

Article

Experimental Core Loss Characterization of a Switched Reluctance Motor Stator

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Abstract

Core loss significantly impacts the efficiency of a switched reluctance motor (SRM), especially under high-frequency operation. Conventional measurement methods are often inadequate for capturing losses in assembled stator cores, where flux paths are complex, and excitation waveforms are non-sinusoidal. In this paper, a core loss measurement method is introduced for SRM based on transformer induction theory, enabling time-domain reconstruction of magnetic field quantities from voltage and current waveforms. The method is validated using a ring core and SRM stator geometry with finite element analysis (FEA), replicating practical excitation conditions. An experimental setup is built with a fractional SRM stator core and custom-designed magnetizing yokes. The experimental setup captures induced voltage and current waveforms from the stator. The proposed method is experimentally validated over a wide frequency range from 100 Hz to 10,000 Hz and under varying flux density levels, confirming the applicability for practical core loss characterization of SRM stators.

Keywords: B-H loop; core loss characterization; eddy current loss; finite element analysis; hysteresis loss; inverter-fed excitation; stator core

1. Introduction

Switched reluctance motors (SRMs) have attracted significant attention due to their simple construction, lack of rare earth magnets, and high robustness under fault conditions [1–3]. However, to fully realize the potential of SRMs in high-efficiency applications, it is essential to address core challenges [4]. The accurate quantification and reduction in motor core losses with proper material characterization [5,6], alongside the experimental evaluation of flux linkage characteristics [7], are key challenges. Steinmetz and Bertotti pioneered in classical loss segregation theories [8,9]. Stator core loss in SRMs plays a critical role in determining the system efficiency, especially under high-frequency, inverter-fed excitations. These excitations are typically non-sinusoidal and contain significant harmonic content, which complexifies the magnetic behavior and substantially exacerbates core losses [10,11]. The core loss is sensitive to manufacturing-induced variations, such as mechanical stress from punching, welding, or press-fitting processes, which alter the magnetic behavior of electrical steel lamination [12]. Designers often apply corrective coefficients during the modeling process to account for these manufacturing effects.

Core loss characterization methods work by subjecting the core material to an alternating magnetic field and recording the resulting energy dissipation [13,14]. Electrical steel manufacturer datasheets generally provide specified core loss figures in W/kg as a



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function of magnetic flux density (B) and frequency, all derived from standardized testing procedures. Both the American Society for Testing and Materials (ASTM) and the International Electrotechnical Commission (IEC) have developed standards governing these measurements, with testing instruments employing various approaches including the single sheet measurement [15,16], Epstein frame [14,15,17], and ring core or toroidal measurement setups [15,18,19].

The toroidal core characterization approach is based on a toroidal transformer induction method incorporating two or more windings wrapped around the test core to experimentally evaluate these magnetic properties [19–21]. To measure specific core loss, the core sample is shaped into a ring-like specimen, which is then wound with secondary and primary coils to provide voltage measurement and excitation respectively [22]. One of the key advantages of this method is its ability to capture the influence of lamination short circuits and thermal stresses with compressive stresses that arise during stacking. However, a notable limitation is that the standard toroidal core geometry does not replicate the actual stator core shape, which can result in core loss values being underestimated relative to those of the real stator core.

The Epstein frame operates on the principle of an unloaded transformer with two windings [23]. Rather than testing assembled cores, it measures core loss in loosely stacked lamination strips. Thus, its results reflect material behavior prior to core assembly. As a result, the Epstein frame cannot account for the influence of manufacturing processes such as welding, stacking, and other production steps on core loss [17].

The Single Sheet Tester (SST) evaluates an individual sheet of electrical steel by clamping it between two yokes. These yokes are constructed from ferromagnetic material with low loss [24]. Despite testing only one sheet at a time, the SST accommodates a broad range of excitation frequencies through a relatively simple testing procedure. Like the Epstein frame, it is better suited to characterizing the properties of the base material than to capturing the effects introduced by manufacturing processes. Nonetheless, the SST proves valuable as a quality control instrument and benchmark reference, helping to ensure consistent production of high-grade materials for use in motor manufacturing [25].

In the slot-wound stator core technique, both secondary and primary windings are threaded through the slots and around the stator back iron, generating an alternating flux that circulates through the back iron [26]. A key limitation of this approach is that it captures core loss only in the stator yoke, leaving the losses occurring in the stator teeth unaccounted. In the C-shape magnetizing coil technique, the yokes that excite produce varying magnetic field intensities, resulting in a non-uniform flux distribution across the stator cross-section [27]. Since the excitation cores are placed within the stator bore, the teeth are subjected to higher flux density levels than the stator back iron, which more closely mirrors actual motor operating conditions. The rotating magnetic field excitation method employs a magnetic excitation core whose configuration closely resembles that of an outer rotor AC motor stator [27,28]. This generates a rotating magnetic field for core loss evaluation. A significant drawback of this method, however, is that the large volume of the excitation core introduces considerable losses of its own, which can inflate the measured core loss figures. To compensate, the excitation core iron loss must be independently quantified using a separate evaluation technique and subsequently subtracted from the total loss.

A switched reluctance motor (SRM) has DC current excitation, typically with higher harmonic content. The flux path of a switched reluctance motor is also different than AC motors requiring advanced magnetic modeling to account for highly localized saturation [29]. This paper presents the design of an experimental setup for the core loss measurement of a stator core of an SRM with custom magnetizing yokes. The core loss

characteristics of an SRM core is measured experimentally over a wide frequency range from 100 Hz to 10,000 Hz and under varying flux density levels. The core loss characteristics of the SRM stator core are measured both with self- and mutual-flux excitation.

The remainder of this paper is organized as follows. Section 2 introduces the fundamental modeling of the core loss measurement method based on single-phase transformer induction and validates the time-domain extraction equations using a ring core Finite Element Analysis (FEA). Section 3 details the geometric design of the reduced scale 4/18 fractional SRM stator segment and the custom primary and mutual coupling excitation yokes. In Section 4, a 2D FEA model of this specific stator setup is conducted to validate the estimation of magnetic field quantities and to establish an effective volume calibration method for non-uniform flux distributions. Section 5 describes the physical construction, winding configurations, and instrumentation of the inverter-fed experimental test rig. Section 6 presents the experimental results, including the reconstructed dynamic B-H loops and an analysis of the impact of chemical etching on manufacturing-induced core losses. Finally, Section 7 provides the conclusions of the paper.

2. Modeling of Core Loss Measurement

In the proposed method, core loss estimation is performed using a measurement approach derived from the single-phase transformer operation principle. The primary winding is excited by alternating current. Secondary winding measures the voltage induced by the alternating magnetic field in the core material. The method is first applied to a ring specimen in an FEA model for validation. The inner radius of the ring core, R_1 , is 120 mm, and the outer radius, R_2 , is 140 mm. The stack length of the ring core is 10 mm. In the finite element model, Nippon 50H700 electrical steel material is assigned to the ring core. The time-domain waveforms of the applied current to the primary winding and the resulting induced voltage in the secondary winding are plotted for each simulation.

The methodology for extracting magnetic field quantities and computing core losses from terminal waveforms follows a consolidated, four-step algorithm applied to both virtual and experimental data:

Step 1: Data Acquisition & Filtering. Primary current $i_p(t)$ and secondary voltage $v_s(t)$ are sampled at 10 MS/s. A moving-average filter is applied to mitigate high-frequency noise, ensuring stability during numerical differentiation.

Step 2: Magnetic Flux Density, $B(t)$. Obtained via time integration of the filtered secondary voltage (Faraday's law):

$$B(t) = \frac{1}{N_s A_{eff}} \int v_s(t) dt \quad (1)$$

where N_s is number of turns, and A_{eff} is effective area.

Step 3: Magnetic Field Strength, $H(t)$. Estimated by mapping $B(t)$ through the inverse of the material's initial DC magnetization curve, preserving measured phase shifts:

$$H(t) = f_{BH}^{-1}(B(t)) \quad (2)$$

Step 4: Total Core Loss Computation. Specific total core loss is calculated by numerically integrating the area enclosed by the dynamic B-H loop over one cycle, which inherently captures hysteresis, eddy currents, and excess losses:

$$P_{total} = f \oint H(t) dB(t) \quad (3)$$

The 2D transient magnetic FEA simulations were configured with a high-density mesh (approximately 15,000 elements) to accurately capture localized flux variations, utilizing a maximum time step of 1 μ s to ensure strict numerical stability up to the 10 kHz excitation range. To rigorously validate the analytical extraction method, excitation current amplitudes were strategically selected (e.g., 5 A and 50 A) to represent both the linear, unsaturated magnetic region and the nonlinear, saturated region.

