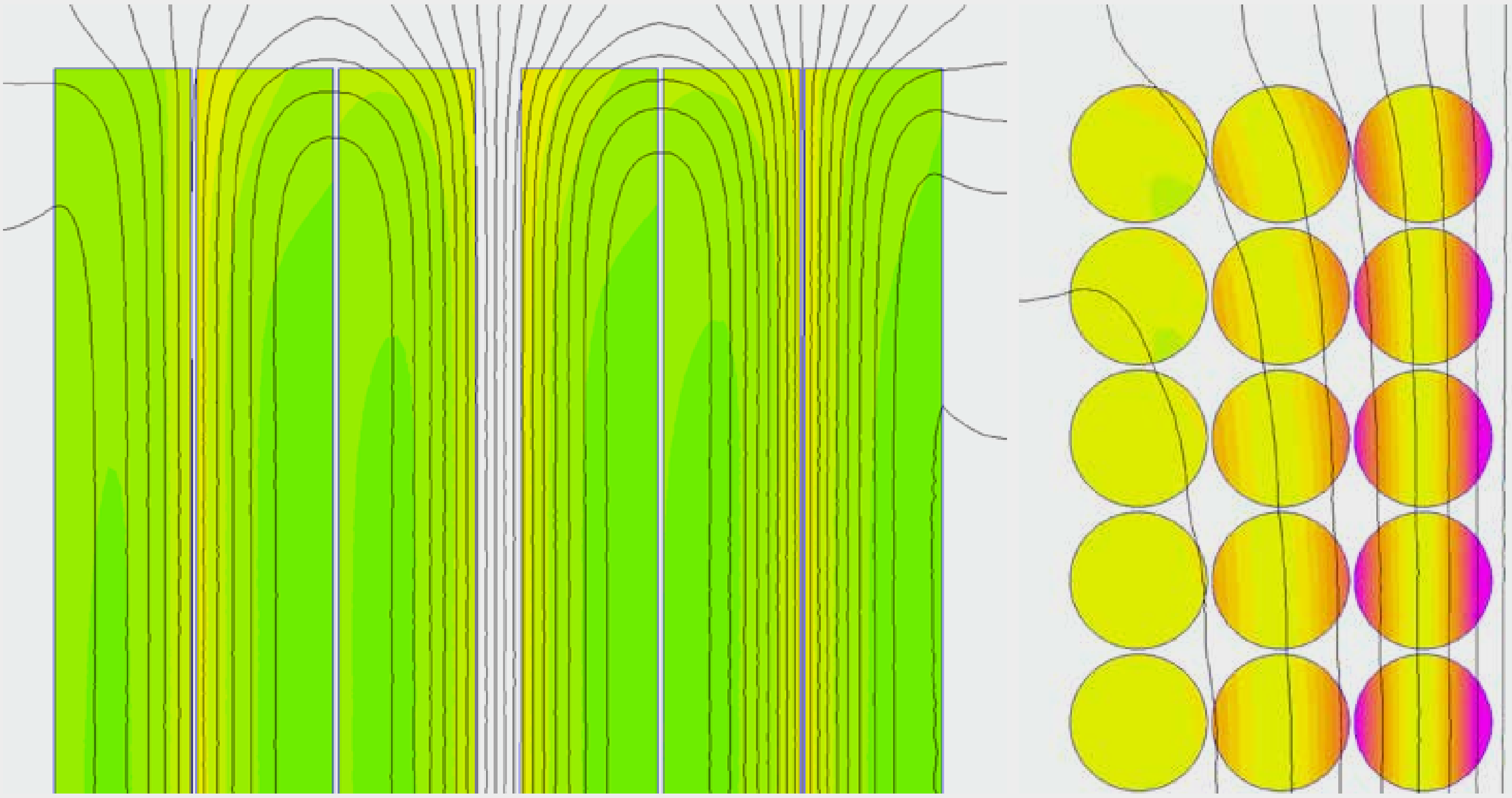


Qualitative FEM study of proximity loss reduction by various winding configurations – Part I

ABSTRACT

Skin depth and proximity effects in transformer windings are important phenomena influencing the design even at power frequencies (50-60 Hz). However, they become critically important at elevated frequencies, especially for high-frequency transformers, operating in switched-mode power supplies for example, at any power level. This article presents a numerical study of optimum winding configurations which can drastically reduce the proximity effects. It is possible to make transformers operating even at 1 MHz without the use of very expensive Litz wire.

