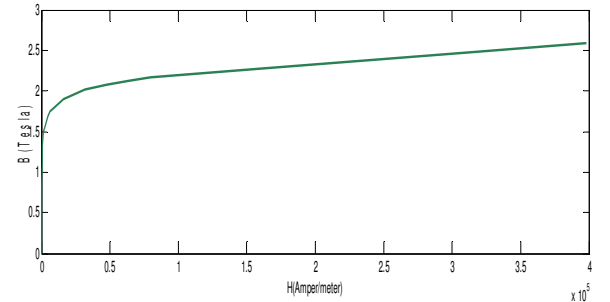


Fig. 9. Actual B-H characteristic curve for steel 1008

Step 4: Define Local Saturation Factors

Once the magnetic field intensity in each tooth and yoke section is known, the next step is to quantify the effect of the iron on the overall magnetic circuit. This is done by defining a local saturation factor for each region. The saturation factor compares the magnetomotive force consumed by the iron path to that consumed by the airgap.

$$K_{sat,stator,i} = 1 + \frac{H_{total,stator,i} \cdot l_{stator,i}}{H_{gap,i} \cdot g} \quad (27)$$

$$K_{sat,rotor,i} = 1 + \frac{H_{total,rotor,i} \cdot l_{rotor,i}}{H_{gap,i} \cdot g} \quad (28)$$

$$K_{sat,yoke} = 1 + \frac{H_{total,yoke} \cdot l_{yoke}}{H_{gap} \cdot g} \quad (29)$$

A factor greater than 1 means the iron is taking up a significant part of the driving force, which reduces the effective flux in the airgap.

Step 5: Correcting the Air-Gap Flux Density

Once the local saturation factors are determined, the linear flux density is adjusted accordingly. The correction is applied by dividing the linear value by the product of the relevant saturation factors. This is carried out for both the radial and tangential components, using the same set of factors, since saturation is a material property that affects the field regardless of its direction.

For the radial component:

$$B_{r_sat}(\theta) = \frac{B_{r,linear}(\theta)}{K_{sat,stator}(\theta)K_{sat,rotor}(\theta)K_{sat,yoke}(\theta)} \quad (30)$$

For the tangential component:

$$B_{\theta_sat}(\theta) = \frac{B_{\theta,linear}(\theta)}{K_{sat,stator}(\theta)K_{sat,rotor}(\theta)K_{sat,yoke}(\theta)} \quad (31)$$

Step 6. Check for Convergence

As the flux density changes after correction, the saturation factors also alter. so, the total algorithm is repeated to converge the results.

$$B_{r_sat,k+1}(\theta) = \frac{B_{r,linear}(\theta)}{K_{sat,stator,k}(\theta)K_{sat,rotor,k}(\theta)K_{sat,yoke,k}(\theta)} \quad (32)$$

$$B_{\theta_sat,k+1}(\theta) = \frac{B_{\theta,linear}(\theta)}{K_{sat,stator,k}(\theta)K_{sat,rotor,k}(\theta)K_{sat,yoke,k}(\theta)} \quad (33)$$

The iteration runs until the relative change in flux density falls below a specified tolerance (ϵ):

$$|B_{r_sat,k+1}(\theta) - B_{r_sat,k}(\theta)| < \epsilon \quad (34)$$

$$|B_{\theta_sat,k+1}(\theta) - B_{\theta_sat,k}(\theta)| < \epsilon \quad (35)$$

