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Design and Evaluation of a Short-Circuit Rotor-Ring for Enhanced Self-Sensing Capability in a Slotless PM Motor

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Abstract—This paper deals with the design and evaluation of a short-circuit rotor-ring for enhanced self-sensing capability in a 2-pole slotless permanent-magnet motor. High-frequency rotor anisotropy is enabled by adding a short-circuited rotor-ring. Optimal ring-dimensions are determined by parametric numerical studies. An effective high-frequency saliency of up to 1.8 is demonstrated. The steady-state estimation error is within $\pm 2^\circ$, practically unaffected by current level (in contrast to slotted self-sensing electrical machinery) due to the large effective air-gap. The proposed observer performs properly also during dynamic load changes. Fall-times from 3000 to 0 rpm in less than 50 ms with negligible overshoot are demonstrated. Also parasitic effects such as power losses, heating and torque ripple are analyzed.

Index Terms—Rotor position estimation, saliency-based sensorless control, self-sensing, slotless permanent-magnet motor.

I. INTRODUCTION

SLOTLESS permanent magnet motors are preferred within power tools, machine spindles, turbine generators, compressors and turbo-machinery due to their high rotational speed [1]–[4], subsequently enabling high power density. Accurate control of such motors, using a power electronic inverter, requires knowledge of the rotor position and speed [5]. This is typically achieved by using a mechanical position sensor (see Fig. 1). An alternative is to adopt a sensorless control algorithm [6]–[9], so as to eliminate the position sensor and subsequently reduce cost, increase robustness and reliability, and reduce size [10].

While back-EMF based methods [11]–[14] provide sufficient control capability in the medium to high speed range, the low speed range requires active position detection methods [15], [16]. A prerequisite for low-speed self-sensing capability using signal injection is an adequate level of magnetic

anisotropy [8], [17]. Furthermore, the 180° ambiguity of estimation for injection-based methods also requires an algorithm for initial polarity detection [18]–[20].

Many motor types, such as the interior and inset PM motor, have a natural magnetic anisotropy due to rotor structure [21]–[23] which enables usage of active position detection methods. Unfortunately, Surface-mounted PM (SPM) motors such as the machine type under investigation, exhibit no magnetic anisotropy based on structure. Instead, such motors can utilize the position dependent inductance due to core saturation and anisotropic permeability of permanent magnets. However, this effect is very limited in slotless motors (typically less than 1 %) compared to slotted ones [24], [25], resulting in a low signal-to-noise ratio. It can be improved by oversampling the phase currents during their transient response to voltage pulses over several PWM periods [26], enabling position estimation in slotless PM motors within $\pm 30^\circ$ for rotor speeds below 200 rpm [25]. However, high-dynamic control of slotless PM motors requires significantly improved position estimation capability. Another way to enhance magnetic anisotropy in slotted SPM motors is to utilize a rotor winding [27]–[30], enabling an asymmetric high-frequency signal response. The method not only improves anisotropy but also the sensitivity to saturation and d and q -axis cross-coupling [28], [31] at the expense of additional rotor losses [32]. Compared to conventional slotted machinery, the slotless motor type has significant advantages in terms of negligible amount of space harmonics [33], [34] and very limited saturation effects [24]. Therefore, this paper presents the previously unprecedented analysis of a short-circuit rotor-ring design methodology for a 2-pole slotless PM motor, enabling accurate self-sensing control capability in the full speed-range.

The investigated machine and the application is presented in Section II. The ring design is made in Section III.

The motor control scheme is described in Section IV. An estimation error analysis, accompanied by the proposed compensation method, is presented in Section V. The power losses and resulting thermal influence, as well as the torque ripple due to the rotor-ring are considered in Section VI. Finally, conclusions are drawn in Section VII.

II. INVESTIGATED MOTOR

The investigated motor, sketched in Fig. 1 and Fig. 2a, belongs to an industrial nutrunner. Deep analysis of such

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Fig. 1. Motor module radial cross-section.

