LOSSES INFLUENCED BY THIRD-HARMONIC FLUX IN INDUCTION MOTORS

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Abstract

The trend toward increasing flux density in modern induction motors results in the increase of third-harmonic flux in the magnetic paths. This study shows the effects of third-harmonic flux on energy losses in induction motors.

The shapes of flux distributions in various sections of an induction motor are interrelated. The most saturated section normally has a flattened shape of flux density distribution. From the third-harmonic flux standpoint this study compares the effects of the shapes of flux density distributions on losses. The losses influenced by the third-harmonic flux may be controlled through lamination punching designs and winding connections.

Introduction

In modern induction motor design there is a trend towards increasing flux densities. As the permissible temperatures of insulating materials used for both the windings and laminations go up, the operating flux and current densities of contemporary induction motors also become progressively higher. The higher flux densities cause magnetic saturation that produces third-harmonic flux [1] in the magnetic circuit. The third-harmonic-flux content of a sample motor in Y-connection reaches 18% of the fundamental flux under saturated conditions at no load, and 11% at full load. The effects of third-harmonic flux in magnetic cores have been studied in detail [2,3,4]. This paper not only discusses the core losses affected by the third-harmonic flux in the magnetic materials of induction motors, but also attempts to look into the stray-load losses influenced by the shape of the air-gap flux. The shape of the air-gap flux is shown later to be determined mainly by the relative phase angle between the third-harmonic and the fundamental fluxes.

It is commonly agreed that the investigation of stray-load loss in rotating machines is far from complete. Most motor designers rely on empirical equations for the estimation of stray-load losses. The accuracy of the empirical equations depends on the understanding of the various sources of stray-load losses. As a result of the complexity of the subject, the International Electrotechnical Commission (IEC) recommends 0.5% of the output to be the stray-load loss of induction motors, and the American National Standards Institute (ANSI) suggests 1.2% for small motors (motor rating less than 2,500 hp) and 0.9% for large motors. These numbers are estimates and do not help to improve the real efficiency of induction motors.

The subject of the origin and evaluation of stray-load loss in induction motors has an extensive literature. Dating back to 1912, Olin [5] has suggested estimating the stray-load loss by relating it empirically to the no-load loss. Our study shows that for small motors, the no-load current is relatively closer to the full-load current and the no-load stray loss plays an important role in the overall efficiency of an induction motor. Jimoh, Findlay, and Poloujadoff [6,7] have outlined the state of the art in understanding and measuring stray-load loss in induction motors with a great number of references. Christofides [8,9] and Adkins [9] have taken account of the effect of magnetic saturation of main flux on the core loss by using a weighted factor averaged over a pole pitch, and by selecting a suitable value of incremental permeability, thereby demonstrating that saturation causes a reduction of stray-load loss. With Christofides' [8] method, the damping factor for the differential harmonic leakage flux is assumed to be independent of the amplitude of the main flux density. Spooner [10] has experimentally investigated the variation of the incremental permeability, both with the magnitude of the harmonic-flux pulsation and with the fundamental field at the instant when the pulsations occur. Smith [11] has shown that if the pulsations occur at high frequencies, skin effects reduce the value of incremental permeability considerably. Rao and Butler [12] have investigated the tooth-body and surface losses assuming that the effect of the tooth overhang angle is negligible. In the absence of main-flux saturation, Angst [13] has suggested an analytical method of calculating the saturated leakage reactance for different stator and rotor slot shapes. The saturation factors can be expressed as a function of slot current, and the reactance can be determined by a process of iteration. Alger [14] has shown that the stray-load loss can be calculated through a detailed equivalent circuit, consisting of a ladder network. Alger [15,16] and Chalmers [17,18] have discussed the high-frequency losses in induction motors. However, the existing literature has not addressed the losses associated with the shapes of flux density distribution in the core, teeth and air gap, respectively.

This study recognizes that the core and stray losses calculations rely heavily on empirical equations. Most practical designers use specific equations obtained statistically according to their individual manufacturing situations. In order to choose the independent variables of the empirical equations, it is important to know the physical nature of these losses. This paper discusses more qualitatively and experimentally than quantitatively on the losses caused by the third-harmonic flux. It is shown that the losses are affected by the relative magnitude and phase angle of the third-harmonic flux to the fundamental values, and the third-harmonic flux can be controlled through lamination punching design and winding connection. Other superharmonics in the air-gap flux (such as the tooth and the fifth harmonics caused by slot permeance and saturation) are also briefly included in the discussion.

