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Postprint

This is the accepted version of a paper presented at *2019 22nd International Conference on Electrical Machines and Systems (ICEMS)*.

Citation for the original published paper:

Bitsi, K., Wallmark, O., Bosga, S. (2019)

An Induction Machine with Wound Independently-Controlled Stator Coils

In: Harbin, China

<https://doi.org/10.1109/ICEMS.2019.8921779>

N.B. When citing this work, cite the original published paper.

Permanent link to this version:

<http://urn.kb.se/resolve?urn=urn:nbn:se:kth:diva-266350>

An Induction Machine with Wound Independently-Controlled Stator Coils

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Abstract—In this paper, a novel induction machine topology with wound, independently-controlled stator coils is presented. The introduced configuration of the stator-winding enables the individual energization and control of the coils in each stator slot. Therefore, the possibility of changing the number of poles and active phases in the stator winding during operation is explored in this study.

Keywords— Asynchronous rotor, independently-controlled stator coils, phase-changing capability, pole-changing capability

I. INTRODUCTION

The deployment of electric vehicles in the transportation sector is considered a viable solution towards reducing the global greenhouse-gas emissions and fossil-fuel consumption [1]. In order to produce highly efficient and economically feasible electrical drive systems, the selection of appropriate electrical-traction motor designs is a key step. Taking into account the disadvantages of the use of permanent magnets in synchronous motors (including high costs, dependence on rare-earth materials and problems with demagnetization), induction motors should always be considered as a reliable and cost-effective alternative.

An innovative induction machine topology is introduced in [2], where an asynchronous rotor is combined with a multi-phase stator cage winding. This winding is composed of solid aluminum bars in every stator slot, which are short-circuited at one axial end of the machine using an end-ring. Thus, each stator slot corresponds to a potentially separate phase winding and can be individually supplied. Consequently, the structure of the stator winding in this machine design facilitates the possibility of changing the number of pole-pairs as well as the number of active phases during operation [2]. However, although the usage of solid conductors in the stator cage offers the benefits of higher fill factor and potentially higher efficiency and better thermal behavior, this configuration results in very high current levels in the DC supply [3]. Therefore, the design of the converter and the battery configuration needs to be reconsidered, as such high current levels (in the order of 15 kA peak current [3]) are not compatible with the high-voltage battery systems that are currently available in the electric-vehicle market [4].

In order to decrease the high current levels rising in the aforementioned topology, the solid bars in each stator slot can be replaced with multi-turn coils that are wound axially along the stator core (Gramme-type winding [5]) and are short-circuited at one axial end. In this way, a machine design that is compatible with the present-day high-voltage battery system can be produced, while the useful capabilities of pole-changing and phase-changing of the topology of [2] can still be harnessed. Therefore, in this paper, this induction machine

with wound independently-controlled stator coils (WICSC) is investigated. The developed finite-element (FE) model is shown in Fig. 1a for the WICSC geometry of 36 stator slots (i.e. coils) and 26 rotor bars. The connection between the multi-turn stator coils ($C_{s,1}$ - $C_{s,36}$) that are wound axially around the stator core is visualized in a simplified schematic in Fig. 1b.

Fig. 1. (a) 2-D FE model of the WICSC machine and (b) simplified schematic showing the connection between the stator coils.

This paper is organized as follows: in Section II, the steady-state equivalent circuit of the topology is derived and analytical estimations of the current scaling are provided for implementing pole-changing and phase-changing during operation. Section III focuses on the development of the FE model of the WICSC machine and demonstrates its numerically estimated electromagnetic behavior in pole-changing and phase-changing states.

II. ANALYTICAL APPROACH

A. MMF from a Balanced m -Phase Stator Winding

In order to generalize the machine model regardless of the number of phases (odd or even), the stator current excitation is expressed based on a methodology that utilizes pole symmetry instead of stator-periphery symmetry (i.e. symmetric with respect to π instead of 2π). This concept is introduced as pole-symmetric fundamental winding configuration in [6], while it is also mentioned in [7] as conventional winding arrangement. For this study, the terminology used in [6] will be employed. Examples of converting the 3-phase and the asymmetrical 6-phase stator currents from the practical (customary) winding configuration to the fundamental winding configuration are shown in Fig. 2.