TABLE I
MOTOR DATA

Quantity	Symbol	Value
Peak torque	$T_{\max}$	1.15 Nm
Maximum speed	$n_{\max}$	36 krpm
Peak phase current	$I_{\max}$	20 A
Nominal DC-bus voltage	U_{DC}	325 V
Motor active length	L_a	64.5 mm
Stator outer radius	r_{lam}	15.5 mm
Rotor radius	r_r	7.5 mm
Air-gap length	δ	0.5 mm
Magnet remanent flux density	B_r	1.3 T

machinery can be found in for example [1]–[4], [33], [34]. The absence of slots eliminates cogging torque and provides a practically linear current versus torque relationship [24], subsequently improving position control capability. The rotor comprises a solid shaft carrying axially segmented ring NdFeB-magnets with parallel magnetization. The stator consists of an overlapping winding inside a cylindrical stator yoke of axially laminated stator steel.

Fig. 2. Axial cross-section of a) The slotless motor ($i_{a,b,c} = 0$ A, i.e., field lines only due to rotor PMs); b) The proposed rotor structure.

Original motor data (maintained throughout the study) are provided in Table I. The proposed rotor-ring is illustrated in Fig. 2b. The typical drive cycle is shown in Fig. 3 characterized by the high-speed rundown at negligible load, followed by a fast braking and a low-speed tightening until the target torque. From a sensorless control point of view, the challenges are the

Fig. 3. Typical load profile of an electric motor for handheld nutrunners: a) rundown ($T_{\text{rundown}} \approx 1$ s); b) tightening ($T_{\text{tight}} \approx 0.2$ s).

low speed operation at load and the fast braking capability, requiring high dynamic speed control.

The small-signal flux density distribution for d-axis AC current at different levels of q-axis DC-current have been simulated for $\{\xi = 60^\circ, \delta_{\text{ring}} = 0.3 \text{ mm}\}$ in Fig. 4 using the frozen permeability method (i.e., the flux density distribution from permanent magnets and the fundamental q-axis DC-current determines the small-signal AC frozen permeability map. As can be seen in Fig. 4a–b, the ring has a negligible effect on q-axis AC flux, but significantly counter-acts d-axis AC flux. This anisotropic reaction yields an effective AC rotor saliency which can be utilized for accurate rotor position detection. Fig. 4c shows the small signal flux pattern under heavily loaded conditions ($i_q = 60$ A). While no significant reduction of the small-signal AC flux densities can be observed, an increased level of d and q-axis cross-coupling can be seen (due to core saturation), resulting in an estimation error. The error angle θ_{error} for $i_d = 1$ A has been quantified for a wide range of i_q using:

$$\theta_{\text{error}} = \arctan\left(\frac{\Psi_q}{\Psi_d}\right) \quad (1)$$

where Ψ_d and Ψ_q are the small-signal d and q-axis flux linkages obtained by 2D FEA at 3125 Hz. The results are presented in Fig. 5, showing only 0.7° estimation error at $i_q = I_{\max}$ which can be explained by the large relative air-gap of the investigated machine. At 3 times the max current ($i_q = 60$ A), the error is around 6.3° . Using Eq. (2), the influence of saturation on saliency was investigated; at max current ($i_q = 20$ A) the saliency decreased with less than 0.2 % compared to the no-load case. At three times the maximum current ($i_q = 60$ A) the saliency decreased with less than 4 %.