KEYWORDS

copper loss, proximity effect, proximity loss, skin effect, eddy currents, windings

How to make a 1 MHz transformer without expensive Litz wire

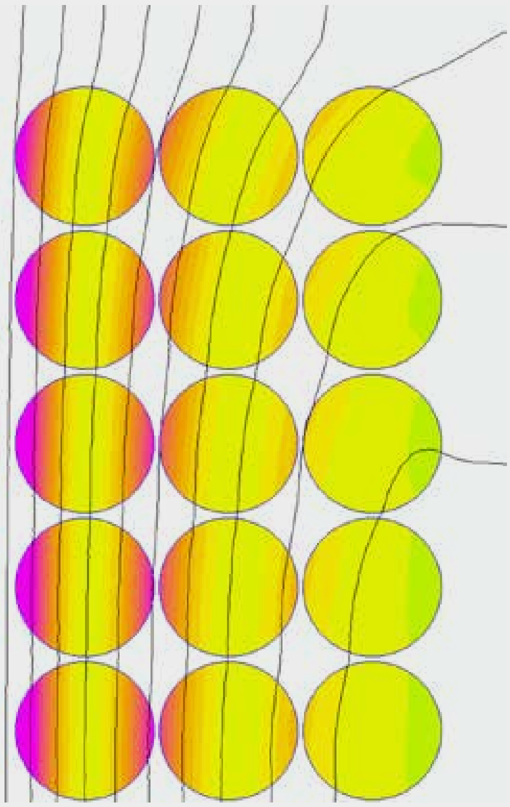
1. Introduction

Efficiency of energy transformation is affected by multiple loss components occurring in transformers. These components depend on the given application, topology of conversion (conventional or switched mode), frequency of operation, energy required for forced cooling, etc.

Loss in the windings (copper loss) is a significant part of the total loss in most power transformers (e.g. as opposed to signal transformers). Electrical resistance of a given winding has an immediate impact on the total efficiency at full load due to Joule's heating. Conductors with greater effective cross-sectional area must be used to reduce the losses and improve the efficiency.

However, transformer designers also have to take into account other, more complex phenomena, such as the *skin effect*. Elec-

With an appropriate core-winding configuration it is possible to use solid wire instead of expensive multi-strand Litz wire, even at 1 MHz



Efficiency of energy transformation is affected by multiple loss components occurring in transformers, which depend on the given application, topology of conversion, frequency of operation, etc.

tromagnetic wave penetrating a conducting material (copper) induces surface eddy currents which oppose the penetrating field, thus attenuating the internal electromagnetic field. The attenuation follows an exponential function and at a distance of one skin depth the field is attenuated to 37 % of the surface value. Therefore, the material deeper inside the conductor carries less current and the copper is used less inefficiently.

Even at 50 Hz or 60 Hz the skin depth in copper is around 9 mm (Fig. 1a, b). This is partly the reason for the high-current busbars in low-voltage switchgear to be made with copper bars not thicker than 10 mm (Fig. 1c).

For higher frequencies the skin depth reduces proportionally to the square root of frequency. This becomes an important limitation for High-Frequency (HF) transformers, because the conductors cannot be thicker than around double the thickness of the skin depth in order to use the copper efficiently. For this reason many HF transformers use thin but wide copper tape in order to maintain the effective cross-sectional area of the conductor without increasing its thickness.

HF transformers are widely used in switched mode power supplies (SMPS) from a sub-W to hundreds of kW. The currents usually have very high harmonic content (triangular, trapezoidal or rectangular pulses) which further exacerbates the losses due to the fact that skin depth becomes shallower for higher frequencies. The problem is described in more detail elsewhere [1, 2]. This paper focuses only on sinusoidal currents.

2. Proximity loss

There is also a phenomenon called *proximity effect* which can have even more severe consequences than the skin depth. The proximity loss depends strongly not only on the frequency, but also on a number of layers in a given set of conductors. The name "proximity" comes from the fact that the effect is caused by the presence of other conductors with currents in the immediate vicinity of each other [3, 4].

Any electric current generates magnetic field around itself and this field penetrates any neighbouring conductor. According to the Faraday's law, if the magnetic field varies then a voltage will be induced in a conducting loop directly, proportionally to the amplitude and frequency of the field penetrating it. Such induced voltage gives rise to eddy current, whose position and direction is such as to oppose the magnetic field generating it (as defined by the Lenz's law).

For neighbouring conductors the effect leads to "side" eddy currents (Fig. 2). The amplitudes of the eddy currents add/subtract accordingly with the direction of the main current and hence the localised

