Optimal Choice for Number of Strands in a Litz-Wire Transformer Winding

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Abstract—The number and diameter of strands to minimize loss in a litz-wire transformer winding is determined. With fine stranding, the ac resistance factor can be decreased, but dc resistance increases as a result of the space occupied by insulation. A power law to model insulation thickness is combined with standard analysis of proximity-effect losses to find the optimal stranding. Suboptimal choices under other constraints are also determined.

Index Terms—Eddy currents, litz wire, magnet wire, power electronics, power transformers, proximity-effect losses, skin effect, transformer windings.

I. INTRODUCTION

A SALIENT difficulty in the design of high-frequency inductors and transformers is eddy-current effects in windings. These effects include skin-effect losses and proximity-effect losses. Both effects can be controlled by the use of litz wire—conductors made up of multiple individually insulated strands twisted or woven together. (Sometimes the term litz wire is reserved for conductors constructed according to a carefully prescribed pattern, and strands simply twisted together are called bunched wire. We will use the term litz wire for any insulated grouped strands.)

This paper addresses the choice of the degree of stranding in litz wire for a transformer winding. The number of turns and the maximum winding cross-sectional area are assumed to be fixed. Under constraints on maximum number of strands or minimum wire diameter, the best solution may not fill the allocated window space fully. However, as will be shown in Section IV, with those constraints removed, the optimum solution does fill the allocated space. In this case, the cross-sectional area of each turn is fixed, and as the number of strands increases, the cross-sectional area of each strand must be decreased. This typically leads to a reduction in eddy-current losses. However, as the number of strands increases, the fraction of the window area that is filled with copper decreases and the fraction filled with insulation increases. This results in an increase in dc resistance. Eventually, the eddy-current losses are made small enough that the increasing dc resistance offsets any further improvements in eddy-current loss, and the total losses start to increase. Thus, there is an optimal number of strands that results in minimum loss.

This paper presents a method of finding that optimum, using standard methods of estimating the eddy-current losses.

Optimizations on magnetics design may be done to minimize volume, loss, cost, weight, temperature rise, or some combination of these factors. For example, in the design of magnetic components for a solar-powered race vehicle [1] (the original impetus for this work) an optimal compromise between loss and weight is important. Although we will explicitly minimize only winding loss, the results are compatible with and useful for any minimization of total loss (including core loss), temperature rise, volume, or weight. This is because the only design change considered is a change in the degree of stranding, preserving the overall diameter per turn and overall window area usage. This does not affect core loss or volume and has only a negligible effect on weight. However, the degree of stranding does significantly affect cost. Although we have not attempted to quantify or optimize this, additional results presented in Section V are useful for cost-constrained designs.

The analysis of eddy-current losses used here does not differ substantially from previous work [2]–[18] ([15] gives a useful review). Although different descriptions can be used, most calculations are fundamentally equivalent to one of three analyzes. The most rigorous approach uses an exact calculation of losses in a cylindrical conductor with a known current, subjected to a uniform external field, combined with an expression for the field as a function of one-dimensional (1-D) position in the winding area [17]. Perhaps the most commonly cited analysis [16] uses “equivalent” rectangular conductors to approximate round wires and then proceeds with an exact 1-D solution. Finally, one may use only the first terms of a series expansion of these solutions, e.g., [14].

All of these methods give similar results for strands that are small compared to the skin depth [17]. (See Appendix B for a discussion of one minor discrepancy.) The solutions for optimal stranding result in strand diameters much smaller than a skin depth. In this region, the distinctions between the various methods evaporate, and the simplest analysis is adequate. More rigorous analysis (e.g., [17]) is important when strands are not small compared to a skin depth. In this case, losses are reduced relative to what is predicted by the analysis used here, due to the self-shielding effect of the conductor.

Previous work, such as [2]–[9] has addressed optimal wire diameter for single-strand windings. The approach in [2]–[9] is also applicable for litz-wire windings in the case that the number of strands is fixed, and the strand diameter for lowest loss is desired [5], [9]. As discussed in Section V, this can be
useful for cost-sensitive applications if the number of strands is the determining factor in cost and the maximum cost is constrained. However, this will, in general, lead to higher loss designs than are possible using the optimal number of strands.