2.1. Evaluation of Eddy Current Loss Results

Instantaneous eddy current loss $p_e(t)$ is calculated using:

$$p_e(t) = \frac{1}{2\pi^2} k_c \left(\frac{dB}{dt} \right)^2 \quad (4)$$

$$k_c = \frac{\pi^2 d^2}{6\rho_e} \quad (5)$$

where d is the lamination thickness, ρ_e is the material resistivity, and B is the magnetic flux density.

Comparisons between the eddy current loss results obtained from direct FEA simulations and those calculated using (4) and (5) from the secondary induced voltage are shown in Figures 1 and 2. The magnetic flux densities and the corresponding instantaneous eddy current losses are compared to evaluate the accuracy and reliability of the proposed approach.

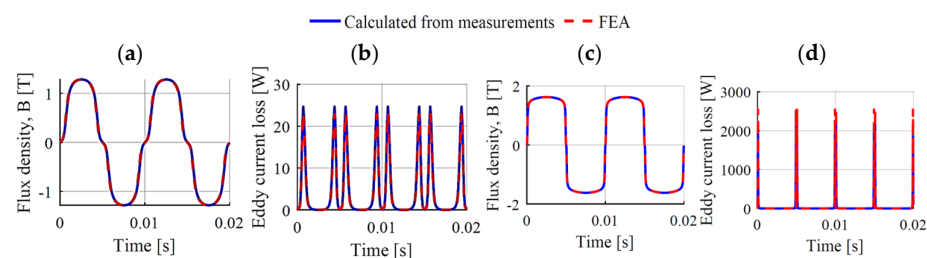


Figure 1. Comparison of FEA solver results and post-processed results based on virtual measurements from FEA at 100 Hz sine excitation in a ring core: (a) magnetic flux density at 5 A, (b) instantaneous eddy current loss at 5 A, (c) magnetic flux density at 50 A, (d) instantaneous eddy current loss at 50 A.

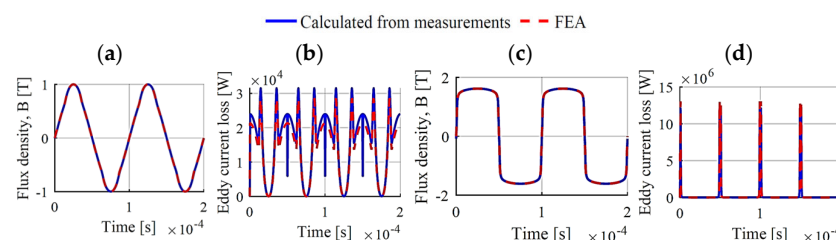


Figure 2. Comparison of FEA solver results and post-processed results based on virtual measurements from FEA at 10 kHz sine excitation in a ring core: (a) magnetic flux density at 3 A, (b) instantaneous eddy current loss at 3 A, (c) magnetic flux density at 50 A, (d) instantaneous eddy current loss at 50 A.

Figure 3 shows the trend of specific eddy current loss with respect to flux density. The results demonstrate good agreement between the proposed post-processing method and the FEA solver, particularly at higher flux levels and lower frequencies. The eddy current loss increases nonlinearly due to its dependence on the square of dB/dt at a given excitation frequency. Minor deviations are observed between the two methods at high frequencies and low flux levels, where numerical differentiation becomes more sensitive

to waveform distortion. The proposed method effectively captures the overall loss trend, validating the robustness across a broad operating range.

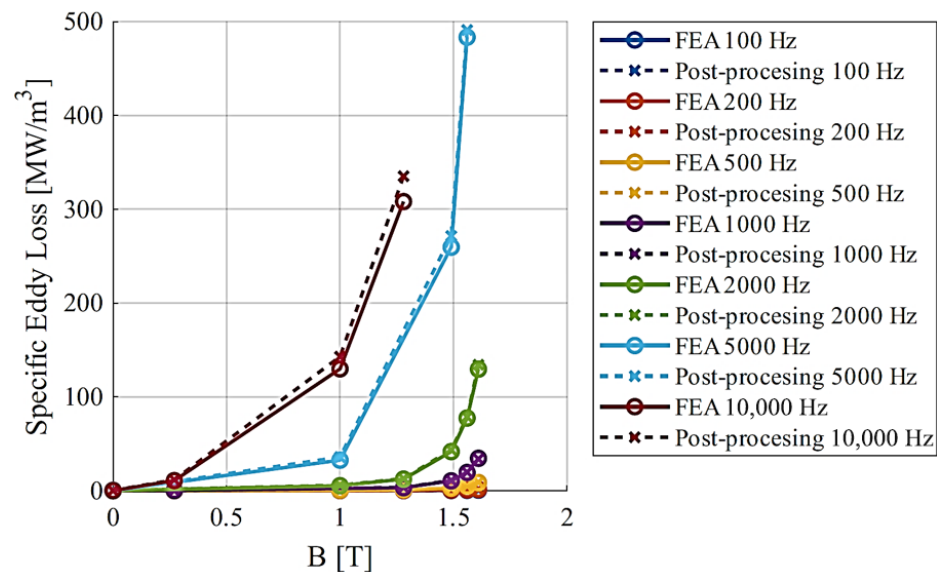


Figure 3. Comparison of specific eddy current loss between FEA solver results and calculated results based on post-processed virtual measurements from FEA for excitation frequencies between 100 Hz and 10 kHz in a ring core.

2.2. Evaluation of Hysteresis Loss Results

Hysteresis loss is determined by computing the energy dissipated from area of the magnetic flux density, B , and magnetic field intensity, H , loops in each electrical cycle. To accurately calculate the hysteresis loop, the initial DC magnetization curve and the coercivity and remanence points of the hysteresis loop are required at different frequencies. The coercivity and remanence values for Nippon Steel 50H700 are extracted from the hysteresis plots provided in the Nippon Steel datasheet [30]. Figure 4 shows the hysteresis loop constructed with those parameters.

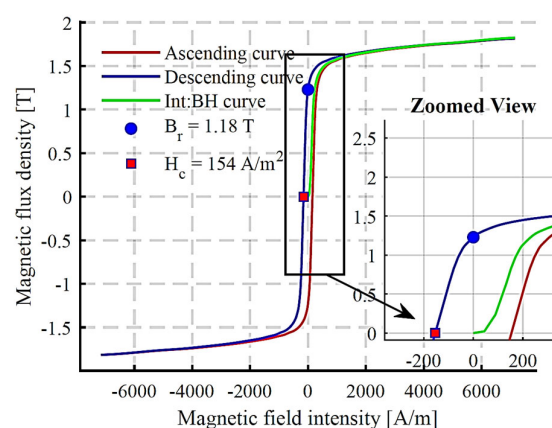


Figure 4. Hysteresis loop of Nippon Steel 50H700 showing coercivity, H_c , and remanence, B_r .

Figures 5 and 6 present comparisons between the magnetic flux density, magnetic field intensity, the hysteresis loop generated directly from the FEA tool, and the same variables calculated by post-processing of the voltage and current waveforms from the same FEA model in a ring core. Time-domain plots of B and H are provided to visualize the agreement between the two methods.

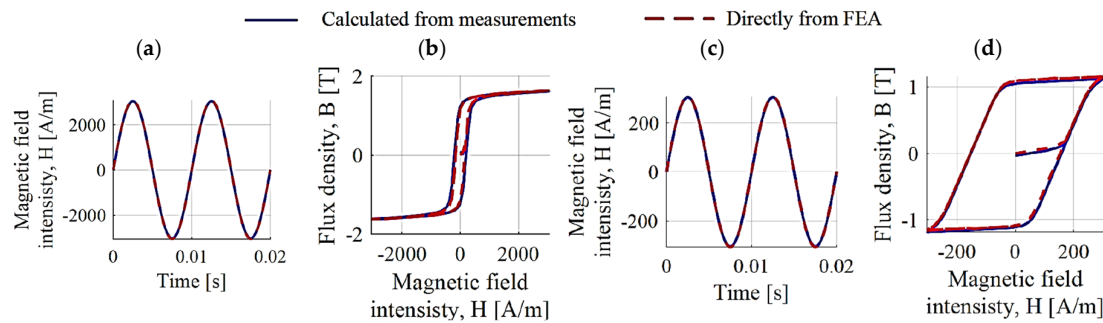


Figure 5. Comparison of FEA solver results and calculated results based on post-processed virtual measurements from FEA at 100 Hz sine excitation in a ring core: (a) field strength at 50 A, (b) B-H loops at 50 A, (c) field strength at 5 A, (d) B-H loops at 5 A.