Third-Harmonic Flux Modulation and Shapes of Flux Density Distribution

Saturation of magnetic materials in induction motors is a nonlinear phenomenon. Fig. 1 shows that a sinusoidally distributed excitation, such as the excitation in a rotating machine, corresponds to a nonsinusoidal flux density distribution and, hence, produces harmonic fluxes. Among those, the third-harmonic flux is the dominating component. This simplified example assumes that the core (or yoke) sections do not consume excitation, the slots are narrow, and the magnetization curve includes the air-gap permeance.

Figure 1. Sinusoidal excitation distribution corresponds to nonsinusoidal flux density distribution

The principle of using third-harmonic flux to modulate the shapes of flux-density distribution is applicable for all magnetic sections, such as the core, the teeth, and the air-gap of an induction motor. Fig. 2 briefly describes this modulation. The horizontal coordinate represents the angular location of a section in the motor. The sinusoidal flux density distribution shown in Fig. 2a can be changed. As shown in Fig. 2b, when a certain third-harmonic flux at 0° phase angle to the fundamental is added, the shape of flux density distribution is flattened. Fig. 2c shows that when the phase angle of the third-harmonic flux density distribution with respect to the fundamental flux density distribution is changed from 0 to 180° (third-harmonic degrees), the shape of flux density distribution is peaked up.

In an induction motor, the flux density distribution in the core, as shown in Fig. 3, is the integration of the flux density distribution in the teeth. If the teeth harmonics viewed from the junctions between the teeth and the core are ignored, and the fundamental component of the teeth flux density distribution at the roots of the teeth is a cosine term included in the following equation,

$$\text{Teeth flux density distribution} = A_1 \cos(\theta - \omega t) + A_3 \cos(3\theta - 3\omega t) \quad (1)$$

then the fundamental component of the core flux density distribution is a sine term derived from the following integration

$$\begin{aligned} \left[\text{core flux density distribution} \right] &= C \int_0^{\frac{\pi}{2}} \left[\text{teeth flux density distribution} \right] \left(\frac{D}{2} \right) d\theta \\ &= C \left(\frac{D}{2} \right) \left[\frac{A_1 \sin(\theta - \omega t)}{1} + \left(\frac{A_3}{3} \right) \sin(3\theta - 3\omega t) \right] \end{aligned} \quad (2)$$

Figure 2. Modulation of flux density distribution
a) sinusoidal flux density distribution
b) flattened flux density distribution
c) peaked up flux density distribution

where the constant C is used to reflect the ratio of cross sectional areas between the teeth and the core.

$$C = \frac{(\text{tooth width at root})}{(\text{tooth pitch at root})(\text{core radial height})}$$

D = diameter at the roots of the teeth (m)

θ = angular location (rad)

ω = angular frequency (rad/s) of the rotating teeth flux

t = time (s)

A₁ = amplitude of fundamental flux density distribution at the roots of teeth (T)

A₃ = amplitude of third-harmonic flux density distribution at the roots of teeth (T)

While A₁ is positive, A₃ can be positive or negative, depending on the shape of the flux. A positive A₃ corresponds to a flattened core flux density distribution, and a negative A₃ corresponds to a peaked-up core flux density distribution.

Equations (1) and (2) show that, first, as mentioned before, if the fundamental teeth flux density distribution is represented in the cosine term, then the fundamental core flux density distribution is in the sine term, and they are 90 electrical degrees apart. Second, the relative phase angles of the third-harmonic flux with respect to the fundamental component changes, and a peaked up flux density distribution in the teeth results in a flattened flux density distribution in the core. Third, the relative amplitude of third-harmonic flux density distribution in the core, with respect to the fundamental in the core, is one third of that in the teeth. Fourth, when the slot leakage flux is neglected, the shape of the air-gap flux density distribution is in line with the shape of the teeth flux density distribution.