II. SKIN EFFECT AND PROXIMITY EFFECT IN LITZ WIRE

Skin effect is the tendency for high-frequency currents to flow on the surface of a conductor. Proximity effect is the tendency for current to flow in other undesirable patterns—loops or concentrated distributions—due to the presence of magnetic fields generated by nearby conductors. In litz-wire windings, skin and proximity effects may be further divided into strand-level and bundle-level effects, as illustrated in Fig. 1. Bundle-level effects relate to current circulating in paths involving multiple strands, whereas strand-level effects take place within individual strands. Bundle-level effects are controlled by the pattern of twisting or weaving—the construction of the litz wire. Simple twisting is adequate to control bundle-level proximity effect loss, whereas more complex constructions are needed to control bundle-level skin effect. Bundle-level effects are not directly affected by the number or diameter of strands; they are determined by the overall diameter and the choice of twisting pattern. Thus, they need not be considered further for the analysis in this paper. At the strand level, proximity effect dominates skin effect in a winding that has many layers. Since a litz winding has a large effective number of layers as a result of the many strands, strand-level skin effects are negligible.

Thus, we need only consider strand-level proximity effect losses for the choice of number of strands. Strand-level proximity effect may be still further divided into internal proximity effect (the effect of other currents within the bundle) and external proximity effect (the effect of current in other bundles) [19], [20]. However, this distinction is useful only as a form of bookkeeping. The actual losses in one strand of a litz bundle are simply a result of the total external field, due to the currents in all the other strands present.

To calculate the total strand-level proximity-effect loss in a litz winding, one can view it as a single winding, made up of \(\eta N\) turns of the strand wire, each with current \(i/\eta\) flowing in it, where \(\eta\) is the number of strands, \(N\) is the number of turns of litz wire, and \(i\) is current flowing in the overall litz bundle. The loss in the litz winding will be the same as in the equivalent single-strand winding as long as the currents flowing in all the strands are equal [6], [21]. Other methods of calculating loss in litz wire also assume equal current in all strands [17], [19], [22]. This assumption is equivalent to assuming that the bundle-level construction has been chosen properly to control bundle-level proximity and skin effects. Note, however, that most of our results remain valid even when there is significant skin-effect loss at the litz-bundle level, for example in a simply twisted bundle. This is because the bundle-level skin-effect loss is independent of the number of strands, and is orthogonal [20] to the strand-level eddy-current losses.

We represent winding losses by

\[
P_{\text{loss}} = F_p I_{\text{ac}}^2 R_{\text{dc}}
\]

where \(F_p\) is a factor relating dc resistance to an ac resistance which accounts for all winding losses, given a sinusoidal current with rms amplitude \(I_{\text{ac}}\). As shown in Appendix A, we can approximate \(F_p\) by

\[
F_p = 1 + \frac{\pi \omega^2 \rho_k^2 N^2 \frac{2 \pi}{\sqrt{3}} k^2}{700 \sqrt{2} \frac{2 \pi}{\sqrt{3}}}
\]

where \(\omega\) is the radian frequency of a sinusoidal current, \(n\) is the number of strands, \(N\) is the number of turns, \(d_c\) is the diameter of the copper in each strand, \(\rho_c\) is the resistivity of the copper conductor, \(b_c\) is the breadth of the window area of the core, and \(k\) is a factor accounting for field distribution in multiwinding transformers, normally equal to one (see Appendix A). For waveforms with a dc component, and for some nonsinusoidal waveforms, it is possible to derive a single equivalent frequency that may be used in this analysis (Appendix C). In an inductor, the field in the winding area depends on the gapping configuration, and this analysis is not directly applicable [23].

The analysis described here considers the strands of all litz bundles to be uniformly distributed in the window, as they would be in a single winding using \(N \eta\) turns of wire the diameter of the litz strands. In fact, the strands are arranged in more or less circular bundles. In this sense, the analysis of [20] may be more accurate, but this difference has very little effect on the results. The most important difference between the model used here and the model in [20] is the greater accuracy of [20] for strands that are large compared to a skin depth. The simpler model is used because it is accurate for the small strand diameters that are found to be optimal, and because its simplicity facilitates finding those optimal diameters. Other models (such as [19] and the similar analysis in [22]) also model large strand diameters and circular bundle configurations accurately, but they fully calculate only internal (not external) proximity effect, and so are not useful for the present purposes.

