IMPROVED GRAPHICAL METHOD FOR CALCULATION OF WINDING LOSSES IN TRANSFORMERS

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We report an improved method for calculating eddy current losses in round wires of magnetic components. The method extends the classical loss presentation by defining a loss coefficient, called eddy current factor k_c. *The effects of the fields of eddy currents in wires and the local fields (not only transverse fields) are considered. We provide a graphical approximation of* k_c *as a function of wire diameter, frequency, layer number, copper packing factors in the direction parallel and perpendicular to the layer. The graphs are obtained by analytical expressions compared with FEM simulations. A reference diameter, equivalent frequency and resistivity are used to unify the approach for different cases.*

1. INTRODUCTION

In most of the actual designs in power conversion the eddy current effects can not be neglected. Eddy current losses, including skin and proximity effects in transformers are discussed by Dowell [1] and many newer papers [2],[3], which are related to some extent to Dowell's interpretation and results. New methods have been leading to minimal core and copper losses by Hurley [4],[5], Petkov [6]. The proposed design procedures provide satisfactory results, but the loss calculation and optimization procedures are complex and require a long working in time. The traditional area-product method discussed by McLyman [7] does not consider enough eddy current losses in windings.

Here we present a fast and accurate approach for calculating eddy current losses in windings.

2. PROPOSED IMPROVED METHOD FOR CALCULATION WINDING LOSSES IN TRANSFORMERS

To provide an expression for Eddy current calculations, applicable for a wide frequency range, we derived a global loss factor k_c, which represents the ratio between the eddy current losses compared to the losses in the ohmic resistance of the winding of the magnetic component. Using the introduced loss factor, the eddy current losses are given by the following equation

$$P_{eddy} = \left(R_0\, I_{ac}^2\right) k_c \qquad (1)$$

where: the loss factor k_c depends on the operating frequency f, the wire diameter d and the distance between the conductors, presented by the parameter η and the distance between the layers presented by the parameter λ;

I_{ac} is the AC current component;

R_0 is the ohmic resistance of the winding.

To allow an easy use of k_c, we provide here a few graphs. To unify the use of the graphs, we introduce a reference diameter and an equivalent frequency.

Reference wire diameter

The choice of 0.5mm, used in the graphs as a *reference wire diameter*, is done in order to use a typical wire diameter for power electronics. The frequency, for which the penetration depth is equal to the reference diameter $d=\delta$, is 20kHz. The limit of the 'low frequency (LF) approximation' for the reference diameter d=0.5mm is 50kHz, thus LF can be applied below 50kHz for that wire diameter. These values are easy to remember. The diameters of wires in adjacent layers are taken equal and in a square fitting. This is the worst-case design, as a hexagonal fitting usually reduces the losses.

Equivalent frequency calculation

To use the provided graphs, Fig.1, and Fig.2, for any frequency, wire diameter and conductor resistivity, the equivalent frequency of the considered case should be first found:

$$f_{eq} \approx f\left(\frac{d_p}{0.5\,\mathrm{mm}}\right)^2\left(\frac{23\times10^{-9}}{\rho_c}\right) \tag{2}$$

where d_p is the practical wire diameter in [mm];

ρ_c is the conductor resistivity in [Ωm].

Using the proposed graphs, the coefficient k_c is found as:

$$k_c \approx m_E^2 K_{tf} \tag{3}$$

where the value of coefficient K_{tf} is found using Fig.1 and Fig.2;

It is not recommended to use partially filled layers in transformer designs. If anyhow partially filled layers are used, the wires should be equally spread. The effect of the partially filled layers is reduced at high values of m_E.

The graphs shown in Fig.1 and Fig.2 concern a design example with a typical wire diameter of 0.5mm and the normal frequency range for power electronics: 10kHz to 10MHz. For more than two layers (m_E>2), the result is almost independent of the number of layers. The usual values of η (copper fill factor in the direction of the layer) in transformers are between 0.7 (typical for thin wires and Litz wire) and 0.9 (typical for d>0.5mm). For other values of η, a linear interpolation between

Fig.10 and Fig.2 can be done. The additional error due to that interpolation is below 2%.

