Verification of Improved Methodology for Design of Magnetic Components

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Abstract – **In this paper, a verification procedure for a proposed improvement to an existing methodology for design of magnetic components is presented. Two magnetic components are calculated and built. An oscilloscope and calorimeter loss measurement set-upsare used for the verification. The results show good matching between the improved methodology and the real measurements.**

Keywords – **Verification, Magnetic components, Nanocrystalline, Ferrites.**

I. Introduction

For the verification of the proposed design methodology in [1], two transformers are calculated and built. For the magnetic core of the first transformer, ferrite is used, while for the other one a nanocrystalline soft magnetic material is chosen. The transformers are for a typical power electronic device - an inverter welding unit. The input parameters required for the calculation are shown in Table 1. All parameters are the same, except the higher maximum working temperature for the nanocrystalline material.

Table 1. Input parameters for both transformers

Input voltage	U_1	300V
No load output voltage	$U_{2,NL}$	60V
Full load output voltage	$U_{2,FL}$	26V
Maximum output current	$I_{2,FL}$	150A
Working frequency	f	80kHz
Air velocity	v	2,5m/s
Duty ratio of the voltage	D	10÷90%
Maximum ambient temperature	T_{amb}	30°C
Maximum working temperature for ferrite transformer	$T_{w,F}$	90°C
Maximum working temperature for nanocrystalline transformer	$T_{w,N}$	110°C

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II. Design Procedure for Ferrite Transformer with Forced Cooling

The transformer is calculated using the “Fast design approach” [2] and the proposed improvements in [1].

First step is to determine the largest dimension of the component - a_{ch}. The transformer is with only one secondary winding. Usually the transformers that are used in power electronics have efficiency higher than 95%. This allows to simplify the calculation by assuming that the primary power is equal to the secondary one:

$$S_{tot} = \sum_{all\ windings} U_{rms} I_{rms} = 300.13 + 26.150 = 7800W \quad (1)$$

$$a_{ch} = \left(\frac{S_{tot}}{A}\right)^{1/\gamma} = \left(\frac{7800}{20.10^6}\right)^{1/3,16} = 0,0724m = 7,2cm \quad (2)$$

The parameters $\gamma$and A are calculated by the equations presented in [2]. Using these equations the value for $\gamma=3,16$ is calculated and $A=20\cdot10^6$ is chosen. As a result core EE80/38/20 is selected with largest dimension $a_{ch}=80mm$ and ferrite type 3F3 by Ferroxcube ([5]).

The next step in the design is to calculate the heat dissipating capability., For T_{amb}=30°C and $T_{w,F}$=90°C, by using the proposed improvements in [1] the value for $P_h=49,4W$ is calculated. Most often, the heat generated by the core and the windings is equally divided [3], [4]:

$$P_{h,cu} = P_{h,fe} = \frac{P_h}{2} = \frac{49,4}{2} = 24,7W \quad (3)$$

Using Eq. (3) the specific core losses can be calculated:

$$P_{fe,sp,v} = \frac{P_{fe}}{V_e} = \frac{24,7}{72300.10^{-9}} = 342kW/m^3 \quad (4)$$

The effective volume of the core $V_e=72300mm^3$ is taken from the datasheet.

By using the parameters from Table 1 and core loss model, described in [6], the corresponding peak induction $B_p=0,17T$ is calculated. This is lower than the saturation peak induction $B_{sat}=0,33T$ at 100°C.

To calculate the number of turns, first the peak-to-peak magnetic flux linkage must be found:

$$\psi_{pp} = V_{in}\frac{T}{2} = \frac{V_{in}}{2f} = \frac{300}{2.80.10^3} = 1,875.10^{-3}Wb \quad (5)$$

Then the number of turns for the primary and the secondary are:

$$N_1 = \frac{\psi_p}{\Phi_p} = \frac{\psi_p}{A_e.B_p} = \frac{1,875.10^{-3}}{392.10^{-6}.0,17} = 14,0\,\text{turns} \quad (6)$$

$$N_2 = N_1.\frac{V_2}{V_1} = 14.\frac{60}{300} = 2,8\,\text{turns} \quad (7)$$

The number of turns for the secondary are rounded up to 3. This leads to slight increase in the secondary voltage $U_{2,NL}=64,3V$.