III. DC RESISTANCE FACTORS

The fraction of the window area occupied by copper in a litz-wire winding will be less than it could be with a solid-wire winding. This leads to higher dc resistance than that of a solid wire of the same outside diameter. A cross section of litz wire is shown in Fig. 2, with the various contributions to cross-sectional area marked. In addition to the factors shown in this diagram, the twist of the litz wire also increases the
dc resistance. In order to find the optimal number of strands for a litz winding, it is necessary to quantify how the factors affecting dc resistance vary as a function of the number of strands.

A. Serving

Typically, litz bundles are wrapped with textile to protect the thin insulation of the individual strands. This serving adds about 0.06 mm (2.5 mil) to the diameter of the bundle. For a given number of turns filling a bobbin, or a section of a bobbin, the outside diameter of the litz wire must be fixed. The area devoted to serving will then also be fixed, independent of the number of strands.

B. Strand Packing

Simply twisted litz wire comprises a group of strands bunched and twisted into a bundle. More complex constructions begin with this step, and then proceed with grouping and twisting the subbundles into higher level bundles. Particular numbers of strands (1, 7, 19, 37, etc.) pack neatly into concentric circular arrangements. However, with large numbers of strands (e.g., >19), and/or very fine strands [e.g., 44–50 American Wire Gauge (AWG)], it is difficult to precisely control the configuration, and the practical packing factor becomes an average number approximately independent of the number of strands. Since the optimal strand diameter is typically much smaller than a skin depth, but the lowest level bundle can be near a skin depth in diameter, in most cases we can assume that there is a large number of strands in the innermost bundle. Thus, this packing factor is independent of the number of strands.

C. Bundle Packing and Filler

The way the strands are divided into bundles and subbundles is chosen based on considerations including bundle-level skin-effect losses, flexibility of the overall bundle, resistance to unraveling, and packing density. In some cases, a nonconducting filler material may be used in the center of a bundle in place of a wire or wire bundle that would, in that position, carry no current because of skin effect.

A typical configuration chosen to avoid significant bundle-level skin-effect losses should have a carefully designed and potentially complex construction at the large-scale level where bundle diameters are large compared to a skin depth. However, because the optimal strand diameter will be small compared to a skin depth, a simple many-strand twisted bundle may be used at the lowest level. If the overall number of strands is increased, the number of strands in each of these low-level bundles should be increased, but the diameter of each low-level bundle should not be changed, nor should the way they are combined into the higher level construction be affected. Thus, for our purposes, the bundle packing factor is independent of the number of strands.

D. Turn Packing

The way turns are packed into the overall winding is primarily a function of winding technique, and it is assumed not to vary as a function of the stranding. However, note that loosely twisted litz wires can deform as the winding is constructed, allowing tighter packing. Another option providing tight turn packing is rectangular-cross-section litz wire. In addition to its turn-packing advantage, it has tighter strand and bundle packing, as a result of the mechanical compacting process that forms it into a rectangular cross section.

E. Twist

The distance traveled by a strand is greater in a twisted bundle than it would be if the strands simply went straight, and so the resistance is greater. An additional effect arises from the fact that a cross section perpendicular to the bundle cuts slightly obliquely across each strand. Thus, the cross section of each strand is slightly elliptical. This reduces the number of strands that fit in a given area, and so effectively increases the resistance. An extreme case of this is illustrated in Fig. 3. The choice of the pitch of the twist (“lay” or length per twist) is not ordinarily affected by the number of strands in the lowest level bundle, and so for the purpose of finding the optimal number of strands, we can again assume it is constant.

F. Strand Insulation Area

Thinner magnet wire has thinner insulation. However, the thickness of the insulation is not in direct proportion to the wire diameter. Thinner wire has copper in a smaller fraction of the overall cross-sectional area and insulation in a larger fraction.
where \( d_t \) is the overall diameter, including the insulation thickness, \( d_c \) is the diameter of the copper only, and \( d_r \) is an arbitrarily defined reference diameter used to make the constants \( \alpha \) and \( \beta \) unitless. The parameters found for single-build insulation wire are \( \beta = 0.97 \) and \( \alpha = 1.12 \) for \( d_r \) chosen to be the diameter of AWG 40 wire (0.079 mm). For heavy-build insulation, \( \beta = 0.94 \) and \( \alpha = 1.24 \). Note that although (4) provides an accurate approximation for wire in the range of 30–60 AWG, its asymptotic behavior for large-strand diameters is pathological. Insulation thickness goes to zero around 6 AWG and is negative for larger strands.

