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Citation for published version (APA):

Smeets, J. P. C., Encica, L., & Lomonova, E. (2010). Comparison of winding topologies in a pot core rotating transformer. In *Proceedings of the 2010 12th International Conference on Optimization of Electrical and Electronic Equipment (OPTIM)*, 20-22 May 2010, Basov (pp. 103-110). Institute of Electrical and Electronics Engineers. <https://doi.org/10.1109/OPTIM.2010.5510563>

DOI:

[10.1109/OPTIM.2010.5510563](https://doi.org/10.1109/OPTIM.2010.5510563)

Document status and date:

Published: 01/01/2010

Document Version:

Publisher's PDF, also known as Version of Record (includes final page, issue and volume numbers)

Please check the document version of this publication:

- A submitted manuscript is the version of the article upon submission and before peer-review. There can be important differences between the submitted version and the official published version of record. People interested in the research are advised to contact the author for the final version of the publication, or visit the DOI to the publisher's website.
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Comparison of Winding Topologies in a Pot Core Rotating Transformer

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Abstract—This paper discusses the comparison of two winding topologies in a contactless energy transfer system from the stationary to the rotating part of a device. A rotating transformer, based on a pot core geometry, is proposed as a replacement for wires and slip rings. An electromagnetic and a thermal model of the rotating transformer are derived. The models are combined and used in a multi-objective optimization. A Pareto front, in terms of minimal volume and power losses, is derived to compare both winding topologies. Finally, the optimization algorithm is used to design a prototype transformer for each winding topology, which are manufactured using a commercially available pot core.

I. INTRODUCTION

In many modern technetronic systems, the transfer of power to rotating parts plays an important role, for example, in robotic applications [1] and in industrial electronics whith rotating electronics. Nowadays, wires and slip rings are used to transfer power to the rotating part. Disadvantages of wires are a limited rotation angle and an increased stiffness. To overcome the problem of limited rotation, slip rings are used. Despite the significant amount of research and development of reliable and durable slip rings, the lifetime is limited by contact wear as well as vibration and frequent maintenance is required [2].

A solution to overcome the disadvantages of wires and slip rings is a contactless energy transfer (CET) system that uses a rotating transformer. That is a transformer with an airgap between the primary and secondary side, where one side can rotate with respect to the other. An extra advantage could be the freedom in winding ratio to transform the primary voltage level to the requirements of the load.

The axial and pot core transformer geometries, both shown in Fig.1, have the possibility to rotate one side with respect to the other side and, therefore, can be used as a rotating transformer. Both geometries are compared in terms of optimal volume and efficiency in [3]. The pot core transformer geometry gives better performance indices compared to the axial geometry. Inside the pot core rotating transformer two winding topologies can be used. The adjacent winding topology, shown in Fig. 2a, where each winding is placed in an own core half and the coaxial winding topology, shown in Fig. 2b, where the secondary winding is place around the primary winding [4].

In this paper both winding topologies are compared in terms of minimal power losses and volume using an optimization algorithm. For this purpose, transformer models are derived

Fig. 1. Axial (a) and pot core (b) rotating transformers.

Fig. 2. Winding topologies for the pot core rotating transformer, (a) adjacent and (b) coaxial.

based on the electromagnetic and thermal behavior and combined in a optimization procedure. A multi-objective optimization is conducted to define the optimal winding topology [5]. Finally, for each winding topology a rotating transformer with minimal power losses is designed and manufactured using a commercially available pot core. The prototypes are used to verify the derived transformer models.

II. ROTATING POT CORE TRANSFORMER

A detailed drawing of the geometry of the rotating pot core transformer is shown in Fig. 3, the corresponding parameters are listed in Table I. Based on Faraday's law of induction and Ampere's circuital law, an initial design expression for the power transfer in the transformer can be given by

$$P = \pi J S k_f f B_{peak} A_e, \quad (1)$$

where J is the current density, S is the winding area, f is the frequency of the applied voltage, k_f is the filling factor of the winding, B_{peak} is the peak flux density and A_e is the cross section of the inner core. Equation (1) shows that the power transfer is depending on the geometric parameters, frequency and flux density.