Based on the above assumptions, if a balanced, sinusoidal m -phase current excitation is considered in the stator winding, the current in phase n can be expressed as:

$$i_{s,n} = \hat{i}_s \cos \left(\omega_s t - (n-1) \frac{\pi}{m} \right) \quad (1)$$

where $\hat{i}_s$ is the stator current amplitude and ω_s the angular electrical frequency.

This research project is supported in part by the Swedish Energy Agency.

Fig. 2. Transformation from the practical winding configuration to the fundamental winding configuration for (a) a 3-phase current excitation and (b) an asymmetrical 6-phase current excitation.

Supposing that the coils of each phase are series connected, the resulting magnetomotive force (MMF) in the air-gap due to the stator current in phase n is found as:

$$F_{w,n} = \frac{n_s i_{s,n} Q_s}{\pi m} \sum_h \frac{k_h}{h} \sin \left(h \left(\theta - (n-1) \frac{2\pi}{m p} \right) \right) \quad (2)$$

where Q_s is the number of stator slots, n_s the number of winding conductors per stator slot, k_h the winding factor for harmonic h , and θ a mechanical angle spanning along the air-gap. Thus, the total MMF waveform in the air-gap due to the m -phase current excitation in the stator can be obtained as follows:

$$F_s = \sum_{n=1}^m F_{w,n} =$$

$$\frac{n_s i_s Q_s}{2\pi m} \sum_{n=1}^m \sum_h \frac{k_h}{h} \sin \left(h \theta + \omega_s t - (n-1) \frac{\pi}{m} \left(\frac{h}{p/2} + 1 \right) \right) +$$

$$\frac{n_s i_s Q_s}{2\pi m} \sum_{n=1}^m \sum_h \frac{k_h}{h} \sin \left(h \theta - \omega_s t - (n-1) \frac{\pi}{m} \left(\frac{h}{p/2} - 1 \right) \right) \quad (3)$$

B. MMF from the Induced Rotor-Bar Currents

The flux-density distribution in the air-gap caused by the stator excitation is assumed to induce sinusoidal currents in the rotor bars with the angular electrical frequency ω_{slip} . This frequency is linked with the electrical frequency ω_s through the slip, s of the machine: $\omega_{slip} = s\omega_s$. The rotor bar current in bar v can, hence, be expressed as:

$$i_{r,v} = \hat{i}_r \sin \left(\omega_{slip} t - (v-1) \frac{\pi p}{Q_r} \right) \quad (4)$$

where $\hat{i}_r$ is the rotor-bar current amplitude and Q_r the number of rotor bars.

It is evident from (4) that the rotor can be practically considered as a poly-phase winding that is composed of Q_r/p phases. Assuming an angle spanning along the air-gap fixed

to the rotor, θ_m , the resulting air-gap MMF distribution due to the rotor-bar current $i_{r,v}$ can be found as:

$$F_{b,v} = \frac{p i_{r,v}}{\pi} \sum_h \frac{1}{h} \sin \left(h \left(\theta_m - (v-1) \frac{2\pi}{Q_r} \right) \right) \quad (5)$$

Consequently, the total MMF waveform in the air-gap due to the Q_r/p -phase rotor-bar excitation can be obtained as follows:

$$F_r = \sum_{v=1}^{Q_r/p} F_{b,v} =$$

$$\frac{p i_{r,v}}{2\pi} \sum_{v=1}^{Q_r/p} \sum_h \frac{1}{h} \cos \left(h \theta_m - \omega_{slip} t - (v-1) \frac{\pi}{m} \left(\frac{h}{p/2} - 1 \right) \right) -$$

$$\frac{p i_{r,v}}{2\pi} \sum_{v=1}^{Q_r/p} \sum_h \frac{1}{h} \cos \left(h \theta_m + \omega_{slip} t - (v-1) \frac{\pi}{m} \left(\frac{h}{p/2} + 1 \right) \right) \quad (6)$$

