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Core Loss Estimation for an Inverter-fed Induction Motor With More Accurate Realisation of Material Non-linearity and Impact of Hysteresis Minor Loops

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Abstract—The work presents a fast and accurate core loss evaluation method for inverter-fed induction machines. The proposed method takes into account the influence of time variation of magnetic permeability on the eddy current loss. Furthermore, the proposed method presents an accurate approach to precisely evaluate the hysteresis loss. The results obtained from the proposed method are compared with the existing methods for a 1.5kW inverter-driven induction motor. The method shows sufficiently high accuracy with reduced complexity and implementation-time for a wide range of inverter switching frequency and motor's loading.

Index Terms—Induction motor, core loss, eddy current loss, hysteresis loss, motor efficiency, inverter, pulse-width modulation (PWM) supply, finite element.

I. INTRODUCTION

Evaluation of core-loss (CL) is one of the critical challenges to precisely determine the efficiency of an inverter fed induction motor (IF-IM). Inverter power supply harmonics increase the rate of change of magnetic field densities at different segments of the iron core and, consequently, the net CL. This increment goes up to 35% compared to the same in a sinusoidally excited motor of the same fundamental frequency [1]. Moreover, applications like electric traction need a wide range of field weakening (a high frequency-operation), which furthermore increases the CL proportion. A precise determination of the CL is hence a prime requirement to design an efficient IF-IM.

One of the accurate techniques to evaluate the CL at the design stage is the finite element analysis (FEA). In FEA, the CL is calculated in the post-processing of the FEA results. Many time and frequency domain models have been proposed for this post-process calculation [2], [3]. These models are modified for improved versions of either the Steinmetz equation or the classical two-term (eddy current loss term and hysteresis loss term) CL model [3], [4]. The modification passes from the addition of excess loss term [5] to making the loss coefficients variables [6]–[11] and improvements in the Jiles-Atherton (JA) model [12], [13]. All these approaches have their tradeoffs in terms of characterization of loss coefficients [11], lack of flexibility [14], computation time [14], limited application domain [15], and complex realization [13]. In a benchmark development, [16], [17] present fast methods by introducing hysteresis and eddy current loss coefficients. However, the formulation is valid as long as the supply voltage is equal to the magnetizing voltage, *i.e.*, high switching frequency and small load. Other concerns with FEA based techniques are

the selection of time step and geometry segmentation. Numerical errors due to inadequate time-stepping lead to a high deviation in the results. [18] presents essential computational requirements for these FEM models, avoiding which lead to erroneous core loss results.

There is a need of a comprehensive technique to determine the CL during the design stage, and hence the objectives of this work are:

- Provide a comprehensive analysis of FEA based CL evaluation approaches for a wide range of three operation indices of an IF-IM viz. Switching frequency and, loading of the motor. This encases the entire range of motors power supply dynamics with several control techniques.
- Propose an accurate and easy to implement CL model, which can be used at the design stage of the motor. The proposed formulation presents a more accurate way to take into account the variation in lamination thickness with time in the existing methods. Alongside, the work presents an easy-to-implement approach for the evaluation of hysteresis loss with good accuracy.

The paper's organisation is as follows: A qualitative analysis of various CL evaluation methods is given in Section II. Section III presents the proposed CL method. The application of the proposed method for an inverter-fed IM is given in Section V. Section VI shows the comparison of the proposed model and the existing CL models with the measurements. The conclusions are drawn in Section VII.

TABLE I
COMPARISON OF METHODS OF CL EVALUATION

		SBM		NSBM			
		CC	VC	CC	VC	HLM	Hybrid
Accuracy							
Low B_m , low f	P_{Eddy}	L	H	L-M	H	NA	VH
	P_{Hys}	M	L	L-M	L	VH	VH
High B_m , High f	P_{Eddy}	L	M	L-M	M	NA	VH
	P_{Hys}	L	L	L-M	L	VH	VH
Computation time		S	M	S	M	VH	VH
Model development		NA	M	NA	M	H-VH	VH
Flexibility		VH	H	VH	H	L-M	L-M
Application range		S	M	S-M	M	H-VH	H-VH

CC: Constant Coefficients; VC: Variable Coefficients; S: Small; L: Low; M: Medium; H: High; VH: Very High; HLM: Hysteresis Loss Models; NA: Not Applicable.