III. RING DESIGN

In this section, the ring dimensions are determined by small-signal harmonic finite element analysis, deploying the frozen permeability method (i.e., the flux density distribution due to permanent magnets determines the frozen permeability map). The goal is to maximize the saliency, defined as:

$$\frac{Z_q(\omega)}{Z_d(\omega)} = \frac{|Z_{q,\text{active}}(\omega) + Z_{\text{ew}}|}{|Z_{d,\text{active}}(\omega) + Z_{\text{ew}}|} \quad (2)$$

where $Z_{d,\text{active}}$ and $Z_{q,\text{active}}$ are the d and q-axis active length impedances, obtained by FEA. $Z_{\text{ew}} = R_{\text{ew}} + j\omega L_{\text{ew}}$ is the

Fig. 4. Small-signal flux density distribution at 3125 Hz using the frozen permeability method for a) $\{i_{d,q}(\text{DC}) = 0 \text{ A}, i_q(\text{AC}) = 1 \text{ A}\}$; b) $\{i_{d,q}(\text{DC}) = 0 \text{ A}, i_d(\text{AC}) = 1 \text{ A}\}$; c) $\{i_q(\text{DC}) = 60 \text{ A}, i_d(\text{AC}) = 1 \text{ A}\}$.

Fig. 5. Influence of q-axis current on angular error ($i_d = 0 \text{ A}$). Solid line shows the mean error while dashed lines show max and minimum values obtained over one mechanical revolution.

analytically computed end-winding impedance [1]. Two parametric studies have been carried out at no-load (i.e., $i_q = 0$):

- Influence of ring angular width ξ (see Fig. 6)
- Influence of ring thickness δ_{ring} (see Fig. 7)

Within the investigated frequency range, results suggest an angular width not exceeding 60° , while no upper limitation can be seen from thickness point of view.

Fig. 6. Influence of ring arc angle ξ on saliency (for $\delta_{\text{ring}} = 0.3 \text{ mm}$).

Fig. 7. Influence of ring thickness δ_{ring} on saliency (for $\xi = 60^\circ$).

A. Semi-analytical approach

As demonstrated in [29], the d and q -axis active part impedances $Z_{d,\text{active}}$ and $Z_{q,\text{active}}$ can be expressed analytically:

$$Z_{x,\text{active}} = \underbrace{\left[R_a + \frac{\omega^2 R_{\text{rx}} M_{\text{rx}}^2}{R_{\text{rx}}^2 + \omega^2 L_{\text{rx}}^2} \right]}_{R_{\text{eq},x}} + j\omega \underbrace{\left[L_a - \frac{\omega^2 L_{\text{rx}} M_{\text{rx}}^2}{R_{\text{rx}}^2 + \omega^2 L_{\text{rx}}^2} \right]}_{L_{\text{eq},x}} \quad (3)$$

where R_a and L_a are the active length phase-to-phase resistance and inductance, respectively. R_{rx} represents the ring resistance and L_{rx} the ring inductance. M_{rx} is the ring to stator winding mutual coupling obtained using the power-invariant Clarke transformation. The deployed parameter values are summarized in Table II. The phase-to-phase quantities have been measured, while ring quantities were extracted from FEA, assuming the ring current paths shown in Fig. 8.

Fig. 8. Assumed ring current paths for calculation of: a) d -axis; b) q -axis parameters.

TABLE II
MOTOR PARAMETERS

Quantity	Symbol	Value
Active length phase-to-phase resistance	R_a	1.55Ω
End-winding phase-to-phase resistance	R_{ew}	0.4Ω
Active phase-to-phase inductance	L_a	0.95 mH
End-winding phase-to-phase inductance	L_{ew}	0.26 mH
Ring d -axis mutual inductance	M_{rd}	$8.81 \mu\text{H}$
Ring q -axis mutual inductance	M_{rq}	$2.26 \mu\text{H}$
Ring d -axis resistance	R_{rd}	$1.7 \text{ m}\Omega$
Ring q -axis resistance	R_{rq}	$3.4 \text{ m}\Omega$
Ring d -axis inductance	L_{rd}	118.6 nH
Ring q -axis inductance	L_{rq}	58.2 nH

