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Contactless Power Transfer to a Rotating Disk

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Abstract—This paper discusses a power transfer system from the stationary to the rotating part of a device, by means of contactless energy transfer. A rotating transformer is proposed as a replacement for wires and slip rings. A pot core geometry is used for the rotating transformer and two different winding topologies are compared. The transformer is analyzed in the electromagnetic and thermal domain. An analytic model for each domain is derived. The validity of the analytical models is confirmed with both 2D and 3D FEM simulations and measurements. Two prototype rotating transformers are designed for the transfer of 1 kW peak, rotating at 6000 rpm. The prototypes are manufactured using commercially available pot cores and tested in an experimental setup.

I. INTRODUCTION

In many modern mechatronic systems, the transfer of power to rotating parts plays an important role, for example, in robotics and in industrial applications where power needs to be transferred to a rotating part. Nowadays, wires and slip rings are used to transfer power to the rotating part. Disadvantages of wires are a limited rotation angle and increased stiffness. Despite the significant amount of research and development of reliable and durable slip rings, contact wear as well as vibration limit the lifetime, and frequent maintenance is required [1]. Furthermore, contact wear creates dust particles, which are unwanted in cleanroom and vacuum applications.

A solution to overcome the disadvantages of wires and slip rings is a contactless energy transfer (CET) system that uses a rotating transformer. The transformer converts power across an airgap, a physical separation which provides the ability to rotate the secondary side of the transformer. An extra advantage could be the freedom in winding ratio, to transform the primary voltage level to the requirements of the load.

The contactless transfer of energy by means of a rotating transformer is under investigation since the 1970's [2]. Later, the concept of a rotating transformer is used in applications such as the transcutaneous energy transmission for pacemakers [3] and inductive charging [4], both cases benefit from the CET. Rotating transformers can be used for the transfer of power and data signals to the moving part simultaneously, by using an extra inductive or capacitive coupling [5].

The axial rotating and pot core transformer geometry can be used for a rotating transformer. Both are investigated by [6] in terms of total volume and efficiency. The pot core geometry, shown in Fig. 1, gives better performance indices in terms of flux density, magnetic coupling and losses. Therefore, this topology is further investigated in this paper.

This paper presents the design of a rotating transformer for a power transfer of 1 kW peak to a load, rotating at

Fig. 1. Pot core rotating transformer.

6000 rpm. The electronics on the load require an input DC voltage of 50 V. First, the geometry of the rotating transformer is analyzed. Second, analytical models are derived for the electromagnetic and thermal behavior of the transformer. Finally, two prototype transformers are designed and manufactured to verify the analytical models.

II. ENERGY TRANSFER TOPOLOGY

The working principle of a rotating transformer can be obtained from Faraday's law and Ampere's circuital law. Applying Lenz's law and assuming a sinusoidal excitation, yields to an equation for the induced voltage over an N-turn winding and an expression for the transferred power, independent of the number of turns

$$V_{rms} = \sqrt{2}\pi f N B_{peak} A_e, \quad (1)$$

$$P = \pi J S k_f f B_{peak} A_e, \quad (2)$$

where f is the frequency of the applied voltage, B_{peak} is the peak flux density and k_f is the filling factor of the windings. The cross section of the inner core, A_e , is the minimal core area to guide the flux and S is the area available for windings, as defined in Fig. 2. A top view and cross section of a rotating pot core transformer are shown in Fig. 2a and Fig. 2b, respectively. The corresponding geometric parameters are listed in Table I. These expressions can be used to determine the core geometry and main parameters at the start of the design of a rotating transformer.

In each core an indentation can be found, to guide the wires of the winding out the core, which creates an incomplete axisymmetric layout. The effect of the indentation on the power transfer during rotation is investigated by a 3D FEM model [7]. Figure 3a-d show the response of the secondary

Fig. 2. Geometry of the pot core, (a) top view and (b) cross section.

TABLE I
GEOMETRICAL PARAMETERS OF FIG. 2 AND FIG. 4

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

voltage for a changing load resistance for different relative positions of the indentations in the core halves. In each figure an extra curve is inserted and identical responses for the different angular positions have been found. Concluding that an axisymmetric geometry can be assumed for further analysis.

The rotating transformer is a part of a DC-DC power conversion system. On the primary side of the rotating transformer, a DC-voltage is converted to a high frequency voltage by a half bridge converter. This reduces the size of the transformer and maximizes the power transfer, as shown in (2). On the secondary side of the transformer, the high frequency voltage is rectified and supplied to the load.