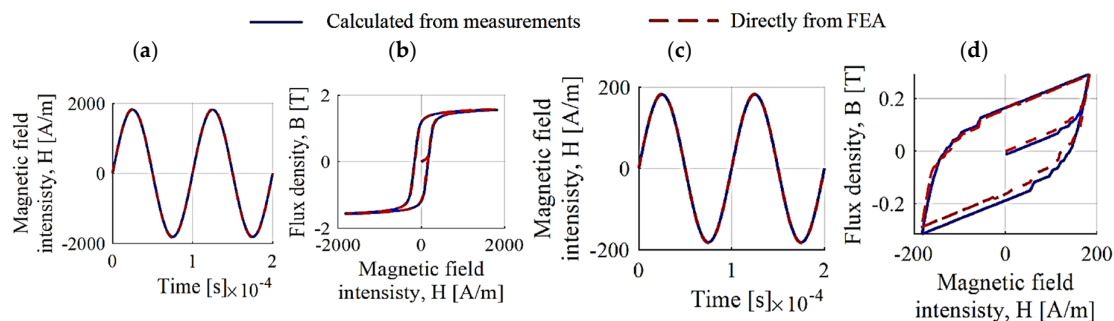


Figure 6. Comparison of FEA solver results and calculated results based on post-processed virtual measurements from FEA at 10 kHz sine excitation in a ring core: (a) field strength at 30 A, (b) B-H loops at 30 A, (c) field strength at 3 A, (d) B-H loops at 3 A.

A comparison of hysteresis loss is conducted across various frequencies and flux density levels. Figure 7 shows the trend of specific hysteresis loss with respect to flux density and further validates the proposed hysteresis loss estimation method. The results from the proposed hysteresis loss modeling method closely follow the trend of the FEA-calculated results, accurately capturing the linear increase in loss with both frequency and flux density.

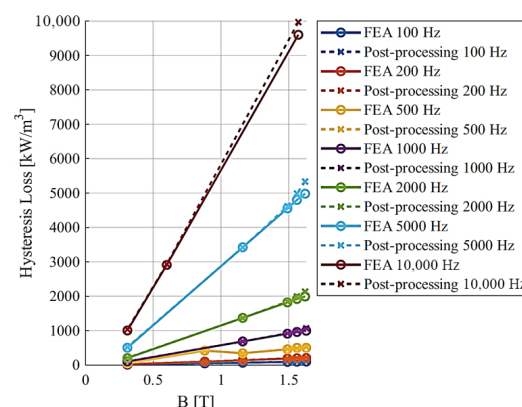


Figure 7. Comparison of specific hysteresis loss from FEA solver results and calculated results based on post-processed virtual measurements from FEA for excitation frequencies between 100 Hz and 10 kHz in a ring core.

3. Experimental Setup for Core Loss Measurement of a Switched Reluctance Machine Stator Core

A geometrically scaled-down stator core is used for experimental core loss characterization. A larger core requires higher magnetic flux to reach the high flux density across the core, which demands higher current, and results in higher induced voltage at high frequencies and requires higher power. A scaled down stator enables practical, representative high-frequency tests with manageable voltage and current levels. Figure 8 shows the complete 18-slot stator and the 4-slot fractional geometries. Table 1 summarizes the details of the stator geometry.

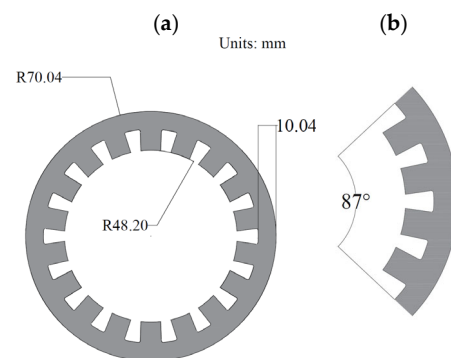


Figure 8. SRM stator geometries: (a) complete 18-slot stator, (b) 4/18 fractional stator segment with 4 slots.

Table 1. Reduced-scale SRM stator geometry parameters.

Parameter	Value [mm]
Tooth width at the air gap	11.06
Tooth height	11.8
Back iron thickness	10.04
Stack height	25

While the initial 2D FEA validation on the ring core utilized Nippon 50H700 to establish a theoretical benchmark using a standard, widely documented material, the scaled stator experimental prototype utilizes JFE 35JN210 silicon steel. The material 35JN210 is a non-oriented electrical steel and is commonly used in high-speed rotating machines, making it the appropriate choice for evaluating the practical applicability of the proposed core loss method in an actual SRM architecture. The major material properties are shown in Table 2.

Table 2. Material properties of JFE Steel 35JN210.

Property	Value
Electrical conductivity (σ)	1.7×10^6 S/m
Lamination thickness (d)	0.35 mm
Relative permeability (μ_r , initial)	2000
Saturation flux density	1.9 T
Density (ρ)	7650 kg/m ³
Maximum specific core loss	2.10 W/kg at 1.5 T, 50 Hz

The excitation setup is designed to closely emulate the magnetic conditions in an SRM stator during operation. One yoke is positioned to excite a stator region corresponding to the case, where magnetic flux typically passes through two adjacent stator teeth of an active

phase (e.g., Phase A), as shown in Figure 9a. This represents the primary excitation condition. An additional yoke is introduced to account for the influence of magnetic coupling between neighboring phases (e.g., Phases B and C). Figure 9b shows the configuration of the mutual excitation yoke.

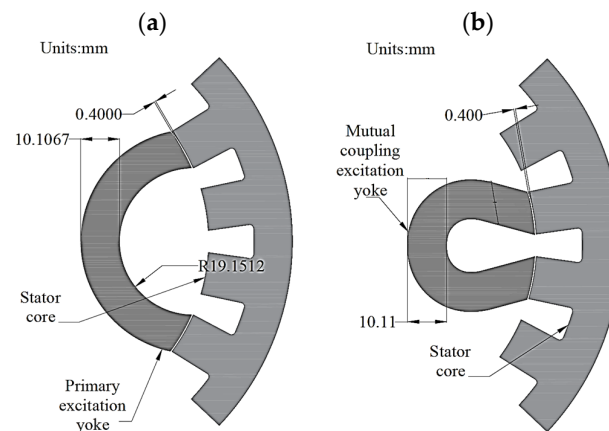


Figure 9. Excitation yoke configurations used in core loss measurements: (a) primary excitation of a single phase, and (b) mutual coupling.

In a functioning 18/12 SRM, exciting a phase generates six symmetrical magnetic poles around the stator circumference. During phase commutation, two phases can be energized simultaneously, which alters the localized flux path. To accurately capture this overlapping magnetic behavior, a four-pole stator segment was utilized instead of a three-pole geometry, paired with two distinct excitation yokes. We evaluate the core loss by applying a time-varying field via a sinusoidal current to these yokes rather than using a rotating magnetic field. Rotating magnetic field core measurement techniques are complex methods with limited applicability to stator core iron loss measurements. Furthermore, relying on a rotating assembly can lead to inaccurate measurements due to the large volume of the required excitation core. Therefore, the mutual coupling yoke is specifically designed to isolate and characterize these overlapping flux paths directly, ensuring high measurement accuracy without the structural complexities and inherent errors of a rotating setup.

Figure 10 shows the primary winding and secondary winding arrangement for both the primary yoke and mutual coupling yoke configurations. In the primary excitation yoke, primary windings are implemented with 16 American Wire Gauge (AWG) magnet wire. The primary excitation yoke uses six strands per turn and nine turns in total, providing sufficient ampere-turns to generate the required flux level. The secondary yoke emulates the mutual coupling effect. A 17 AWG wire size is considered with four strands and nine turns due to space constraints and lower current requirement. Secondary windings are configured similarly in both excitation yoke configurations. The secondary windings are designed with 20 AWG wire with a single strand and nine turns to measure the induced voltage. A single-phase two-leg unipolar inverter provides excitation voltage using Sinusoidal Pulse Width Modulation (SPWM) to the excitation windings. Unipolar SPWM is chosen for its lower harmonic distortion, which reduces filtering requirements.

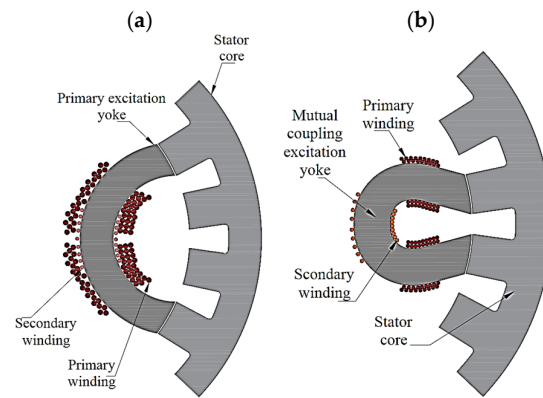


Figure 10. Excitation coil configurations for the yokes used in the measurement setup, (a) primary excitation yoke (6 strands, 9 turns) and (b) mutual coupling yoke (4 strands, 9 turns).