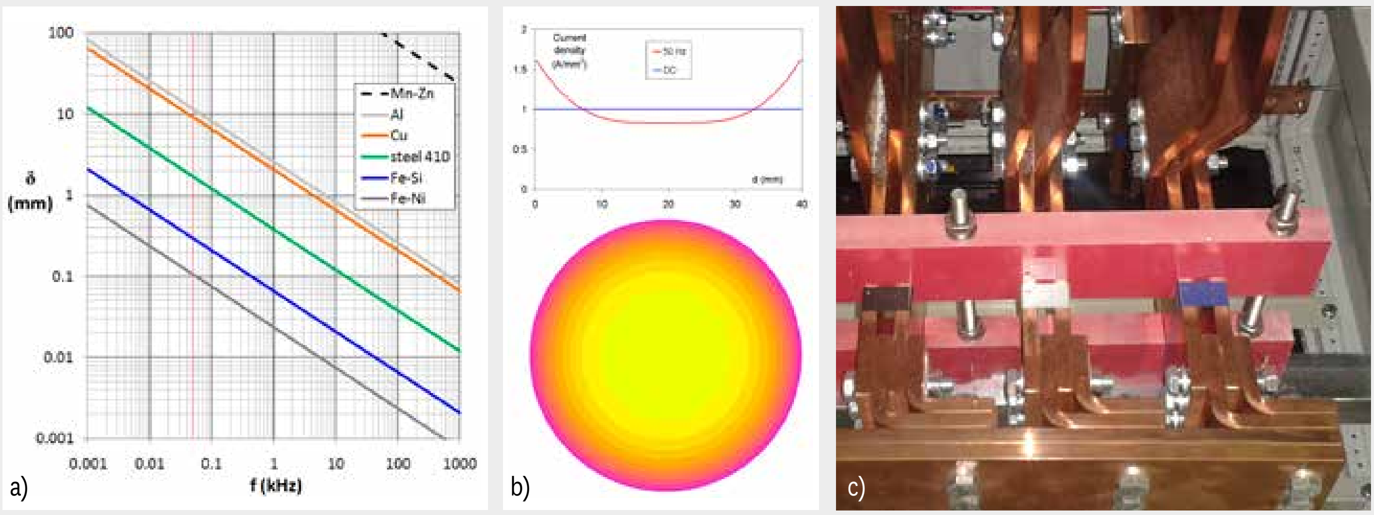


Figure 1. (a) Skin depth for various conductors and frequencies;
(b) Hypothetical current distribution in 40 mm diameter copper rod at DC and 50 Hz;
(c) LV distribution board with 10 mm thick copper busbars

The proximity loss strongly depends not only on the frequency, but also on the number of layers in a given set of conductors

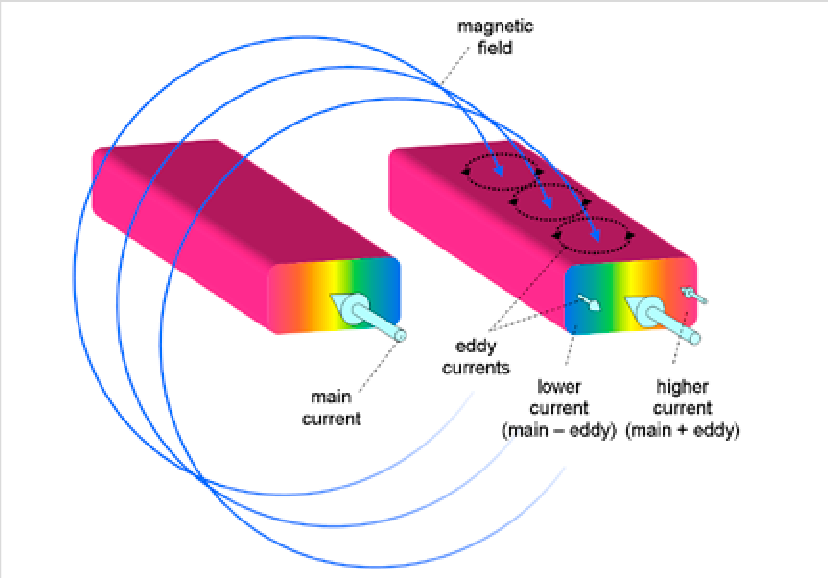


Figure 2. Illustration of the proximity effect [2]

current density in the conductor is not uniform. As a result, the effective resistance of the wire is increased for alternating currents (so-called AC resistance) and thus the copper losses can be elevated significantly.

The distribution of current density depends on many factors, for instance, on the number of adjacent conductors. The effect is schematically illustrated in Fig. 3. A hypothetical simplified single-phase transformer has the tape windings made as three layers of primary and three layers of secondary (each winding is connected in series so that there are three turns in each coil). At full load and with the turn ratio of 1:1 the currents flowing in primary and secondary can be assumed equal.

The magnetic field strength *H* builds up with each layer of the primary winding, to reach the maximum between the windings. The current in the secondary winding flows in the opposite direction and thus *H* is reduced with each layer of secondary.

However, the localised eddy currents are stronger in the presence of higher *H*, and the power *P* lost in the windings is proportional to the square of the eddy currents. *H* is much lower in the outer conductors so the power loss is at similar level as it would be for DC current ($P = 100$ %). But the inner conductors are exposed to much greater amplitudes of *H* and thus *P* can grow disproportionately large (over 600 % peak in Fig. 3).

The situation shown in Fig. 3 is for conductor thickness equal to the skin depth, which is usually sufficient for minimising problems with just the skin effect. However, as shown in Fig. 3, the proximity effect can produce significantly elevated losses even for such conditions. This is a real problem for high-frequency applications like HF transformers for SMPS (Fig. 4). The problem occurs also in small electronic transformers [5], medium-frequency transformers [6], rotating machines [7], as well as three-phase busbars whose thickness is usually limited to 10 mm for this reason [3].

Usually, proximity loss calculations are too complex to be solved analytically [8]. Therefore, many studies are carried out numerically, by employing FEM or Finite Element Modelling [9, 10]. This is not always practical, because 3D problems are very computationally intensive and multi-strand wires cannot be fully modelled, especially for higher frequencies or larger components, in which each turn of the Litz wire could be composed from hundreds of strands (Fig. 5).

There are many technical and scientific papers describing various ways of calculating the losses due to proximity effect. A well-known design principle is to follow the so-called Dowell's curves, which are generalised and normalised curves helping in the assessment of the apparent increase in AC resistance over the DC resistance (coefficient k in Fig. 4) [11].

