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Toroidally-wound permanent magnet machines in high-speed applications

A. Borisavljevic, S. Jumayev, E. Lomonova

Abstract—The paper studies potentials and limits of permanent magnet machines with toroidal windings in high speed applications. Three designs in different applications are used as test cases. The analyses based on analytical and FE models illustrate both merits and weaknesses of toroidally-wound machines that have not been addressed in literature. Strong external leakage of the armature field leads to relatively high inductance for a slotless machine and, yet, to susceptibility to losses in the housing. Losses in copper are, arguably, the decisive factor for suitability of this type of machine for a particular application; in particular, the machine is hardly suitable to applications requiring high power density. The paper finally demonstrates the importance of having a flawless winding process in order to avoid excessive core losses.

Keywords—high-speed machines, toroidally-wound machines, toroidal windings, ring windings, permanent magnet machines, inductance calculation, copper loss, housing

I. INTRODUCTION

Using toroidal (also: *ring-type* or *cup-shaped*) windings can be beneficial for electrical machines which require very short axial length. Such machines are often used for propulsion and traction [1] or as direct-drive generators [2], [3]. In recent years, toroidally-wound permanent magnet machines were suggested in several applications requiring very high speeds [4]–[7].

In addition to standard advantages correlated with slotless machines - no cogging and high efficiency [8] - this way of winding stators is very useful when the axial length is restricted since the end windings do not increase the machine axial length. Furthermore, the machine may also benefit in terms of operating temperature as the windings can be in a direct contact with the housing and/or coolant. Relatively high leakage inductance can also be an advantage for a slotless machine whose inductance would, otherwise, be very low often entailing the need for additional chokes at the inverter output. Finally, low armature-reaction field and distributed windings are expected to result in low rotor losses.

Besides potential merits, there are also some evident problems involved with the use of toroidal windings. Compared to other windings for slotless machines, these windings usually have extended conductor length and, in turn, potentially high copper losses. Because of the pronounced leakage field, eddy currents can be expected in machines surroundings/housing. In this paper, also other risk factors connected with the manufacturing process will be highlighted.

The objective of the paper is to evaluate toroidally-wound PM machines and assess their potentials and limits for high-speed (industry) applications. Comprehensive analysis of such machines has not been reported in literature so far.

The evaluation of toroidally-wound machines will be made in terms of their general applicability to different high-speed applications, susceptibility to (excessive) losses

and manufacturability. The study will apply some existing electromagnetic models to analyze the given machines.

Winding configuration is the main peculiarity of toroidally-wound machines; this study is, therefore, mainly concerned with effects connected with winding (armature) field, such as inductance, losses induced by the armature field, losses in copper; as well as with some practical aspects of winding production.

II. EXAMPLE MACHINES

Three machine design examples for different high-speed applications will be used as test cases throughout the paper. All three designs represent two-pole radial-flux permanent magnet machines with diametrically-magnetized magnet on the rotor. Relevant properties of these machines are given in Appendix.

Machine A represents a small high-speed machine with a disk-shaped rotor designed for a micro-milling spindle; design of this machine is thoroughly described in [9]. For a motor with very small stack length compared to its diameter, toroidal windings represent a sensible solution. The motor is designed for 200.000 rpm and modest power of 150 W.

Machine B represents a small motor for an air compressor of a respiratory system. The machine is designed for highly-dynamic operation with the speed rapidly changing in the range of 30.000-90.000 rpm and with the maximum continuous torque of 4.5 mNm. Toroidal windings in this design are chosen because of their simplicity and potentially low price; besides, relatively high inductance compared to other slotless machines of similar sizes is favorable for filtering high-order harmonics of the stator currents.

Machine C represents a gas-turbine generator within a micro Combined Heat and Power (micro-CHP) system. In contrast to the previous two, this is a highly power-dense application in which the generator is supposed to deliver 3.7 kW of electrical power at the nominal speed of 240.000 rpm. In the current system a water-cooled slotted machine with distributed winding is used with nominal efficiency between 91 and 92%. A prospective for using toroidal windings was found in very simple stator construction, which would facilitate inexpensive production of low-loss iron cores, and in the possibility to directly cool the windings of the machine.