Fig. 3. Geometry of the pot core rotating transformer, (a) top view and (b) cross section.

TABLE I
GEOMETRICAL PARAMETERS OF FIG. 2 AND FIG. 3

Parameter	Description
r_1, r_2, r_3, r_4	Radius of the different core parts
r_{cin}	Length of the inner core part
r_{cout}	Length of the outer core part
h_{out}	Outer height of a core half
h_{in}	Height of the winding area S
h_c	Thickness of the horizontal core part
l_{ag}	Length of the airgap
A_e	Effective core area
S	Winding surface
N_p	Number of turns on primary side
N_s	Number of turns on secondary side

The rotating transformer is part of a dc-dc power conversion system, which consists of a half bridge converter connected to the primary side of the transformer to create a high frequency voltage and a diode rectifier connected to the secondary side of the transformer to rectify the voltage back to a dc-voltage.

III. ANALYTICAL MODELS

In this section the electromagnetic and thermal model of the rotating transformer are described. The models will be combined for further analysis.

A. Magnetic model

A magnetic model is derived to calculate the inductances of the transformer. The magnetizing inductance, L_m , is calculated using a reluctance model. The model is shown in Fig. 4, where $\mathcal{R}$ presents the reluctance of the magnetic path and the subscripts c , ag and lk indicate the flux path in the core, airgap and leakage, respectively. The magnetizing inductance is calculated by

$$L_m = \frac{N_p^2}{2(\mathcal{R}_{c_a} + \mathcal{R}_{c_b} + \mathcal{R}_{c_c}) + \mathcal{R}_{ag_a} + \mathcal{R}_{ag_b}}. \quad (2)$$

The leakage flux lines in the rotating transformer do not have an a priori known path, therefore, it is inaccurate to model them with a reluctance network as well. The leakage inductance, L_{lk} , is calculated by the energy of the magnetic field in the winding volume

$$\frac{1}{2} L_{lk} I^2 = \frac{1}{2} \int_v \mathbf{B} \cdot \mathbf{H} dv, \quad (3)$$

Fig. 4. Reluctance model of the rotating transformer with adjacent winding topology.

Fig. 5. Electric equivalent circuit of the rotating transformer.

which is equal to the magnetic energy of the leakage inductance [6]. An expression for the magnetic field strength is found by the magnetic circuit law. In the case of the adjacent winding topology, the magnetic field strength is expressed for the primary winding as function of the axial length

$$H(z) = \frac{N_p i_p}{(r_3 - r_2) h_{wp}} z, \quad (4)$$

where h_{wp} is the height of the primary winding. A similar expression can be derived along the secondary winding. In the airgap a uniform mmf is assumed, defining the magnetic field strength by

$$H = \frac{N_p i_p}{l_{ag}}. \quad (5)$$

Combining (3)-(5), results in an expression for the total leakage inductance of the transformer seen from the primary side

$$L_{lk} = \mu_0 N_p^2 \frac{2\pi}{\ln(r_3/r_2)} \left(\frac{h_{wp} + h_{ws}}{3} + l_{ag} \right). \quad (6)$$

B. Electric model

An electric equivalent circuit of the rotating transformer is derived to calculate the power losses in the transformer. The model is shown in Fig. 5. In the circuit the rotating transformer is represented by the magnetizing and leakage inductances and a lossless transformer with winding ratio $a = N_p/N_s$ and coupling factor, k . Furthermore, winding resistance and resonance capacitances are inserted. The circuit is connected to a square wave input voltage source and an equivalent load resistance.