**IV. NUMBER OF STRANDS FOR MINIMUM LOSS**

With no constraints on number of strands or strand diameter, the minimum-loss design will be with a full bobbin. Any design that does not fill the bobbin can be improved by increasing the number of strands by a factor \( s \), and decreasing the strand diameter \( d_c \) by \( 1/\sqrt{s} \). This keeps the dc resistance constant and decreases ac resistance, as shown by (2). This improvement can be continued until insulation area increases result in a full bobbin. Thus, the minimum-resistance solution fills the bobbin, and we can find this solution by analyzing a full bobbin.

For a full bobbin, the outside diameter of the complete litz bundle is

\[
d_t = \sqrt{\frac{F_p b_h h}{N}}
\]

where \( b_h \) is the breadth of the bobbin, \( h \) is the height allocated for the particular winding under consideration, \( N \) is the number of turns in that winding, and \( F_p \) is a turn-packing factor for turns in the winding, expressed relative to perfect square packing (for \( F_p = 1 \), the litz bundle would occupy \( \pi/4 \) of the window area).

Assuming a factor \( F_{lb} \) accounting for serving area, bundle packing, any filler area, strand packing, and the effect of twist on diameter, we can find the outside diameter of a single strand

\[
d_t = \sqrt{\frac{F_p F_{lb} b_h h}{n N}}
\]

where \( n \) is the number of strands in the overall litz bundle.

The diameter of the copper in a single strand can then be written using (4)

\[
d_c = d_r \alpha \left( \frac{d_c}{d_r} \right)^\beta
\]

We now define a total resistance factor \( F_r' \)

\[
F_r' = F_{dc} F_r = \frac{\text{ac resistance of litz-wire winding}}{\text{dc resistance of single-strand winding}}
\]

where \( F_{dc} \) is the ratio of dc resistance of the litz wire to the dc resistance of a single strand winding, using wire with the same diameter as the litz-wire bundle. Using (6) and (7), we can show

\[
F_{dc} = n^{1/\beta} F_{lb}^{1/\beta}
\]
Combining (2), (8), and (9) results in

$$F_r' = F_p'^{1/\beta} [n^{1/\beta - 1} + \gamma n^{(1-2/\beta)}]$$  \hspace{1cm} (10)

where

$$\gamma = \frac{\pi^2 N^2 \omega^2 \rho_0^2 d_r^{6/\beta} F_r^{6/\beta} \alpha^{6/\beta} (F_p' F_p b_h h / N)^{3/\beta} k}{768 \rho_{dc}^2 \mu_0}.$$  \hspace{1cm} (11)

Equation (10) can now be minimized with respect to $n$ to find the optimal number of strands

$$n_{opt} = \left( \frac{2/\beta - 1}{1/\beta - 1} \right)^{1/(3/\beta - 2)}.$$  \hspace{1cm} (12)

This will give nonintegral numbers of strands; the nearest integral number of strands can be chosen to minimize ac resistance.

V. DESIGN EXAMPLES AND SUBOPTIMAL STRANDING

For a design example, we used a 14-turn winding on an RM5-size ferrite core. The breadth of the bobbin is 4.93 mm, and the breadth of the core window 6.3 mm. A height of 1.09 mm is allocated to this winding. Based on an experimental hand-wound packing factor $F_p$ = 0.85 and litz packing factor $F_l$ = 0.66, unserved, plus a 32-$\mu$m (1.25 mil) layer of serving, the above calculation indicates that, for a frequency of 375 kHz, 130 strands of number 48 wire gives minimum ac resistance, with a total resistance factor of $F_r' = 2.35$, ac resistance factor $F_r = 1.03$, and dc resistance factor $F_{dc} = 2.29$.