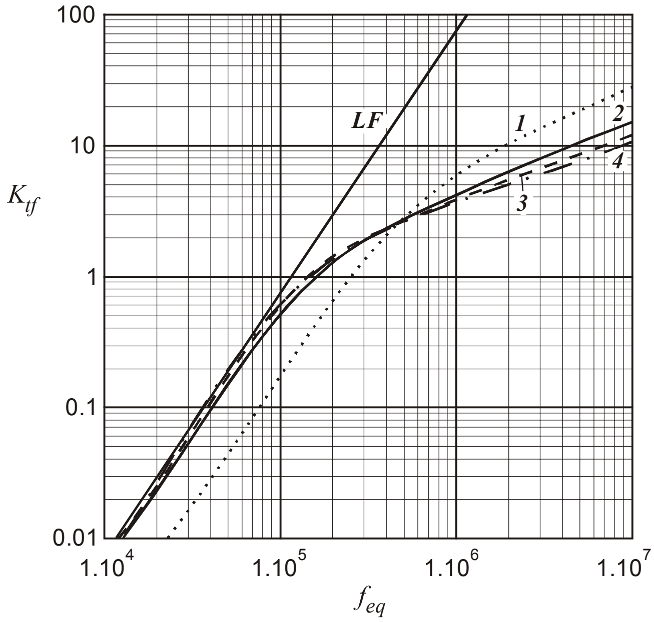

Fig.1 Typical transformer factor k_{tf} for d=0.5mm, η=0.9, ρ=23×10^{-9} and λ=0.5,
1) dotted line: half layer, m_E=0.5; 2) solid line: single layer, m_E=1; 3) dashed: two layers, m_E=2;
4) dash-dot: three or more layers, m_E>2. *LF* – low frequency approximation.

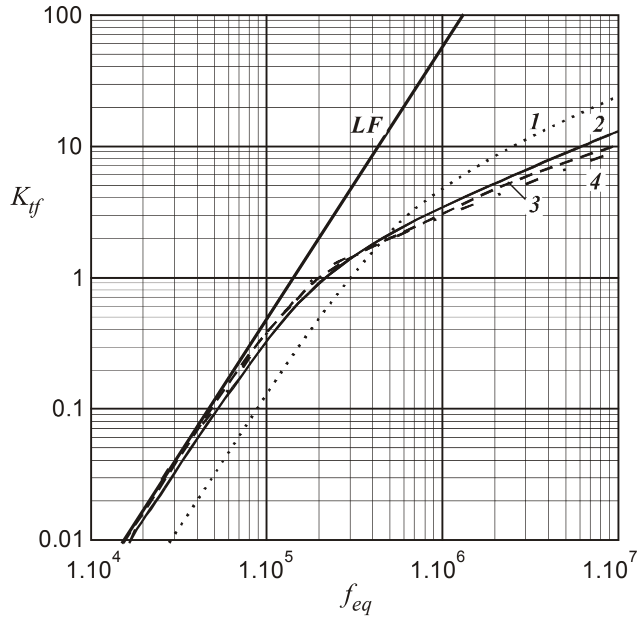

Fig.2 Typical transformer factor k_{tf} for d=0.5mm, η=0.7, ρ=23×10^{-9} and λ=0.5,
1) dotted line: half layer, m_E=0.5; 2) solid line: single layer, m_E=1; 3) dashed: two layers, m_E=2;
4) dash-dot: three or more layers, m_E>2. *LF* – low frequency approximation.

3. EXPERIMENTAL RESULTS

3.1 Design constructions and specifications

The presented method is applied in the design of welding transformers. We designed and built 2 transformers. The specifications are given in Table 1.

Table 1 The specifications of the designed and built transformers.

	Core	Primary Winding		Secondary Winding	
		N	d, [mm]	N	d, [mm]
Transf. 1	PM 74/59	10	1.25	2	0.3 - foil
Transf. 2	U 80/40/25	20	1.18	5	2

In Fig.3 and Fig.4 the arrangement of the windings are shown:

Transformer 1: Primary winding includes two windings in parallel (each winding is one layer) and 2 wires in parallel in each layer. Secondary winding (2 turns) is a foil winding, sandwiched between the two primaries.