The winding resistance, R_p, R_s , consists of a dc and ac-resistance. An expression for the winding resistance in case of non-sinusoidal waveforms is derived in [7], based on Dowell's formula for AC-resistances. The effective winding resistance is calculated by

$$R_{eff} = R_{dc} + \frac{\Psi}{3} \Delta^4 R_{dc} \left(\frac{I'_{rms}}{2\pi f \cdot I_{rms}} \right)^2, \quad (7)$$

where Ψ is a correction factor for the number of layers, Δ is the winding thickness of a layer when it is converted to an equivalent foil-type winding divided by the skin depth, I'_{rms} is the rms-value of the derivative of the current waveform and ω is the angular frequency.

To improve the power transfer of the transformer, resonant techniques are used [8]. A resonant capacitor is placed in series on both sides of the transformer:

- On the primary side, to create a zero crossing resonance voltage and thereby allowing the use of a half bridge inverter.
- On the secondary, to overcome the voltage drop across the leakage inductance and thereby improving the power transfer.

Furthermore, by applying series resonance on the secondary side, the primary side is made insensitive for coupling changes, for example caused by vibration during rotation. This can be illustrated by calculating the value of the primary resonance capacitance for a series and parallel resonance on the secondary side, respectively

$$C_{p \text{ series}} = \frac{1}{\omega_{res}^2 (L_p)}, \quad (8)$$

$$C_{p \text{ parallel}} = \frac{1}{\omega_{res}^2 (L_{lk_p} - M^2/L_{lk_s})}, \quad (9)$$

where $L_p (= L_m/k)$ and $L_s (= L_m/a^2 k)$ is the self inductance of the primary side and secondary side of the rotating transformer, respectively, and $M (= k\sqrt{L_p L_s})$ is the mutual inductance of the rotating transformer. The results of equation (8) and (9) is shown in Fig. 6 for an increasing magnetic coupling. A constant primary resonance capacitance can be obtained by applying series resonance on the secondary side,

The resonance technique creates a band pass filter around the resonance frequency to filter-out unwanted harmonics and thereby decreasing the AC-losses in the windings. The quality of this filter depends on the resonance frequency, leakage inductance and load resistance of the transformer and is defined by

$$Q = \frac{2\pi f_{res} L_{lk}}{R_{load}}. \quad (10)$$

Using resonance capacitors, the primary voltage at resonance, V_p , can be calculated with [9]

$$V_p = \left(R_p + \frac{\omega_{res}^2 M^2}{R_s + R_{load}} \right) I_p, \quad (11)$$

$$(12)$$

Fig. 6. Influence of the magnetic coupling on the primary resonance capacitance.

and the current density in the winding is calculated, based on (1)

$$J_n = \frac{I_n N_n}{S k_f}. \quad (13)$$

The conduction and core losses are the main power losses in the rotating transformer. The conduction losses, P_{cond} , are calculated by

$$P_{cond} = I_{p_{rms}}^2 R_p + I_{s_{rms}}^2 R_s, \quad (14)$$

where $I_{p_{rms}}$ is the primary rms-current, which consists of the magnetizing current and the reflected load current. The core losses, P_{core} , are calculated by the Steinmetz equation

$$P_{core} = C_m C(T) f_{res}^x B^y V_{core}, \quad (15)$$

where C_m , x and y are material specified constants (for example $C_m=7$, $x=1.4$ and $y=2.5$ for the 3C81 core material). $C(T)$ is a temperature depending constant and is equal to 1 if the core temperature is $\pm 20^\circ$ around the ideal working temperature, which is $60^\circ C$ for the 3C81 core material. For a constant power transfer, the flux density can be calculated as a function of frequency as indicated in (1). By varying the frequency an optimal working point with minimal losses can be found (shown in Fig. 7).

C. Thermal model

It is important to estimate the core temperature since the core and conduction losses cause a temperature rise in the core material, which has an optimal working temperature with minimal power losses. A thermal equivalent circuit of the core, shown in Fig. 8, is made using a finite-difference modeling technique, where the thermal resistance concept is used for deriving the heat transfer between the nodes [10].