Figure 3. Flux density distribution in core is the integration of flux density distribution in teeth

Fig. 4 shows the flux distributions of the air gap and the core of a Y-connected motor under no load (Fig. 4a) and loaded (Fig. 4b) conditions. The amplitudes of the third harmonic teeth flux density are 18% of fundamental flux at no load, and 11% at full load. The lower third-harmonic flux content at full load is caused by the higher third-harmonic current in the rotor at full-load slip.

Experimental Setup

In order to observe the difference in magnetic saturation between stator core and teeth, a two-pole winding is wound in a stator which was originally used for a four-pole, 5-hp motor. By doing so, the stator core of the two-pole sample motor becomes much more saturated than the teeth. A full-pitch search coil located in the slots close to the air gap is used for the air gap (or the teeth) flux density distribution measurement. A conductor wrapped around the core is used for the core flux density distribution measurement. Fig. 5 shows the experimental setup. A torque gauge is coupled between the shafts of the test motor and a synchronous machine. When a no-load test is conducted, the synchronous motor drives the test motor at synchronous speed, and the no-load mechanical input power is indicated by the torque gauge. When a load test is conducted, a synchronous generator is used to provide the load. Hot resistance values of the stator winding are recorded for stator copper loss calculations. Rotor copper loss is estimated from

$$\text{rotor copper loss} = \left(\frac{\text{load power}}{+ \text{friction and windage loss}} \right) \frac{S}{1-S} \quad (3)$$

where

S = slip

Although there are various methods for measuring the stray-load loss, the high resolution digital meters with accuracy rated at $\pm 0.25\%$ of reading and torque gauge with 0.4 in.-lb resolution used for the test in this study give reasonably accurate results. The core and stray losses are obtained from

$$\begin{bmatrix} \text{core and} \\ \text{stray losses} \end{bmatrix} = \begin{bmatrix} \text{input} - \text{output} - \text{stator copper loss} \\ - \text{rotor copper loss} - \text{friction and windage loss} \end{bmatrix} \quad (4)$$

Figure 4. Air gap and core flux density distributions
a) no load b) loaded

No-Load Basic Core Loss and Stray Loss

Unlike the transformer core loss, the rotating-machine core loss is more complex. The core loss per unit volume under a given flux density obtained from an Epstein test normally is too low for the evaluation of no-load core loss of a rotating machine. There are two types of basic core losses associated with Epstein test, namely the eddy-current and hysteresis core losses. The former one is proportional to the flux density square, and the latter one to the K th power of flux density, where the exponent K varies from 1.4 to 1.8 or from 1.5 to 2.5 for modern materials [1]. This study assumes that the basic eddy-current and hysteresis core losses are proportional to the square of fundamental flux density. The rest of the core and stray losses at no load defined by Equation (4) is termed as no-load stray loss.

Since the stray loss is low when the motor is not saturated, Fig. 6 shows that the basic core loss curve can be derived from the point at low flux density, and be extended according to the relationship of flux density square. The no-load stray loss is thus obtained by subtracting the basic core loss from the no-load core and stray losses. Even at no load the stray loss goes up rapidly as the flux density goes up.

6401.0079

Figure 5. Experimental setup

6401.0080

Figure 6. Basic and no load stray losses with delta-connected winding

A separate test on the relationship between flux waveforms and losses is conducted. The result [1] given in Fig. 7 shows the experimental core losses at various fundamental flux densities with different third-harmonic fluxes. The top curve with third-harmonic flux at 180° phase angle refers to the "peaked-up" situation. The naturally compensated

6401.0081

Figure 7. Core losses vs. fundamental flux densities (for 0.3048-mm thick laminations)

saturated curve refers to the "flattened" situation where the third-harmonic flux is produced at 0° phase angle by magnetic saturation. The optimum curve refers to the situation with more third-harmonic flux injected to the magnetic path at 0° phase angle with respect to the fundamental flux. This optimum curve does not show a significant difference from the naturally compensated curve. The core loss curve for the sinusoidal flux without third-harmonic flux is located in the middle. As the fundamental magnetic flux density goes up, the additional loss caused by third-harmonic flux in the magnetic material goes up under the "peaked-up" situation (i.e., with third-harmonic flux at 180° phase angle). Under the "flattened" situation (i.e., with third-harmonic flux at 0° phase angle) the core loss is less than that of the sinusoidal situation without third-harmonic flux. These curves indicate that under saturated conditions, for most thin laminations the "naturally compensated" flattened flux corresponds to the saturated section of the magnetic path. The core loss of core material carrying a flattened shape of flux density distribution is less than that of the same material carrying the fundamental flux alone.