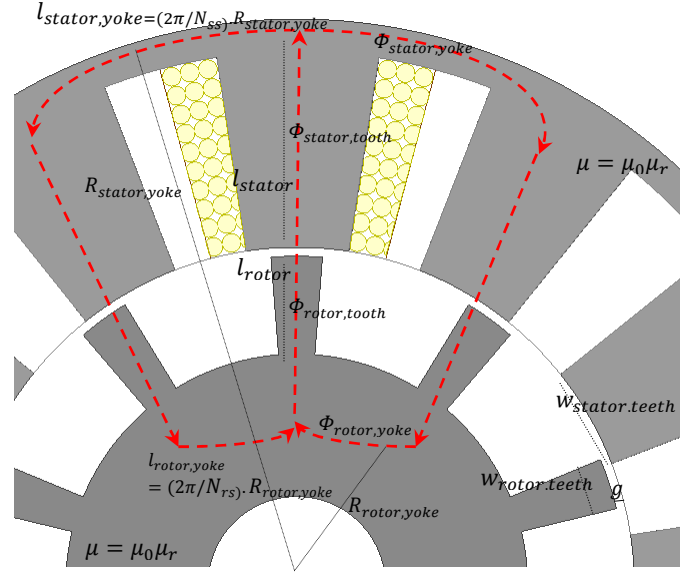


Fig. 10. Direction of magnetic flux

III. RESULTS

This section presents the magnetic field distribution obtained from the analytical model for the three phase configurations under study: 3-phase, 4-phase, and 6-phase SRMs. The results are compared at three levels: the linear analytical solution, the nonlinear saturated solution using the proposed Two-Component Saturation Correction Model, and the finite element method (FEM) simulations, which are considered as the reference. To ensure that the machine operates under saturated conditions and to properly evaluate the performance of the proposed Two-Component Saturation Correction Model, a current density of 30 A/mm² is applied to the phase windings. This value is deliberately chosen to push the iron core into deep saturation, particularly in the stator and rotor teeth, where the flux density is expected to exceed the knee point of the B-H curve of the lamination steel. Operating under such conditions allows a thorough assessment of the saturation correction capability of the nonlinear model, as the linear solution is expected to deviate significantly from the actual field distribution. The relationship between the current density and the phase current is given by:

$$k_f J = \frac{I_{phase}}{A_{slot}} \quad (36)$$

Where A_{slot} is the cross-sectional area of the stator slot and k_f is filling factor. For a given slot fill factor and conductor cross-section, the phase current is adjusted accordingly to achieve the desired current density. The phase currents are assumed to be sinusoidal and concentrated in the stator slots. The excitation follows a sequential energisation pattern based on rotor position, with each phase shifted by a specific electrical angle. The general form of the current for phase k is:

$$I_k(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + \theta_k) \quad (37)$$

Where I_{peak} , θ , θ_k are respectively, peak current amplitude, mechanical rotor position and phase shift angle for phase k . Three-scenario to be compared are energized as follow:

1. 3-Phase SRM

For a three-phase system, the phase currents are shifted by 120electrical degree:

$$I_1(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta) \quad (38)$$

$$I_2(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + 2\frac{\pi}{3}) \quad (39)$$

$$I_3(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + 4\frac{\pi}{3}) \quad (40)$$

2. 4-Phase SRM

For a four-phase system, the phase currents are shifted by 90electrical degree:

$$I_1(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta) \quad (41)$$

$$I_2(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + \frac{\pi}{2}) \quad (42)$$

$$I_3(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + \pi) \quad (43)$$

$$I_4(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + \frac{3\pi}{2}) \quad (44)$$

3. 6-Phase SRM

For a six-phase system, the phase currents are shifted by 60electrical degree:

$$I_1(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta) \quad (45)$$

$$I_2(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + \frac{\pi}{3}) \quad (46)$$

$$I_3(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + \frac{2\pi}{3}) \quad (47)$$

$$I_4(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + \pi) \quad (48)$$

$$I_5(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + \frac{4\pi}{3}) \quad (49)$$

$$I_6(\theta) = I_{peak} \cdot \sin(N_{rs} \cdot \theta + \frac{5\pi}{3}) \quad (50)$$

The accuracy and reliability of the proposed analytical model, including the Two-Component Saturation Correction Model, are validated against finite element (FEM) simulations. The validation is carried out for the three phase configurations (3-phase, 4-phase, and 6-phase) using the geometric parameters specified in Table 3. The same rotor/stator pole combination (12/10) and identical material properties are used across all cases to ensure a consistent basis for comparison.

For each phase configuration, the following quantities are evaluated and compared:

- Radial flux density distribution in the airgap,
- Tangential flux density distribution in the airgap,
- Electromagnetic torque,
- Torque ripple,
- Radial force distribution.

The linear analytical model, the nonlinear analytical model, and the FEM simulations are compared side-by-side. The linear model assumes infinite permeability and serves as a baseline to highlight the improvements introduced by the saturation correction. The nonlinear model incorporates the iterative saturation compensation described in previous sections, while the FEM results are treated as the reference solution.