Transformer 2: The transformer contains 2 coils connected in parallel. Primary winding includes two windings in parallel (each winding is one layer). Secondary winding (5 turns, 2 wires in parallel) is sandwiched between the two primaries.

The build transformers are pictured in Fig.5 and Fig.6.

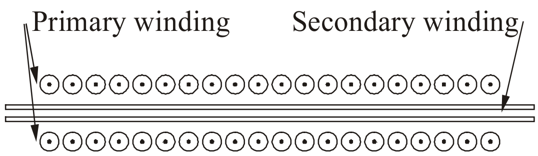

Fig.3 Transformer 1 - winding arrangement.

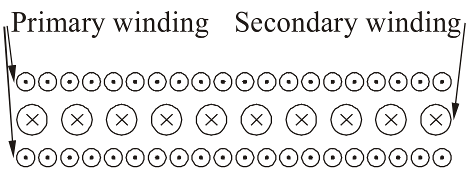

Fig.4 Transformer 2 winding arrangement.

Fig.5 Transformer 1, PM core, N87 ferrite.

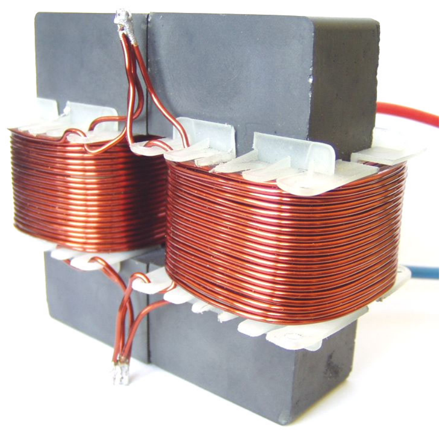

Fig.6 Transformer 2, UU core, N67 ferrite.

3.2 Calculations

Transformer 1

Primary Winding -A single layer winding, 2 wires in parallel, wire diameter of 1.25 mm, the frequency is 100kHz, the copper resistivity is $\rho = 21\times10^{-9}\ \Omega$ m.

We have $\eta = d\,N\,p/w = 1.25mm \times 10\times 2/30\,\mathrm{mm} = 0.833$. We have to keep the same diameter/penetration depth ratio, i.e. to find the equivalent frequency

$$f_{eq,1} = 100\,\mathrm{kHz}\left(\frac{1.25}{0.5}\right)^2\left(\frac{23}{21}\right) = 684.5\,\mathrm{kHz}$$

Then we find $k_{tf,0.9}$=3.3 from Fig.1 for η=0.9 and $k_{tf,0.7}$=2.5 form Fig. 2 for η=0.7. The actual value of η for the primary is η_1=0.833. Using the found two values and interpolating for η_1=0.833, we obtain $k_{tf}\approx$3.033. It is a single layer transformer, so $m_E = 1$ and we obtain k_c=k_{tf}=3.033.

Transformer 2

Primary Winding -A single layer winding, wire diameter of 1.18 mm, the frequency is 60kHz, the copper resistivity is $\rho = 21\times10^{-9}\ \Omega$ m.

We have $\eta = d\,N\,p/w = 1.18mm \times 20\times 1/26\,\mathrm{mm} = 0.908$. We have to keep the same diameter/penetration depth ratio, i.e. to find the equivalent frequency

$$f_{eq,1} = 60\,\mathrm{kHz}\left(\frac{1.18}{0.5}\right)^2\left(\frac{23}{21}\right) = 366\,\mathrm{kHz}$$

Then we read k_{tf1}=2.15 from Fig. 1 for η=0.9. The actual value η_2=0.908 is very close to η=0.9. It is a single layer transformer, so $m_E = 1$ and we obtain k_c=k_{tf}=2.15.

Secondary Winding -A half layer transformer design $m_E = 0.5$ (the considered single layer secondary is sandwiched between two primaries), 2 wires in parallel, wire diameter of 2 mm, the frequency is 60kHz, the copper resistivity is $\rho = 21\times10^{-9}$ Ω m.