The basic core and stray losses at no load of the sample motor, under three shapes of air-gap flux density distribution: the peaked-up shape with Y-connected winding, the sinusoidal shape with delta-connect winding, and the flattened shape with the injected third-harmonic flux to peak up the core flux density distribution: are shown in Fig. 8. The third-harmonic current cannot flow in a Y-connected winding; the third-harmonic flux is therefore preserved. The delta-connected winding tends to destroy the third-harmonic flux by allowing the third-harmonic current to flow. The worst case for this particular sample motor among these three shapes is the flattened shape with injected third-harmonic flux to flatten the air-gap flux density distribution and in the meantime to peak up the core flux density. The results shown in Fig. 7 explains why the Y-connected situation of the sample motor has the least no-load core and stray losses.

Fig. 9 shows the no-load stray loss vs. the rms coil current of the sample motor under a nonsaturated (air-gap flux density is 0.27 T) and a saturated (0.53 T) conditions. Two coils are connected in parallel for each phase. The magnitudes of the no-load stray losses are significantly different among the three shapes of the air-gap flux density distribution under the saturated condition, but practically not different under nonsaturated condition. The curves show that the no-load stray losses go up when the flux den-

Figure 8. Basic core and no load stray losses vs. amplitude of fundamental air gap flux density

Figure 9. No load stray loss vs. rms coil current at 0.53 T and 0.27 T

sity goes up. The Y-connected winding with a less saturated teeth flux density distribution and a peaked-up air-gap flux density distribution of the sample motor has the least no-load stray loss.

Conventional Stray-Load Losses

Fig. 10 shows the stray-load losses according to the conventional definition (i.e., the no-load core and stray losses are subtracted from the core and stray losses under loaded condition). The magnitudes of the stray-load losses corresponding to the three shapes of the air-gap flux density distribution of the sample motor are slightly, but clearly, different under the saturated condition at 0.53 T. The conventional stray-load loss goes down when the air-gap flux density increases. This trend is true for all three shapes of air-gap flux density distribution (i.e. Y-connected for a peaked-up air-gap flux density distribution, delta-connected for a sinusoidal flux density distribution, and the flattened shape of flux density distribution).

When the shape of the air-gap flux is flattened, the stray-load loss becomes the smallest among the three different shapes. It is opposite to that shown in Fig. 9. However, regardless what definition is used, the objective of this

Figure 10. Conventional stray load loss vs. rms coil current

investigation is to reduce the overall loss, which includes the losses affected by the third-harmonic flux at no load.

The reason for the Y-connected winding (i.e. peaked-up air-gap flux density distribution) of the sample motor to have the highest stray-load loss is investigated further. Fig. 11 shows the slot harmonic flux contents computed through finite-element method when the rotor cage is open. The 36 stator slots per pole pair give the 35th and the 37th slot harmonic fluxes. These slot harmonic fluxes go up while the air-gap radial fundamental flux density goes up. When the shape of the air-gap flux density distribution is changed from a peaked-up to a flattened shape and the magnitude of the fundamental flux density in the air gap is maintained the same, the differences of these slot harmonics for the sample motor are not significant.

The air-gap harmonic flux contents for both the no-load and loaded conditions, respectively, are also measured by a spectrum analyzer. The magnitudes of the measured slot harmonics are small when they are compared with those of the open-cage results obtained from the finite element method. Under the test conditions, the harmonic currents in the rotor cage tend to reduce the magnitude of the resultant slot harmonic fluxes. Again, the magnitudes of the slot harmonics of the sample motor do not change much among the three shapes of air-gap flux density distributions. This observation holds true for either no-load or load-

Figure 11. Slot harmonic flux content

ed conditions. Therefore, the differences in stray-load losses among the three shapes of air-gap flux density distribution have to come from other sources for the sample motor.