The phase currents of an SRM are typically non-sinusoidal. The phase current waveforms change based on the operating condition and they can have relatively high-order harmonic content. For this reason, the proposed experimental setup is designed to characterize the core loss of an SRM stator at high flux density and high frequency. By modeling the core loss characteristics at high frequency and high flux density for self- and mutual-flux paths, the core loss in the dynamic operation with varying current and flux levels can be calculated [7].

4. 2D Finite Element Modeling and Estimation of Magnetic Flux Density and Field Strength

Figure 11 shows a representation of the combined model after integrating the FEA model for core loss measurement with the single-phase inverter model. The total magnetomotive force (MMF) is not entirely dropped across the stator core due to the presence of an air gap between the excitation yoke and the stator. Part of the MMF is distributed across the air region, while the rest is shared by the magnetic materials (yoke and stator). The total MMF (F) can be expressed as:

$$F = H_{yoke}l_{yoke} + H_{stator}l_{stator} + H_{air}l_{air} \quad (6)$$

where H and l are the magnetic field strength and path length for each region respectively.

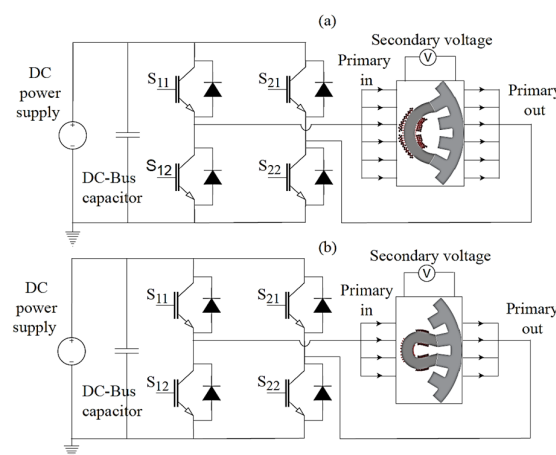


Figure 11. Integrated stator core excitation setup with the single-phase inverter model: (a) primary excitation yoke, (b) mutual coupling yoke.

A substantial portion of the total MMF drops across the air gap because the air gap has lower permeability than the core material. Isolation of the H_{stator} and H_{yoke} from the measured current might not be practical, since the magnetic flux density, $B(t)$, within the core is already estimated from the integrated secondary voltage. The $H(t)$ in the core is derived using the inverse of the initial magnetization curve, as given by (2). A comparison between FEA and B-H curve-based results is shown in Figure 12 to illustrate the estimation approach for magnetic field strength. The $B(t)$ and $H(t)$ are obtained from the FEA simulation, and selected $B(t)$ values are mapped to $H(t)$ using the static B-H curve of the core material.

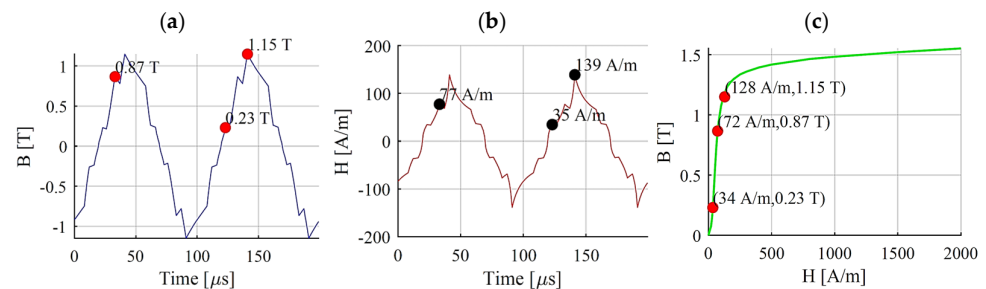


Figure 12. Illustration of magnetic field intensity estimation: (a) FEA-derived flux density $B(t)$, (b) FEA-derived field strength $H(t)$, and (c) $H(t)$ estimation based on the inverse B-H curve using measured $B(t)$.

4.1. Validation of Flux Density and Magnetic Field Intensity

Using the time-domain virtual measurement approach defined in Section 2, the calculated $B(t)$ and estimated $H(t)$ derived from the simulated terminal waveforms are compared with the direct internal field results obtained from the FEA model under identical excitation conditions. The comparison focuses on the yoke region where the coil is placed. Figure 13 shows the selected location within the yoke where time-varying magnetic field data is extracted from the FEA model. This point is positioned directly beneath the secondary winding and represents the region of interest for comparing estimated and simulated field quantities. This virtual measurement-based estimation verifies the accuracy of the mathematical reconstruction and evaluates its ability to capture complex magnetic behavior under non-sinusoidal, inverter-fed excitation prior to physical experimentation. Validation is performed across a frequency range of 100 Hz to 10 kHz under varying excitation levels.

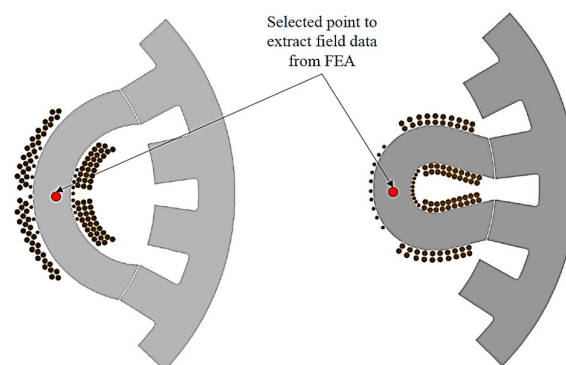


Figure 13. Integrated excitation setup with the single-phase inverter model.

Figures 14–16 show the waveforms extracted from FEA, and comparison of B and H waveforms under the primary excitation yoke configuration with time-domain FEA results at the selected point. Two excitation frequencies, 100 Hz and 10 kHz, are applied, and

different voltage levels are used to demonstrate the accuracy of the proposed method. The corresponding excitation voltages and coil currents are also included for reference.

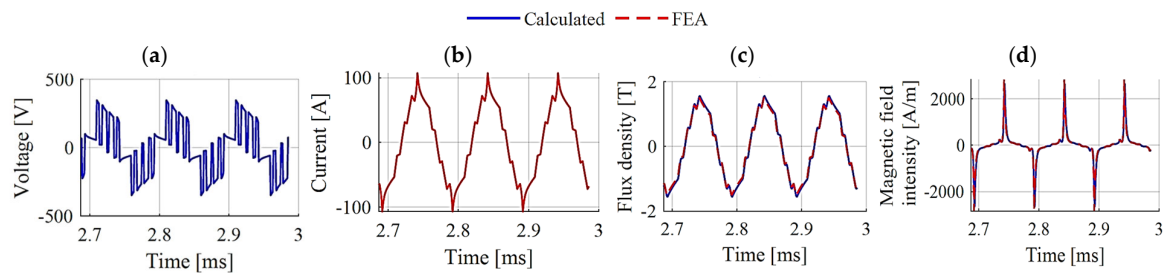


Figure 14. Waveforms at 10 kHz fundamental, 300 Vdc, and 0.8 modulation index in the primary yoke configuration: (a) FEA-extracted induced voltage in the secondary winding, (b) FEA-extracted current in the primary winding, (c) comparison of calculated and FEA-simulated flux density (B), and (d) comparison of calculated and FEA-simulated magnetic field strength (H).

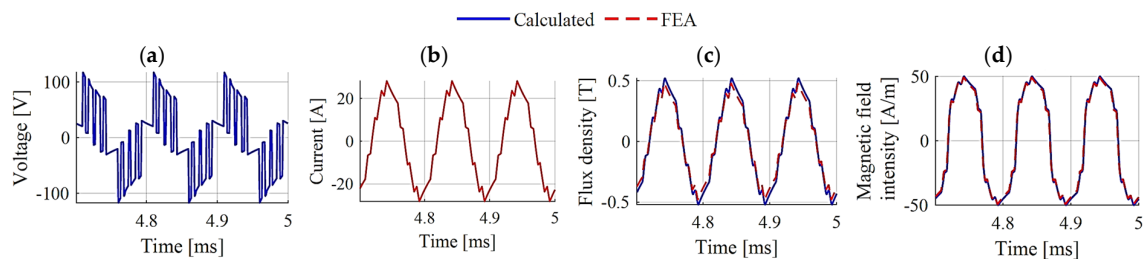


Figure 15. Waveforms at 10 kHz fundamental, 100 Vdc, and 0.8 modulation index in the primary yoke configuration: (a) FEA-extracted induced voltage in the secondary winding, (b) FEA-extracted current in the primary winding, (c) comparison of calculated and FEA-simulated flux density (B), and (d) comparison of calculated and FEA-simulated magnetic field strength (H).