III. TOROIDALLY-WOUND MACHINES GENERAL

Figure 1 shows a cross-section of a typical high-speed toroidally-wound machine. It is represented to be a two-pole machine as most of high-speed machines have two poles including the test machines in this study. The rotor generally consists of an iron shaft, permanent magnet and non-magnetic retaining sleeve; machines B and C, however, have no iron in the rotor in which case $r_1 = 0$. Magnetization is shown to be diametrical as it is in all the test machines; however, other magnetization types are possible too.

Six coils are wound around a toroidal iron core. Each coil spans over the angle $\theta_c \leq 60^\circ$ and opposite coils belong to the same phase, as indicated in the picture. Usually, coils are mutually separated, either by protrusions

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Fig. 1. Two-pole toroidally-wound permanent magnet machine – a cross-section

of the stator core or housing or an additional material: in machine A they are extensions of the stator core, while in machine B they are made out of plastic material. These protrusions help fitting the stator into the housing. The housing is made of conductive material (e.g. aluminium); it is sometimes in a direct contact with the windings, in which case $r_8 = r_7$, or there is some additional gap between the windings and housing as shown in the picture.

IV. INDUCTANCE

Because conductors are wound around the stator core, significant leakage of the armature field is expected. In order to give a reasonable estimation of the phase inductance, a 3D model of the toroidally-wound machine is needed.

TABLE I. SYNCHRONOUS INDUCTANCE CALCULATED BY DIFFERENT MODELS

	2D, magnetizing	2D		3D	
	analytical	anal.	FEM	anal.	FEM
Machine A [μH]	43	64	64	187	280
Machine B [μH]	24	49	49	82	101

In Table I synchronous inductance calculations are shown for different models applied to machines A and B. Magnetostatic model has been used to analytically model the machines in 2D and the method of images was used for 3D analytical calculations as explained in [10]. In the first case, the inductance values represent only the *magnetizing* inductance i.e. the values are analytically calculated in 2D by disregarding the external leakage of the armature field. In the second case, 2D calculations which also include the leakage outside of the stator core are shown. Finally, 3D model results are shown with 3D FEM results being the most representative of the actual inductance. It is evident that the contribution of the external leakage inductance to the overall synchronous inductance is immense.

In [10] inductances of different slotless machines with same air-gap dimensions and permanent-magnet flux linkage (and torque) are compared using the method of images for calculations. The machine with toroidal windings has significantly higher inductance than other slotless machines, namely those with inserted skewed and concentrated windings. This is evident from the graphs in Figures 2 and 3 in which self and mutual inductance are plotted with respect to the machine length.

High inductance is potentially a very beneficial feature for this type of machines. Namely, slotless machines regularly have very small inductance and, often, an additional choke needs to be added on the inverter outputs to filter the current. With toroidally-wound machines, however, this is not *per se* needed.

Fig. 2. Self inductance of a phase in a slotless machine with toroidal, skewed and concentrated windings vs. stack length [10]

Fig. 3. Mutual inductance of a slotless machine with toroidal, skewed and concentrated windings vs. stack length [10]

V. LOSSES IN HOUSING

The external leakage of the armature field of toroidal windings is apparently rather high and it will inevitably induce eddy currents in neighboring conductive bodies. Since machine housings are commonly made of materials, such as aluminum, which are both good thermal and electrical conductors, one can expect significant losses in the housing. Moreover, for good cooling of a machine, the housing must be in contact with or, at least, in close proximity of the stator core which, consequently, increases susceptibility of housing to induced losses from neighboring toroidal coils.

The example of Machine A illustrates this effect very well. Figure 4 shows measurements of ac resistance of one phase in the toroidal winding against electrical frequency for the cases of the stator being inside and outside of the aluminum housing. It is evident that the losses in the housing, in this example, have a dominant portion of the armature-induced losses. Actually, in this particular design the losses in the housing take up as much as about 50% of all losses in the machine [9].

To accurately model these losses, one must take into account fields of all external conductor parts of the toroidal windings as well as the particular geometry of the housing. Still, some general conclusions can be drawn using relatively simple 2D models that take into account eddy currents, such as the one explained in [11]. One can solve the diffusion equation:

$$-\nabla^2 \mathbf{A} + \mu\sigma \frac{\partial \mathbf{A}}{\partial t} = \mu \mathbf{J}_{s,ext} \quad (1)$$

outside of the stator core ($r \geq r_6$ in Figure 1), apply the

Fig. 4. Normalized AC resistance of the machine A measured in the locked-rotor test: with and without housing [9]

boundary conditions for magnetic field ($B_{r,i} = B_{r,i+1}$, $H_{\theta,i} = H_{\theta,i+1}$) and, finally, calculate losses in the housing as the flux of the Poynting vector through the inner surface of the housing [12].