The thermal model is derived by dividing the upper half of the geometry into six regions, where regions I till V represent the core and region VI represents the transformer winding. Five nodes are defined for each region and the heat transfer between the nodes is modeled by a thermal resistance. Conduction resistances are used to model heat transfer inside

Fig. 7. Power losses as a function of the frequency.

Fig. 8. Thermal equivalent circuit of the transformer.

the regions and convection resistances are used to model the heat transfer between the border of the regions and the air. The conductive thermal resistance in z - and r - direction are calculated by

$$R_{th_z} = \frac{\Delta z}{\pi(r_o^2 - r_i^2)k}, \quad (16)$$

$$R_{th_r} = \frac{\ln(r_o/r_i)}{2\pi k \Delta z}, \quad (17)$$

where k is the thermal conductivity, equal to 4.25 and 394 $\text{Wm}^{-1}\text{K}^{-1}$ for the ferrite core and copper windings, respectively. The convective thermal heat resistance is calculated by

$$R_h = \frac{1}{hA}, \quad (18)$$

where h is the heat transfer coefficient obtained from the Nusselt-number, which is equal to 12.7 and 8.5 $\text{Wm}^{-2}\text{K}^{-1}$ for the axial and radial boundaries of the pot core.

No heat transfer is assumed at left and lower boundary of the model, assuming a worst-case thermal situation. The power losses in each region are presented by a heat source and inserted in the middle node of region. By calculating the heat transfer between each node, the temperature in the middle

TABLE II
LIMITS OF THE OPTIMIZATION VARIABLES.

min	var.	max
r_1	r_2	r_{max}
r_2	r_3	r_{max}
0	h_{in}	h_{max}
0.5 mm	l_{ag}	2.0 mm
1	N_p	N_{max}
1	N_s	N_{max}
1 kHz	f_{res}	200 kHz

of each region is obtained. An ambient temperature of 20°C is assumed.

IV. OPTIMIZATION ALGORITHM

The analytical models are implemented in MATLAB and used in an optimization procedure to find the optimal transformer design in terms of both minimal volume and power losses for a constant power transfer of 1 kW and a secondary voltage of 50 V. A sequential quadratic programming algorithm is used to find the minimal Pareto front of the two objective functions [11]. Therefore, the weighted sum method for multi-objective problems is used

$$\begin{cases} \min & F(x) = \sum_{m=1}^{N_{obj}} w_m f_m(x) & m = 1, \dots, N_{obj} \\ & g_j(x) \leq 0 & j = 1, \dots, J_{neq} \\ & h_k(x) = 0 & k = 1, \dots, K_{eq} \\ & x_i^l \leq x_i \leq x_i^u & i = 1, \dots, N_{var} \end{cases} \quad (19)$$

The weights $w_m \in [0, \dots, 1]$ are selected such that the sum of the weighting coefficients is always $\sum_{m=1}^{N_{obj}} w_m = 1$. This function finds the minimum of the objective functions subjected to the inequality, g_j , and equality constraints, h_k , within the lower and upper boundaries of the variables x_i . In the next sections the variables, constraints and objective functions are explained in more detail.

A. Variables

As shown in (1), the core dimensions, length of the airgap, number of turns and frequency are parameters which have influence on the design of the rotating transformer. The lower and upper value of those variables is specified in Table II. Where N_{max} is the maximum number of turns, defined by

$$N_{max} = \frac{Sk_f}{A_{wire}}. \quad (20)$$

Parameters r_{max} and h_{max} limit the maximum core dimensions and thereby, reduce the calculation time. Furthermore, the ratio between the inner and outer radial length and the thickness of the horizontal core part are fixed, based on existing pot cores dimensions [12]

$$r_{c_{out}} = 0.55r_{c_{in}}, \quad (21)$$

$$h_c = 0.65r_{c_{in}}, \quad (22)$$

$$r_1 = 2.7 \text{ mm}. \quad (23)$$

With constraint (23) the inner radius of the core is set to obtain a minimal hole in the middle of the transformer to mount the core. Other geometric parameters such as core

Fig. 9. Pareto front before (a) and after (b) normalization.

material specifications and wire parameters are given as input parameters for the optimization function.