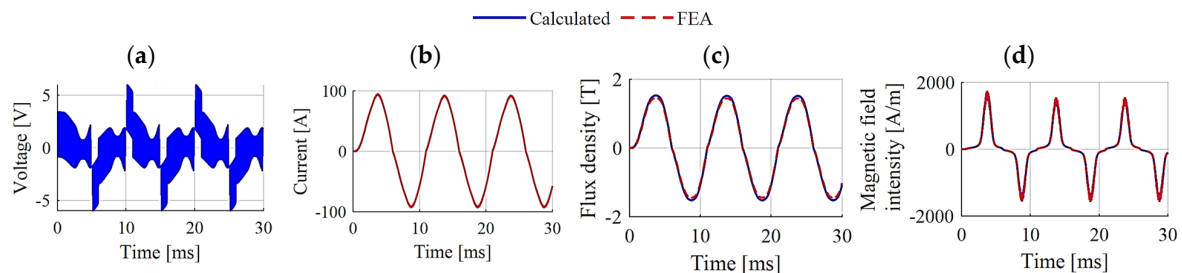


Figure 16. Waveforms at 100 Hz fundamental, 10 Vdc, and 0.8 modulation index in the primary yoke configuration: (a) FEA-extracted induced voltage in the secondary winding, (b) FEA-extracted current in the primary winding, (c) comparison of calculated and FEA-simulated flux density (B), and (d) comparison of calculated and FEA-simulated magnetic field strength (H).

Figures 17–19 present the virtually extracted waveforms from FEA, and the comparison of B and H waveforms, with the corresponding excitation voltages and coil currents obtained under the mutual excitation yoke configuration with time-domain FEA results at the selected point. Two excitation frequencies, 100 Hz and 10 kHz, are considered, and various voltage levels are used to validate the accuracy of the proposed method.

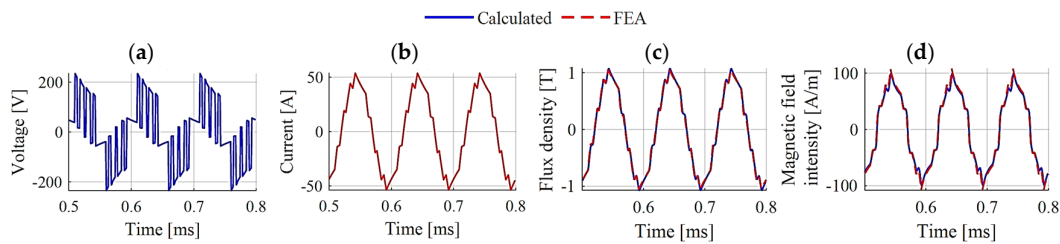


Figure 17. Waveforms at 10 kHz fundamental, 200 Vdc, and 0.8 modulation index in the mutual excitation yoke configuration: (a) FEA-extracted induced voltage in the secondary winding, (b) FEA-extracted current in the primary winding, (c) comparison of calculated and FEA-simulated flux density (B), and (d) comparison of calculated and FEA-simulated magnetic field strength (H).

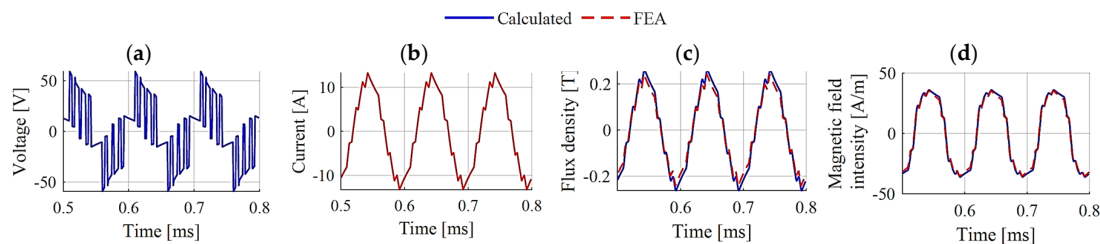


Figure 18. Waveforms at 10 kHz fundamental, 50 Vdc, and 0.8 modulation index in the mutual excitation yoke configuration: (a) FEA-extracted induced voltage in the secondary winding, (b) FEA-extracted current in the primary winding, (c) comparison of calculated and FEA-simulated flux density (B), and (d) comparison of calculated and FEA-simulated magnetic field strength (H).

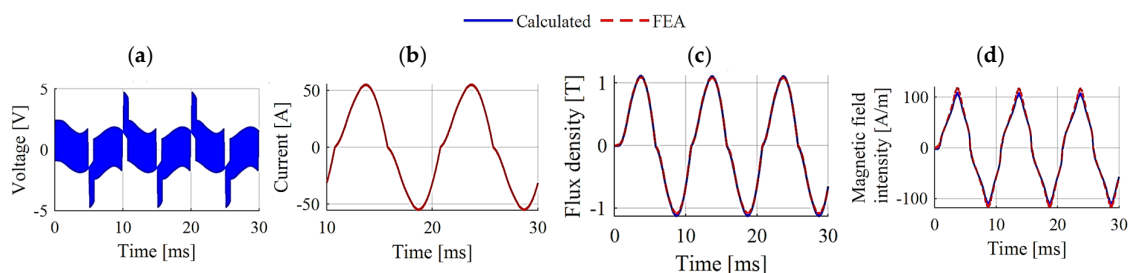


Figure 19. Waveforms at 100 Hz fundamental, 7.5 Vdc, and 0.8 modulation index in the mutual excitation yoke configuration: (a) FEA-extracted induced voltage in the secondary winding, (b) FEA-extracted current in the primary winding, (c) comparison of calculated and FEA-simulated flux density (B), and (d) comparison of calculated and FEA-simulated magnetic field strength (H).

The waveform comparisons in Figures 14–19 demonstrate good agreement between the estimated and FEA-simulated magnetic quantities under various excitation conditions. Flux density, B , and magnetic field intensity, H , show consistent waveforms and magnitudes across the selected voltage levels and frequency settings, confirming the accuracy of the proposed estimation method in capturing the magnetic behavior under inverter-fed excitation.

4.2. Flux Distribution and Effective Volume

The calculated magnetic flux density in the stator yoke does not represent the overall magnetic behavior across the entire stator due to geometric non-uniformities and localized excitation effects. To address this, a flux density estimation methodology is proposed to estimate the internal flux distribution and define an effective magnetic volume.

As shown in Figure 20, the stator is divided into twelve distinct regions based on the geometry and expected variations in the magnetic flux path. These regions are labeled as

A_1 to A_{12} , with A_1 corresponding to the excitation yoke region under the primary yoke configuration. Flux density measurements are extracted from this region, and the ratio of flux density in each area is calculated relative to A_1 to characterize the spatial variation.

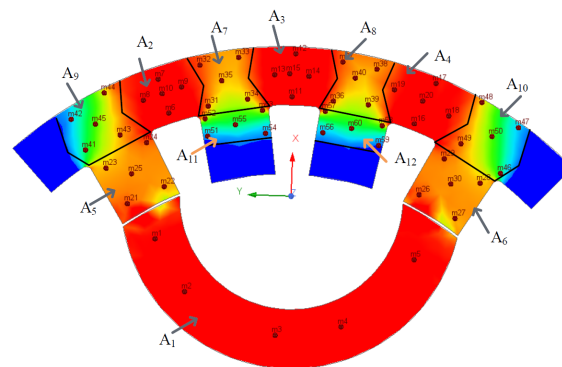


Figure 20. Flux density distribution and the segmented areas of the stator showing the selected points for the flux density mapping.

Figure 21 presents the flux density ratios for a 10 kHz excitation frequency. The ratio of flux density in any selected region compared to A_1 remains nearly constant across different excitation current levels. This consistency allows the use of a normalized mapping approach, where a reference measurement from A_1 can be scaled to estimate the flux density distribution in the entire stator.

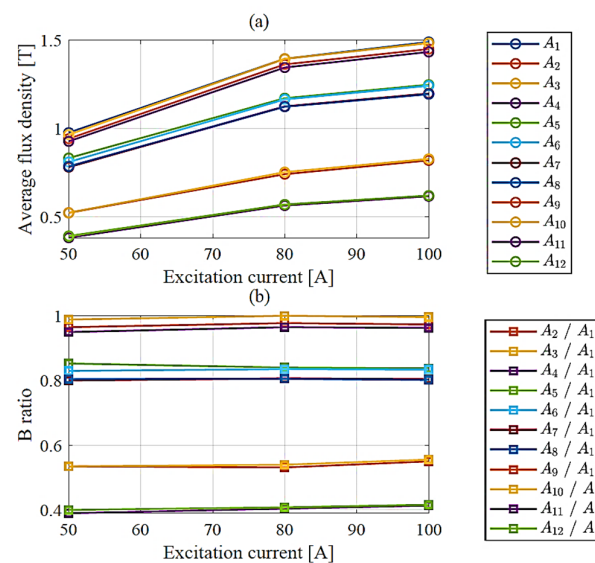


Figure 21. Flux density distribution analysis under different excitation levels at 10 kHz frequency: (a) average flux density in different areas at multiple excitation currents, and (b) ratio of average flux density compared to the reference area (A_1).