TABLE III
SPECIFICATIONS OF THE SRM

Parameters	Value
Maximum current density of inner winding, J	30A/mm ²
Saturation flux density	1.7T
Number of stator slots, N_{ss}	12
Number of conductors per each inner slot, N_c	20
Number of phases, N_{ph}	3, 4, 6
Number of rotor slots, N_{rs}	10
Width of rotor slot, w_{rs}	26.99 π /180 rad
Width of stator slot, w_{ss}	12.5 π /180 rad
Rotor yoke radius, R_{ry}	26mm
Stator yoke radius, R_{sy}	60mm
Inner radius of stator, R_{ss}	38mm
Outer radius of rotor, R_{rs}	37mm
Stack length, L_s	100mm
Width of stator yoke, w_{sy}	3.63mm
Filling factor, K_f	1
Rotor speed	600 rpm
Tolerance, ϵ	0.01
Middle of airgap, R	37.5mm

A. RADIAL AND TANGENTIAL FLUX DENSITY

The validation of proposed model is carried out at $\theta = 0$ that distribution of current for 3-phase, 4-phase and 6-phase are illustrated in Figures 11, 12, 13 respectively that (*out*) is the direction of current density outward and (*in*) is the direction of current density inward.

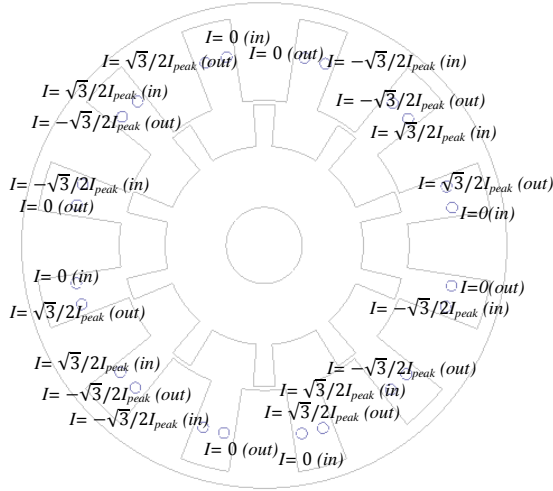


Fig. 11. Layout of winding at $\theta = 0$ for 3-phase

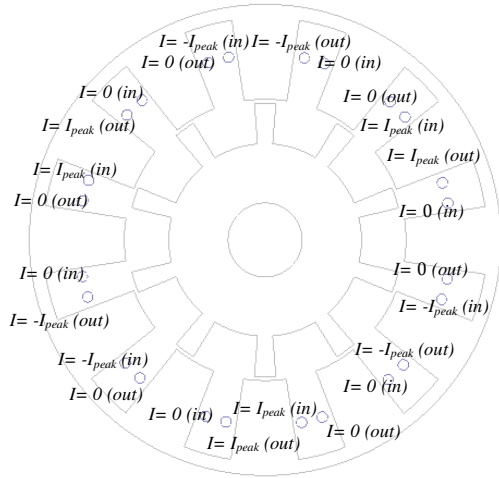


Fig. 12. Layout of winding at $\theta = 0$ for 4-phase

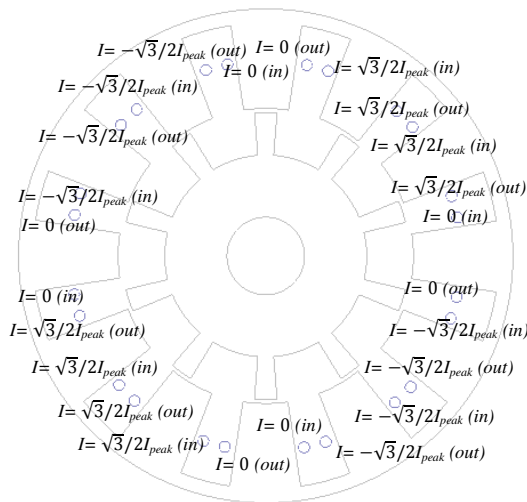


Fig. 13. Layout of winding at $\theta = 0$ for 6-phase

Figure 14 presents a comparative validation of the magnetic flux density distribution for the three phase configurations under study: 3-phase, 4-phase, and 6-phase SRMs. The results are shown for both B_r and B_θ components of the flux density in the airgap, plotted against the mechanical angle over a range of 0 to 180 degrees.

For each phase configuration, three sets of results are displayed simultaneously:

- The linear analytical solution (dashed blue line), which assumes infinite permeability in the iron and serves as a baseline,
- The nonlinear analytical solution (solid red line), obtained using the proposed Two-Component Saturation Correction Model,
- The FEM simulations (solid black line), treated as the reference solution.