B. Constraints

For the electromagnetic and thermal properties of the rotating transformer, a number of constraints is introduced. Firstly, from a magnetic point of view, saturation in the core should be avoided and the coupling should be larger than 60%, i.e.

$$B_{core} \leq B_{sat}, \quad (24)$$

$$k \geq 0.6. \quad (25)$$

Secondly, from an electrical point of view, the input voltage is limited by the maximal output voltage of the dc-voltage source, the maximal current density is limited by the wire properties and the quality factor of the resonance circuit should be larger than 1 to filter-out higher harmonics

$$V_p \leq V_{dc_{max}}, \quad (26)$$

$$J_n \leq J_{n_{max}}, \quad (27)$$

$$Q \geq 1. \quad (28)$$

Finally, from a thermal point of view, the core temperature should stay below 100°C , because up to this temperature the core losses are almost constant,

$$T_{core} \leq 100^\circ\text{C}. \quad (29)$$

C. Objective functions

The design optimization is conducted in terms of minimal volume and power losses, using the following objective functions

$$f_1(x) = \pi r_4^2 \cdot 2h_{out} \quad (30)$$

$$f_2(x) = P_{cond} + P_{core}. \quad (31)$$

Both objectives are normalized by defining the two limits of the Pareto front, resulting in parameter sets x^{1*} and x^{2*} for the individual minimization of $f_1(x)$ and $f_2(x)$, respectively [13] (see Fig. 9). The normalized objective functions are

$$f_{1n}(x) = \frac{f_1(x) - f_1(x^{1*})}{f_1(x^{2*}) - f_1(x^{1*})}, \quad (32)$$

$$f_{2n}(x) = \frac{f_2(x) - f_2(x^{2*})}{f_2(x^{1*}) - f_2(x^{2*})}. \quad (33)$$

The normalization allows an equal comparison of both winding topologies.

Fig. 10. Pareto front for optimal design in terms of volume and power losses.

V. DISCUSSION OF THE OPTIMIZATION RESULT

By applying different combinations of weighing factors, a minimal Pareto front is found for both topologies, shown in Fig. 10. The Pareto front shows that the adjacent winding topology obtains lower power losses for the same core volume compared the coaxial winding topology. In the Pareto front two asymptotes can be obtained. A vertical asymptote for the minimal required core volume, limited by the maximal allowable core temperature, since the losses are increasing dramatically for a small core with a high frequency and high magnetic flux density. And a horizontal asymptote for the minimal power losses, which is based on an optimum in magnetic flux density, frequency and volume, comparable as shown in Fig. 7.

Detailed transformer parameters are given for two realistic extreme optimization cases for the coaxial and adjacent winding topology in Table III and IV, respectively. The objective functions are defined as $90\%f_{1n}(x) + 10\%f_{2n}(x)$ for case 1 and $10\%f_{1n}(x) + 90\%f_{2n}(x)$ for case 2. In other words, the volume is minimized in case 1 and the power losses are minimized in case 2. The letter A and C before the case numbers indicate the adjacent and coaxial winding topology, respectively. The upper half of the cross section of two coaxial cases is shown in Fig. 11. The core dimensions of the adjacent winding topology are almost identical to the coaxial winding topology and therefore not shown.

Comparing the four cases, the following observations are made:

- In case C1, a small core radius with a relative large winding area is obtained and, in case C2, a large core radius and a smaller winding area can be found. The total volume of the adjacent winding topology is slightly lower, because the winding area is used more efficient.
- In all four cases the airgap is minimized to the minimal realizable mechanical airgap.
- The magnetizing inductances of both winding topologies are comparable for the different cases.
- The leakage inductance of the coaxial winding topology is approximately 15 times lower compared to the adjacent

Fig. 11. Optimized core dimensions for the coaxial winding topology, (a) minimal volume and (b) minimal losses.