As seen from Fig. 4 the proximity loss can be reduced in two ways. The first route is

The localised eddy currents are stronger in the presence of a higher *H*, and the power *P* lost in the windings is proportional to the square of the eddy currents

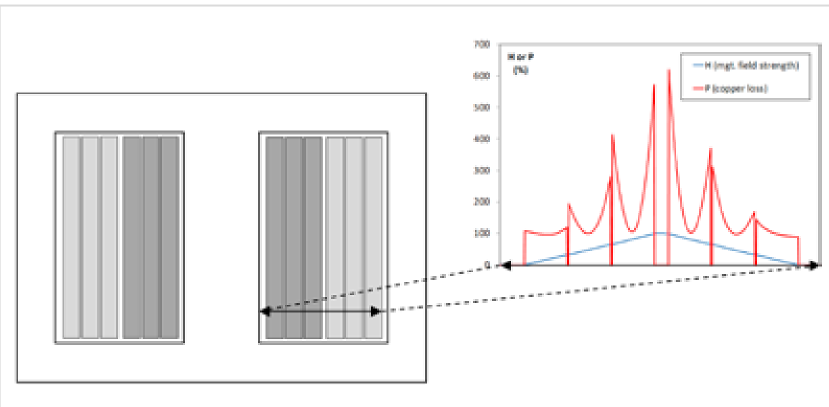


Figure 3. Illustration of proximity loss in single-phase transformer windings with 3-layer primary (dark grey) and 3-layer secondary (light grey); each winding is connected in series

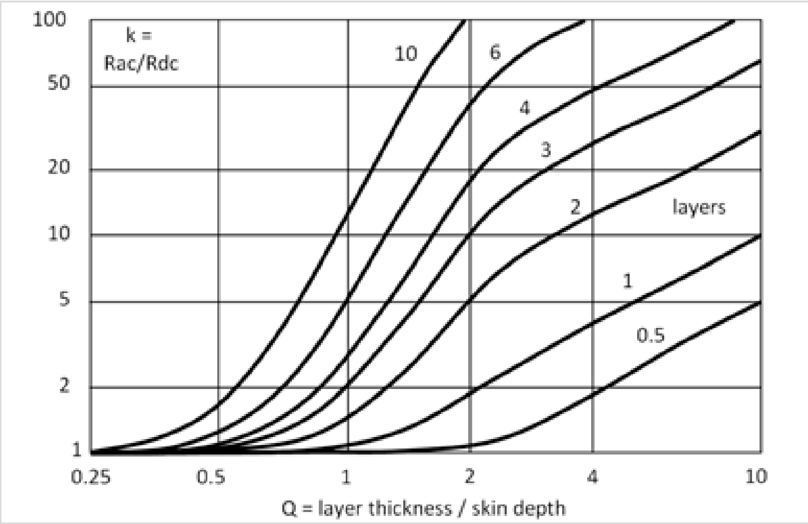


Figure 4. Dowell's curves for tape conductors [11]

The proximity loss can be reduced by reducing the wire thickness and the number of layers

to reduce the wire thickness (which is proportional to the ratio of layer thickness to the skin depth). This can be achieved by using a multi-strand Litz wire (each strand is separately isolated). However, at the same time there will be more layers, because of the increased number of strands. So the improvement is not straightforward, but eventually it can be achieved by using sufficiently thin strands, which is costly and more difficult in production. This approach is frequently taken for high-frequency transformers because off-the-shelf components can be used (Fig. 5).

Reducing the number of layers can also produce improvement of performance (Fig. 4). For instance, the above mentioned hypothetical three-layer transformer with Q = 1 (conductor thickness equal to skin depth) will suffer from k = 2, so the proximity loss would be responsible for additional loss of 200 %.

The layer thickness is the most important parameter for proximity effects

The Q value can be reduced by using thinner conductors (like in Litz wire). If the conductors in the hypothetical transformer had half the thickness, then the Q value would be reduced to 0.5. But then each turn would be constructed from two sub-layers so that the total number of layers would be 6. For Q = 0.5 this would produce lower AC resistance than the original configuration (Fig. 4), but it still might be an insufficient improvement and the "strand" thickness would have to be reduced further. This carries a penalty in terms of the associated costs of thinner

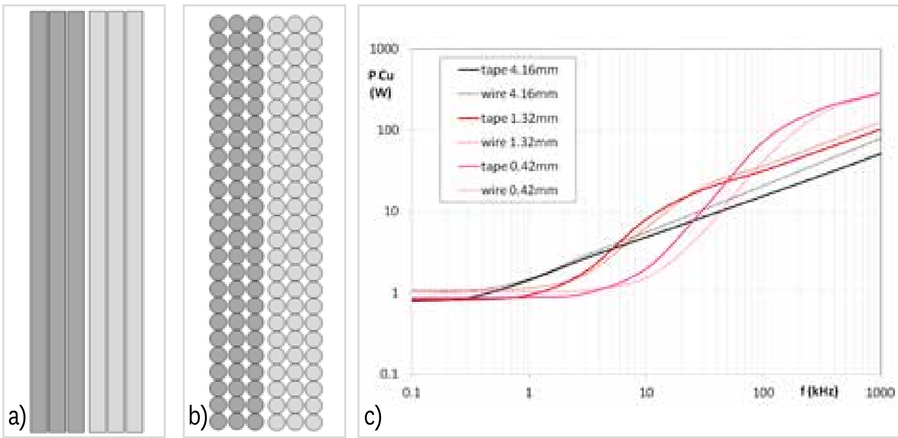


Figure 6. Comparison of tape (a) and wires connected in parallel (b), and their winding performance (c)