The approach of mapping the spatial flux density distribution across the stator core relies on estimating local flux densities for different regions. As shown in Figure 21b, once the flux density in the reference area A_1 is measured, the flux densities in the other areas can be determined using the geometry-dependent ratios. This assumption is reasonable because the spatial pattern is mainly governed by the stator and yoke geometries under similar excitation. Figure 22 illustrates the complete workflow of the equivalent effective volume calibration and validation process.

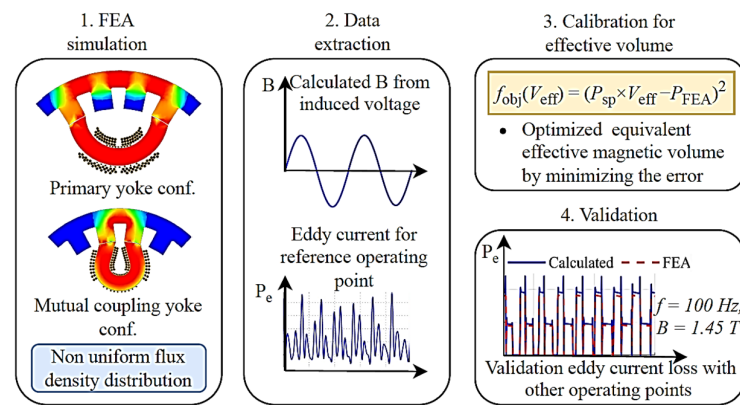


Figure 22. Workflow of the equivalent effective volume calibration and validation process.

To determine the equivalent effective magnetic volume (V_{eff}), an optimization process was established to minimize the error between the calculated losses and the FEA-simulated losses. The objective function for this calibration is defined as:

$$f_{obj}(V_{eff}) = (P_{sp} \times V_{eff} - P_{FEA})^2 \quad (7)$$

where P_{sp} is the specific eddy current loss derived from post-processed virtual measurements, and P_{FEA} is the total eddy current loss obtained directly from the FEA solver.

The optimization algorithm evaluated a bounded range of volumetric fractions to find the optimal equivalent distribution for the non-uniform stator regions. Minimizing the objective function yielded a final calibrated effective stator volume of $1.98 \times 10^{-5} \text{ m}^3$ for the primary excitation configuration and $9.76 \times 10^{-6} \text{ m}^3$ for the mutual coupling configuration. These calibrated volumes correspond to approximately 67.3% and 61.4% of the physical flux-distributed stator volumes ($2.94 \times 10^{-5} \text{ m}^3$ and $1.59 \times 10^{-5} \text{ m}^3$), respectively. Because the flux density distribution within the excitation yoke regions is assumed to be uniform, their equivalent effective volumes remained equal to their actual physical volumes. To ensure the robustness of this calibration, the calibrated V_{eff} was subsequently validated at varying operating points, such as 100 Hz and 1.45 T, confirming high estimation accuracy across different excitation conditions without requiring recalibration.

5. Experimental SRM Core Loss Measurement Setup

Applying the time-domain extraction methodology and effective volume calibration validated in the previous sections, this section details the physical construction of the test rig. This section presents the development of the experimental setup of the proposed core loss measurement method. The experimental setup replicates the same stator and excitation yoke geometry used in simulation, with mechanical and electrical implementation to facilitate core loss estimation under inverter-fed excitation.

5.1. Stator and Yoke Construction

The stator core and custom excitation yokes are constructed using JFE Steel 35JN210 electrical steel laminations, with a thickness of 0.35 mm and a stacking factor of 0.95. Wire electrical discharge machining (EDM) is used for cutting the laminations. A 4/18 fractional stator segment is implemented. Two excitation yokes are fabricated. One is for primary excitation, and the other to emulate mutual coupling from adjacent phases. The manufactured yokes and stator segments are shown in Figure 23.

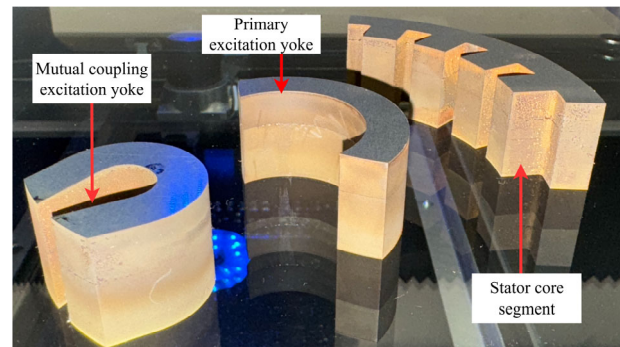


Figure 23. Fabricated excitation yokes and stator core segment before the chemical etching process.

Custom-designed mounts made of PV6 material are used to hold the stator and yokes in place during testing. The material was chosen for its non-magnetic properties and mechanical rigidity to withstand the magnetic attraction during excitation. Figure 24 shows the mechanical design of two yokes and the stator core assembly.

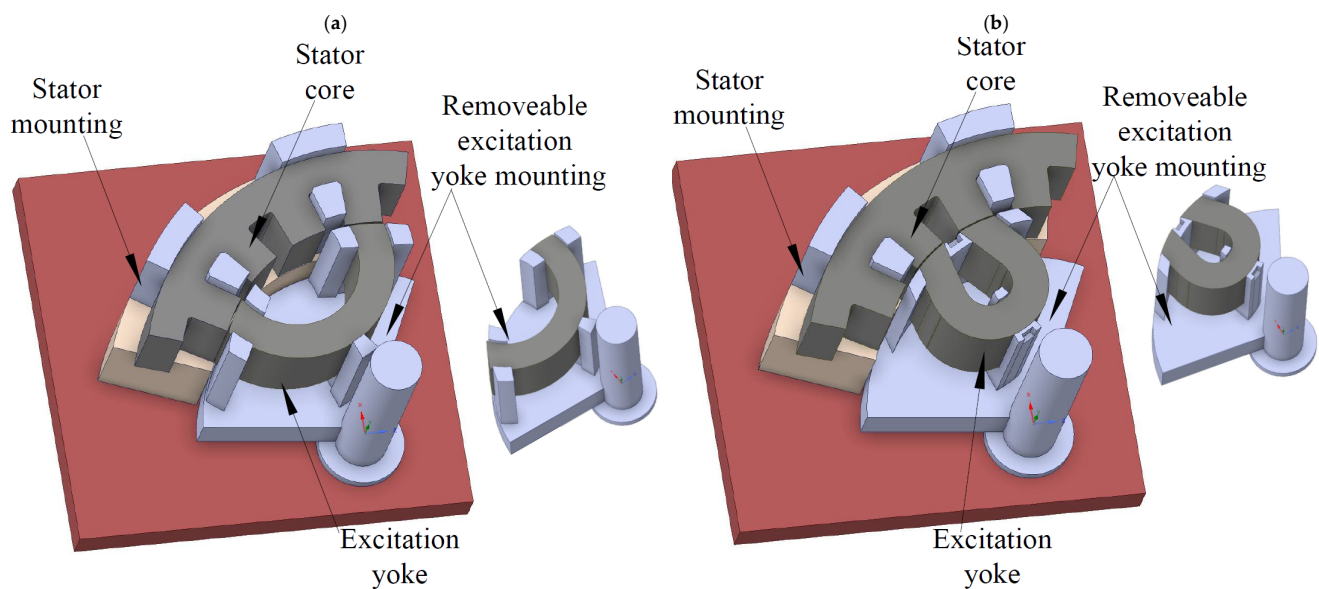


Figure 24. CAD model of stator excitation yoke, and mounting assembly with 0.4 mm air gap: (a) primary excitation yoke, and (b) mutual coupling excitation yoke.

5.2. Winding Configuration

Custom bobbins are designed to fit the cross-section of the excitation yokes and facilitate efficient coil winding. Figure 25a shows the designed two bobbins for the primary excitation yoke. Figure 25b shows the non-removable winding retention method for the mutual coupling excitation yoke. Table 3 presents the winding specifications and the current density for each excitation yoke configuration.

Table 3. Winding specifications and current density.