To distribute the allowed total copper losses $P_{h,cu}$ among the windings, the coefficient α_i is used.

$$\alpha_1 = \frac{N_1 .I_{rms,1}}{N_1 .I_{rms,1} + N_1 .I_{rms,1}}; \quad \alpha_1 = 0,29; \quad \alpha_2 = 0,71; \qquad (8)$$

$$P_{h,cu,1} = \alpha_1 .P_{h,cu}\,; \quad P_{h,cu,1} = 0,29.24,7 = 7,11W \qquad (9)$$

$$P_{h,cu,2} = \alpha_2 .P_{h,cu}\,; \quad P_{h,cu,2} = 0,71.24,7 = 17,59W \qquad (10)$$

The diameter of the copper wire used for the primary and the secondary windings is calculated by using Eq. (11) – Eq.(12).

$$d_1 \geq \frac{2}{\sqrt{\pi}}.I_1.\sqrt{\frac{\rho_c.l_{T1}.N_1}{P_{h,cu,1}}} = 1,17mm \qquad (11)$$

$$d_2 \geq \frac{2}{\sqrt{\pi}}.I_2.\sqrt{\frac{\rho_c.l_{T2}.N_2}{P_{h,cu,1}}} = 3,97mm \qquad (12)$$

Standard enamel copper wires type ПЕТ-2F with temperature grade+155°C are selected. The diameter for the primary windings is $d_{1,p}=1,30mm$ and for the secondary $d_{2,p}=5,00mm$.

With already selected wires, the copper ohmic and eddy current losses can be calculated:

$$P_{cu,ohm1} = \rho_c .l_{T1}.N_1\left(\frac{4}{\pi .d_{p,1}^2.p_1}\right)I_1^2 = 5,74W \qquad (13)$$

$$P_{cu,ohm2} = \rho_c .l_{T1}.N_1\left(\frac{4}{\pi .d_{p,1}^2.p_1}\right)I_1^2 = 11,07W \qquad (14)$$

One of the advantages of the “Fast design approach” is the calculating of the eddy current losses. The coefficient k_c is introduced. It shows how much the eddy current losses are larger than the ohmic losses:

$$P_{cu,eddy1} = k_{c,1}.P_{cu,ohm1}\,;\; P_{cu,eddy2} = k_{c,2}.P_{cu,ohm2} \qquad (15)$$

For calculating k_c Eq. (16) is used:

$$k_c = m_e^{\,2}.k_{tf} \qquad (16)$$

where

m_e is the equivalent number of layers;

k_{tf} – a coefficient used in the “Fast design approach”.

The coefficient k_{tf} is determined by the graphs presented in [2]. To do this, first the equivalent frequency f_{eq} must be found.

$$f_{eq,1} = f.\left(\frac{d_{p1}}{0,5\,\text{mm}}\right)^2.\left(\frac{23.10^{-9}}{\rho_c}\right) = 540,8kHz \qquad (17)$$

$$f_{eq,2} = f.\left(\frac{d_{p2}}{0,5\,\text{mm}}\right)^2.\left(\frac{23.10^{-9}}{\rho_c}\right) = 8,0MHz \qquad (18)$$

The two coefficients can be found from the graphs. As a result the following values are obtained:

$$k_{c,1} = 0,98; \quad k_{c,2} = 5,51 \qquad (19)$$

The results for the eddy current losses are:

$$P_{cu,eddy1} = k_{c,1}.P_{cu,ohm1} = 0,98.5,74 = 5,63W \qquad (20)$$

$$P_{cu,eddy2} = k_{c,2}.P_{cu,ohm2} = 5,51.11,07 = 61,00W \qquad (21)$$

Summing all results together give the total copper losses:

$$\begin{aligned} P_{tot} &= P_{cu,ohm1} + P_{cu,eddy1} + P_{cu,ohm2} + P_{cu,eddy2} = \\ &= 5,74 + 5,63 + 11,07 + 61,00 = 83,44W \end{aligned} \qquad (22)$$

This is more than three times the maximum allowed copper losses 24,7W, calculated by Eq.(3).