We have $\eta = d\,N\,p/w = 2mm \times 5\times 2/26\,\mathrm{mm} = 0.769$. The equivalent frequency is

$$f_{eq,1} = 60\,\mathrm{kHz}\left(\frac{2}{0.5}\right)^2\left(\frac{23}{21}\right) = 1.051\,\mathrm{MHz}$$

We find $k_{tf,0.9}$=6 from Fig. 1 for η=0.9 and $k_{tf,0.7}$=4.7 form Fig. 2 for η=0.7. The actual value of η for the secondary is η_2=0.769. Using the found two values and interpolating for η_2=0.769, we obtain $k_{tf}\approx$5.149. It is a half layer transformer, so $m_E = 0.5$ and we obtain $k_c = 0.5^2 k_{tf} = 1.287$.

The calculated results are tabulated in Table 2.

Table 2 Calculated results for the experimental transformers.

		k_{tf}	k_c	P_{ohm} [W]	P_{eddy} [W]	P_{total} [W]
Tr.1	primary	3.033	3.033	0.504	1.529	2.033
	secondary	-	-	1.393	0.32	1.713
Tr.2	primary	2.15	2.15	1.455	3.129	4.584
	secondary	5.149	1.287	2.059	2.65	4.709

Remark: The losses of the secondary winding, Tr1, are calculated using the known expressions for rectangular conductors [1].

3.3 Measurements

We measured the winding losses of the built transformers (short circuit test, primary current 10A) and the results are shown in Table 3.

Table 3 Measured winding losses under short circuit test, 10 A.

	Measured results	Calculated results		
	P_{cu} [W]	$\mathrm{P_{cu}}$ [W]	P_{end} [W]	$\mathrm{P_{total}}$ [W]
Tr.1	4.7	3.746	0.68	4.426
Tr.2	9.37	9.293	0.11	9.403

Remark: In Table 3 P_{end} are the losses due to the ends of the secondary winding.

The calculated values for the copper losses 4.426 and 9.403W are very close to the measured values of 4.7 and 9.37W. The remaining difference can be attributed to mechanical tolerances, and the typical accuracy of about 3% of the proposed wide frequency method.

The conclusion of the short circuit test is that the calculations and the measurements give almost the same results.

4. CONCLUSION

The advantages of the proposed approach are the fast and straight calculation combined with high accuracy and wide application. The method is applicable for a wide variety of transformers with different frequencies, wire diameters and conductor fittings.

The proposed method is used in designing several welding transformers. The target power of the transformers is 2.5 kW. The experiments show good matching with the calculations.

REFERENCES

[1] Dowell P.L., "Effects of eddy currents in transformer windings", *Proc. Inst. Elect. Eng.*, vol. 113, No.8, August 1966, pp.1387-1394.

[2] Sullivan C. R., "Computationally Efficient Winding Loss Calculation with Multiple Windings, Arbitrary Waveforms, and Two-Dimensional or Three-Dimensional Field Geometry", *IEEE Transactions on Power Electronics,* vol. 16, No 4, January 2001, pp. 142-150.

[3] Murgatroyd P.N., 'Calculation of proximity losses in multistrand conductor bunches', *IEE Proceedings - B*, 1992, vol.139, No 1, pp. 21-28.

[4] Hurley W.G., E. Gath and J. G. Breslin, "Optimized Transformer Design: Inclusive of High-Frequency Effects", *IEEE Transactions on Power Electronics,* vol. 13, No 4, Yuly 1998, pp. 651-658.

[5] Hurley W.G., E. Gath and J. G. Breslin, "Optimising the AC Resistance of Multilayer Transformer Windings with Arbitrary Current Waveforms", *IEEE Transactions on Power Electronics,* vol. 15, No 2, March 2000, pp. 369-376.

[6] Petkov R., "Optimum Design of a High-Power, High-Frequency Transformer", *IEEE Transactions on Power Electronics,* vol. 11, No 1, January 1996, pp. 33-42.

[7] Colonel Wm. McLyman, *'Transformer and inductor design handbook'*, New York, Marcel Dekker, 1988.