C. Stator and Rotor Inductances

In order to derive the equivalent-circuit WICSC machine model, the coupling between a single stator phase to a single rotor bar is considered. The flux linkage per stator phase, ψ_s , can be defined as:

$$\psi_s = L_{self,s} i_{s,1} + L_{\lambda,s} i_{s,1} + L_{mut,s-r,1} i_{r,1} \quad (7)$$

where $L_{self,s}$ is the per stator-phase self inductance, $L_{\lambda,s}$ the per stator-phase leakage inductance and $L_{mut,s,1-r,1}$ the per stator-phase mutual inductance resulting in stator flux due to the rotor bar current.

In a similar way, the flux linkage per rotor bar, ψ_r , can be expressed as:

$$\psi_r = L_{self,r} i_{r,1} + L_{\lambda,r} i_{r,1} + L_{mut,r-s} i_{s,1} \quad (8)$$

where $L_{self,r}$ is the per rotor-bar self inductance, $L_{\lambda,r}$ the per rotor-bar leakage inductance and $L_{mut,r-s}$ the per rotor-bar mutual inductance resulting in rotor flux due to the stator current.

The air-gap flux density of the main harmonic ($h = p/2$) due to the stator currents, $B_{s,p/2}$ can be expressed as:

$$B_{s,p/2} = \frac{\mu_0 F_{s,p/2}}{\delta} = \frac{\mu_0 n_s \hat{i}_s Q_s k_{p/2}}{\pi p \delta} \sin \left(\frac{p}{2} \theta - \omega_s t \right) \quad (9)$$

where δ is the air-gap length of the WICSC machine.

In order to express the air-gap flux density of the main harmonic due to the rotor-bar excitation, $B_{r,p/2}$, an angle fixed to the stator should be used. This can be achieved by replacing in our analysis θ_m with $\theta - \omega_{mech} t$, where ω_{mech} is the angular mechanical frequency of the rotor:

$$\omega_{mech} = \frac{\omega_s - \omega_{slip}}{p/2} \quad (10)$$

Following the concept that was explained above, $B_{r,p/2}$ can now be expressed as:

$$B_{r,p/2} = \frac{\mu_0 F_{r,p/2}}{\delta} = \frac{\mu_0 \hat{i}_r Q_r}{\pi p \delta} \cos \left(\frac{p}{2} \theta - \omega_s t \right) \quad (11)$$

The stator and rotor inductances can be found based on the following equations [8]:

$$L_{\text{self},s} = \frac{L_a r_r}{i_{s,1}^2} \int_0^{2\pi} B_{s,p/2} F_{w,1,p/2} d\theta = \frac{2\mu_0 n_s^2 Q_s^2 L_a r_r k_{p/2}^2}{\pi \delta p^2 m} \quad (12)$$

$$L_{\text{self},r} = \frac{L_a r_r}{i_{s,1}^2} \int_0^{2\pi/p} B_{r,p/2} F_{b,1,p/2} d\theta = \frac{2\mu_0 Q_r L_a r_r}{\pi \delta p^2} \quad (13)$$

$$L_{\text{mut},s-r} = \frac{L_a r_r}{i_{s,1} i_r \sin(\omega_s t)} \int_0^{2\pi} B_{r,p/2} F_{w,1,p/2} d\theta = \frac{2\mu_0 n_s Q_s Q_r L_a r_r k_{p/2}}{\pi \delta p^2 m} \quad (14)$$

$$L_{\text{mut},r-s} = \frac{L_a r_r}{i_{s,1} i_r \sin(\omega_s t)} \int_0^{2\pi/p} B_{s,p/2} F_{b,1,p/2} d\theta = \frac{2\mu_0 n_s Q_s L_a r_r k_{p/2}}{\pi \delta p} \quad (15)$$

where L_a is the active axial length and r_r the rotor radius.