Two different winding topologies can be placed in the rotating pot core transformer. The first topology is the adjacent winding topology, which is shown in Fig. 4a, where each winding is placed in a separate core half. Therefore, one side of the transformer can be completely isolated from the other side, and for example placed in vacuum. The second topology is the coaxial winding topology, which is shown in Fig. 4b, where the windings are placed around each other. This topology requires the use of an extra winding bobbin, which reduces the effective winding area. Because both windings rotate around each other with a small gap in between, vibration due to rotating can easily damage the windings. In this paper, both winding topologies are compared and the differences from a magnetic and electrical point of view are identified.

III. ANALYSIS

The design of a rotating transformer requires modeling in the electromagnetic and thermal disciplines.

A. Magnetic model

An axisymmetric magnetic reluctance model has been derived to calculate the inductances of the transformer. The

Fig. 3. Secondary voltage characteristics for different relative angular positions of the indentations in the core.

Fig. 4. Winding topologies, (a) adjacent and (b) coaxial.

magnetic flux paths, shown in Fig. 5, have been identified by a 2D FEM model and based on the physical layout a reluctance model has been created. The model is shown Fig. 6a, for the adjacent winding topology. $\mathcal{R}$ represents the reluctance and the subscripts c , ag and lk indicate the flux paths through the core, airgap and leakage paths, respectively.

Combining the reluctances of each half of the core and the airgaps, results in the reluctance network as shown in Fig. 6b, which can be rewritten as an equivalent electric circuit, shown in Fig. 6c. Where, L_m presents the magnetizing inductance, L_{lk_p} and L_{lk_s} presents the leakage inductance on the primary and secondary side, respectively.

1) *Magnetizing inductance:* The magnetizing inductance has been calculated by

$$L_m = \frac{N_p^2}{2(\mathcal{R}_{c_a} + \mathcal{R}_{c_b} + \mathcal{R}_{c_c}) + \mathcal{R}_{ag_a} + \mathcal{R}_{ag_b}}, \quad (3)$$

where the path of the mutual flux lines has been assumed through the both half cores and the airgaps. The reluctances for the pot core are determined by

$$\mathcal{R}_{c_a} = \mathcal{R}_{c_c} = \frac{\Delta z}{\mu_0 \mu_r \pi (r_o^2 - r_i^2)}, \quad (4)$$

$$\mathcal{R}_{c_b} = \frac{\ln(r_o/r_i)}{2\mu_0 \mu_r \pi \Delta z}, \quad (5)$$

Fig. 5. Flux lines in (a) adjacent and (b) coaxial winding topology.

Fig. 6. Reluctance modeling for the adjacent winding topology, (a) identifying the magnetic flux paths, (b) reluctance model, (c) equivalent electrical circuit.

where r_o and r_i are the outer and inner core radius of the part, respectively and Δz is the height of the core part. Due to the fringing flux around the airgap, an extra fringing flux factor, F_f , has been added to calculate the airgap reluctance [8]

$$\mathcal{R}_{ag} = \frac{l_{ag}}{\mu_0 \pi (r_o^2 - r_i^2)} \frac{1}{F_f}, \quad (6)$$

$$F_f = 1 + \frac{l_{ag}}{\sqrt{\pi (r_o^2 - r_i^2)}} \ln \left(\frac{4h_{in}}{l_{ag}} \right). \quad (7)$$

2) *Leakage inductance*: In the rotating transformer there are various leakage flux lines, that do not link both windings. Because those flux lines do not have an a priori known path, it is inaccurate to model them with a reluctance network as well. A different approach is to calculate the leakage inductance by the stored energy in the winding volume. The magnetic energy of the leakage flux can be expressed by

$$\frac{1}{2} L_{lk} I^2 = \frac{1}{2} \int_v \mathbf{B} \cdot \mathbf{H} dv, \quad (8)$$

which is equal to the energy of the magnetic field in the winding volume [8]. An expression for the magnetic field strength can be found by Ampere's circuital law. In the case of the adjacent winding topology, the magnetic field strength can be expressed for the primary winding as function of the axial length

$$H(z) = \frac{N_p i_p}{r_3 - r_2} \frac{z}{h_{in}}. \quad (9)$$

In the airgap, the magnetic field strength can be defined by assuming a uniform mmf

$$H = \frac{N_p i_p}{l_{ag}}. \quad (10)$$

Along the secondary winding, the magnetic field strength can be expressed similarly as (9). As the secondary winding space is traversed, the mmf linearly falls to zero, since $N_p i_p = -N_s i_s$. Solving the integral, (8), yields

$$L_{lk} = \mu_0 N_p^2 \frac{2\pi}{\ln(r_o/r_i)} \left(\frac{h_{w_p} + h_{w_s}}{3} + l_{ag} \right), \quad (11)$$

where L_{lk} is the total leakage inductance seen from the primary side. A similar expression for the leakage inductance can be derived for the coaxial winding topology, where the magnetic field strength should be expressed as function of the radius.