Fig. 5 shows the total calculated resistance factor and its components as a function of number of strands. The figure and the numbers confirm the intuition that because $\beta$ is close to one and the dc resistance increases only very slowly, the decrease in resistance using finer strands outweighs the decreased cross-sectional area until the ac resistance factor is brought very close to one. Note that although the factor $F_{dc}$ is large, only a factor of 1.18 is due to the change in wire insulation thickness. The remaining factor of 1.95 is due to the dc resistance factors that do not vary with number of strands, such as serving area and strand packing.

The optimization leads to choosing a large number of fine strands, which will often mean a high cost, and will sometimes require finer strands than are commercially available. From Fig. 5, one can see that a decrease from the optimum of 130 to about 50 strands entails only a small increase in ac resistance. Consideration of the cost tradeoff for a particular application becomes necessary.

Given a suboptimal number of strands, chosen to reduce costs, a full bobbin may no longer be best. The problem of choosing the optimal strand diameter for a fixed number of strands has been addressed by many authors [2]–[5], [7]–[9], [14]. Although this is typically only used for single strands, the analysis also can be applied for more than one strand by simply using the product of the number of turns and the number of strands $N_{str}$ in place of the number of strands $N$. The result [7]–[9], [14] that $F_{opt} = 1.5$ holds, and

$$d_{opt} = \left( \frac{384 \rho_{dc}^2 \rho_0^2}{\pi^2 \omega^2 F_r^{6/\beta} \alpha^{6/\beta} N^{3/\beta}} \right)^{1/6}.$$  \hspace{1cm} (13)

In many practical cases, cost is a stronger function of the number of strands than of the diameter of the strands. In the range of about 42–46 AWG, the additional manufacturing cost of smaller wire approximately offsets the reduced material cost. Thus, designs using the diameter given by (13) often approximate the minimum ac resistance for a given cost.

Fig. 6 shows total resistance factor as a function of the number of strands for the same example design, but at 1 MHz, where the optimal stranding is a difficult and expensive 792
strands of AWG 56 wire, and so analysis of alternatives is more important. The solid line is for a full bobbin, and the dashed line is for the same number of strands, but with the diameter chosen for minimum losses rather than to fill the bobbin. Where the two lines meet, the optimal diameter just fills the bobbin. Beyond that point it would not fit, and the line is shown dotted.

The example can be understood more completely by examining contour lines of total resistance factor $F^r$, as a function of number of strands and diameter of strands for the example discussed in the text at 1 MHz. The diagonal dashed line indicates a full bobbin. The valley at the upper right is the minimum loss. The minimum loss without overfilling the bobbin is marked by an “x.” Contour lines are logarithmically spaced.

One could also consider a constraint for minimum wire diameter. Many manufacturers cannot provide litz wire using strands finer than 48 or 50 AWG. In Fig. 7, the minimum resistance for 50 AWG stranding is with a full bobbin, but for 40-AWG wire, the minimum ac resistance can be seen to occur with fewer than the maximum number of strands. This situation can be analyzed by considering (2) with all parameters fixed except for the number of strands, such that

$$F^r = 1 + \zeta n^2$$  \hspace{1cm} (14)

where $\zeta$ is a constant obtained by equating (2) and (14). The total resistance factor is then

$$F^r_\text{dc} = F^r_{\text{dc}1} \cdot \frac{(1 + \zeta n^2)}{n}$$  \hspace{1cm} (15)

where $F^r_{\text{dc}1}$ is the dc resistance factor with a full bobbin, for the fixed strand diameter. The value of $n$ that minimizes this expression is $n = \sqrt{1/\zeta}$, such that $F^r = 2$. This will be the optimal number of strands, given a fixed minimum strand diameter, unless this is too many strands to fit in the available window area.

The above analysis shows how to find the optimal stranding, given a constraint on either strand diameter or number of strands, both of which are important in determining cost. More explicit analysis of cost is addressed in [25].

VI. EXPERIMENTAL RESULTS

The designs specified in the preceding section were constructed with two types of litz wire: 130 strands of 48 AWG and 50 strands of 44 AWG. The primary and secondary windings were made from a single length of litz wire, wound on the bobbin in opposite directions. This is magnetically equivalent to having a shorted secondary, but it reduces potential problems with interconnect resistance. In order to evaluate skin effect in the absence of external proximity effect, litz wire was also measured outside of a winding. The resistance was measured with an HP 4284A LCR meter, using a custom-built test jig for low-impedance measurements. The measurements are shown in Fig. 8.