B. Ring assembly considerations

The glued joints between the rotor and copper sleeves have to withstand the centrifugal forces due to rotation $\sigma_{\text{centrifugal}}$ and shear stress due to thermal expansion σ_{shear} according to:

$$\sigma_{\text{centrifugal}} = \frac{m\omega^2 r_r}{A_{\text{contact}}}, \quad \sigma_{\text{shear}} = \frac{E_{\text{Cu}} A_{\text{cross}} \alpha_{\text{diff}} \Delta T}{A_{\text{contact}}} \quad (4)$$

where m is the sleeve mass, ω the rotor angular speed and A_{contact} is the contact area between the rotor and sleeve. E_{Cu} is the Young's modulus of copper and A_{cross} the axial cross-sectional area. $\alpha_{\text{diff}} = \alpha_{\text{Cu}} - \alpha_{\text{NdFeB}}$ is the differential coefficient of thermal expansion and ΔT the temperature difference. The centrifugal force at 36 krpm and shear stress at $\Delta T = 150 \text{ K}$ result in 287 kPa tensile stress and 360 kPa shear stress in the glued joint. While both of these values are relatively low, the thermal cycling and class H temperature requirements limit the choice of adhesives to mainly special types of urethane modified acrylic compounds and epoxies [35].

C. Prototype evaluation

A ringed-pole rotor [28] was built using commercially available copper sheet with 0.3 mm thickness (nominal composition 99.9 % copper and 0.042 % phosphour). Two pieces of 100 mm long strips (see Fig. 9a), each spanning $\xi = 60^\circ$ of the rotor surface, were symmetrically assembled around the d -axis using glue. The copper sheets were connected in each end using two parallel conductors of 1.5 mm^2 cross-sectional area, as shown in Fig. 9b. The prototype saliency was evaluated by connection to a linear amplifier cascaded with a signal generator. The d and q -axis impedances were computed in the frequency range 1–15 kHz, deploying Fourier analysis. A LEM PR30 with 100 kHz bandwidth was used for the current measurement, while voltage was acquired with a SI-9002 differential probe (25 MHz bandwidth). The experimental results are compared to simulated and semi-analytically computed values in Fig. 10, showing satisfactory agreement.

IV. SENSORLESS MOTOR CONTROL

The motor was connected to a MACNO TDE OPD F32A three-phase inverter. The DC-bus voltage was reduced to 42 V (using a transformer) in order to protect the hardware from over-currents, yet enabling the intended low-speed tests. Inverter data can be found in Table III. The manufactured

Fig. 9. Rotor prototype a) Before assembly; b) Fully assembled rotor.

Fig. 10. Experimental evaluation using signal generator.

prototype was evaluated under d and q -axis locked-rotor inverter supply in Fig. 11, showing a saliency of $\frac{Z_q}{Z_d} = 1.45$ which is very close to the experimental results of Fig. 10. The

TABLE III
INVERTER DATA

Quantity	Symbol	Value
DC-bus voltage	V_{DC}	42 V
Switching frequency	f_{sw}	25 kHz
Dead-time	T_{d}	0.1 μs
Signal injection frequency	f_{inj}	3125 Hz
Current sensor range	I_{max}	$\pm 70 \text{ A}$

inverter was controlled by a dSpace 1104, deploying 3 external current sensors (LA 55-P) with 200 kHz bandwidth and 0.65 % accuracy at rated current. The prototype was assembled in a test bench and the rotor shaft was connected to an incremental encoder (Kübler 5882) with 2048 increments/revolution. The

Fig. 11. Inverter locked-rotor test at 3125 Hz ($\hat{U}_h = 20$ V).

test bench is shown in Fig. 12. The complete motor control

Fig. 12. The motor test-bench a) incremental encoder; b) self-sensing slotless motor.

algorithm shown in Fig. 13 is described in [19] and consists of a conventional cascaded PI-controller, controlling the stator currents and the rotational speed. The current and speed control loop bandwidths were set to 200 and 40 Hz, respectively.