The comparison clearly shows that the linear model consistently overestimates the flux density, particularly at peak regions where the iron is deeply saturated. This deviation is more pronounced in the 3-phase configuration, where the flux density reaches higher values due to the concentrated winding arrangement. In contrast, the nonlinear model follows the FEM results with excellent agreement across all three configurations. The improvement is especially noticeable in the radial component, where the saturation correction reduces the peak error significantly.

For the tangential component, the differences between the linear and nonlinear models are smaller in magnitude but still meaningful. The tangential flux density, though lower than the radial component, plays an important role in torque production and local saturation. The proposed model captures its variation accurately, confirming that both components must be corrected simultaneously for a physically consistent representation.

Overall, Figure 14 demonstrates that the proposed Two-Component Saturation Correction Model successfully bridges the gap between the simple linear analytical approach and the more accurate but computationally expensive FEM. The results confirm that the model is reliable across all three phase configurations, making it a practical tool for design and optimization studies where both speed and accuracy are required.

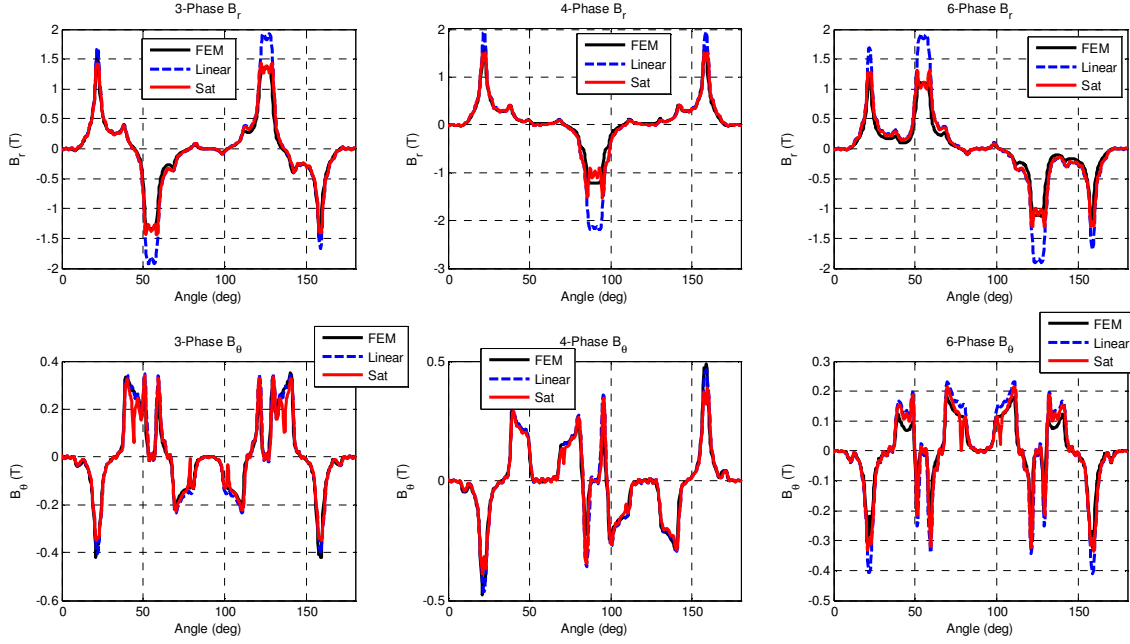


Fig. 14. Validation of proposed analytical model in 3-4-6 phases of concentrated winding

B. ELECTROMAGNETIC TORQUE AND RADIAL PRESSURE

The electromagnetic torque is computed from the magnetic field distribution obtained from the subdomain method. Once B_r and B_θ are known in the airgap, the local stress on the rotor surface can be calculated using the Maxwell stress tensor. The radial, tangential components of the stress and unbalanced magnetic force resultant from radial, tangential components of the stress (index to determine of eccentricity of rotor that F_x, F_y are total force resolved into x and y direction), are given by:

$$f_r = \frac{(B_r^2 - B_\theta^2)}{2\mu_0} \quad (51)$$

$$f_\theta = \frac{B_r B_\theta}{\mu_0} \quad (52)$$

$$F_{UMF} = \sqrt{F_x^2 + F_y^2} \quad (53)$$

The tangential stress f_θ is directly responsible for the electromagnetic torque. By integrating this stress over the rotor surface, the total torque is obtained:

$$T(\theta) = R^2 \cdot L_{stack} \int_0^{2\pi} f_\theta d\theta \quad (54)$$

Torque ripple refers to the periodic fluctuation of the electromagnetic torque during rotor rotation. In switched reluctance motors, this fluctuation is inherent due to the discrete nature of phase excitation and the doubly salient structure of the machine. When one phase is turned off and next is turned on, the torque does not remain perfectly constant it rises and falls as the rotor moves through the

commutation zones. In most industrial applications, torque ripple is undesirable because it causes:

- Mechanical vibration and acoustic noise
- Speed fluctuations
- Increased mechanical stress on bearings and couplings
- Reduced control precision

In sensitive applications such as blood pumps, where the motor is implanted inside the body and direct contact with blood exists, these effects are even more critical. Excessive vibration can increase the risk of hemolysis (damage to red blood cells) and thrombosis (blood clot formation). Therefore, minimizing torque ripple is not just a performance issue it is a safety requirement.

Torque ripple is typically expressed as the peak-to-peak variation of the torque waveform, normalized by the average torque that is the value between 0 and 1:

$$T_{ripple} = \frac{T_{Max} - T_{Min}}{T_{Avg}} \quad (55)$$

Where $T_{Max}, T_{Min}, T_{Avg}$ are maximum instantaneous torque, minimum instantaneous torque, average torque over one electrical cycle, respectively.

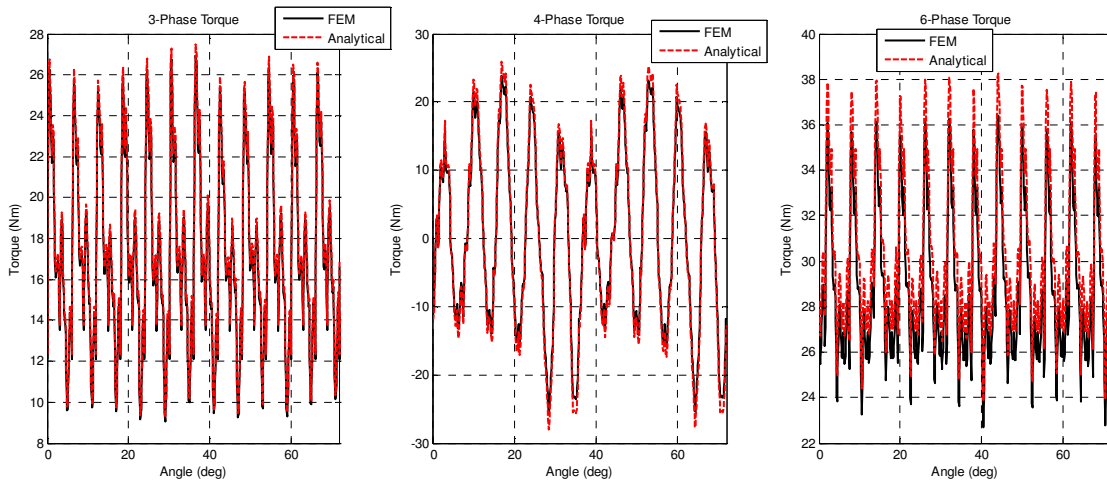


Fig. 15. Validation of Electromagnetic torque in 3- 4 - 6 phases of concentrated winding considering saturation

Figure 15 shows the comparison between the FEM and the proposed analytical model reveals an excellent agreement across the entire angular range. The nearly identical torque value confirm that the analytical model, which incorporates the Two-Component Saturation Correction, is capable of predicting the electromagnetic torque with remarkable accuracy. This close match is particularly important for applications such as blood pumps, where even small deviations in torque prediction can lead to significant errors in vibration and hemolysis estimation. The consistency of the results also validates the underlying assumptions of the model, including the treatment of magnetic saturation and the inclusion of both radial and tangential flux density components. Such a high level of agreement not only demonstrates the reliability of the proposed model but also highlights its potential as a computationally efficient alternative to time-consuming FEM simulations, especially during the early stages of motor design and optimization. Key observation of Figure 15 shows:

1. **3-Phase Configuration:** The torque waveform exhibits a relatively smooth profile with ripple of 0.88 and average torque of 18*N.m* approximately.
2. **4-Phase Configuration:** This case shows a significantly different behaviour. The torque waveform is highly oscillatory, with severe ripple and no clear average value. This confirms that the 12/10 structure is inherently unsuitable for four-phase excitation, as discussed earlier.
3. **6-Phase Configuration:** The torque waveform is the smoothest among the three, with the lowest ripple of 0.44 and average torque of 31*N.m* approximately. The reduced ripple and stable average torque make the 6-phase configuration the most suitable for low-vibration applications such as blood pumps.