Although the overall litz-wire diameter was small enough to limit bundle-level skin-effect losses to a few percent, the fine strands in the optimal solution also limit proximity-effect losses to similar levels, so it is necessary to separate the two effects in order to judge the accuracy of the proximity-effect calculations. Fortunately, the losses are orthogonal [20], and the skin effect losses (for a litz wire outside of the winding) can simply be subtracted from the measured losses in the transformer in order to isolate proximity-effect losses. Accurate prediction of the bundle-level skin effect was found to be difficult, in part because the details of the bundle constructions were not well known. However, if the experimentally measured bundle-level skin effect is subtracted from the total measured losses, the result matches the proximity-effect losses predicted by (2) very closely, as can be seen in Fig. 8. This confirms the validity of the model used in the optimization.

VII. CONCLUSION

The number of strands for a minimum-loss litz-wire winding may be found by evaluating the tradeoff between proximity-effect losses and dc resistance. Of the factors leading to increased dc resistance in a litz-wire winding, only the space allocated to strand insulation varies significantly with the number of strands in a well designed construction. A power law can be used to model insulation thickness in the region of interest. Combining this with standard models for eddy-current loss results in an analytic solution for the optimal number of strands. The simplest model for loss, using only the first terms of a series expansion, can be used because good designs use strands that are small compared to a skin depth. Experimental results correlate well with the simple model.

Stranding for minimum loss may lead to many strands of fine wire and thus excessive expense. Minimum loss designs constrained by minimum strand size or maximum number of strands have also been derived.
field is not zero at one edge of the winding, a factor $k = (1 - \varphi^3)/(1 - \varphi^3)$ is used to account for the resulting change in losses, where $\varphi = B_{\text{min}}/B_{\text{max}}$ [14].

APPENDIX B

COMPARISON WITH EXPANSION OF DOWELL SOLUTION

Equation (2) is similar to the expression for the first terms of a series expansion of the exact one-dimension solution

$$F_r = 1 + \frac{5p^2 - 1}{45} \psi^4$$  (17)

where $p$ is the number of layers and $\psi$ is the ratio of effective conductor thickness to skin depth. For a large number of layers (equivalent to the assumption, given above, of a trapezoidal field distribution), this reduces to $F_r = 1 + (p^2/9) \psi^4$. The usual expression for $\psi$ is

$$\psi = \sqrt{F_r h_{\text{eq}}/\delta}$$  (18)

where $F_r = N_h h_{\text{eq}}/b_h$, $h_{\text{eq}}$ and $b_{\text{eq}}$ are the height and breadth of an “equivalent” rectangular conductor and $N_h$ is the number of turns per layer. Based on equal cross-sectional area, $h_{\text{eq}} = h_{\text{eq}} = \sqrt{\pi/4} d_c$. This results in

$$\psi^4 = \frac{(\pi/4)^2 h_{\text{eq}}^2 N_h}{84b_h b_c}$$  (19)

where $h_b$ is the height of the bobbin area allocated to this winding. The number of layers is $p = \sqrt{n/N_h b_h/b_c}$. Substituting these expressions for $p$ and $\psi$ into the simplified version of (17) and using $\delta = \sqrt{2\delta_c/(\pi h_{\text{eq}})}$, we obtain

$$F_r = 1 + \frac{\pi^2 h_{\text{eq}}^2 r_c^2 N_h^2 h_{\text{eq}}^2}{3 \times 784 r_c^2 b_c^2}$$  (20)

the same as (2), except for the substitution of $b_h$ for $b_c$ and the addition of a factor of $\pi/3$. This discrepancy, which was first noted in [6], can be explained by comparing (16) to the equivalent expression for a rectangular conductor

$$P = \frac{\omega^2 B^2 s^4}{24 \rho_c}$$  (21)

where $s$ is the side of a square conductor. Equating these two, we obtain $s = (3\pi/16)^{1/4} d_c$. Thus, it appears that using an equivalent square conductor with sides equal to $s = (3\pi/16)^{1/4} d_c$ for proximity-effect loss calculations would be a more accurate approximation than the equal area approximation that is usually used [16].