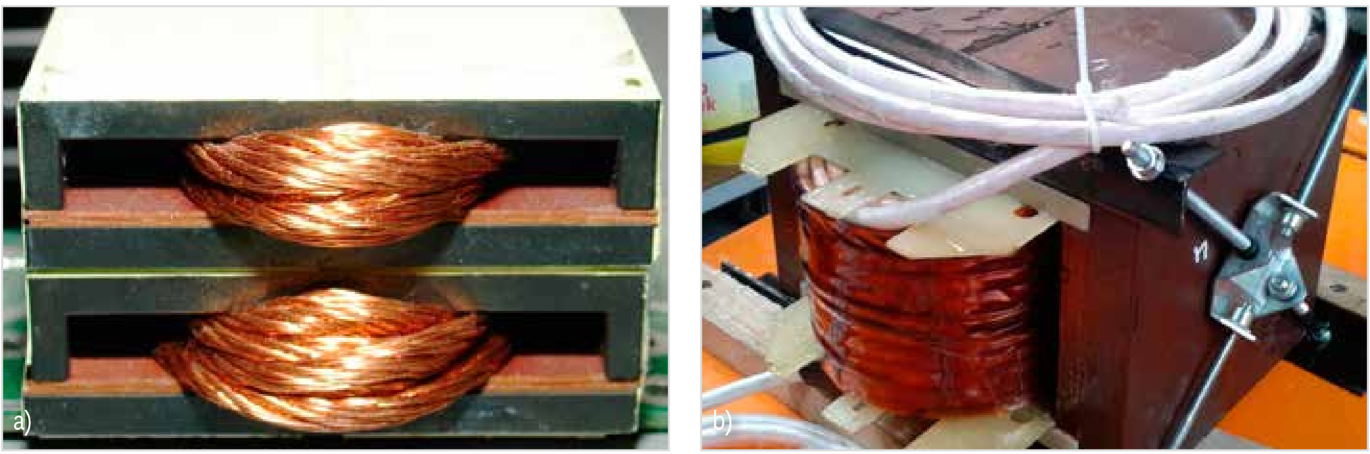


Figure 5. Examples of HF transformers made with Litz wire: a) 500 W rated with a ferrite core (Litz strands visible); b) 50 kW rated with a nanocrystalline core (Litz wire with 564 strands, 75 A)

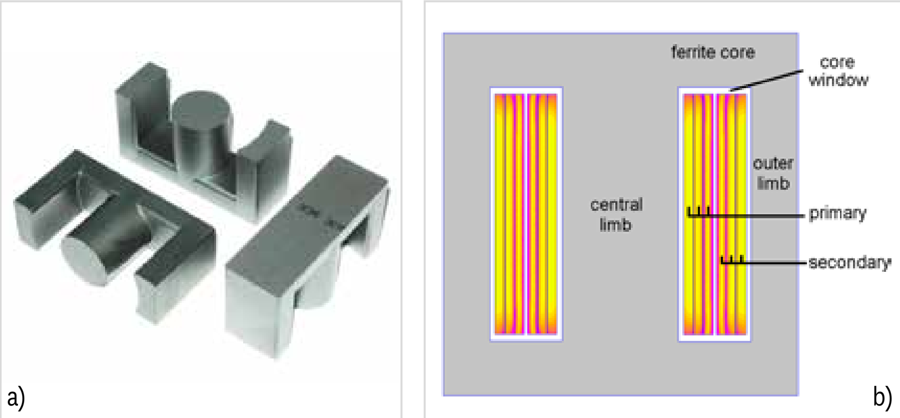


Figure 7. Studied model: a) ferrite core ETD49; b) example of calculated results for three-layer primary and three-layer secondary windings (compare with Fig. 3)

conductors, assembly of the winding and lower copper fill factor of the available winding window.

3. Study of proximity effects

It should be noted here that it is the layer thickness which is the most important parameter. The effects are similar, especially from qualitative viewpoint, for a round wire and a corresponding tape thickness [10]. At low frequencies the resistance of tape is lower due to the π/4 factor resulting from the round shape of the wire. But at high frequencies the proximity effect dominates and the order-of-magnitude changes are mostly driven by the basic thickness of the conductor, as well as by other parasitic effects like distribution of the leakage inductance within the core window, distance between the core and the winding, wire spacing within the same layer, etc. [12]. Therefore, for investigating qualitative behaviour computer simulations can be carried out for tapes instead of individual wires, as this greatly reduces the computation effort (Fig. 6).

The study presented below is based on dimensions of an EE-type off-the-shelf ferrite core ETD49 (Fig. 7) [13]. The core was modelled without an air gap, and the core losses were not considered as they are irrelevant for the proximity loss. The total copper loss including eddy currents, skin and proximity effects were calculated automatically by the software [14]. All simulations were performed for sinusoidal currents.