Rotor position estimation is achieved by injection of a high-frequency sinusoidal voltage into the stator windings according to [36]. Using the estimated rotor reference frame d^x and q^x , the injected signals can be expressed as follows [19]:

$$u_{hd}^x = \hat{U}_h \cos(\omega_h t) \quad \text{and} \quad u_{hq}^x = 0. \quad (5)$$

The resulting high-frequency q -axis current is processed by the **Demodulator** into a DC-quantity (γ_{LPF} in Fig. 13), containing the angular information. The rotor position and speed are finally determined by the PI-observer block (**Estimator**), deploying feed-forward of the q -axis current.

Since a small rotor movement during startup is acceptable in the application under study, the rotor is simply aligned to pre-determined, DC current-vector. The injection signal frequency is preferably chosen as high as possible for achieving the highest saliency (as shown in Fig. 10), low settling time of estimation, and minimal frequency interaction between injected and fundamental signals [37]. However, the deployed hardware limits the switching frequency to 25 kHz and subsequently the signal frequency to 3125 Hz that is, to 8 current samples per signal period. The signal amplitude is set to 5-10 % of the nominal voltage, resulting in a signal current amplitude of about 3-7 % of the rated motor current.

A. Observer parameters

Three different observer parameter setups have been evaluated (**Observers A–C**, see Table IV). The results are presented

TABLE IV
OBSERVER DATA

	k_p	k_i	k_f
Observer A	5000	1200000	0
Observer B	15000	1200000	0
Observer C	15000	1200000	250000

in Fig. 14, where ω_{est} is the estimated speed, ω_{ref} is the reference speed and ω_{me} is the measured speed from the encoder. As can be seen, **Observer A** exhibits fast transient speed response and short settling time. However, **Observer A** also gives uprise to a steady-state second harmonic speed estimation error. To overcome this issue, **Observer B** is proposed, unfortunately also resulting in a significantly slower transient performance. As a compromise, **Observer C** is finally proposed, utilizing feed-forward of the low-pass filtered q -axis current to achieve fast transient response. **Observer C** exhibits the overall best performance and is therefore used in the following sections.

V. ESTIMATION ERROR ANALYSIS

Even with an ideal observer, the estimation error is still aggravated by imperfections in every stage of the drive line [38]–[41]. The main error sources and their resulting harmonic order of distortion are summarized in Table V. Conventional self-

TABLE V
HARMONIC IMPACT OF POSSIBLE ERROR SOURCES

		Harmonic order		
		1 st	2 nd	6 th
Inverter	Dead-time	-	-	x
	V_{CE} , V_{diode}	-	-	x
Current sensors	DC-offset	x	-	-
	Gain	-	x	-
Motor design	Winding	-	x	-
	Ring	-	-	-
	Rotor eccentricity	-	-	-

sensing slotted electrical machinery suffers from an increasing estimation error with load due to saturation effects. Therefore, the influence of saturation and cross-coupling on estimation error have gained considerable attention in recent years [8], [31], [42]. However, the simulations conducted in Section II demonstrate very limited impairment of position estimation in the machine under consideration due to cross-coupling between d and q -axis over a wide range of fundamental q -axis current (see Fig. 5), owing to its large relative air-gap. Within sensorless control of electric machinery, it is widely known that inverter non-idealities, such as dead-times, delays and voltage drops over power electronics components primarily contribute to the 6th harmonic estimation error [37]–[40]. Fig 15 shows the influence of dead-time T_d on estimation error in the machine under investigation at 1200 rpm.

As can be seen, longer dead-time T_d significantly increases the 6th harmonic of the estimated angle error while having negligible impact on any lower order harmonic amplitude. The results emphasize the importance of careful choice of dead-time, as well as considering implementation of adequate compensation algorithms. Unexpectedly, Fig. 15 also shows

Fig. 13. Control scheme.