II. CL EVALUATION METHODS

This section presents a modified CL model in the review of the limitations of existing CL models. The section comprises 3 subsection. Subsection-A gives an overview of existing CL models. A qualitative analysis of these models is presented in subsection B.

where,

$$\varepsilon = \tau \sqrt{\mu f \sigma \pi} \quad (7)$$

The firsthand computation of (5) is difficult because it uses $B(t)$ as primary variable and, frequency dependant permeability expression *i.e.* (6). To simplify (5) and, use it for CL estimation, the following modifications have to be done:

1. Changing of the equation's domain (*i.e.* from $B(t)$ -variable form to time-variable form).
2. Assuming the permeability time-invariant for the range of integral.

In the the following subsections, the use of equation (5) for CL estimation is explained.

A. Eddy current loss(P_{Eddy})

Rewriting the integral term of P_{Eddy} (5) in time-variable form gives

$$P_{Ed}(f, \mu) = \frac{1}{T} \frac{\sigma \tau(f, \mu)^2}{12} \int_0^T \left(\frac{dB(t)}{dt} \right)^2 dt \quad (8)$$

Evaluation of (8) for a sinusoidal field variation($B(t)$) with time-independent τ *i.e.* $\tau(f, \mu)$ provides

$$P_{Ed}(f, \mu, B_{max}) = \left(\frac{\sigma \tau(f, \mu)^2}{6} \right) B_{max}^2 f^2 \quad (9)$$

To apply (9) for a nonsinusoidal field variation($B(t)$), magnitudes of all the harmonic orders are calculated and net P_{Eddy} is determined as sum of P_{Eddy} s of all harmonics *i.e.* for

$$B(t) = \sum_{i=1}^n B_{max_i} \cos(\omega_i t + \phi_i) \quad (10)$$

$$P_{Eddy} = \sum_{i=1}^n P_{Ed}(f_i, \mu_i, B_{max_i}) \quad (11)$$

Generally, while solving (8) to get (9), the material permeability (μ) is assumed constant for a particular value of B_{max} and f . This, in fact simplifies the evaluation of (8) treating $\tau(f, \mu)$ a constant. In [20], this constant μ is taken corresponding to the B_{max} whereas, [21] suggests the μ at average value of excitation.

However, the value of μ depends upon the magnetisation (as shown in Fig.2-(b)), and hence, it does not remain constant for all the instants of a time period. Fig. 2(c) shows the variation of μ over a time period *i.e.*, $\mu(B(t))$. Value of μ is minimum at the maximum value of the excitation. Hence, taking the μ of maximum magnetisation overestimates the lamination thickness (as per (6)) and consequently gives higher P_{Eddy} value(as per (9)). Also, it is imperative to mention here that the variation in μ within the cycle increases with the increase in the maximum magnitude of the waveform (B_{max}). The Fig.2(c)-(d) show its variation for both higher and smaller magnetisation B_{max} . As a consequence, τ also modulates along a fixed value $\tau(f, \mu)$ (as shown in Fig.2(e)) calculated by (6)-(7). As mentioned in Section II-C, the variation in τ has been

one of the major sources of error in P_{Eddy} estimation. Hence, the expression (8) needs to be evaluated with taking the time-variation of τ into account as

$$P_{Ed}(f_i) = \frac{1}{T_i} \frac{\sigma}{12} \int_0^{T_i} \tau(f_i, \mu(B(t)))^2 \left(\frac{dB(t)}{dt} \right)^2 dt \quad (12)$$

In a simplest form, the variation in τ (as shown in Fig.2-(e)) can be written as

$$\tau(f, t) = \tau_{0f} + \Delta \tau_f |\sin(\omega t)| \quad (13)$$

where,

$$\tau_{0f} = \frac{\tau(f, \mu_{max}) + \tau(f, \mu_{min})}{2} \quad (14)$$

$$\Delta \tau_f = \tau(f, \mu_{min}) - \tau(f, \mu_{max}) \quad (15)$$

Solving the expression (12) and (13) for sinusoidal $B(t)$ gives

$$P_{Ed}(f, B_{max}, \tau') = \left(\frac{\sigma \tau'(f, \mu)^2}{6} \right) B_{max}^2 f^2 \quad (16)$$

where, $\tau'(f, \mu)$ comes out as

$$\tau'(f, \mu) = \left(\tau_{0f}^2 + \left(\frac{\Delta \tau_f}{4} \right)^2 \right)^{1/2} \quad (17)$$