Transformer with such losses will overheat and consequently be damaged, an optimization must be carried out in order to decrease the copper losses. Some of the possibilities are [7]:

- Increasing the wire diameter;
- Connecting several wires in parallel;
- Using Litz wire;
- Using copper foil or wires with rectangular cross sections.
- Use larger core

Increasing the diameter of the wire is used for partially filled layers. The diameter is increased until the layer is filled completely. For low number of turns (<5), this is not practical. In this case, it is best to use the second technique – wires in parallel. Usually 2÷4 wires are put in parallel. This allows the same cross section area to be obtained by using smaller wires, and hence smaller copper losses. The results from such optimizations are presented in Table2.

Case 0 is the default – with no optimizations. In case 1, the first two presented optimizations are used (increasing the wire diameter and connecting several wires in parallel). For Case 2, Litz wire is used. This results in lowering the copper losses up to the maximum value. However, because of the additional insulation, the cooling properties are worsen. Together with the fact that this design is “just on the limit”, gives enough reasons to continue with the optimization.

Two sets of EE80/38/20 are used for cases 3 and 4. The MLT and the copper wire resistance are calculated more precisely, by taking into account difference in the MLT for the primary and the secondary windings and the resistance temperature dependence. This leads to lowering the losses bellow the maximum value safety margin of about 15% in case 4. With this step the optimization procedure for the ferrite transformer is over.

Similar design and optimization are carried out for the nanocrystalline core transformer. The results are presented in case 5. Two sets of the cut core F3CC0010 are used with nanocrystalline material FT-3M. The maximum working temperature for this material is 110°C. The temperature dependence of the wire resistance and different MLT are also taken into account.

Cases 4 and 5 use wires with square cross section. This results in easier winding and better utilizing of the window area. To use such wires, a conversion from rectangular to equivalent square wires is used. Different techniques exist for such conversions -[2] and [3].

The results from such optimizations are presented in Table2.

TABLE2. DESIGNED TRANSFORMERS

Parameter	Winding	Symbol	Unit	Ferrite core					Nanocrsytalline
				No optimizations	Wires in parallel	Litz wire	2 x E80	2 x E80 and square wires	
				Case0	**Case1**	**Case2**	**Case3**	**Case4**	**Case5**
Number of turns	primary	N_{1p}	-	14			10		11
	secondary	N_{2p}	-	3			2		2
Magnetic induction		B	T	0,17			0,13		0,26
Diameter of the wire (with insulation)	primary	$d_{1,p}$	mm	1,30 (1,41)	3,55 (3,68)	0,20 (2,65)	1,00 (1,09)	2,0 (2,11)	1,50 (1,63)
	secondary	$d_{2,p}$	mm	5,00 (5,14)	4,00 (4,13)	0,20 (4,97)	4,5 (4,64)	3x7,5 (3,8x7,9)	0,7x32 (1,3x32,6)
Wires in parallel	primary	p_1	-	1	1	100	1	1	2
	secondary	p_2	-	1	4	23350	1	1	1
Equivalent number of layers	primary	m_{E1}	-	1	1	10	1	1	1
	secondary	m_{E2}	-	1	1	26,5	1	1	2
Equivalent number of turns in a layer	primary	n_{E1}	-	14	14	140	10	10	22
	secondary	n_{E2}	-	3	12	79,4	2	4,43	46
Copper fill factor in the direction of the layer	primary	η_1	-	0,34	0,93	0,53	0,19	0,38	0,89
	secondary	η_2	-	0,28	0,93	0,3	0,17	0,28	0,98
Copper fill factor in the direction perpendicular to the layer	primary	λ_1	-	0,07	0,18	0,1	0,05	0,10	0,12
	secondary	λ_2	-	0,26	0,21	0,27	0,23	0,17	0,12
Equivalent frequency	primary	$f_{eq,1}$	kHz	540	3920	12,8	320	1280	719
	secondary	$f_{eq,2}$	kHz	8000	5120	12,8	6480	3670	199
Eddy current coefficient	primary	k_{c1}	-	1,0	9,4	0,4	0,4	2,3	3,4
	secondary	k_{c2}	-	5,5	10,6	1,0	4,2	3,7	1,4
Ohmic resistance	primary	$R_{dc,1}$	mΩ	34,0	4,7	15,1	23,2	5,9	10,0
	secondary	$R_{dc,2}$	mΩ	0,5	0,2	0,5	0,3	0,2	0,3
Ohmic losses	primary	$P_{cu,ohm,1}$	W	5,7	0,8	2,6	3,9	1,0	1,5
	secondary	$P_{cu,ohm,2}$	W	11,1	4,3	10,7	5,8	2,3	6,5
Eddy current losses	primary	$P_{cu,eddy,1}$	W	5,6	7,4	1,1	1,4	4,7	5,2
	secondary	$P_{cu,eddy,2}$	W	61,0	45,9	10,5	24,3	17,4	9,0
Copper losses	both	P_{cu}	W	83,8	58,5	24,8	35,5	25,4	22,2
Maximum allowed copper losses	both	P_h	W	24,7			29,8		25,9