The measured data of the 5th and 7th harmonics from the spectrum analyzer is given in Table 1. The magnitudes of the 5th and 7th harmonics under loaded conditions are not significantly different among the three shapes of air-gap flux distributions, but the 5th and 7th harmonic flux of the Y-connected motor under no-load conditions is the lowest among the three shapes of air-gap flux density distribution. They are nearly double in magnitude from the no-load values. This makes the peaked-up shape of the Y-connected case have the highest stray-load loss among the three shapes of air-gap flux density distribution. However, for the sample motor, although the Y-connected winding corresponds to the highest conventional stray-load loss, the overall efficiency is the highest one because its no-load loss is the lowest one.

Table 1. 5th and 7th harmonic fluxes (Φ represents flux)

Shape of Air Gap Flux Density Distribution	$\frac{\Phi_5}{\Phi_1}$		$\frac{\Phi_7}{\Phi_1}$	
	No Load	Loaded	No Load	Loaded
Sinusoidal (Δ)	0.0293	0.0315	0.0160	0.0160
Peaked-up (Y)	0.0115	0.0215	0.0054	0.0127
Flattened	0.0449	0.0473	0.0199	0.0206

Output Power vs. Total Losses

Fig. 12 shows that when the sample motor is not saturated (air-gap flux density is 0.27 T) the differences in total losses at a given output power among three shapes of air-gap flux density distribution (i.e. peaked-up with Y-connection, sinusoidal with delta-connection, and flattened) are not significant. However, when the machine becomes saturated (air-gap flux density is 0.53 T), the Y-connected motor has the lowest loss among the three situations. For small motors, the no-load current is relatively closer to the full-load current, and the no-load stray loss plays an important role in the overall efficiency of an induction motor. For large motors, the peaked-up or flattened air-gap flux distributions might affect the tooth-harmonic and the high-frequency losses more significantly. This topic is being pursued.

Punching Design Philosophy

The shaft height of a given frame of an induction motor is normally standardized. Consequently, the outer diameter of a stator punching is basically fixed for a given frame. If the stator inner bore diameter is more or less fixed, there are two practical approaches to accommodate more copper. The designer can make the slot deeper; this means a more saturated core but less saturated teeth. The designer can also make the slot wider; this means a less saturated core but more saturated teeth.

Different numerical or analytical approaches can be used to obtain the flux density distributions in various sections of the magnetic path of an induction motor. For instance, the finite-element method that we use, or the Cistelecan and Onica [19] method of determining the air-gap flux density distribution and the third-harmonic flux

Figure 12. Output power vs. total loss with different shapes of air gap flux density distribution

contents can be used. Once the shapes of flux density distributions in a motor are obtained, the most saturated section can be identified.

The qualitative conclusions for punching design are:

(1) Nonsaturated core and teeth: The core and teeth fluxes are sinusoidally distributed. The excitation current is sinusoidal. No circulating third-harmonic current flows in the delta-connected winding, or in the rotor cage under loaded condition. Y-connected windings perform the same as delta-connected windings.

(2) A saturated core with less-saturated teeth: The flux density distribution in the core is flattened. Consequently, the teeth, and the air-gap flux distributions are peaked up. The volume of the teeth is relatively small as compared with that of the core. The increase in total loss caused by the peaked-up teeth flux density may not be too bad. Although the peaked-up teeth flux density distribution does not show a significant change in the teeth harmonics from other shapes of air-gap flux density distribution of the sample motor, it is felt that further investigation on this particular topic is needed. Since the core is saturated, the third-harmonic flux flattens the core flux density distribution. Consequently, the core loss in the core compared with having the fundamental flux component alone is greatly reduced. The designer would have to consider this effect and the possible change of teeth harmonics caused by the peaked-up air gap flux density distribution according to the specific dimensions and materials used in the motor. A delta-connected winding would reduce the content of third-harmonic flux. Therefore, if one is relying on the flattened core flux density to reduce the total loss, delta-connections should not be used.