Yoke	Wire Strands	Current [Arms]	Current Density [A/mm ²]
Primary	6 × 16 AWG (6 × 1.31 mm ²)	70	8.9
Mutual	4 × 17 AWG (4 × 1.04 mm ²)	36	8.65

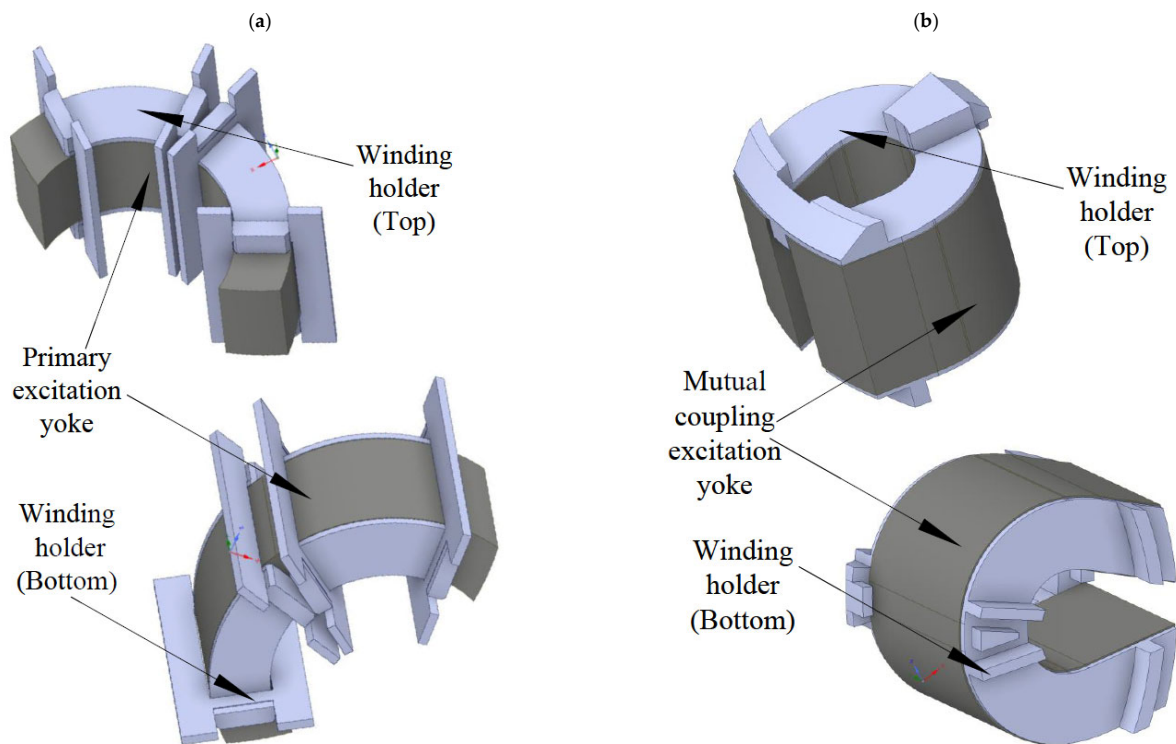


Figure 25. Custom-designed bobbins for excitation yoke windings: (a) primary excitation yoke, and (b) mutual coupling excitation yoke.

Figure 26 shows the excitation yokes with windings for the experimental setup. In the primary yoke configuration, each bobbin contains three strands of 16 AWG wire. The total number of turns is nine, with each strand having an approximate length of 1 m. The current rating of the primary excitation yoke is approximately 70 arms. For the mutual coupling excitation yoke, windings are implemented on both sides of the yoke using two strands of 17 AWG wire per side, totaling nine turns. The rated excitation current is approximately 36 Arms. The secondary windings are identical in both yokes and consist of a single strand of 20 AWG wire with nine turns. The secondary winding, which functions as the search coil for measuring the induced voltage, is wound close to the yoke and directly beneath the excitation winding to minimize flux leakage. All current densities are kept below 10 A/mm^2 , which is acceptable for short-duty testing.

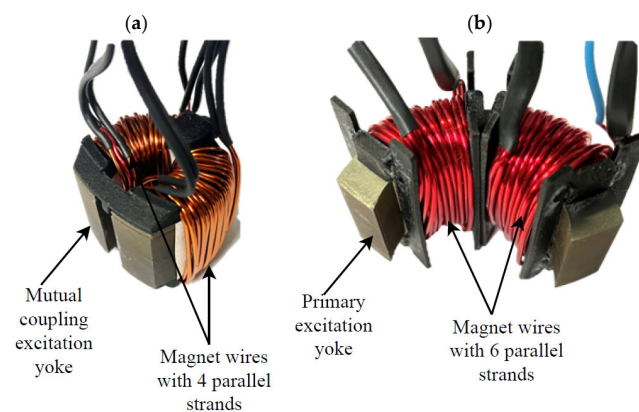


Figure 26. Excitation yokes with excitation windings and search coil: (a) mutual coupling excitation yoke, and (b) primary excitation yoke.

6. Experimental Results

The magnetic loss in the SRM stator core was evaluated through reconstruction of the hysteresis loop using measured voltage and current waveforms. The magnetic flux density $B(t)$ was obtained by time integration of the secondary induced voltage, while the magnetic field strength $H(t)$ was estimated using the inverse of the initial magnetization curve. This initial magnetization curve was not sourced from manufacturer datasheets. Instead, it was experimentally obtained directly from the fabricated core. This was achieved by mounting the excitation assembly with a zero-air gap condition against the stator core and applying a low-frequency (100 Hz) quasi-static excitation. The initial rising magnetization curve was then extracted by evaluating the first quarter-cycle of the acquired voltage and current waveforms. Mapping $B(t)$ through this static inverse curve yields a base field strength that is perfectly in phase with the flux density. However, under high-frequency inverter-fed operation, eddy currents and dynamic hysteresis cause the dynamic $H(t)$ to lag $B(t)$. To accurately reproduce this dynamic hysteresis loop, the phase difference $\phi = \angle v(t) - \angle i(t)$ between the measured secondary induced voltage and the primary excitation current was extracted. This measured phase difference was then utilized to introduce the corresponding temporal phase shift between $B(t)$ and the estimated $H(t)$, effectively widening the reconstructed loop to reflect the true dynamic energy dissipation. The specific hysteresis loss was then calculated from the area enclosed by the reconstructed B–H loop multiplied by the excitation frequency. This approach enables direct evaluation of magnetic loss under inverter-fed excitation without relying on manufacturer material data, which may not reflect the effects of fabrication and assembly. The reconstructed B–H loops therefore represent the actual magnetic behavior of the processed stator core.

The reconstructed B–H loops obtained using the mutual-coupling excitation configuration are shown in Figure 29 for excitation frequencies ranging from 100 Hz to 10 kHz. As the frequency increases, the loop area expands, indicating higher hysteresis loss due to increased dynamic effects. In addition, slight distortion of the loop shape can be observed at higher frequencies, reflecting the influence of non-sinusoidal excitation and waveform harmonics.

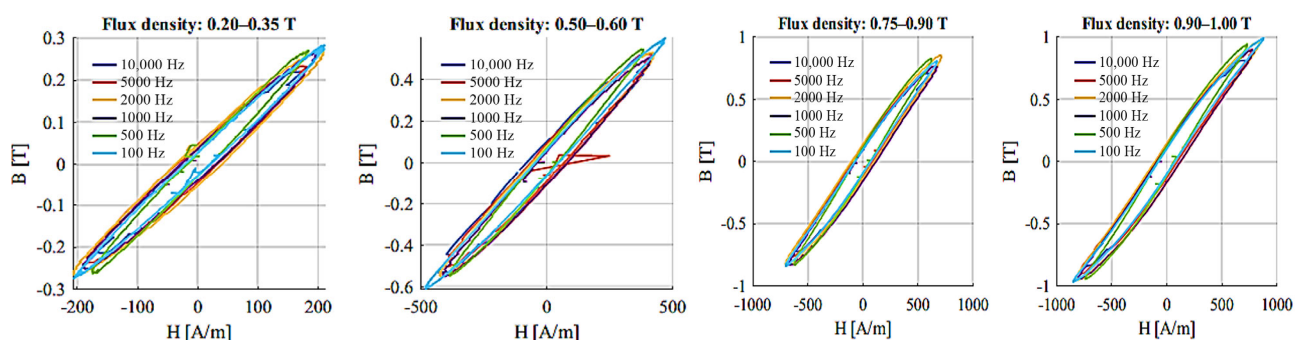


Figure 29. B–H loops from experimental measurements at different excitation frequencies and peak flux density ranges for the mutual coupling.

Similarly, Figure 30 presents reconstructed B–H loops for the primary excitation configuration. Compared to the mutual-coupling case, the primary excitation allows higher achievable flux-density levels, resulting in wider loops and increased loss. The comparison between the two excitation configurations highlights the influence of flux distribution and excitation path on the measured magnetic response.