In equation (1) $\mathbf{A}$ is magnetic vector potential, $\mathbf{J}_{s,ext}$ is the current density of the external conductor parts, and μ and σ are magnetic permeability and electrical conductivity of the given region.

The model is applied to idealized stator geometries of Machine A and B (according to Figure 1) with full pitch of toroidal coils ($\theta_c = 60^\circ$) and aluminum as the housing material at room temperature. Losses in the housing per mm length of the lateral stator conductors (only) have been calculated for machines' rated electrical frequencies. The results are shown in Figures 5 and 6 for different thicknesses of the housing wall and its distance from the windings.

Fig. 5. Loss in the housing of the machine A as a function of housing thickness for different distances from the windings d ; δ represents the skin depth in aluminum at 3.5 kHz

From the graphs one can conclude that the losses can be very high if the housing thickness is smaller than the material skin depth at the given frequency. If the thickness is increased beyond the skin depth, the minimum loss is first reached, and, with further increasing the loss growth is insignificant: as can be expected, most of the losses occur in the region close to the inner surface of the housing.

Logically, as the housing is placed further away from the stator, the losses decrease. However, moving the housing away from the conductors can have serious consequences to the temperature in the machine and, apparently, the losses

Fig. 6. Loss in the housing of the machine B as a function of housing thickness for different distances from the windings d ; δ represents the skin depth in aluminum at 1.5 kHz

themselves would not even be significantly reduced.

In conclusion, housing losses are a great point of concern for toroidally-wound machines and must be taken into account in the design phase. The losses can be alleviated by proper design of the housing and appropriate system for cooling the potential advantage of the machine is that the bulk of losses occur in the regions which can be actively cooled.

VI. COPPER LOSSES

Losses in copper are another important factor for performance of toroidally-wound machines. Namely, these machines have relatively high inactive conductor length and conduction (I^2R) losses are expected to be relatively high compared to other electrical machines with the same torque. Besides, as in other slotless machines, the parts of the windings residing in the air-gap are susceptible to eddy-current losses induced by the pulsating field of the rotor magnet(s). For a given magnitude and frequency of the excitation field, losses in wires increase with 4th power of the wire diameter [13], [14]. Since the two loss components are conflicting in terms of their dependence on size and number of conductor wires, it is essential to adequately size the wires and to find a balance between conduction and eddy-current losses.

The conduction loss representation - I^2R - should, in principle, take into account the reduction of effective wire cross-section as a result of the skin-effect. However, in electrical machines wire diameter is regularly much smaller than the skin depth, and the wire-level skin effect can be neglected [13], [14]. Therefore, for a conductor consisting of n_{st} parallel strands of diameter d_{st} dc conduction loss in one conductor turn yields:

$$P_{dc} = I^2 R_{dc} = \frac{4I^2 \rho l_{tot}}{\pi n_{st} d_{st}^2}, \quad (2)$$

where l_{tot} is the total length of one conductor turn and ρ is resistivity of copper.

An exact solution of eddy-current loss in round wires can be found in [15]; however, for wire diameters larger than the skin depth, the power dissipated in a single bundle of strands lain in the air gap can be closely approximated by [13], [14]:

$$P_{eddy} = \frac{\pi B_m^2 \omega^2 l_{ag} n_{st} d_{st}^4}{128 \rho}, \quad (3)$$

where B_m is the amplitude of sinusoidal flux density in the air gap at the position of the conductor and l_{ag} is the air gap length.

In addition to these loss components, *bundle-level* skin and proximity effect (involving multiple strands in a conductor *bundle*) also contribute to the overall copper loss. However, bundle-level skin effect is practically independent from the strand sizing [14] as the conductor optimization tends to fill up the available winding area. Bundle-level proximity effect is, on the other hand, adequately controlled by twisting or weaving the wires [14], [16].

A procedure for optimal sizing of conductor wires is explained in [14]; following this procedure it is possible eventually to assess the minimum possible copper loss using equations 3 and 2:

$$P = P_{dc} + P_{eddy}. \quad (4)$$

Here, the procedure from [14] will be shortly recounted.

