

Article

Current Harmonics Controller for Reduction of Acoustic Noise, Vibrations and Torque Ripple Caused by Cogging Torque in PM Motors under FOC Operation

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Abstract: This article presents an effective algorithm to reduce acoustic noise, vibrations and torque ripple caused by cogging torque in three-phase Permanent Magnet (PM) motors under Field Oriented Control (FOC) operation. Cogging torque profile is suitably included into q -axis current reference, which must be then precisely tracked in order to mitigate acoustic noise, vibrations, torque ripple and speed ripple caused by cogging torque. Conventional FOC structure has been extended by a Current Harmonics Controller (CHC) to achieve precise control of dq current harmonics for all operation speeds, which is crucial to reduce impact of cogging torque and increase performance of electric drive with PM motor. Effectiveness of proposed control technique is experimentally verified by vibrations and acoustic noise measurements.

Keywords: cogging torque; acoustic noise; vibrations; torque ripple; current harmonics controller; cogging torque reduction; PM motor

1. Introduction

High-performance robotic applications, some automotive applications or home applications require smooth torque, low vibrations and quiet operation. Low-cost PM motors used in this kind of applications can significantly decrease operation quality and comfort of the users. However, price reduction in industry leads to more frequent use of low-cost PM motors with simple and low construction quality. Harmonic magnetic forces caused by cogging torque generate radial vibrations in the stator, which results in undesirable vibrations and acoustic noise in medium-low power PM motors [1–5]. Cogging torque is also the source of torque ripple and produces jerky rotation of the shaft at lower speeds [6]. A few strategies have been proposed in the past in order to minimize the cogging torque. Undesirable impact of the cogging torque on the performance of the PM motor can be suppressed by modifications in the motor construction [7] or by current injection [8,9].

Motor with higher construction quality allows for achieving better operation in terms of noise, vibrations and torque ripple. Most of the common techniques to suppress the cogging torque are associated with the skewing of the stator slots or magnet segments [10–12]. Magnet pole arc can have a huge impact on the cogging torque amplitude [10,13,14], as well as the fractional number of slots per pole or suitable magnet shifting [10,12,13]. Most significant harmonic component in the cogging torque profile may be attenuated through adding notches into the stator teeth [10]. However, these advanced design approaches are worthless in low-cost PM motors. Considerable torque ripple, vibrations and noisy operation are therefore often observed in these motors. These are the reasons why reduction of

the cogging torque by advanced control algorithms seems to be the cheapest and in many cases also the only feasible option to reduce the effects of the cogging torque.

Frequency of torque ripple, speed ripple, vibrations and acoustic noise caused by cogging torque is a few times higher than fundamental stator electrical frequency f_e . However, speed PI controller of conventional FOC has a strongly limited bandwidth and that causes significant amplitude attenuation and phase delay during regulation of alternating disturbance signals, like cogging torque. These are the reasons why conventional FOC is not able to reduce impact of cogging torque sufficiently even at lower speeds, where the effect of the cogging torque on the ripple in speed is most significant.

Most of the control strategies to minimize impact of the cogging torque by modifying control algorithm require cogging torque identification. Map of the cogging torque is usually added into control as a map of the q -axis current depending on the position of the rotor. Precisely tracked q -axis current reference ensures production of electromagnetic torque, which counteracts the cogging torque in order to ensure smoother operation of the motor [8,15,16]. However, this conventional FOC extension is possible to apply only at lower speeds because required current harmonics added into q -axis current reference usually have a few times higher frequency than f_e . Bandwidth of conventional current PI controllers can be too tight for such high frequencies of an input variable and then the compensation becomes insufficient [17] because a standard PI controller is designed to be effective with DC quantities, while producing delays and amplitude errors when elaborating alternating quantities. Frequency of added q -axis current reference to reduce impact of cogging torque depends on frequency of its harmonic components. Accurately tracked dq current trajectories are essential in order to reduce acoustic noise, vibrations and torque ripple of the motor and based on the frequency of tracked dq current trajectories can be required to use advanced control strategies.