TABLE III
TRANSFORMER PARAMETERS FOR TWO CASES WITH COAXIAL WINDING TOPOLOGY.

Parameter	Case: C1	Case: C2	Unit
r_{cin}	5.6	16.2	mm
r_{cout}	3.1	10.5	mm
r_4	23.9	42.6	mm
h_{out}	28.6	31.0	mm
l_{ag}	0.5	0.5	mm
S	311	195	mm ²
A_e	194	1456	mm ²
V	102	354	cm ³
N_p	96	62	turns
N_s	10	6	turns
B_{core}	294	116	mT
f_{res}	20.4	10.7	kHz
L_{mp}	3.38	8.63	mH
L_{Lkp}	0.05	0.05	mH
L_{Lks}	0.55	0.48	μ H
k	0.98	0.99	
P_{loss}	10.7	4.1	W
T_{core}	48.6	30.7	$^{\circ}$ C

winding topology. This is because both windings of the coaxial winding topology share an identical magnetic flux path, which is not the case in the adjacent winding topology.

- Less leakage result in a higher coupling coefficient, which is obtained for the coaxial winding topology.
- The winding ratio is the same in the four cases, because of the fixed secondary voltage and the maximized primary voltage. The optimization algorithm maximizes the primary voltage, to reduce the primary current and thereby the losses.
- In case 2, a lower frequency and magnetic flux density is obtained compared to case 1, corresponding to the relation between the geometry, frequency and flux density as given in (1).
- The power losses in case A1 are 20% higher as in case C1, corresponding to a 21% smaller volume in case A1 compared to case C1.
- In case 2 both topologies have almost equal power losses.
- The temperature is depending on the power losses and core volume and is thus higher in the adjacent winding topology compared to the coaxial winding topology.

TABLE IV
TRANSFORMER PARAMETERS FOR TWO CASES WITH ADJACENT WINDING TOPOLOGY.

Parameter	Case: A1	Case: A2	Unit
r_{cin}	4.7	16.1	mm
r_{cout}	2.6	8.9	mm
r_4	20.7	37.6	mm
h_{out}	31.2	31.0	mm
l_{ag}	0.5	0.5	mm
S	302	203	mm ²
A_e	149	1087	mm ²
V	84	274	cm ³
N_p	99	70	turns
N_s	10	7	turns
B_{core}	317	131	mT
f_{res}	24	11.2	kHz
L_{mp}	2.70	8.35	mH
L_{Lkp}	0.83	0.65	mH
L_{Lks}	8.45	6.55	μ H
k	0.76	0.93	
P_{loss}	12.9	4.2	W
T_{core}	52.8	32.5	$^{\circ}$ C

TABLE V
P66/55 POT CORE DIMENSIONS.

Parameter	Dimension	Unit
r_{cin}	10.8	mm
r_{cout}	5.9	mm
r_4	33.2	mm
h_{out}	28.7	mm
l_{ag}	0.5	mm
S	286	mm ²
A_e	583	mm ²
V	199	cm ³

Overall, minimal losses can be obtained in a relative larger core. The adjacent winding topology is favorable because it uses the winding area more efficient, resulting in a lower magnetizing current and thereby, lower losses, as shown in the Pareto front.

VI. EXPERIMENTAL VERIFICATION

For each winding topology a rotating transformer is designed using the optimization algorithm. The optimization is conducted for fixed power transfer of 1 kW, obtaining minimal power losses, using the commercially available P66/56 pot core from Ferroxcube. Thereby, the core dimensions are fixed and they are specified in Table V. The core consist of the material 3C81 [12], a special developed Mann ferrite for high power applications below a frequency of 200 kHz, with minimal power losses around 60 $^{\circ}$ C. The material has a low saturation level, hence in this paper a saturation level of 350 mT is assumed. The manufactured rotating transformers are shown in Fig. 12. The corresponding parameters are specified in Table VI and VII for the adjacent and coaxial winding topology, respectively. The parameters are compared with FEM simulations [14] and inductances are measured with the HP 4194A impedance analyzer, a maximum error of 8% is obtained.