which is the root mean squared (*rms*) value of $\tau(f, t)$. The proposed method hence suggests the evaluation of (12) with taking the *rms* of the τ . This *rms* value of $\tau(f, t)$ is calculated using (6), (14) and (15). The proposed modification does not impose the restriction on applying the superposition and, the P_{Eddy} for a non-sinusoidal $B(t)$, with it is calculated as

$$P_{Eddy} = \sum_{i=1}^n P_{Ed}(f_i, B_{max_i}, \tau'_i) \quad (18)$$

where, $P_{Ed}(f, B_{max}, \tau')$ is calculated using (16). For numerical values of B-data obtained by FEA, the discrete form of the proposed evaluation (18) is used as

$$P_{Ed}(f_i) = \frac{1}{N_i} \frac{\sigma}{12} \sum_{k=1}^{N_i} \tau'(f_i, \mu(B(k)))^2 \frac{(B_{k+1} - B_k)^2}{t_{k+1} - t_k} \quad (19)$$

B. Hysteresis loss P_{Hys}

The term for P_{Hys} in (5) is f -times the area of the hysteresis loop in $B-H$ plane. In inverter fed motors, flux density waveforms (of different motor segments) consist of higher-order time harmonics, which results in several minor loops. Net hysteresis loss in these cases is taken to be the sum of areas of all the loops[11]-[12]. However, this method results in higher P_{Hys} because the sum of areas of the loops created by each harmonic independently is greater than the actual hysteresis area. The reason for the higher estimation of P_{Hys} is mainly due to the following three reasons:

- The addition of a harmonic in a sinusoidal B-waveform doesn't always create a minor loop. As shown in Fig.3 (a and b), both the B-waveforms (of equal fundamental component B_{max1}) comprise equal magnitudes of 3rd and 5th harmonics (B_{max3}, B_{max5}) but their B-H loops are

Fig. 2. (a) B-H curve of the material 50C350; (b) Variation of differential permeability(μ) with $B(t)$; (c) Variation in (μ) for sinusoidal excitation of higher B_{max} ; (d) Variation in (μ) for sinusoidal excitation of smaller B_{max} ; (e) Variation in τ for the μ of (c); (f) The impact of DC-offset on the area of a minor loop.

different. The B-H loop of B-waveform (a) does not form any minor loop whereas, the B-H loop of B-waveform(b) results in two minor loops, each corresponding to the one harmonic component.

- the area of the minor loop (if generated) for a particular harmonic is not exactly equal to the area of the hysteresis loop created by that harmonic independently with the same magnitude. The minor loop area is affected by the DC-bias offered at its location by the major loop. Fig.2(f) shows this impact taking DC offset 0T and, 0.8T.
- it is not necessary that the difference of time instants of maxima and minima of a minor loop provides the time period of a particular harmonic. The superposition of multiple harmonics with different phase-angles may shift the location of these maxima and minima.

The above limitations result in the inaccurate prediction of the hysteresis loss for a distorted B-waveform using an empirical expression. Modelling of the hysteresis loop gives a good solution. The Preisach model, the energetic model, and the Jiles-Atherton model are the prominent hysteresis loop models. These models use H (or B) as an independent variable and compute the corresponding B (or H) value. HL is created using these B-H values for one complete cycle.

The proposed method puts forward a new alternate time-domain approach to determine the area of HL for a distorted

Fig. 3. (a) Nonsinusoidal waveform-A; (b) Nonsinusoidal waveform-B; (c) Hysteresis loop for waveform-A; (d) Hysteresis loop for waveform-B.

B-waveform without evaluating the H (i.e. without forming the major and minor HL). The method uses the following two data: a).measured values of HL areas(dynamic) with sinusoidal excitations as a prerequisite which are evaluated using Epstein frame as detailed in section III. b).time-instant B values of the distorted B-waveform which can be directly retrieved from the magnetic vector potential based formulation in FEM. Due to the symmetry of B-waveform, only the half-cycle data is used here. In execution of the proposed method, time spaced half cycle B-data is examined sequentially for its rate of change to get all its local and global extrema($B_1, B_2, B_3...B_M$ as shown in Fig.3-(b)) as points of zero slope. From the sequentially stored values of these extrema, the net HL area is evaluated using the measured HL areas for sinusoidal magnetisation of different magnitudes as described below:

- area of major loop(A_m)- the maximum absolute value of the stored extrema(B_{max}) corresponds to the major loop. Hence, the loop area measured for maximum magnetisation = B_{max} for sinusoidal excitation is taken as an area of major loop.
- the area of a minor loop(A_i)- the difference between consecutive maxima and minima ($\Delta B_i = B_n - B_{n-1}$) gives the height of minor loop in B-H plane. The measured HL area corresponding to the maximum B value equal to the half of the ΔB_i is taken as the area of the minor loop(with zero DC bias)i.e.