For a fixed number of winding turns N and cross-sectional area in the air-gap designated for the winding A_w , the maximum bundle diameter is defined by:

$$d_b \leq d_{b,max} = \sqrt{\frac{K_{pack} A_w}{N}}, \quad (5)$$

where K_{pack} is the packing factor of round wires which is strongly influenced by the particular winding process.

By taking into account packing factor of wires on both strand and conductor level, and the insulation build of the wire, one can express the correlation between the strand diameter and number of strands:

$$d_{st} = d_r^{1-1/\beta} \alpha^{-1/\beta} \left(\frac{K_{pack} K_{pack,st} A_w}{n_{st} N} \right)^{1/(2\beta)}. \quad (6)$$

where d_r , α and β are coefficients of an empirical correlation between the total insulated-wire diameter and nominal (copper) diameter:

$$d_{st,tot} = d_r \alpha \left(\frac{d_{st}}{d_r} \right)^\beta. \quad (7)$$

After substituting d_{st} in the expression for total copper loss (4), one can find the optimal number of strands from equation:

$$\frac{\partial P}{\partial n_{st}} = 0. \quad (8)$$

The explicit formula yields:

$$n_{st,opt} = \left(\frac{d_{b,max}}{d_r^{1-\beta} \alpha p_f} \right)^\gamma \left(\frac{512 I^2 \rho^2 l_{tot} (1-\beta)}{\pi^2 B_m^2 \omega^2 l_{ag} (2-\beta)} \right)^{\frac{\beta}{6}\gamma}, \quad (9)$$

where:

$$\gamma = \frac{6}{3-2\beta}. \quad (10)$$

Naturally, the optimal number of strands is the integer number nearest to the value defined by (9). Although optimal, the previous method usually results in a very high number of strands. Not only that this may be inconvenient from the production point of view, but it can also be unjustified given the actual reduction of losses. For this reason, another approach to adequately size the conductors is also presented.

If the number of strands is fixed, the optimal conductors no longer necessarily fill the available winding space. For the given number of strands n_{st} , the optimum strand diameter is obtained from:

$$\frac{\partial P}{\partial d_{st}} = 0 \quad (11)$$

to be:

$$d_{st,opt1}(n_{st}) = 2 \cdot \sqrt[3]{\frac{2 I \rho l_{tot}}{\pi B_m \omega n_{st} \sqrt{l_{tot} l_{ag}}}}. \quad (12)$$

However, strands of the optimum diameter might not fit the available area for the conductor bundle (5). The maximum nominal strand diameter $d_{st,max}$ is actually defined by equation (6). Therefore, the optimal strand diameter for a fixed number of strands n_{st} is defined by:

$$d_{st,opt}(n_{st}) = \min(d_{st,opt1}(n_{st}), d_{st,max}(n_{st})). \quad (13)$$

The maximum number of strands $n_{st,max,cost}$ is determined by cost factors, and the optimal number of strands $n_{st,opt}$ is defined by (9). The smaller of these two values represents the definite limiting value for the number of strands:

$$n_{st,max} = \min(n_{st,max,cost}, n_{st,opt}). \quad (14)$$

A practical approach to optimally sizing the wires (considering cost constraints) can be established as follows: increase the number of strands until either maximum number of strands $n_{st,max}$ has been reached or the corresponding copper loss P has begun to converge (i.e. increasing the strand number does not lead to a significant loss reduction).

Machines A and C have been used to demonstrate the approaches for conductor sizing and the extent of copper losses in different toroidally-wound machines. In both cases, amplitude of flux density B_m in the winding region is calculated using the 2D analytical model mentioned in Section V, this time applied to the area of the machine inside the stator core. For both machines the optimal strand number $n_{st,opt}$ is calculated according to equation (9). After that, optimal diameters and corresponding total copper loss are calculated for different strand numbers starting from $n_{st} = 1$ until $n_{st} = n_{st,opt}$. The results are displayed in Figures 7 and 8.

Fig. 7. Minimum total copper loss vs. number of parallel conductor strands for the machine A

From both figures it is evident that using the truly optimal numbers of strands is unjustifiable given the actual reduction in overall loss. It is particularly noticeable with the machine A where the loss corresponding to the optimal strand number of 13 is just a fraction lower than the minimum loss generated when only 4 strands are used.

More importantly, the results of this analysis indicate also that toroidally-wound machines are hardly suitable for highly power-dense applications. Namely, the test design C would result in roughly 10 times higher power density (i.e. 20 times higher power at roughly twice the air-gap volume) than the design A and, at the same time, more than 50 times more losses would be generated in copper. The bulk of these

Fig. 8. Minimum total copper loss vs. number of parallel conductor strands for the machine C

losses would actually be generated in the air-gap rendering the design impractical.