Control strategies for identification and reduction of cogging torque in low-cost BLDC (BrushLess Direct Current) drives were presented in [8]. Authors added cogging torque map based on rotor position into the control structure and reached higher performance and lower torque ripple in robotic applications compared with BLDC motors with better quality and higher price. This approach was successful only at low-speed applications while authors in [18] designed very complex high-bandwidth current control based on a dead-beat controller, together with a self-commissioning method to identify cogging torque and shape of Back-EMF (ElectroMotive Force) in order to reduce their impact. Authors were able to achieve stable and precise regulation of current even at high frequencies with this dead-beat controller extended by current prediction. Another approach with repetitive current controller placed parallel to conventional current PI controller was investigated in [19] and successful rejection of periodic disturbances was achieved; however, this approach works well only with disturbances period and integral multiply of sampling time. Different extensions of conventional current PI controller were introduced in [20,21], where multiple reference frame approach was utilized for control of current harmonics in higher speeds and thus torque ripple reduction, while, in [22], authors used a resonant controller in cooperation with a standard PI controller (PIR controller), which increased a controller's ability to reduce periodic variations in control error, which has the same frequency as resonant frequency of a resonant controller. However, authors mentioned side effects of PIR controller during high dynamic operation. Another technique for torque ripple reduction with a single layer artificial neural network was introduced in [23,24]. This technique used only speed ripple or ripple in estimated torque to form compensation voltage. Despite of these great results, these methods are quite difficult and need significant computational time.

Many researchers in the past focused on the reduction of torque ripple in PM motors by modifying the motor design and also by different advanced algorithms mentioned above. Only few authors focused on the minimization of vibrations and acoustic noise, which are very important parameters for automotive or home applications, where the comfort of the user is in the first place [3,4,7,9]. Authors of this paper have focused on the identification of various sources of torque ripple in previous work [6]. Then, an adaptive algorithm has been applied to reduce torque ripple, speed ripple and acoustic noise in low-cost PM motors caused mainly by non-sinusoidal Back-EMF and by cogging torque

under FOC operation. The adaptive method was based on reduction of speed ripple harmonics. Considerable speed ripple must be visible for proper functionality of this adaptive method, which is a disadvantage [25]. Authors of this paper have also focused on the control strategies for the identification of cogging torque in PM motors and for reduction of speed ripple by tracking the cogging torque profile, where also limits of conventional method for cogging torque reduction are presented [17]. This article presents simple and effective modification of FOC structure in order to reduce detrimental effects of cogging torque like torque ripple, vibrations and acoustic noise. Based on the above-mentioned analyses and experiences, a novel Current Harmonics Controller (CHC) is presented. CHC has the capability to control dq current harmonic components, which is crucial in order to reduce impact of cogging torque. Precise tracking of cogging torque map suitably added into a q -axis current reference must be ensured. Successful experimental verification of proposed method is included in the article, where measurements of acoustic noise and vibrations are performed.

2. Mathematical Model of a Three-Phase PM Motor

Voltage equations of three-phase PM motor in dq frame are as follows [26]:

$$\begin{aligned} u_d &= R_s i_d + \frac{d\psi_d}{dt} - \omega_e \psi_q = R_s i_d + L_d \frac{di_d}{dt} - \omega_e L_q i_q, \\ u_q &= R_s i_q + \frac{d\psi_q}{dt} + \omega_e \psi_d = R_s i_q + L_q \frac{di_q}{dt} + \omega_e L_d i_d + \omega_e \Psi_{PM}, \end{aligned} \quad (1)$$

where u_d and u_q are dq voltages, R_s is stator resistance, L_d and L_q are dq inductances, i_d and i_q are dq currents, ψ_d and ψ_q represent dq linkage fluxes, ω_e is electric angular velocity and Ψ_{PM} is the flux of PM.

Figure 1 shows Back-EMF voltage of PM motor ACT 57BLF02, which is used in this paper. Back-EMF contains dominant 5th and 7th harmonic components in abc reference frame, which leads to dominant 6th harmonic component in dq reference frame. Dominant 5th and 7th harmonic components in Back-EMF voltage will lead to production of 6th harmonics component in dq currents, which must be suppressed. Non-sinusoidal Back-EMF also leads to 6th harmonics component in produced electromagnetic torque T_e . This must also be considered in the motor's model and that's why the rotational induced voltages in Equation (1) must be adjusted to involve non-sinusoidal shape of Back-EMF [27]:

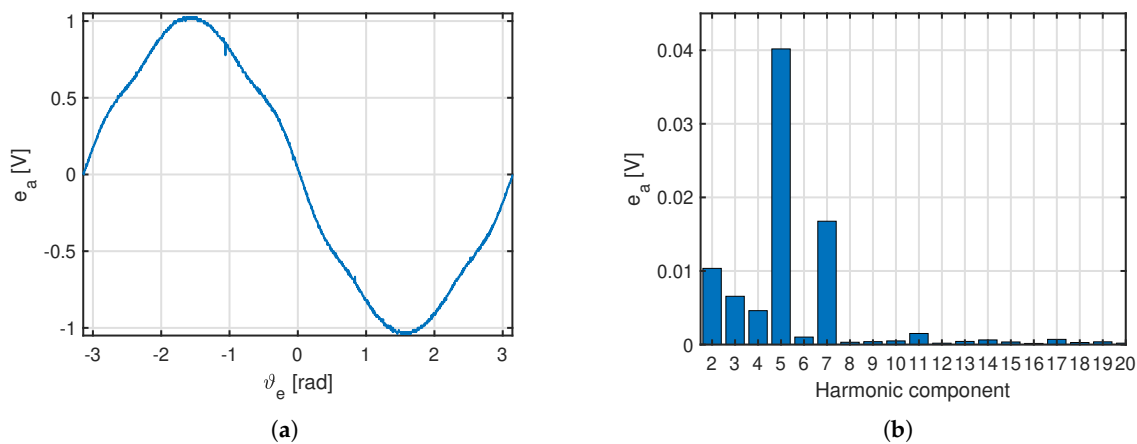


Figure 1. (a) Measured phase Back-EMF voltage, (b) FFT analysis of Back-EMF voltage, $n_r = 750$ rpm.

$$\omega_e \begin{bmatrix} -\psi_q \\ \psi_d \end{bmatrix} = \omega_e \begin{bmatrix} L_q \\ L_d \end{bmatrix} \begin{bmatrix} -i_q \\ i_d \end{bmatrix} + P(\vartheta_e) \begin{bmatrix} e_a \\ e_b \\ e_c \end{bmatrix} = \omega_e \begin{bmatrix} L_q \\ L_d \end{bmatrix} \begin{bmatrix} -i_q \\ i_d \end{bmatrix} + \omega_e K_e \begin{bmatrix} f_d(\vartheta_e) \\ f_q(\vartheta_e) \end{bmatrix}, \quad (2)$$

where e_a , e_b and e_c describe phase Back-EMF in abc reference frame, K_e is voltage constant, $f_d(\vartheta_e)$ and $f_q(\vartheta_e)$ describe shape of Back-EMF in dq reference frame depending on the electrical position of the rotor ϑ_e . Transformation matrix abc frame to dq frame is represented by $P(\vartheta_e)$:

$$P(\vartheta_e) = \frac{2}{3} \begin{bmatrix} \cos(\vartheta_e) & \cos(\vartheta_e - \frac{2\pi}{3}) & \cos(\vartheta_e + \frac{2\pi}{3}) \\ -\sin(\vartheta_e) & -\sin(\vartheta_e - \frac{2\pi}{3}) & -\sin(\vartheta_e + \frac{2\pi}{3}) \end{bmatrix} \quad (3)$$

Voltage equations of Permanent Magnet Synchronous Motor (PMSM) in dq reference frame (1) which include non-sinusoidal rotational induced voltages (2) will be adjusted as follows [28]:

$$\begin{aligned} u_d &= R_s i_d + L_d \frac{di_d}{dt} - \omega_e L_q i_q + \omega_e K_e f_d(\vartheta_e) \\ u_q &= R_s i_q + L_q \frac{di_q}{dt} + \omega_e L_d i_d + \omega_e K_e f_q(\vartheta_e) \end{aligned} \quad (4)$$

Mechanical equations of three-phase PM motor are as follows [26]:

$$\frac{d\omega_r}{dt} = \frac{1}{J} (T_e - T_{load} - T_{cogg}) \quad \frac{d\vartheta_r}{dt} = \omega_r \quad (5)$$

where J is inertia, T_e is electromagnetic torque, T_{load} is load torque, T_{cogg} is cogging torque and ϑ_r is the mechanical angular position of the rotor. Electromagnetic torque T_e for three-phase PM motor with sinusoidal Back-EMF voltages is described as follows:

$$T_e = \frac{3}{2} p (\psi_d i_q - \psi_q i_d) = \frac{3}{2} p (\psi_{PM} i_q + (L_d - L_q) i_d i_q) \quad (6)$$

where $\psi_{PM} i_q$ is proportional to synchronous torque, while $(L_d - L_q) i_d i_q$ is proportional to reluctance torque, whereas T_e for three-phase PM motor with non-sinusoidal waveform of Back-EMF voltages respecting their shape functions is described as follows:

$$T_e = \frac{e_a i_a + e_b i_b + e_c i_c}{\omega_r} = \frac{3}{2} p (K_e f_d(\vartheta_e) i_d + K_e f_q(\vartheta_e) i_q + (L_d - L_q) i_d i_q) \quad (7)$$

Any misalignment between stator currents and Back-EMF voltages leads to ripple in produced electromagnetic torque T_e according to Equation (7), which is also shown in Figure 2.