The last step in the design of the transformer is to check, is there enough space to place physically all the windings in the transformer window:

$$k_{cu} = \frac{\sum_{i=1}^{n} p_i . N_i . \frac{\pi . d_{i,p}^2}{4}}{W_a} \qquad (23)$$

where:

$p$is number of wires in parallel (or the litz wire strands);

k_{cu} – filling factor;

W_a – window area.

For round wires $k_{cu}=0,5\div0,8$ and for litz wires it should be $k_{cu}=0,3\div0,5$.

For all of the designed transformers, this coefficient fits to the reference values.

For conducting real experiments, the transformers from case 4 and 5 are built and can be seen in Fig.1.

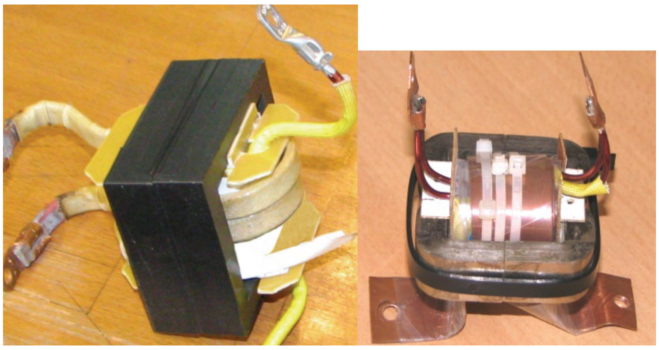

Fig.1Transformers from Case 4(left) and Case 5(right)

Short circuit and no load tests are carried out with the transformers. The system described in [8] is used for the loss measurements. The results are presented in TABLE3:

TABLE3. CALCULATED AND MEASURED LOSSES

	Winding losses				Core losses			
	calculated	measured	difference		calculated	measured	difference	
	W	W	W	%	W	W	W	%
Case 4	25,38	24,71	-0,67	-2,7	29,80	28,64	-1,16	-4,1
Case 5	22,19	21,04	-1,15	-5,5	25,91	25,8	-0,11	-0,4

One can see, that the measured values are about 3% less than calculated. This can compensate for some small manufacturing tolerances and rounding of numbers. This also means that with the proposed improvements it is possible to design magnetic componentsfor the typical conditions met in power electronics.

A comparison between the two transformers is done in TABLE4.

TABLE4. COMPARISON BETWEEN THE TWO TRANSFORMERS

		Case 4	Case 5	Difference	Improvement
Mass	**kg**	1,176	0,572	0,604	51,4%
Area	**mm2**	6080	4608	1472	24,2%
Volume	**mm3**	340480	211968	128512	37,7%
Price	**€**	2x(6,00÷7,00)	2x(7÷30)		0 ÷ -400%

It can be seen from the results in TABLE4, that using nanocrystalline core leads to significant improvement in the mass, area and the volume of the core. The higher price remains a disadvantage. However there are now some Chinese manufacturers (like Foshan Huaxin Microlite Metal Co Ltd 0) of nanocrystalline soft magnetic materials that offers several times cheaper price compared to the leading manufacturers (Hitachi metals - [10], Vacuumschmelze - [11]). This will only lead to decrease the prices of nanocrystalline cores in global aspect.

III. CONCLUSION

This paper gives a verification of the proposed “Improved methodology for design of magnetic components”. Several transformers are calculated and two of them are actually built. Some experiments are conducted and the results show matching with the measured data better than 5%.

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