(3) A less-saturated core with saturated teeth: The flux density distributions in the teeth and in the air gap are flattened. The flux density distribution in the core is peaked up. Because the volume of the core is normally greater than that of the teeth as compared with having the fundamental flux component alone in the core, the peaked-up flux density may increase the core loss significantly. Again, although the flattened teeth and air-gap flux distributions do not practically affect the teeth harmonics in the sample motor, further study is needed for a more general conclusion on this particular topic. Because the volume of core is normally greater than that of the teeth, the tradi-

tional design of induction motors chooses a significantly lower core flux density than the tooth flux density. The forementioned observation suggests that from the third-harmonic flux standpoint the traditional design philosophy might not be the most economical and efficient approach. Under this situation the delta connection tends to destroy the existence of third-harmonic flux. It should be noted that under loaded condition, the rotor runs at a certain slip. There is a certain third-harmonic current flowing in the rotor even though the stator winding is connected in Y. The third-harmonic effects would be somewhat reduced. The sample motor shows that the third-harmonic flux in this case drops from 18% at no load to 11% at full load.

(4) Saturated core and teeth: This is a point corresponding to the situation when the teeth are saturated, the normally flattened teeth flux density distribution requires that the flux density distribution in the core to be peaked up. However, because the core is also saturated, it tends to flatten the core flux and peak up the teeth flux. These contradicting tendencies allow only sinusoidally distributed core and teeth flux densities. If both the flux density distributions in the core and in the teeth are sinusoidal and the magnetic path is nonlinear under saturated conditions, the corresponding rotating magnetomotive force (MMF) must be nonsinusoidally peaked up, and contain a large third-harmonic component.

$$\text{MMF} = \text{MMF}_1 \cos(\theta - \omega t) - \text{MMF}_3 \cos(3\theta - 3\omega t) \quad (5)$$

where

MMF_1 = amplitude of fundamental MMF
 MMF_3 = amplitude of third-harmonic MMF

When the winding is connected in Y, the third-harmonic current cannot flow, and neglecting the higher order harmonics, the stator current is sinusoidal. Under this situation the only way to provide third-harmonic rotating MMF component in both stator core and teeth depends on the cage current at a certain slip (i.e., at asynchronous speed). In other words, for a Y-connected motor under saturated condition, regardless how equally saturated the teeth and the core sections are, the situation with sinusoidally distributed flux densities in the core and the teeth at synchronous speed does not exist. When the winding is connected in delta, the single phase, third-harmonic current produces a pulsating, third-harmonic component in the MMF. The forward rotating component of this pulsating field meets the requirement of the third harmonic MMF. Consequently, a nearly sinusoidal flux density distribution is possible.

Conclusion

(1) The trend towards increasing flux density in modern induction motors is accompanied by increasing third-harmonic flux in the magnetic path. This study shows that the core losses of induction motors are closely related to the magnitude and phase angle of the third harmonic flux. Proper use of third-harmonic flux can actually reduce the core loss.

(2) The shapes of flux density distributions in various sections of an induction motor are interrelated. The most saturated section normally has a flattened shape of flux density distribution. From the third-harmonic-flux standpoint this study compares the effects of the shapes of flux

density distribution on losses. Traditional design of induction motors normally chooses a lower core flux density than the teeth flux density. The reason behind it is that the core normally has a greater volume than the teeth, therefore, a lower core flux density may reduce the total core loss even though the tooth flux density has to be pushed up. This study shows that from a third-harmonic flux standpoint, the effect of the saturated teeth peaks-up the flux density distribution in the core, the original thought of reducing the core loss in the core may not be the most economical and efficient approach.

(3) The complexity of the stray-load loss evaluation is well recognized by motor designers. The suggested stray-load losses recommended by IEC and ANSI are estimated and used only when test data is not available. In order to estimate the real efficiency of induction motors, most motor designers rely on empirical equations to calculate stray-load loss. This study also attempts to investigate the influence on teeth harmonics and stray-load loss due to the shape of air-gap flux density distribution.

(4) Although the sample motor does not show significant difference in the magnitudes of teeth harmonics among the three shapes of air-gap flux density distribution, further study on this particular topic is being pursued for a more general conclusion.

(5) Under saturated condition, the stray-load loss according to the conventional definition goes down while the rms value of the stator current remains unchanged. The stray loss at no-load goes up under saturated condition. The total core and stray-load loss of a motor goes up when it is saturated.

(6) The circulating current in the delta-connected winding strongly diminishes the third-harmonic flux. Therefore a delta-connected winding may increase the losses if the third-harmonic flux is acting to reduce the losses.

(7) This study indicates that the losses influenced by the shapes of flux density distributions in a motor may be controlled by the third-harmonic flux through lamination punching designs and winding connections.

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