By utilizing the above stator and rotor inductances' equations, (7) and (8) can be finally re-expressed as:

$$\psi_s = L_{\lambda,s} i_{s,1} + L_m (i_{s,1} + i'_{r,1}) \quad (16)$$

$$\psi'_r = L'_{\lambda,r} i'_{r,1} + L_m (i_{s,1} + i'_{r,1}) \quad (17)$$

where L_m is the magnetizing inductance and together with $i'_{r,1}$, ψ'_r and $L'_{\lambda,r}$ are defined as follows:

$$L_m = \frac{2\mu_0 n_s^2 Q_s^2 L_a r_r k_{p/2}^2}{\pi \delta p^2 m} \quad (18)$$

$$i'_r = \frac{Q_r}{n_s Q_s k_{p/2}} i_r \quad (19)$$

$$\psi'_r = \frac{n_s Q_s k_{p/2}}{m} \psi_r \quad (20)$$

$$L'_{\lambda,r} = \frac{n_s^2 Q_s^2 k_{p/2}^2}{Q_r m} L_{\lambda,r} \quad (21)$$

D. Steady-state Equivalent-circuit WICSC Machine Model

With the help of (16) and (17), the stator-phase voltage phasor, $\bar{V}_s$, and the rotor-bar voltage phasor in the stator reference frame, $\bar{V}_r$, of the WICSC machine can be expressed using the following equations:

$$\bar{V}_s = R_s \bar{I}_{s,1} + j\omega_s L_{\lambda,s} \bar{I}_{s,1} + j\omega_s L_m (\bar{I}_{s,1} + \bar{I}'_{r,1}) \quad (22)$$

$$\bar{V}_r = \frac{n_s Q_s k_{p/2} R_{r,1}}{Q_r s} \bar{I}'_{r,1} + \frac{j\omega_{\text{slip}}}{s} \bar{\psi}'_{r,1} \frac{m}{n_s Q_s k_{p/2}} \quad (23)$$

where $\bar{I}_{s,1}$ and $\bar{I}'_{r,1}$ are the stator-phase and rotor-bar current phasors respectively, R_s the per-phase stator resistance and R_r the resistance of a single rotor bar (taking into account the corresponding short-circuit ring segments). It should be mentioned that the length of the end winding in the WICSC machine can be easily determined due to the simplicity of the used Gramme-type winding. Thus, contrary to the case of the conventional induction machine, R_s can be estimated with great accuracy by taking into account the axial-end effects.

Since the rotor bars are short circuited by the two short-circuit rings on each axial end of the rotor, $\bar{V}_r = 0$. If the following quantity is also defined:

$$R'_r = \frac{n_s^2 Q_s^2 k_{p/2}^2}{Q_r} R_r \quad (24)$$

the corresponding voltage equation for the rotor (23) can eventually be restated as:

$$0 = \frac{R'_r}{s} \bar{I}'_{r,1} + j\omega_s L'_{\lambda,r} \bar{I}'_{r,1} + j\omega_s L_m (\bar{I}_{s,1} + \bar{I}'_{r,1}) \quad (25)$$

The final equation in order to adequately describe the equivalent-circuit of the WICSC machine model, shown in Fig. 3, is the stator-impedance per phase $\bar{Z}_s$:

$$\bar{Z}_s = \frac{\bar{V}_s}{\bar{I}_{s,1}} = R_s + j\omega_s L_{\lambda,s} + \frac{j\omega_s L_m (\frac{R'_r}{s} + j\omega_s L'_{\lambda,r})}{\frac{R'_r}{s} + j\omega_s (L'_{\lambda,r} + L_m)} \quad (26)$$

Fig. 3. The steady-state equivalent-circuit of the WICSC machine model.