Fig. 14. Experimental results for 1200 rpm step response using Observers A–C.

a significant second harmonic in the estimation error. It can be shown that an error in any of the current sensor offsets or gains would result in an estimation error of first and second

Fig. 15. Influence of dead-time T_d on estimation error at 1200 rpm using $\hat{U}_h = 15$ V.

order, respectively. However, analysis of the sum of currents $i_{\text{sum}} = i_a + i_b + i_c$ rejects both of these potential error sources (all three phase currents were measured). The influence of possible motor asymmetries due to non-ideal manufacturing have also been analyzed. The absence of stator slots in slotless machinery enables an asymmetric winding assembly, making it an important candidate for further evaluation. Another aspect that requires attention is the influence of an asymmetrically assembled rotor ring. The influence of rotor eccentricity also

deserves attention. FEA results indicate that neither rotor eccentricity nor an asymmetrically assembled rotor ring have any influence on the second harmonic. On the contrary, an asymmetrically assembled winding appears to be the only possible explanation to the second harmonic estimation error. Winding asymmetry may be efficiently compensated by using the following modification of the abc to $\alpha\beta$ transformation:

$$T = \begin{bmatrix} \sin(\varphi_a) & \sin(\varphi_b - 2\pi/3) & \sin(\varphi_c + 2\pi/3) \\ \cos(\varphi_a) & \cos(\varphi_b - 2\pi/3) & \cos(\varphi_c + 2\pi/3) \end{bmatrix} \quad (6)$$

where $\varphi_{a,b,c}$ are the displacements of phases a,b,c due to non-ideal manufacturing (ideally $\varphi_a = \varphi_b = \varphi_c = 0$).

A. Compensation of winding asymmetry

The influence of a promising compensation parameter setup identified by manual tuning ($\varphi_a = -1.1^\circ$; $\varphi_b = 0^\circ$; $\varphi_c = 1^\circ$) has been evaluated in Fig. 16. The compensation reduces the second harmonic estimation error with 80 % at 300 rpm.

Fig. 16. Experimental results for angular estimation error at 300 rpm using $\hat{U}_h = 15$ V: a) Time domain; b) Fourier domain.

B. Influence of torque load

The influence of torque load on estimation error has been evaluated at 600 rpm. Results are reported in Fig. 17, where $i_q = 8$ A corresponds to a high load. As can be seen, the investigated slotless machine exhibits negligible impair in position estimation at torque load (in contrast to conventional slotted electrical machinery). Moreover, the results in Fig. 17b validates the suggested compensation also under load.

C. Influence of signal amplitude

Using the suggested compensation, the influence of signal amplitude $\hat{U}_h$ on estimation error during braking (speed-step from 3000 to 0 rpm) has been investigated. The results are shown in Fig. 18. As can be seen, the transient estimation error during step response can be reduced by more than two thirds by increasing $\hat{U}_h$ from 5 to 20 V.

Fig. 17. Experimental results for angular estimation error during torque load at 600 rpm for $\hat{U}_h = 10$ V.

VI. PARASITIC EFFECTS DUE TO SIGNAL INJECTION

In this section, the heating and torque ripple due to signal injection are analyzed in sections VI-A and VI-B, respectively.

A. Power losses and thermal evaluation

The additional power losses due to a rotor ring can be subdivided into: a) the losses generated by the injection signal itself; b) the time-harmonic losses due to inverter switching; and c) the losses due to the space harmonics caused by the stator winding [32]. However, unlike slotted machines, the investigated slotless machine exhibits negligible levels of MMF space-harmonics [34]. Therefore, only power losses generated by the signal injection and time harmonics (a–b) are treated below.

a) *Signal injection losses*: The additional power losses due to signal injection have been evaluated using 2D FEA for different signal amplitude levels in the frequency range 1–15 kHz. The harmonic impedances, necessary to determine the current levels, were obtained with Eq. (3). Stator copper and rotor-ring losses have been considered, while iron losses have been neglected. The results are shown in Fig. 19. As expected, power loss minimization is not only favoured by a low signal amplitude, but also high signal frequency (due to the increasing skin-effect in the rotor ring). The stator power losses dominate until 1.5 kHz, whereafter rotor ring losses become the predominant loss source.

b) *Time-harmonic losses*: In this section, the additional time-harmonic motor losses due to presence of the rotor-ring are simulated for two different switching frequencies (f_{sw}): 25 and 50 kHz. The typical inverter comprises a 2-level three-phase inverter with a 325 V DC-bus, deploying symmetrical regular sampled PWM with third-harmonic injection of one-sixth of the fundamental magnitude. The typical no-load time-

Fig. 18. Experimental results for angular estimation error during braking using different signal amplitudes.