$$A_{0i} = A_{measured} \left(\frac{\Delta B_i}{2} \right) \quad (20)$$

where $A_{measured}(B_i)$ is the measured value of the HL area with the sinusoidal magnetization of the maximum magnitude of B_i without DC bias. The DC bias offered at the location of a minor loop is equal to $B_{DCi} = (B_n + B_{n-1})/2$ (Fig.3(b)). The area of minor loop with this DC bias is calculated as [22]

$$A_i = A_{0i} (1 + k_{DC} B_{DCi}^\alpha) \quad (21)$$

where k_{DC} and α are the constants representing the variation in HL area with DC bias. They are obtained from the measurements as per [22] and, there values for the steel used in this work are 0.67 and 2 respectively.

- total area of the HL and, the corresponding specific hysteresis loss is determined as

$$P_{Hys_S} = \left(A_m + \sum_{i=1}^L A_i \right) f \quad (22)$$

where L and f are total number of the minor loops and fundamental frequency respectively.

The proposed method uses hysteresis loss data with sinusoidal excitation as a prerequisite which are evaluated as detailed in Section IV.

IV. CL CHARACTERISATION OF MATERIAL

The IM used in this work is made of ferromagnetic core material 50C350. The Loss characterisation of this material is done with 352 turns Epstein frame. The setup consists of primary and secondary windings mounted uniformly on the four sides of the frame inside which, the sample iron-sheets are placed. A pulsating (of sinusoidal time variation) magnetic field is set up in the sample iron sheets by exciting the primary side from a 3kW power amplifier. The THD in voltage (or B-waveform) is limited to 0.5%. The field strength $H(t)$ inside the sample sheet is determined from the primary current as:

$$H(t) = \frac{N_P I(t)}{l_{mean}} \quad (23)$$

where, l_{mean} is mean length of the magnetic flux path, N_P is the number of turns in primary winding. The flux density B is defined via measurement of the voltage induced in the secondary winding ($v_S(t)$) as

$$B(t) = \frac{\phi(t)}{A_{core}} = \frac{1}{A_{core} N_S} \int_0^t v_S(t) dt \quad (24)$$

where, A_{Core} is the crosssection area for the flux, N_S is the number of secondary windings. The total iron loss for the volume of sample iron-sheets is obtained as shown

$$P_{Core} = \frac{N_P}{N_S} (P_{in} - I^2 R_P) \quad (25)$$

where, P_{Core} is the total iron loss of the core sample, P_{in} is the input power for the primary coil, R_P is the primary coil resistance. Total core loss estimated with (25) is sum of two losses viz. eddy current loss and hysteresis losses,

$$P_{Core} = P_{Eddy} + P_{Hyst}. \quad (26)$$

The components of field strength $H(t)$ in (23) corresponding to these losses are,

$$H(t) = H(t)_{Eddy} + H(t)_{Hyst}. \quad (27)$$

The instantaneous value of $H(t)_{Eddy}$ is determined from the instantaneous flux density in (24) as

$$H(t)_{Eddy} = \frac{\sigma}{12} \tau(f, \mu)^2 \frac{dB(t)}{dt} \quad (28)$$

Using (7) and (10) for the effective $\tau(f, \mu)$ lamination thickness, field strength corresponding to the magnetic hysteresis $H(t)_{Hyst.}$ is obtained as

$$H(t)_{Hyst.} = H(t) - \frac{\sigma}{12} \tau(f, \mu)^2 \frac{dB(t)}{dt} \quad (29)$$

The area under the $B(t)$ versus $H(t)$ curve, i.e.,

$$A_{Net} = \oint H(t) dB(t) \quad (30)$$

gives the net core loss P_{Core} of a cycle, whereas the area under the $B(t)$ versus $H(t)_{Hyst.}$ curve

$$A_{Hyst.} = \oint H(t)_{Hyst.} dB(t) \quad (31)$$

evaluates the hysteresis loss. Fig.4 shows both these areas for 1.3T, 100Hz. Fig.5(a)-(d) show the separated losses (P_{Eddy} and $P_{Hyst.}$) for the range of and magnetisation strength at low and high frequencies.

Fig. 4. B-H curves corresponding to the net and magnetising field intensities at 100Hz, 1.3T.