APPENDIX C

NON-SINUSOIDAL CURRENT WAVEFORMS

Non-sinusoidal current waveforms can be treated by Fourier analysis. The current waveform is decomposed into Fourier components, the loss for each component is calculated, and the loss components are summed to get the total loss

$$P = \sum_{j=0}^{\infty} F_r(\omega_j) R_{dc}$$  (22)
where $I_j$ is the rms amplitude of the Fourier component at frequency $\omega_j$. From (2), it can be seen that

$$F_r(\omega) = 1 + (F_R(\omega_1) - 1) \frac{\omega_1^2}{R_{dc}}. \quad (23)$$

Defining $F_{T\text{-tot}}$ by $P = E_{T\text{-tot}} R_{dc}$ leads to

$$F_{T\text{-tot}} = 1 + \frac{(F_R(\omega_1) - 1) \sum_{j=0}^{\infty} I_j^2 \omega_j^2}{R_{dc}}. \quad (24)$$

This can also be written as

$$F_{T\text{-rms}} = 1 + (F_R(\omega_1) - 1) \frac{\omega_{eff}^2}{\omega_1^2}. \quad (25)$$

where

$$\omega_{eff} = \sqrt{\frac{\sum_{j=0}^{\infty} I_j^2 \omega_j^2}{\sum_{j=0}^{\infty} I_j^2}}. \quad (26)$$

One may calculate this effective frequency for a nonsinusoidal current waveform and use it for analysis of litz-wire losses, or for other eddy-current loss calculations. Note that this applies to waveforms with dc plus sinusoidal or nonsinusoidal ac components. The results will be accurate as long as the skin depth for the highest important frequency is not small compared to the strand diameter.

A triangular current waveform with zero dc component results in an effective frequency of $1.10 \omega_1$, where $\omega_1$ is the fundamental frequency. Once the effective frequency of a pure ac waveform has been calculated, the effective frequency with a dc component can be calculated by a reapplication of (26)

$$\omega_{eff} = \sqrt{\frac{P_{dc}^2 \omega_{eff,dc}}{P_{dc}^2 + P_{ac}^2}}. \quad (27)$$

Finding Fourier coefficients and then summing the infinite series in (26) can be tedious. A shortcut, suggested but not fleshed out in [9], can be derived by noting that

$$\sum_{j=0}^{\infty} I_j^2 \omega_j^2 = \left[ \text{rms} \left\{ \frac{d}{dt} I(t) \right\} \right]^2 \quad (28)$$

so that

$$\omega_{eff} = \frac{\text{rms} \left\{ \frac{d}{dt} I(t) \right\}}{I_{T\text{-tot\text{-rms}}}}. \quad (29)$$

The primary limitation of effective-frequency analysis is that it does not work for waveforms with more substantial harmonic content. For instance, the series in (26) does not converge for a square wave. Similarly, the derivative of a square wave in (29) results in an infinite rms value. A Bessel-function-based description of loss may be necessary. However, in practice leakage inductance prevents an inductive component from having perfectly square current waveforms.

A square wave with finite-slope edges leads to a finite value of $\omega_{eff}$, which can be found from (29) to be

$$\omega_{eff} = \frac{\omega_1}{\pi} \sqrt{\frac{6}{\Delta(3 - 4\Delta)}} \quad (30)$$

where $\Delta$ is the transition time as a fraction of the total period. For $\Delta = 0.5$, the waveform becomes triangular and (30) gives the same value of $\omega_{eff}$ as calculated above. This expression (30) and the calculation and minimization of loss based on (2) is valid as long as there is not significant harmonic current for which the wire diameter is large compared to a skin depth. Based on the rule thumb that the highest important harmonic number is given by $N = 0.35/\Delta$ [27], a rough check on this would be to calculate skin depth for a maximum frequency $\omega_{max} = (0.35/\Delta)$ and compare this to the wire diameter. If there are significant harmonics for which the skin depth is small compared to wire diameter, then the analysis in [27] can facilitate 1-D analysis of nonsinusoidal waveforms, or for more accuracy Bessel-function analysis [17] with a Fourier decomposition of the waveform can be used.

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REFERENCES


SULLIVAN: OPTIMAL CHOICE FOR NUMBER OF STRANDS IN A LITZ-WIRE TRANSFORMER WINDING


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