Fig. 19. Additional power losses due to signal injection from FEA-simulations.

harmonic voltage spectrum for $f_{sw} = 25$ kHz considering the first two harmonic sideband groups is shown in Fig. 20. The figure shows the stator reference frame ($\alpha\beta$) as well as the rotor reference frame (dq) voltage spectrum. Using [43], it can be demonstrated that the harmonic components during no-load interact to create mainly a pulsating q -axis voltage, subsequently creating a predominantly pulsating d -axis current. Utilizing Eq. (3), the current amplitudes can be approximated, and their resulting ring power losses can be evaluated using FEA. For a simplified loss analysis, the resulting rms currents of the two considered sideband groups in Fig. 20 are injected in the d -axis using 2D FEA at their respective multiple of switching frequency. The ring losses at

$f_{sw} = 25$ kHz and $f_{sw} = 50$ kHz are estimated to 3.1 W and 1 W, respectively.

Fig. 20. Phase-to-phase voltage spectrum at 30 krpm showing first two time-harmonic sideband groups for 25 kHz switching frequency.

c) *Thermal evaluation:* Utilizing the axi-symmetric thermal model for slotless motors in [44], which takes into account temperature dependency of materials, the impact of the estimated additional ring power losses have been evaluated. The rotor-ring was represented by a 0.1 mm copper layer at the rotor surface, thus maintaining the same axial cross-sectional ring area. The thermal impact of the time-harmonic and signal injection losses for a set of different configurations are summarized in Table VI. As can be seen, the choice of both the switching frequency and injection signal amplitude can have a significant impact on heating. However, the required experimental evaluation is out of the scope for this paper.

TABLE VI
RING HEATING CONTRIBUTION

$\hat{U}_h$ [V]	f_{sw} [kHz]	ΔT Rotor centre	ΔT Stator surface
20	25	+24%	+20%
10	25	+12%	+11%
10	50	+4%	+5%
5	50	+2%	+3%

B. Torque ripple

The output torque at rated load (T_{max}) has been simulated using time-stepping 2D FEA, assuming no estimation error. As can be seen in Table VII, the resulting torque oscillation is proportional to the signal amplitude, and in the range 0.3-1.6 % of the rated torque. Even though the torque ripple is efficiently filtered out by the rotor inertia, an audible noise remains detectable.

TABLE VII
TORQUE RIPPLE

$\hat{U}_h$ [V]	f_{inj} [Hz]	Torque ripple
20	3125	1.6%
10	3125	0.8%
5	3125	0.4%
5	6250	0.3%
0	-	0.1%

VII. CONCLUSION

This paper presents a 2-pole slotless motor with self-sensing capability. Effective saliencies of up to 1.8 are demonstrated. The manufactured prototype exhibits promising speed and position estimation performance. Utilizing the suggested compensation scheme for winding asymmetry, the steady state estimation error is within $\pm 2^\circ$. The proposed observer also performs properly during dynamic load changes. Fall-times are demonstrated from 3000 to 0 rpm in less than 50 ms with negligible overshoot. In contrast to its slotted counterparts, the investigated slotless machine exhibits negligible impair of position estimation with increased torque load. Also the ring losses and its thermal impact, as well as the torque ripple due to signal injection have been analyzed and successfully quantified. Future work includes tests at full DC-bus voltage and torque load, as well as experimental evaluation of the ring losses.

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