Toroidally-wound machines can be suitable in relatively low-power applications. On the other hand, when high power density is required, the losses in copper of the toroidal windings are prone to be excessive.

VII. WINDING MANUFACTURING CONSIDERATIONS

Toroidal windings are quite common in inductors and machinery for their production is readily available. In principal, using winding machines, high fill factors can be achieved. Still, for machine applications, winding process must be carefully scrutinized.

Figure shows a simplified drawing of Machine A cross-section with indicated flux lines of the phase-*a* coils. If the serially connected coils of the phase are equal i.e. have a same number of turns, the conversely-directed flux lines linked by the two coils will exit the stator core and cross the air-gap; just like in a conventionally-wound machine. However, if one of the coils have an additional turn, the field of the added turn links entirely through the stator iron whose small reluctance allows the resulting field to be very high.

Fig. 9. Motor geometry cross-section: depiction of the effect of an additional turn in one of the phase coils on the field in the stator core [9]

In Figure 10 a flux-density plot and field lines are shown for examples of an evenly wound Machine A and of the same machine with only one extra turn in one of the two toroidal coils of the phase *a* (precisely, 41 turns instead of 40). While adding an extra turn in the left-hand coil of the phase does not increase by much the phase inductance, flux

Fig. 10. Flux-density plots and field lines of the motor with balanced coils (top) and with a single coil having an additional turn (bottom) - 2D FEM [9]

density in the stator iron increases tremendously: 262 mT of the maximum flux-density in the unevenly wound motor against only 20 mT in the balanced machine (simulated for $i_a = 1$ A, $i_b = i_c = -0.5$ A). The losses in the core will quadratically follow the increase in amplitude of the flux density; therefore, an unevenly wound coils will contribute to a significant increase in ac losses in the stator (as also demonstrated in [9]).

Hence, the winding production process for toroidally-wound machines must be highly repetitive and the winding process must be monitored. In turn, highly-repetitive coils require, arguably, a fully-automatized winding process.

VIII. CONCLUSIONS

Using permanent magnet machines with toroidal windings in high-speed industry applications can be beneficial from the point of low manufacturing cost. The stator with toroidal windings inactive axial length due to end windings, the iron core is simple and allows for inexpensive winding and assembly procedures. Furthermore, the machine may also benefit in terms of operating temperature due to the direct contact of the windings with the housing and/or coolant.

This paper critically reviews merits of toroidal windings and also indicates potential risks connected with using such windings. The study is primarily focused on phenomena related to the stator (armature) field, in particular: its influence on inductance and losses.

Contribution of the (external) leakage inductance to the overall synchronous inductance of the machine is very high. Toroidally-wound machines tend to have significantly higher inductance than other slotless machines. Such a high inductance can, actually, be rather beneficial in terms of cost

of the electrical drive as it can eliminate the need for big inductive chokes at the converter output.

On the other hand, the design of a toroidally-wound machine must address eddy-current losses outside of the machine, especially when it comes to the electrically-conductive housing. The study shows that the housing thickness should be somewhat than its material's skin depth at the operating frequency so that the losses are minimized.

Since the external leakage field is immense, an accurate 3D model of the machine is necessary to adequately represent the armature field and calculate inductance and external losses.

Assuming all other geometrical and electromagnetic parameters of the machine to be *a priori* defined, based on modeling of copper losses, the paper establishes different approaches for the sizing of conductors. The total copper loss is, arguably, the decisive factor for the suitability of a toroidally-wound machine for a certain application. The machine can be favorable for high-speed applications in when efficiency is most important design factor. On the other hand, when high power density is (also) required, copper losses in the air-gap conductors become the main obstacle for the performance.

Unevenly wound coils in motor phases causes an unnecessary increase of the armature field and, in turn, losses in the stator core. This phenomenon implies that an automatic and highly repetitive winding process is strongly recommended for toroidal windings.

APPENDIX

TABLE II. PARAMETERS OF THE TEST MACHINES

Parameter	Test machine A	Test machine B	Test machine C
f [Hz]	3500	1500	4000
I [A]	2	2	12
P [W]	150	30	3700
N	40	28	24
l_{ag} [mm]	6	16	39

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IX. BIOGRAPHIES

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