Fig. 5. CL separation: P_{Eddy} and $P_{Hyst.}$ for different frequencies and magnetisations of the rearmost measurements.

V. CL EVALUATION FOR IM

This section presents the application of the proposed method in the evaluation of CL for an inverter fed IM. For this, a 1.5kW three phase IM is selected. Table III shows the

rated parameters of this IM. As mentioned in Section-II C, the proposed method uses one cycle time data of flux density, a 2-D FEA model of the 1.5kW IM is developed in Maxwell Ansys 19.2. The stator windings of the FEA model are excited with measured current waveforms and, the model is run with the anhysteretic magnetisation and active eddy current effect.

Retrieving the B-data for all the nodes of mesh elements and their CL calculation demands requires storage. Hence, the motor geometry is divided into small segments [2], [23], [25] of homogenous flux density. Both the tangential (B_T) and radial (B_R) components of magnetic field density inside a segment are constant [2], [23], [25]. Fig.6 shows the motor cross-section with segments. To apply the proposed method, both B_T and B_R are extracted from the FEA model for one fundamental time period. Using the B_T and B_R , specific CL for the segment evaluated as

a). P_{Eddy} : using the expression (19) i.e. the evaluation with instantaneous flux density dependant permeability (i.e. $\mu(B(t))$). P_{Eddy} with B_T and B_R are determined separately and added to get total P_{Eddy} .

b). P_{Hys} : using the expression (22).

Net CL of the machine is estimated as

$$P_{CL} = \sum_{i=1}^S m_i [P_{Hys_S_i} + P_{Eddy_S_i}] \quad (32)$$

where, S is the total number of geometry segments and, m_i denotes the mass of i-th segment.

The FEA model is run for multiple measurements of motor currents to encase the entire range of the IM's operation (0.5Nm to 2.5 Nm) and carrier frequency variation (300Hz to 5kHz).

Fig. 6. FEA model of the motor and postprocess geometry segments.

The CL for each operating condition is evaluated applying the proposed method.

VI. EXPERIMENTAL MEASUREMENT OF CL

Fig.7 shows the results obtained with the proposed technique. For experimental verification of these results, the IM is operated with inverter power supply. CL is evaluated for a total 50 operating points of the motor as given in Table 2. For all these operating points, the speed of the motor is fixed at 2889rpm by varying its terminal voltage. Fig.9 shows the voltage variation for the two extremes of switching frequencies i.e. 300Hz and 5kHz.

TABLE II
MOTOR OPERATION DOMAIN

	Speed (rpm)	Torque (Nm)	f_{sw}	Power -supply
Upper value	2889	2.5	5000	Sine pulse width
Lower value	2889	0.5	300	modulation
Steps	1	5	10	(SPWM)

TABLE III
RATED PARAMETERS OF THE EXPERIMENTAL IM

Sr.	Parameter	Unit	Value
1	Pole-pair	-	1
2	V_{Rated} -frequency	Volt/Hz	415/50
3	Output power	kW	1.5
4	Core iron-grade		50C350
5	Lamination thickness	mm	0.5
6	Slots: Stator, Rotor	Number	36,34
7	Outer diameter	mm	400
8	Inner diameter	mm	400
9	Stack length	mm	400
10	Air-gap length	mm	0.5

A. Measurements-Set-up and procedure:

Mechanical (Windage and bearing) losses of the IM are first calculated with the sinusoidal power supply as per [24]. Correction for stator resistance for temperature, skin depth and proximity effect is done as [24]:

$$R_S(T, f) = R_{S0}(1 + \alpha(T - 20))(1 + \frac{l_s}{l_w}(k_{nR}(\xi) - 1)) \quad (33)$$

where, l_s is the the stack length and l_w is half the mean length of a stator turn. $k_{nR}(\xi)$ is skin and proximity effect factor depends upon the conductor size, number of strands, conductor per slot and slot dimensions[. Stator copper losses are calculated with the measured values of input *rms* phase currents (I_s) as:

$$P_{CuS} = R_S(T, f) \sum_{i=1}^3 I_i^2 \quad (34)$$

The CL is determined using the measured values of the one cycle-averaged input power (P_{In}) and then, applying the input-output power balance as:

$$P_{Core} = P_{In} - P_{CuS} - s(P_{In} - P_{CuS} - P_{Core}) - P_{Out} \quad (35)$$

$$P_{Core} = P_{In} - P_{CuS} - \frac{P_{Out}}{1 - s} \quad (36)$$

where,

$$P_{Out} = P_{Shaft} + P_{Windage} + P_{Bearing} \quad (37)$$

and,

$$P_{Shaft} = T_{Shaft} \omega_r \quad (38)$$

T_{Shaft} and ω_r are the measured values of shaft torque and rotor speed in rad/s respectively using torque sensor (Kistler 4503B) and CoMo Torque Type 4700 data acquisition system. Other equipments of the test-bench are: a) 2-level, H-bridge three-phase inverter b). 1.5 kW IM motor prototype c). Digital oscilloscope: YOKOGAWA DLM2054 with voltage and cur-