It is important to clarify why the 4-phase configuration has been excluded from the radial force analysis. As demonstrated earlier in Figure 15, the 4-phase excitation of the 12/10 structure produced a highly oscillatory torque waveform with no stable average value. This behaviour

confirms that the 12/10 topology is inherently incompatible with four-phase operation.

Figure 16 shows the radial force distribution over one mechanical cycle. The 3-phase case exhibits significant force fluctuations, while the 6-phase configuration provides a much smoother and more symmetric profile, indicating better magnetic balance and lower vibration potential.

Figure 17 presents the resolved components of the unbalanced magnetic force (F_x and F_y). The 3-phase case shows strong periodic variations, whereas the 6-phase configuration exhibits significantly smaller and more balanced force components, which is critical for bearingless motor applications where rotor stability depends on precise radial force control.

Figure 18 displays the resultant unbalanced magnetic force (UMF). The 6-phase configuration shows the lowest UMF with a smooth profile, confirming that the magnetic forces are well-balanced, leading to reduced vibration and improved rotor stability.

In summary, the nonlinear model shows excellent agreement with FEM across all cases. The 6-phase configuration consistently outperforms the 3-phase design in terms of force smoothness and balance, making it the preferred choice for blood pump applications where low vibration and high reliability are essential.

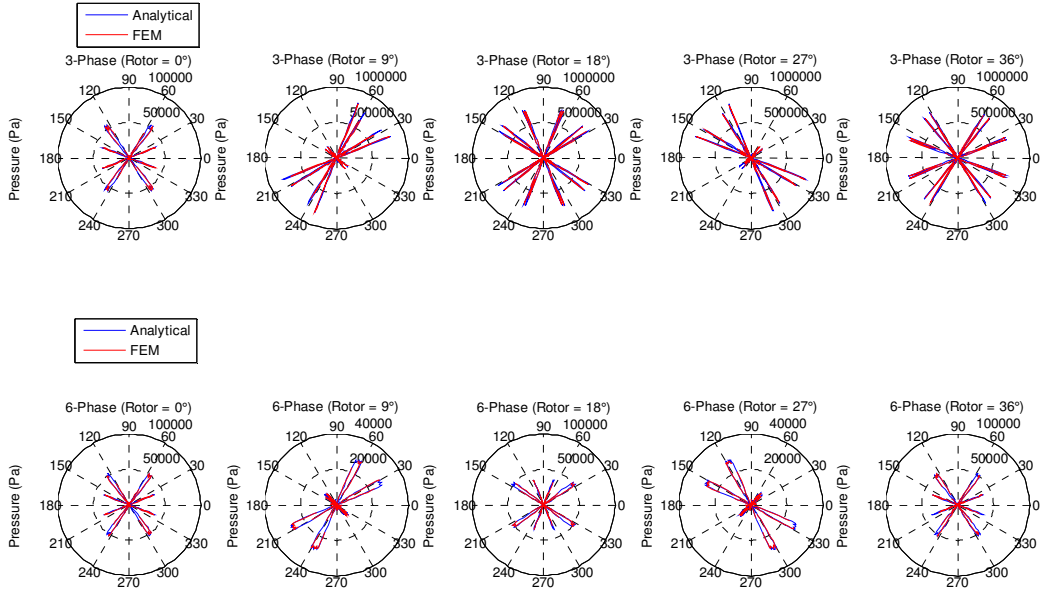


Fig. 16. Validation of radial pressure in 3- 4 - 6 phases of concentrated winding considering saturation around the rotor