3) *Verification*: The inductances of the prototype transformers have been calculated and obtained from 2D FEM simulations and measurements on the prototype transformers (section IV). The inductances for the adjacent and coaxial winding topology are shown in Fig. 7 and 8, respectively. The figures show that by increasing the airgap, the magnetizing inductance and, thereby, also the magnetic coupling decreases. The leakage inductance is almost constant for an increasing airgap and depending on the winding topology. A lower leakage inductance is found in the coaxial winding topology, because both windings share an almost identical flux path.

In this paper an airgap length of 0.5 mm has been assumed. At an airgap of 0.5 mm, a maximal error of 5% can be found between the measured and analytical calculated inductances. It should be noticed that rotating the cores with a small airgap in between, requires an accurate assembly of the transformer.

B. Electric model

To complete the electric equivalent circuit, winding resistances, R_p , R_s and resonant capacitors, C_p , C_s have been added to the circuit, as shown in Fig. 9.

1) *Winding resistance*: The voltage applied from the half bridge converter to the rotating transformer has a square-waveform, which gives rise to the AC-losses due to harmonics. An analytical expression for the wire resistance in case of non-sinusoidal waveforms has been derived by [9], based on Dowell's formula for AC-resistance.

2) *Resonant capacitors*: On both sides of the transformer, resonant capacitors have been added to overcome the voltage drops across the leakage inductance, by locally boosting the voltage and, thereby, increasing the magnetic flux density. Resonance capacitors can be placed in series or parallel to the winding at either side of the transformer.

On the primary side the resonance capacitor has been placed in series to act as a DC-blocking capacitance and to create a zero crossing resonance voltage. This makes it possible to use zero-current switching, to minimize the switching losses. Placing the primary capacitor in parallel would result in a high current in the resonance loop due to the high frequency

Fig. 7. The (a) primary magnetizing inductance, (b) primary leakage inductance and (c) secondary leakage inductance of the adjacent winding rotating transformer.

Fig. 8. The (a) primary magnetizing inductance, (b) primary leakage inductance and (c) secondary leakage inductance of the coaxial winding rotating transformer.

Fig. 9. Complete electric equivalent circuit of the rotating transformer.

Fig. 10. Influence of the magnetic coupling on the primary resonance capacitance.

input voltage. This current would increase the power losses and should therefore be avoided.

On the secondary side the resonant capacitor has been added to boost the power transfer capability. Figure 10 shows the normalized value of C_p for a changing magnetic coupling for series and parallel resonance on the secondary side [10]. To make the resonance capacitor on the primary side insensitive for magnetic coupling changes, which are for example caused by vibration during rotating, the resonance capacitor on the secondary side is placed in series to the secondary winding.

The frequency at which the circuit operates at resonance, f_{res} , can be calculated by

$$f_{res} = \frac{1}{2\pi\sqrt{L_{lk_n}C_n}}. \quad (12)$$

Furthermore, the resonance circuit acts a filter for higher harmonics and, thereby, decreases the conduction losses.

3) *Power losses*: Conduction and core losses are the main power losses in the rotating transformer. The conduction losses, P_{cond} , have been calculated by

$$P_{cond} = I_{p_{rms}}^2 R_p + I_{s_{rms}}^2 R_s, \quad (13)$$

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

$$P_{core} = C_m C(T) f^x B^y V_{core}, \quad (14)$$

where C_m , $C(T)$, x and y are specified material constants and V_{core} is the core volume.

Both the core and the conduction losses are dependent on the frequency. Increasing the frequency under a constant power transfer, boosts the conduction losses because of the rising AC winding resistance and decreases the core losses because of the lower magnetic flux density. For a specific power transfer, an optimum between f_{res} and the magnetic flux density can be found, resulting in minimal core and conduction losses.

C. Thermal model

The core and conduction losses cause a temperature rise in the transformer. It is important to investigate the thermal

Fig. 11. A 2D thermal equivalent model drawn on a quarter cross section of the core.

TABLE II
HEAT TRANSFER COEFFICIENTS

Parameter	Dimension	Unit
$k_{Ferrite}$	4.25	$Wm^{-1}K^{-1}$
k_{Copper}	394	$Wm^{-1}K^{-1}$
h_{axial}	12.7	$Wm^{-2}K^{-1}$
h_{radial}	8.5	$Wm^{-2}K^{-1}$

behavior of the transformer during the design, because the relative permeability of the core material as well as the power losses in the core are temperature dependent. A thermal model allows the estimation of the average winding and core temperature.

A thermal equivalent circuit, shown in Fig. 11, is made using a finite-difference modeling technique, where the thermal resistance concept is used for deriving the heat transfer [11]. The thermal model is derived by dividing the upper half of the geometry into six regions, where region 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 model heat transfer inside the regions and convection resistances are used to model the heat transfer between the border of the regions and the air. No heat transfer is assumed at the left and lower boundary of the model. The power losses in each region are presented by a heat source and inserted in the middle node of each region.