Fig. 7. Measured and calculated CL values at different f_{SW} for T_L equal to (a)2.5 Nm, (b)1.5 Nm, (c)1.0 Nm, (d)0.5 Nm

rent probes d). Voltage and current transducers: LEM sensors LV25-P and LA 15P f). dSpace DS1103 for data acquisition f). Photo-tachometer. Fig.8 shows the experimental setup.

The CL for each operating condition is evaluated with the proposed and other widely followed methods of literature. The other methods include:

- CL evaluation with (2) with constant K_e and K_h values: M-1 ($K_e=3.717W/Kg/Hz^2/T^2$, $K_h=0.0449W/Kg/Hz^\alpha/T^\beta$).
- CL evaluation with (2) with frequency and magnetisation dependant K_e (i.e. $K_e(B_m, f_{eq})$) and constant K_h : M-2.
- CL evaluation with (2) with both K_e and K_h frequency and magnetisation dependant (i.e. with $K_e(B_m, f_{eq}); K_h(B_m, f_{eq})$): M-3.
- CL evaluation with using the analytic expression of eddy current loss and, Jiles-Atherton model for the hysteresis loss: M-4.

Fig. 8. Experimental setup: Schematic and test-bench

B. Results and discussions:

Fig. 9. Variation of motor's (a) phase voltage and (b) stator current with load torque

TABLE IV
ERROR ESTIMATES FOR THE DIFFERENT CL METHODS

Parameter	M-1	M-2	M-3	M-4	Proposed
Max. Error(%)	37.5	25.03	23.11	13.25	16.21
Min. Error(%)	24.43	19.68	16.34	6.62	5.11
σ_{Errors}	8.11	5.30	4.45	1.28	1.32
f_{SW} at E(Max)	300Hz	300Hz	300Hz	300Hz	300Hz
f_{SW} at E(Min)	5Khz	5Khz	5Khz	5Khz	5Khz

Fig.7 shows the measured and estimated CL values with all the techniques mentioned above. These results draw the following conclusions:

- The proposed method ensures the accuracy of CL estimation at par with method M-4. The maximum and minimum error values with the proposed method are 16.21% and 5.11% ,respectively. Whereas, the same with M-4 are 13.25% and 6.62% ,respectively. However, the proposed

method is much easier to implement than M-4 with J-A model.

- For a particular T_L , both the proposed method and M-4 show consistent error (*i.e.* least values of the standard of deviation of errors (σ_{Errors})). For example, σ_{Errors} , at $T_L=2Nm$, for the proposed method and M-4 are 2.6 and 3.13 respectively, which indicates that the evaluation of hysteresis loop area with the proposed method is close to the same evaluated with JA method. However, there is no direct method available to measure the hysteresis loss (and hence, HL area) independently.
- Method M-2 and M-3 show a higher error at low f_{sw} because time harmonics are dominant, and B-waveforms have larger distortion compared to the same at high f_{sw} . The error becomes consistent at higher f_{sw} where, the current waveforms are close to a sinusoid. However, still, the error is high in comparison to the proposed method and M-4.

VII. CONCLUSION

A fast and accurate iron loss calculation method for IM is proposed for inverter-fed motors in this work. The proposed model determines the eddy current loss by taking the time variation of material permeability into account. Also, the method proposes an easy to implement approach for the evaluation of hysteresis loss. The accuracy of the proposed method is verified by comparing it with the existing CL models. For this, an inverter-fed IM is driven for a wide range of switching frequencies and load-torque. The CL of the IM is obtained for all these operating points applying the four existing CL methods. The evaluation of CL using the proposed method resulted in the least estimation error. The results obtained using the proposed method are close to the same obtained by an analytical appended Jiles-Atherton method for hysteresis loss with a differential of maximum and minimum error equal to 2.96% and 1.51%, respectively.

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