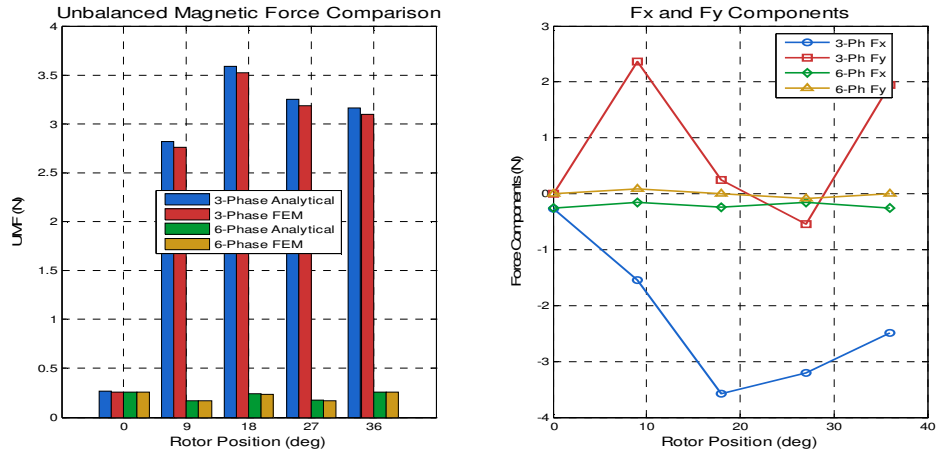


Fig. 17. Validation of F_x and F_y in 3- 4 - 6 phases of concentrated winding considering saturation

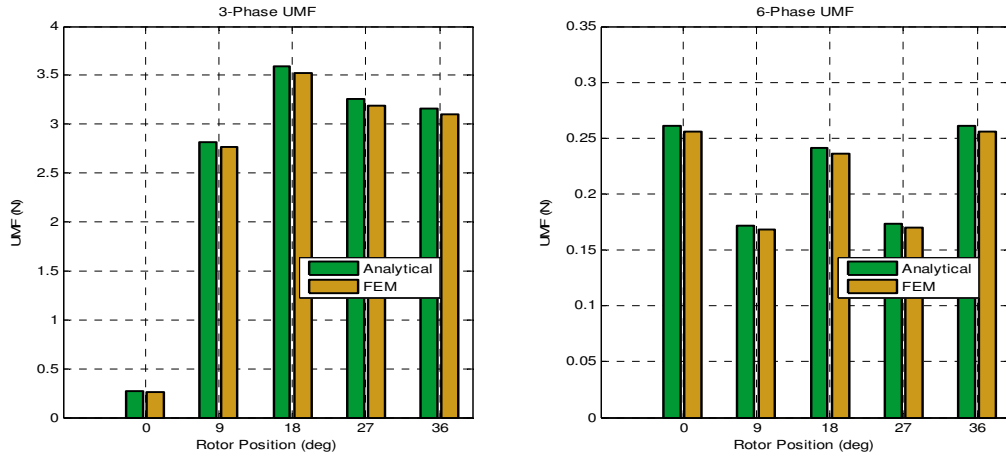


Fig. 18. Validation of proposed in 3- 4 - 6 phases of concentrated winding considering saturation

IV. DISCUSSION

The results presented in Figures 15–18 clearly demonstrate the performance advantages of the 6-phase configuration over the conventional 3-phase design for the 12/10 SRM structure. In terms of torque quality, the 6-phase machine produces a significantly smoother torque profile, with a ripple of 0.44 compared to 0.88 for the 3-phase case. This reduction in torque ripple is a direct consequence of the increased number of phases, which leads to a higher commutation frequency and more frequent torque pulses per rotor revolution. As a result, the overlap between successive phase torques is improved, reducing the dips and peaks that characterize the torque waveform of lower-phase machines.

The radial force behavior follows a similar trend. The 6-phase configuration exhibits a more symmetric and evenly distributed radial force around the rotor circumference, while the 3-phase design shows pronounced force concentrations at specific angular positions. This imbalance in the 3-phase case is a source of mechanical vibration and acoustic noise, as the rotor is subjected to periodic unbalanced magnetic pulls. In the 6-phase design, the higher number of phases allows for a more uniform excitation pattern, which spreads the magnetic forces more evenly and reduces the net unbalanced force acting on the rotor.

The resolved force components F_x and F_y further confirm this improvement. In the 3-phase configuration, both components show significant periodic variations, indicating a strong unbalanced magnetic pull that varies with rotor position. The 6-phase machine, on the other hand, exhibits much smaller fluctuations in both directions, suggesting better rotor stability and reduced bearing stress. This is particularly important for bearingless motor designs, where precise control of radial forces is essential for stable rotor suspension.