E. Derivation of Electromagnetic Torque

The resistance on the rotor side of the equivalent circuit can be divided in two terms as follows:

$$\frac{R'_r}{s} = R'_r + \left(\frac{1-s}{s}\right) R'_r \quad (27)$$

The second term of the above equation represents the mechanical power, P_{mech} , developed by the rotor on the shaft of the WICSC machine. With amplitude-invariant phasor scaling, P_{mech} can therefore be expressed as [9]:

$$P_{\text{mech}} = \frac{m}{2} \left(\frac{1-s}{s}\right) R'_r |\bar{I}'_{r,1}|^2 \quad (28)$$

Hence, the developed electromagnetic torque can be derived based on the previous analysis:

$$T_e = \frac{mp}{4} L_{m,1} \hat{i}_s^2 \frac{\frac{L_m \omega_{\text{slip}}}{R'_r}}{1 + \left(\frac{L_m \omega_{\text{slip}}}{R'_r}\right)^2} \quad (29)$$

F. Analytical Investigation of Pole-Changing and Phase-Changing Capability

In automotive applications, a wide speed-range operation is an important requirement. As already underlined in Section I, the topology of the WICSC machine exhibits the inherent ability of changing the number of poles and active phases during operation. Therefore, by properly adjusting the number of poles or/and phases at different operating speeds, optimum efficiency can be achieved over the entire torque-speed envelope [10].

This useful property will be exemplified for simplicity in the case of attaining the maximum electromagnetic torque,

$T_{e,max}$. According to (29), $T_{e,max}$ occurs when $L_m \omega_{slip} = R'_r$ and is, thus, given by the following equation:

$$T_{e,max} = \frac{\mu_0 n_s^2 Q_s^2 L_{ar} k_{p/2}^2}{2\pi \delta p} \hat{i}_s^2 \quad (30)$$

In the case of achieving the same $T_{e,max}$ with different pole numbers, p and p' in a specific WICSC machine topology, the required stator current amplitude should be scaled as follows:

$$\hat{i}_s' = \frac{n_s k_{p/2}}{n_s' k_{p/2}'} \sqrt{\frac{p'}{p}} \hat{i}_s \quad (31)$$

It is easily understood that for a specific WICSC machine topology, the number of turns n_s cannot be changed during operation.

For acquiring the same maximum torque capability from winding layouts with different number of phases, m and m' , the required stator current amplitude can be computed according to the formula:

$$\hat{i}_s' = \frac{n_s k_{p/2}}{n_s' k_{p/2}'} \hat{i}_s \quad (32)$$

III. NUMERICAL INVESTIGATION

A. Winding Layout

The electrical periodicity of the machine, t_{per} , is defined as the greatest common divisor of the number of stator slots, Q_s and the number of poles, p :

$$t_{per} = \text{GCD}(Q_s, p/2) \quad (33)$$

It can be deduced that t_{per} denotes also the number of winding symmetries.

In this analysis, the winding layout of the WICSC machine is decided with the aim to maximize the MMF main harmonic ($h = p/2$). The applied method is based on the principle of the star of slots, which corresponds to the phasor representation of the main electromagnetic (EMF) harmonic induced in the coils of each stator slot [11]. The diagram of the star of slots includes Q_s/t_{per} spokes, with each spoke consisting of t_{per} phasors, and thus, reduces the machine model to its equivalent 2-pole model by taking advantage of the per-pole-pair symmetry. A modification is introduced in the star-of-slots design rules in order to adapt to the pole-symmetric fundamental winding configuration: the angular displacement between the phases is considered π/m rather than $2\pi/m$. This technique is exemplified in Fig. 4 for a 12-slot, 20-pole and 3-phase (double-layer, fractional-slot concentrated) WICSC machine. The resulting star of slots is presented in Fig. 4a, where the numbers of the stator slots corresponding to each spoke are indicated with dark red color. Furthermore, the stator winding layout along the stator periphery and the produced MMF due to the stator winding configuration are shown in Fig. 4b and Fig. 4c respectively.

B. Numerical Verification of Pole-Changing and Phase-Changing Capability

In Section II-F, the required stator-current was estimated in order to attain the desired torque output while adjusting the

Fig. 4. (a) Star of slots, (b) winding layout and (c) stator MMF distribution of a 12-slot, 20-pole and 3-phase WICSC machine.

number of poles and active phases. That analysis needs to be validated numerically. For this purpose, the WICSC geometry of 36 stator slots and 26 rotor bars, which is depicted in Fig. 1a, was selected and analyzed. The electromagnetic analysis of the FE models is performed using the FE software COMSOL Multiphysics¹, while MATLAB² is used as a scripting tool for processing the COMSOL models.