The core window could be filled with the windings in a number of ways. The base configuration was a single primary turn and single secondary turn (Fig. 8b). The base current density was set arbitrarily to 1.41 A/mm^2, which corresponded to a peak current of 274.6 A (194.1 A rms) in both primary and secondary winding (1:1 transformer at full load). So, for instance, for a three-layer winding (Fig. 8c) the peak current in each conductor was 91.5 A (because 274.6 / 3 = 91.5). The value of current was modified accordingly for each configuration so that the total ampere-turns were identical in every modelled case, for all the wire and tape configurations.

Thickness of the bobbin was assumed as 1 mm, and minimum spacing between the layers due to insulation was 0.05 mm. For the thinnest winding (0.13 mm) this represented a significant fraction of the core window so that it was possible to use only 23 turns (instead of the required 27 in order to maintain the factor of 3 between the subsequent configurations). This was one of the reasons for arbitrarily choosing such a fairly large core size, in order to provide the possibility of modelling multiple-layer windings with a practical insulation thickness.

The total copper loss versus frequency (from 100 Hz to 1 MHz) for these configurations is shown in Fig. 9. It is interesting to see that for the highest frequency the thinnest „strands" result with the highest losses, and this is caused by the proximity effect. For this configuration (Fig. 8e) the Q factor at 1 MHz is 2 and the number of layers is 23. It can be seen from Fig. 4 that for Q = 2 and 10 layers, the AC loss would increase around 100-fold (which

One way of reducing the proximity loss is to reduce the number of effective layers, which can be achieved by interleaving primary and secondary windings so that the peak magnetic field is reduced

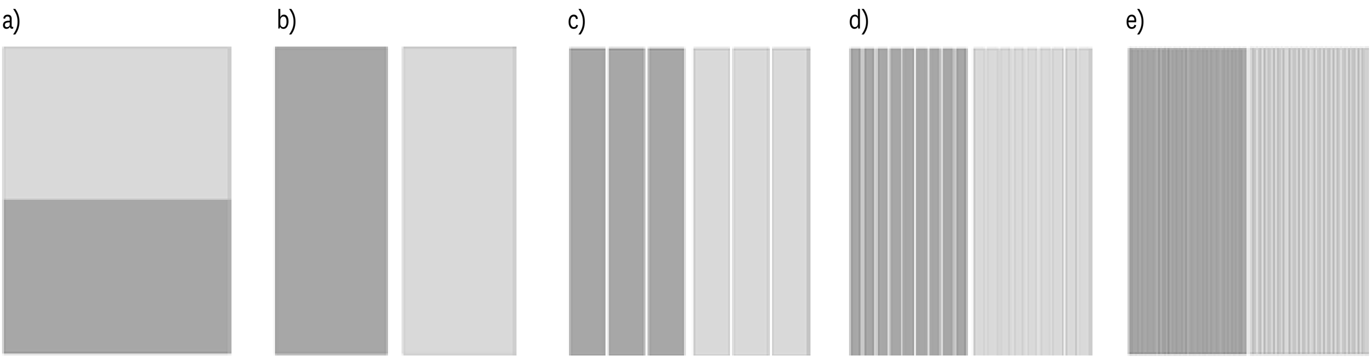


Figure 8. Tape winding configurations with various Cu tape thickness (shown here for the right-side core window only): a) 0.5 layer of 8.42 mm; b) 1 layer of 4.16 mm; c) 3 layers of 1.32 mm; d) 9 layers of 0.42 mm; e) 23 layers of 0.13 mm (for better clarity drawings are not to scale)

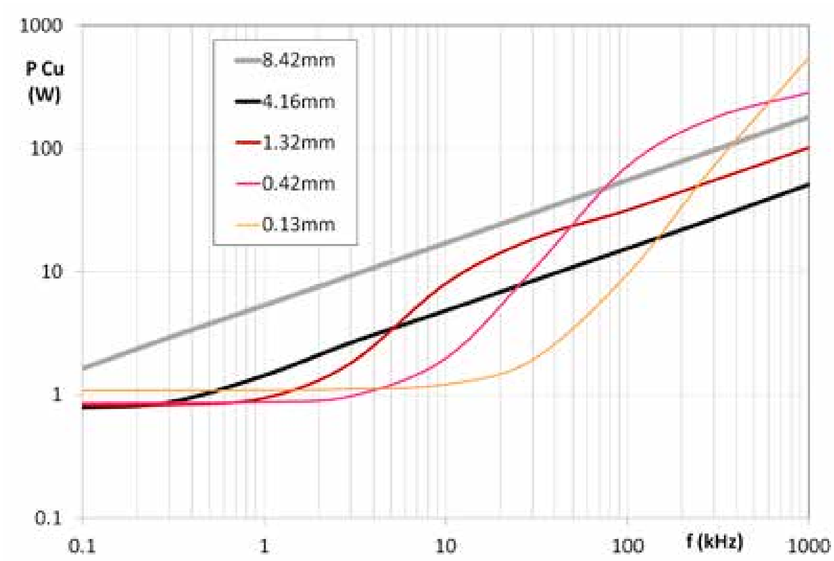


Figure 9. Total copper loss for configurations from Fig. 8

The proximity loss can be reduced drastically, so that the increase in power loss above the DC case is almost negligible as compared to the non-interleaved version

would be disastrous for such winding in practice), and for 23 layers it would be even higher – so the Dowell curves agree qualitatively with the calculated data.