The resultant UMF, which represents the net radial force on the rotor, is also significantly lower in the 6-phase case. A lower UMF means less vibration and reduced mechanical stress on the rotor and bearings, which directly translates to improved reliability and longer device lifetime. For blood pump applications, where the motor is implanted and must operate continuously over long periods, this reduction in vibration is not just a performance improvement it is a safety requirement.

The 4-phase configuration, however, failed to produce a stable torque output, confirming the inherent incompatibility of the 12/10 structure with four-phase excitation. This harmonic mismatch prevents the torque contributions from different phases from overlapping constructively, resulting in destructive interference and a near-zero average torque. As a result, the 4-phase case was excluded from the radial force analysis, as any force comparison under such unstable conditions would be physically meaningless.

In addition to the electromagnetic benefits, the 6-phase configuration also offers practical advantages in terms of control flexibility. The higher number of phases provides more degrees of freedom for current shaping and fault-tolerant operation, allowing the motor to maintain smooth torque delivery even under abnormal operating conditions. This is particularly valuable in medical applications, where system failure is not an option.

Overall, the results confirm that the 6-phase configuration is not only superior to the 3-phase design in terms of torque ripple and radial force balance, but also offers a more robust and reliable solution for vibration-sensitive applications such as blood pumps. The excellent agreement between the proposed nonlinear model and FEM simulations further validates the accuracy of the analytical approach, making it a reliable tool for future design and optimization studies.

Author Contributions

E.S. conceptualized the study, developed the mathematical framework based on the subdomain method, derived the analytical solutions for the switched reluctance motor (SRM) configurations under investigation, and carried out the theoretical analysis. E.S. contributed to the analytical modeling, the development of the Two-Component Saturation Correction Model, finite element analysis (FEA) simulations, validation of the results against FEM, and the comparative analysis of torque ripple and radial pressure for bearingless blood pump applications. The author wrote and reviewed the manuscript. The author read and approved the final manuscript.

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Data Availability

The datasets generated and analyzed during the current study are available from the corresponding author on reasonable request. All numerical results were obtained from analytical modeling and finite element simulations using the mathematical models, formulations, and parameter values reported in this article.

Declarations

Ethics Approval and Consent to Participate

Not applicable. This study did not involve human participants, human data, human tissue, or animals requiring ethical approval and informed consent.

Consent for Publication

Not applicable. This manuscript does not contain any individual person's data in any form.

Competing Interests

The author declares that they have no competing interests.

Clinical Trial Registration

Not applicable. This research does not involve any clinical trial.

V. CONCLUSION

This paper has presented a comparative analytical study of 3-phase, 4-phase, and 6-phase SRM with a 12/10 stator/rotor pole configuration, focusing on torque ripple and radial force characteristics under magnetic saturation. A Two-Component saturation correction model was developed, which accounts for both radial and tangential flux density components in the saturation process, offering a more physically consistent representation of the magnetic field distribution compared to conventional linear models.

The model was validated against finite element simulations, and the results showed excellent agreement across all three configurations. The comparative analysis revealed distinct performance trends:

- The **3-phase configuration** produces the highest average torque but suffers from significant torque ripple and unbalanced radial forces, making it less suitable for vibration-sensitive applications.
-
- The **4-phase configuration** failed to produce a stable torque output, confirming the inherent incompatibility of the 12/10 structure with four-phase excitation due to harmonic mismatch.
-
- The **6-phase configuration** offers the best overall performance, with the lowest torque ripple, the most balanced radial force distribution, and the smallest unbalanced magnetic force, making it the most suitable choice for low-vibration applications.

The radial force analysis, presented for the 3-phase and 6-phase configurations, demonstrated that the higher number of phases in the 6-phase design leads to a more uniform force distribution, reduced unbalanced magnetic pull, and lower vibration potential. These findings are particularly significant for blood pump applications, where mechanical vibration is directly linked to haemolysis and device reliability.

From a clinical perspective, the 6-phase configuration offers a pathway towards quieter, more reliable, and more biocompatible motor drives for implantable blood pumps. The reduced vibration and lower mechanical stress translate directly to improved patient safety and longer device lifetime. The proposed analytical model provides a computationally efficient and accurate tool for predicting torque and radial force behavior in saturated SRMs, making it suitable for early-stage design and optimization studies.

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