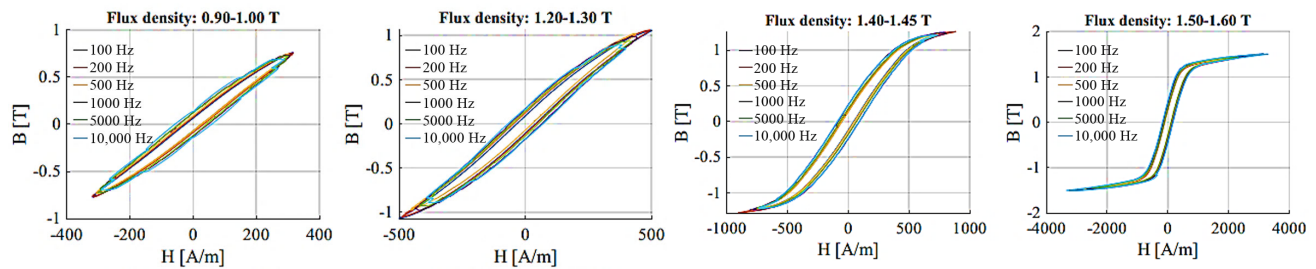


Figure 30. B–H loops from experimental measurements at different excitation frequencies and peak flux density ranges for the primary excitation yoke setup before the etching process.

Across both configurations, the reconstructed loops demonstrate consistent trends over the frequency range of 100 Hz to 10 kHz, confirming the robustness of the proposed measurement and reconstruction method.

To investigate the impact of manufacturing-induced damage on magnetic loss, measurements were conducted before and after chemical etching of the stator core. The etching process removes surface burrs and reduces interlaminar electrical contact, thereby mitigating eddy-current paths introduced during machining.

The effect of etching on the measured core loss for the primary excitation configuration is shown in Figure 31. The results indicate a clear reduction in total core loss after etching across the tested frequency range. The reduction becomes more pronounced at higher frequencies, where eddy-current effects dominate the total loss.

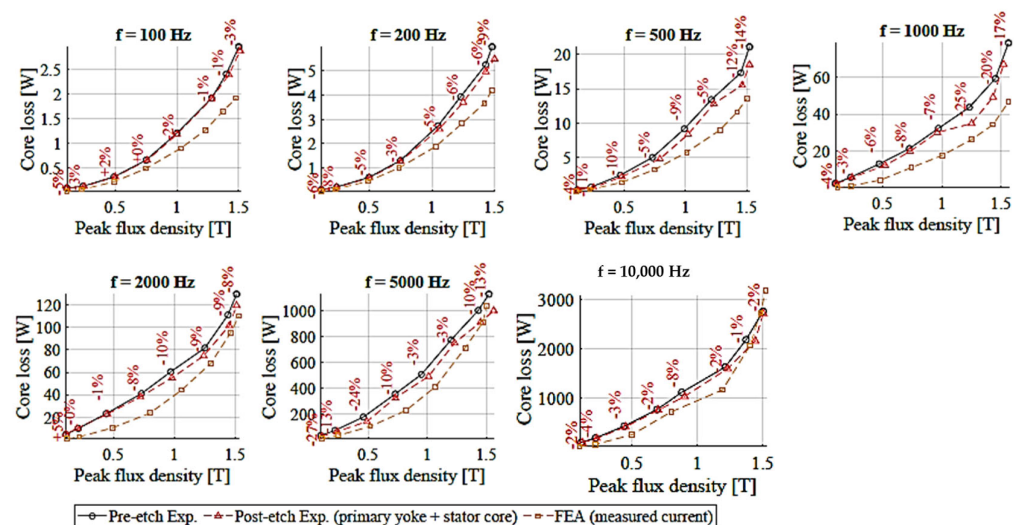


Figure 31. Comparison of experimentally measured core loss before and after full etching (primary excitation yoke + stator core) across different excitation frequencies under different flux levels.

A similar trend is observed for the mutual-coupling excitation configuration, as shown in Figure 32. In this case, the post-etched core consistently exhibits lower loss compared to the pre-etched condition over the entire frequency range. Although the absolute loss levels differ due to the distinct flux paths, the relative reduction confirms that the improvement is not configuration-specific.

The observed reduction in total core loss is primarily attributed to a decrease in the eddy-current component. This is consistent with the physical effect of etching, which removes burrs and reduces partial interlaminar short circuits, thereby increasing the effective insulation between laminations. These results highlight the significant influence of manufacturing processes on the magnetic performance of laminated cores and demonstrate that post-processing treatments can substantially improve core loss characteristics.

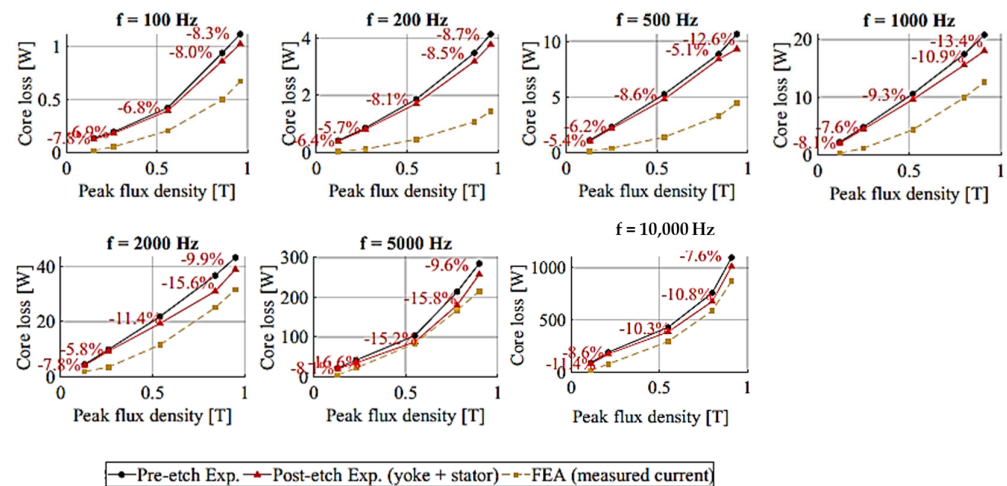


Figure 32. Comparison of experimentally measured core loss before and after full etching (mutual coupling excitation yoke + stator core) across different excitation frequencies under different flux levels.

The strong agreement between the proposed method and the FEA results stem from the method's foundation in core transformer induction theory, which inherently captures the fundamental relationship between induced voltage and internal flux. The minor discrepancies observed at the highest frequencies (e.g., 10 kHz) are primarily attributable to the sensitivity of numerical differentiation to high-frequency waveform distortion. The physical impact of the inverter-fed excitation is clearly visible in the B-H loops; high-frequency switching harmonics introduce localized non-sinusoidal distortions and expand the loop area due to dynamic hysteresis effects. Furthermore, comparing the primary and mutual coupling configurations reveals that mutual coupling alters the active flux path and local saturation levels, resulting in distinctly different absolute loss profiles. Ultimately, these results carry an important implication for SRM design: because manufacturing and inverter-driven harmonics can significantly increase core losses, designers should not rely solely on standard manufacturer datasheets but must instead utilize empirical characterization of the assembled stator to accurately predict machine efficiency.

7. Conclusions

This paper presented an experimental setup and analytical methodology for characterizing core loss in a switched reluctance motor (SRM) stator. A time-domain reconstruction technique based on single-phase transformer induction theory was developed to extract magnetic flux density and field strength directly from induced voltage and current waveforms. The methodology was validated through 2D Finite Element Analysis (FEA) on a ring core and subsequently applied to a physically scaled-down 4/18 fractional SRM stator core constructed from 0.35 mm JFE Steel 35JN210 laminations. Custom-designed primary and mutual coupling excitation yokes were fabricated to emulate the distinct single-phase and coupled flux paths typical in an SRM. Measurements were conducted using a two-level voltage source inverter across a broad frequency range from 100 Hz to 10,000 Hz to accurately capture high-frequency behavior. To address non-uniform flux distribution, a normalized flux density mapping approach was established by dividing the stator into twelve distinct geometric regions. The proposed method enables the reconstruction of dynamic B-H loops directly from measured signals without relying solely on standard manufacturer material data, which often omits manufacturing effects. Experimental comparisons conducted before and after a chemical etching process revealed a clear, measurable reduction in total core losses. The core loss reduction was shown to be highly dependent on excitation

frequency, becoming significantly more pronounced at high excitation frequencies where the eddy-current component dominates the total magnetic loss.

While the proposed methodology provides a practical approach for evaluating manufacturing-induced core losses in assembled stators, some limitations should be noted. The experimental setup uses a scaled down geometry to characterize the core loss at high frequency and high flux density. Conducting the same experiment to a complete stator would require a larger power supply and inverter. The tests were conducted under short-duty cycles to maintain a consistent temperature. Conducting the tests for longer excitation conditions might require active cooling of the excitation coil. The air gap between excitation yoke and the stator is maintained using a specific tool designed to hold the stator and excitation core in place. It might be challenging to design a specific tool for different stators measured. Therefore, a mechanism is needed to position the excitation yoke inside the stator and maintain consistent air gap length.

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