However, for the thinnest winding the thickness of insulation already becomes a problem (23 layers instead of "ideal" 27), so further decrease in Cu tape thickness will be counterproductive at the highest frequencies (above 500 kHz), as the winding might not fit in the available core window and the current density would increase – and so would the copper loss. A larger core would have to be used instead, which would elevate the costs even further.

On the other hand, the 0.5 layer of the thickest winding has elevated losses even at lower frequencies. This is because the skin depth at 100 Hz is 6.52 mm and the conductor is 8.42 mm thick. The Q factor is thus greater than unity and growing with frequency, and it is evident from Dowell curves in Fig. 4 that a significant proximity effect would still take place even for 0.5 layer.

However, as mentioned above, one way of reducing the proximity loss is to reduce the number of effective layers. This can be achieved by interleaving primary and secondary windings so that the peak magnetic field is reduced.

The interleaving is discussed here only from purely magnetic point of view and reduction of proximity effect as such. There are several practical difficulties in making interleaved windings: higher interwinding capacitance, increased risk of insulation failure, added complexity of interconnections between the interleaved layers, etc. [15] However, these are outside of the scope of this paper and will not be discussed here.

Fig. 3 shows that the peak losses can exceed 600 % of the base value. For the same configuration (Fig. 8c), but interleaved primary (P) and secondary (S) so that the winding is made not as PPP-SSS, but as P-SS-PP-S (referred here as 1-2 interleave) the build-up of *H* is caused by just one layer (and not three as in Fig. 3). The peak *H* is thus reduced to 33 % of the previous value, and it reverses between some parts of the windings (compare Fig. 10 and Fig. 3). As a result the proximity loss is reduced drastically, so that the increase in power loss above the DC case is almost negligible as compared to the non-interleaved version.

Of course interleaving helps for any layer thickness and is proportional to the number of effective layers. For instance, as shown in Fig. 11 interleaving of the type PPP-SSSSSS-PPPPPP-SSS (i.e. interleave 3-6) of 0.42 mm tape produces comparable results to 0.13 mm tape for low and medium frequencies.

However, a 1-2 interleave for 0.42 mm thickness is much better for all frequencies than the non-interleaved 0.13 mm winding (Fig. 11) – the reduction is around *100-fold* at 1 MHz. Interestingly, 1-1 interleave (P-S-P-S...) is comparable to 1-2 interleave (not shown for clarity).

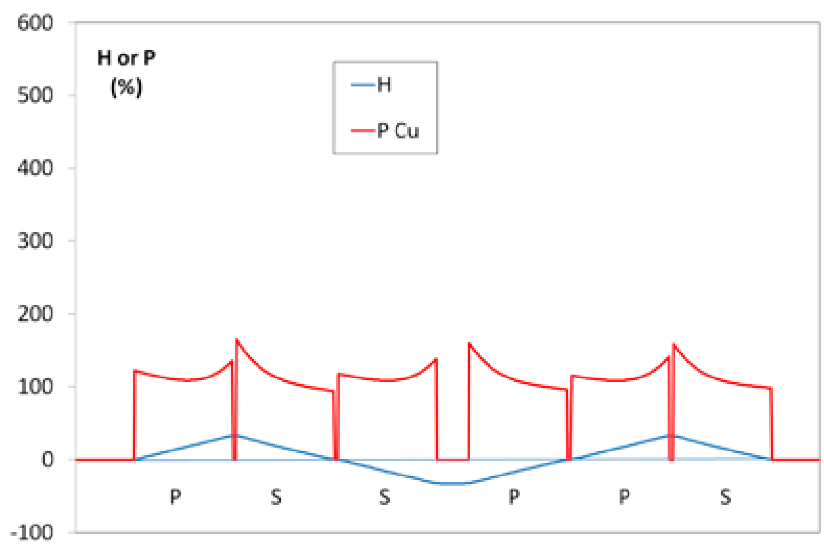


Figure 10. Copper loss for 1.32 mm winding with 1-2 interleave at 2 kHz (compare with Fig. 3)