The average temperature at each node has been calculated by determining the heat transfer between the nodes, expressed by

$$[\mathbf{R}_{th}][\mathbf{T}] = [\mathbf{Q}] \quad (15)$$

where $\mathbf{R}_{th}$ is a matrix which consists of all thermal resistances between the nodes, $\mathbf{T}$ is a vector comprising the temperature at each node and $\mathbf{Q}$ is a vector with all heat energy flowing into the transformer. The thermal resistances are defined using the heat transfer coefficients for conduction and convection, given in Table II.

1) *Verification:* To verify the thermal equivalent model, a 2D thermal finite element model has been created, based on the thermal assumptions. The temperature has been obtained

TABLE III
AVERAGE TEMPERATURE IN EACH REGION ($T_{amb} = 20^\circ C$)

Region	$T_{analytical}(^\circ C)$	$T_{numerical}(^\circ C)$
I	61.4	58.7
II	60.5	58.2
III	60.4	58.2
IV	58.2	57.7
V	60.8	58.3
VI	61.1	58.5

Fig. 12. Prototype rotating transformers, (a) adjacent and (b) coaxial winding topology.

in the center of each region and shown in Table III. The largest error between the analytical and numerical calculated increase compared to the environment temperature of $20^\circ C$ is 6.9%.

IV. PROTOTYPE TRANSFORMERS

To verify the analytical models, two rotating transformers have been designed and manufactured. A picture of the prototypes is shown in Fig. 12. The transformer parameters have been obtained from a sequential quadratic programming algorithm in MATLAB, in which the analytical models have been implemented. The algorithm has been used to find transformer parameters obtaining minimal losses. Since a limited number of cores is commercially available for a power transfer of 1 kW, the pot core P66/56 from Ferroxcube is chosen [12]. The pot core consist of the material 3C81, a special developed MnZn ferrite for high power applications. The pot core parameters are specified in Table IV. In Table V the optimized transformer parameters are specified for both winding topologies.

In Table V the optimized transformer parameters are specified for both winding topologies. This table shows the differences between the winding topologies. First of all the number of turns has been maximized in both winding topologies to decrease the frequency to obtain minimal power losses. More turns fit in the adjacent winding topology, showing a higher winding efficiency for this topology. Second, the higher magnetizing inductance in the adjacent winding topology decreases the magnetizing current and, thereby, the conduction losses. Overall, although a lower magnetic coupling is obtained, the adjacent winding topology is in favorite in terms of minimal power losses.

An experimental setup has been created consisting of a

TABLE IV
POT CORE PARAMETERS

Parameter	Dimension	Unit
r_{cin}	10.8	mm
r_{cout}	5.9	mm
r_4	33.2	mm
h_{out}	28.7	mm
S	286	mm ²
A_e	583	mm ²
B_{sat}	350	mT
μ_r	2700	-
ρ	1	Ωm

TABLE V
PROTOTYPE TRANSFORMER PARAMETERS

Parameter	Adjacent	Coaxial	Unit
l_{ag}	0.5	0.5	mm
N_p	100	83	turns
N_s	10	8	turns
B_{core}	104.3	75.23	mT
f_{res}	18.6	30.8	kHz
L_{mp}	9.72	6.81	mH
k	0.92	0.99	-
k_{favg}	0.22	0.18	-
P_{loss}	9.4	14.5	W

half bridge inverter connected to the primary side of the transformer and a diode rectifier connected to the secondary side. A stationary power transfer of 100 W has been obtained, due to limitation of the half bridge. The voltage across an equivalent load of 2.5 Ohm have been measured and simulated in MATLAB Simulink. The waveforms for the adjacent winding topology are shown in Fig. 13, the measured and simulated voltage have the same amplitude. Ripples can be found on the measured waveforms, an extra filter could be add to remove these ripples. A power transfer of 50 W is measured for different angular velocities, shown in Fig. 14. No significant difference in the power transfer is noticed for an increasing angular velocity.

V. CONCLUSION

In this paper a rotating transformer has been presented as a replacement for slip rings and wires to transfer power from a stationary to a rotating part of a device. Electromagnetic and thermal analytical models of the transformer have been derived with a maximal error of 7% compared to measurements and FEM simulations. Two prototype transformers with minimal losses are manufactured and a stationary power transformer of 100 W is obtained and no influence of the rotation has been found in the power transfer. The adjacent and coaxial winding topology inside a rotating pot core transformer have been compared in terms of minimal power loss. In conclusion, the adjacent winding topology uses the winding area more effectively, which reduces the frequency and magnetizing current to obtain lower power losses compared to the coaxial winding topology.

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Fig. 13. Measured and simulated (a) load voltage and (b) load current for adjacent winding topology.

Fig. 14. Power transfer for different angular velocities.

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