Comparison of the parameters of the prototype transformers shows that minimal losses are obtained in the adjacent winding topology. This can be explained by the different number of turns which fit in the winding area of both topologies. Since

TABLE VI
OPTIMIZED TRANSFORMER PARAMETERS FOR THE ADJACENT WINDING TOPOLOGY.

Parameter	Optimization	FEM	Measurement	Unit
N_p	100	-	-	turns
N_s	10	-	-	turns
l_{ag}	0.5	-	-	mm
f_{res}	18.6	-	-	kHz
B_{core}	104	106	-	mT
L_{m_p}	9.2	10.5	8.8	mH
L_{Lk_p}	0.82	0.89	0.82	mH
L_{lk_s}	8.2	8.9	8.6	μ H
k	0.92	0.92	0.91	
P_{loss}	9.4	10	-	W
T_{core}	59	56	-	$^{\circ}$ C

TABLE VII
OPTIMIZED TRANSFORMER PARAMETERS FOR THE COAXIAL WINDING TOPOLOGY.

Parameter	Optimization	FEM	Measurement	Unit
N_p	83	-	-	turns
N_s	8	-	-	turns
l_{ag}	0.5	-	-	mm
f_{res}	30.8	-	-	kHz
B_{core}	75	74	-	mT
L_{m_p}	6.8	7.3	7.1	mH
L_{Lk_p}	0.09	1.0	1.0	mH
L_{lk_s}	0.9	0.9	0.8	μ H
k	0.99	0.99	0.99	
P_{loss}	14.5	12	-	W
T_{core}	89	85	-	$^{\circ}$ C

the adjacent winding topology uses the winding area more efficiently, a higher number of turns is obtained which increases the magnetizing inductance and simultaneously decreases the magnetizing current. This results in lower conduction losses. Furthermore, since more turns fit in the adjacent winding topology, a lower frequency can be obtained which reduces the core losses.

The stationary performance of the rotating transformers is measured in an experimental setup. A half bridge converter is connected to the primary side of the transformer and a diode rectifier is connected to the secondary side. Resonance capacitances are connected on both sides of the transformer. A 200 VDC input voltage is supplied to the half bridge and an equivalent resistance of 2.5 Ohm is connected to the diode rectifier. The primary voltage waveform is measured after the half bridge and the secondary voltage waveform is measured before the diode rectifier. The waveforms are shown in Fig. 13 for a power transfer of 50 W, since the used half bridge limits the maximal input voltage. The figure shows the primary voltage on the left axis of the graphs. The axis on the right indicates the secondary voltage, approximately 10 times lower in amplitude due to the winding ratio and varies around 0. The amplitude of the secondary voltage of the coaxial winding topology is lower compared to the adjacent winding topology because of the slightly higher winding ratio.

VII. CONCLUSION

In this paper the adjacent and coaxial winding topologies in a rotating pot core transformer have been compared in terms of total core volume and power losses. A multi-objective

Fig. 12. Manufactured transformers (a) adjacent winding topology and (b) coaxial winding topology.

Fig. 13. Measured primary and secondary voltage waveform of the rotating transformer with (a) adjacent and (b) coaxial winding topology.

optimization has been defined, using an electromagnetic and a thermal model of the rotating transformer. The optimization algorithm has been used to derive the minimal Pareto front, which showed that lower power losses could be obtained in the adjacent winding topology. Two prototype transformers have been designed and manufactured to verify the models. Overall, the adjacent winding topology is favorable for a fixed power transfer of 1 kW.

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