At first, the model is simulated for a 3-phase stator winding layout with different number of poles: $p = 2$, $p = 4$ and $p = 6$, while $\hat{i}_s$ is properly adjusted according to (31) (the number of turns n_s is constant). The result of this analysis is shown in Fig. 5, where it is visible that there is a small deviation in the value of the maximum torque computed from the models with different pole number. This deviation can be deemed acceptable, considering the simplifying assumptions of the analytical approach.

Fig. 5. (a) Produced magnetic field density at the maximum torque and (b) output torque versus slip graph for different number of poles in a 36-slot, 3-phase WICSC machine.

In addition, the investigation with different number of phases is performed for 4-pole stator winding layout and $m = 3$ and $m = 6$. This time, $\hat{i}_s$ is scaled according to (32) depending on the calculated winding factors, $k_{p/2}$ of the different winding layouts. As it can be observed in Fig. 6, the

¹Comsol Multiphysics® is a registered trademark of Comsol AB, Stockholm, Sweden.

²MATLAB® is a registered trademark of The Mathworks Inc., Natick, Ma, USA.

two WICSC model produce identical maximum torque. The proper adjusting of phases while maintaining constant torque is, thus, achievable and (32) holds.

Fig. 6. Output torque versus slip graph for different number of phases, $m = 3$ and $m = 6$, in a 36-slot, 4-pole WICSC machine.

IV. CONVERTER OUTPUT VOLTAGE CONSIDERATIONS

In order to properly design and size a WICSC machine, it is crucial to consider also the design of the converter topology. Since each stator coil (i.e., coil corresponding to each stator slot) is controlled independently, it should be connected to an individual converter leg, as it is illustrated in Fig. 7.

Fig. 7. Schematic of an m -phase bridge-type voltage source inverter.

If each slot index is denoted with $\kappa = 1, 2, \dots, Q_s$, the converter output voltage $u_{\kappa,z}$ for slot κ is then:

$$u_{\kappa,z} = s_{\kappa} \frac{u_{dc}}{2} \quad (34)$$

where $s_{\kappa} = \pm 1$, assuming that the dc-link voltage of the converter is grounded in its midpoint z . Moreover, if $u_{\kappa,n}$ is the coil voltage in slot κ , the following expression holds:

$$-\sum_{\kappa=1}^{Q_s} u_{\kappa,z} + \sum_{\kappa=1}^{Q_s} u_{\kappa,n} + Q_s u_{n,z} = 0 \quad (35)$$

The assumption of a balanced current excitation entails that for each stator coil, there is a corresponding coil with equal voltage magnitude and opposite sign. Thus, the following can be derived:

$$\sum_{\kappa=1}^{Q_s} u_{\kappa,n} = 0 \quad (36)$$

$$\sum_{\kappa=1}^{Q_s} s_{\kappa} = 0 \quad (37)$$

From the above, (35) can be re-expressed as:

$$u_{n,z} = \frac{\sum_{\kappa=1}^{Q_s} u_{\kappa,z}}{Q_s} \quad (38)$$

As a result, the coil voltage in slot κ is:

$$u_{\kappa,n} = u_{\kappa,z} - u_{n,z} = \frac{u_{dc}}{2} \left(s_{\kappa} - \frac{\sum_{\kappa=1}^{Q_s} s_{\kappa}}{Q_s} \right) = s_{\kappa} \frac{u_{dc}}{2} \quad (39)$$

Hence, the coil voltage limit of a WICSC machine is:

$$|u_{\kappa,n}| \leq \frac{u_{dc}}{2} \quad (40)$$

V. CONCLUSION

This paper presents the geometry of an induction machine with wound independently-controlled stator coils. The steady-state equivalent-circuit of the WICSC machine model is derived, by expressing the stator excitation based on the pole-symmetric fundamental winding configuration. In addition, the useful capability of changing the number of poles and active phases in the WICSC machine during operation is investigated. The numerical analysis of WICSC FE models for utilizing the pole-changing and phase-changing ability provide results that harmonize considerably with the analytical approach.

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