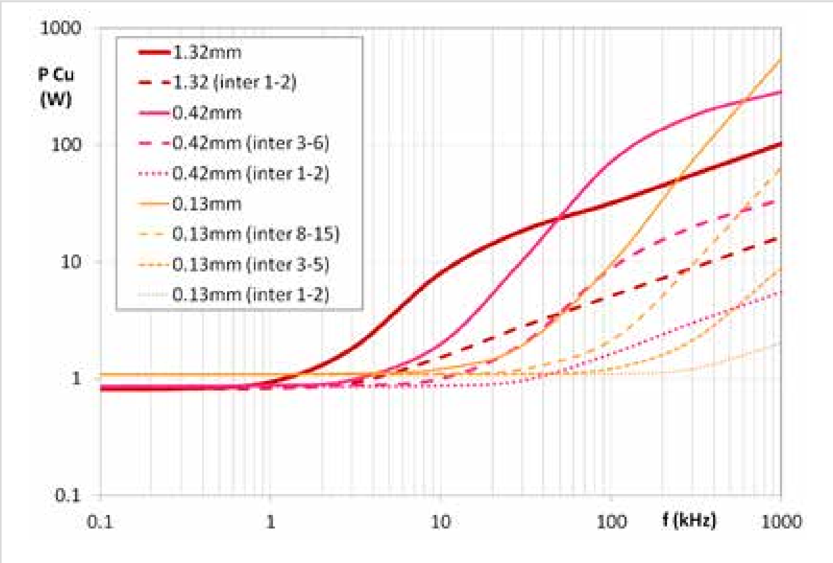


Figure 11. Total copper loss for configurations with interleaving: for 0.13 mm the interleaving 8-15 and 3-5 correspond to 9-18 and 3-6 respectively, and are a consequence of the number of turns reduced from 27 to 23 (see also Fig. 8)

It is also very interesting to see that the 0.13 mm tape would have to be heavily interleaved in order to significantly reduce proximity effect at the highest frequencies. Above 500 kHz the 0.42 mm tape with a 1-2 interleave performs better than the 0.13 mm tape with a 3-5 interleave. This is a direct consequence of the number of effective layers, which is 1 for the 1-2 interleave, and around 3 for the 3-5 interleave.

With such "conventional" interleaving the smallest number of effective layers cannot be reduced below 1, and other techniques must be used, as described below.

In Part 2 of this paper, other, more effective solutions for reducing the proximity effect will be presented and discussed.

Above 500 kHz the 0.42 mm tape with a 1-2 interleave performs better than the 0.13 mm tape with a 3-5 interleave

Bibliography

[1] S. Venkatramen, *Winding eddy current losses in switch mode power transformers due to rectangular wave currents*, Proceedings of Powercon 11, 1984, Sec. AI.1

[2] B. Carsten, *High frequency conductor losses in switchmode magnetics*, High Frequency Power Convener Conference, 1986, p. 155

[3] A. Ducluzaux, Cahier technique no. 83, *Extra losses caused in high current conductors by skin and proximity effects*, Technical Collection, 2012, Schneider Electric

[4] A. Payne, *Skin effect, proximity effect and the resistance of circular and rectangular conductors*, http://g3rbj.co.uk (accessed 2015-09-14)

[5] F. Wong, J. Lu, *Design consideration of high frequency planar transformer*, Intermag Magnetics Conference, 2002

[6] G. Ortiz, J. Biela, J.W. Kolar, *Optimized design of medium frequency transformers with high isolation requirements*, Proceedings of IECON Conference, 2010, p. 631, DOI: 10.1109/IECON.2010.5675240

[7] A. Mlot, M. Korkosz, P. Grodzki, M. Lukaniszyn, *Analysis of the proximity and skin effects on copper loss in a stator core*, Archives of Electrical Engineering, Vol. 63 (2), 2014, p. 211

[8] R. Ridley, *Proximity loss in magnetics windings*, Designers Series XIII, Switching Power Magazine, 2006

[9] C.R. Sullivan, *Optimal choice for number of strands in a Litz-wire transformer winding*, IEEE Transactions on Power Electronics, Vol. 14 (2), 1999, p. 283

[10] D.C. Meeker, *Improved continuum skin and proximity effect model for hexagonally packed wires*, Journal of Computational and Applied Mathematics, Vol. 236 (18), 2012, p. 4635

[11] C.W.T. McLyman, *Transformer and inductor design handbook*, 3rd edition, Marcel Dekker, 2004, ISBN 0-8247-5393-3

[12] B. Carsten, *Calculating the high frequency resistance of single and double layer toroidal windings*, Application notes, Carsten Associates, http://www.micrometals.com/appnotes_index.html (accessed 2015-10-22)

[13] *Ferrites and accessories*, ETD 49/26/16, Data sheet, TDK, 2013

[14] D.C. Meeker, FEMM, *Finite Element Magnetics Method*, http://www.femm.info (accessed 2015-08-30)

[15] S. Smith, *Magnetic Components: Design and Applications*, Springer Science & Business Media, 2013, ISBN 9789401540001

[16] J.G. Halser III, J.G Halser II, *Wide band audio transformer with multifilar winding*, Patent US 5500632, 1996

[17] P.D. Evans, W.J.B. Heffernan, *Transformers and coupled inductors with optimum interleaving of windings*, Patent US5543773, 1996

[18] *Second Decade Electromagnetics*, Orbic Institute, http://www.orbicinstitute.org/second_decade.html (accessed 2015-08-30)

[19] *Orbic Energy Conversions*, http://www.orbicenergyconversion.com (accessed 2015-08-30)

[20] A. Goldman, *Handbook of modern ferromagnetic materials*, Springer Science & Business Media, 1999, ISBN 9780412146619

Author

Stan Zurek completed a PhD in electrical engineering in 2005 and continued research in magnetic materials at Wolfson Centre for Magnetics, Cardiff University, UK. He joined Megger in 2008 and is Manager of Magnetic Development. He is the author and co-author of over 65 scientific publications related to magnetics and magnetic materials. Dr. Stan Zurek is a Senior Member of IEEE Magnetics Society.