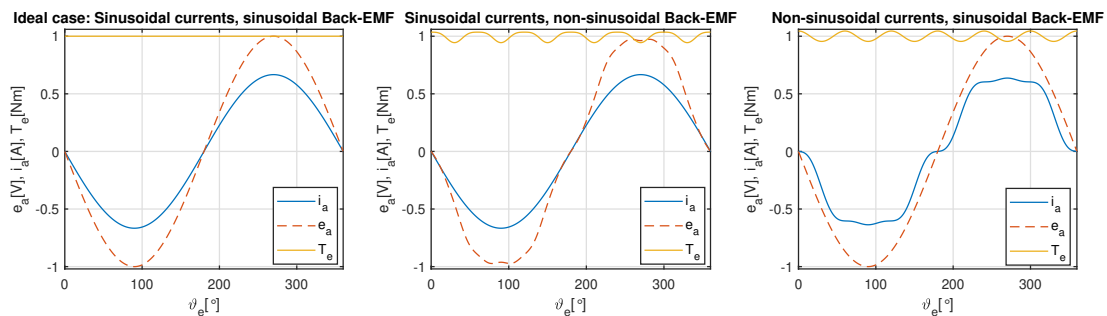


Figure 2. Ripple in produced T_e based on misalignment between stator currents and Back-EMF voltages.

Mechanical torque on the shaft is expressed as:

$$T_{mech} = T_e - T_{cogg} \quad (8)$$

Figure 3 shows fundamentals of torque ripple in PMSM controlled by FOC based on equations above. Stator currents i_{abc} can be distorted by harmonic components in Back-EMF e_{abc} and by inverter output voltages u_{abc}^* distorted by dead-time. Misalignment between stator currents and Back-EMF

voltages leads to ripple in electromagnetic torque T_e . Misalignment between T_e , T_{load} and cogging torque T_{cogg} leads to ripple in produced mechanical torque. This paper is focused on torque ripple caused by cogging torque and its impact on acoustic noise and vibrations.

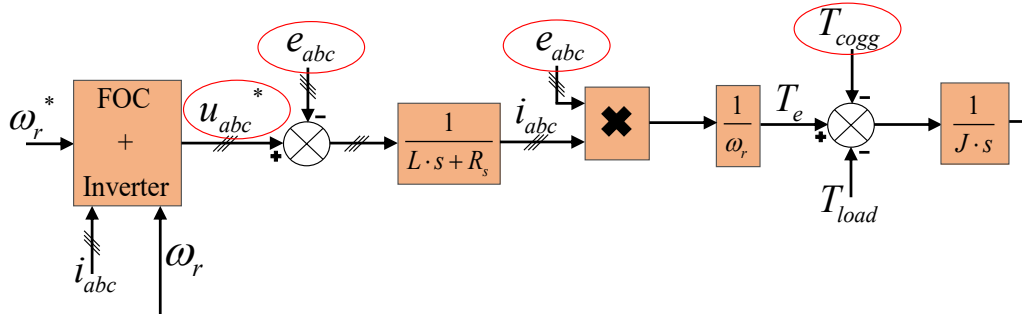


Figure 3. Fundamentals of torque ripple in PMSM controlled by FOC.

Cogging Torque

Cogging torque represents the interaction between magnets on the rotor and stator poles or teeth without electrical excitation. Therefore, it is independent of any stator current and yields to torque ripple and thus to ripple in the rotor speed. This parasitic torque is also the source of vibro-acoustics issues caused by harmonic magnetic forces, which generate radial vibrations in the stator and result in undesirable acoustic noise [1,2,5,29]. Moment of inertia has a filtering impact on the ripple in mechanical speed. Cogging torque can be mathematically described as follows:

$$T_{cogg} = -\frac{1}{2}\phi_\delta^2 \frac{dR}{d\theta_r} \quad (9)$$

where ϕ_δ describes magnet's flux crossing the air-gap and R determines overall reluctance through which the flux flows. When the overall reluctance R does not differ with rotation of the rotor, the derivation in Equation (9) is equal to zero, which leads to zero cogging torque [13,30,31].

Cogging torque can be mathematically represented by Fourier series as follows:

$$T_{cogg} = \sum_{k=1}^{\infty} \left(T_{ksin} \sin(kLCM[Q, 2p]\theta_r) + T_{kcos} \cos(kLCM[Q, 2p]\theta_r) \right) \quad (10)$$

T_{ksin} and T_{kcos} determine the amplitude of k -th harmonic element of sine and cosine components. $LCM[Q, 2p]$ describes the least common multiple of the poles number $2p$ and the slots number Q [13].

Figure 4 shows measured cogging torque profile for one electrical revolution and its FFT analysis at no-load. Measured cogging torque profile has dominant 2nd and 6th harmonic component. Cogging torque can be identified by various methods, but in this paper was extracted from q -axis current during slow rotation of the rotor under FOC operation [17]. Identified cogging torque map can be added into the q -axis current reference and then the impact of cogging torque can be greatly reduced if precise tracking of the q -axis current reference is ensured.

Figure 5 shows almost identical cogging torque profile for one electrical revolution and its FFT analysis measured at constant load (45% of nominal load).

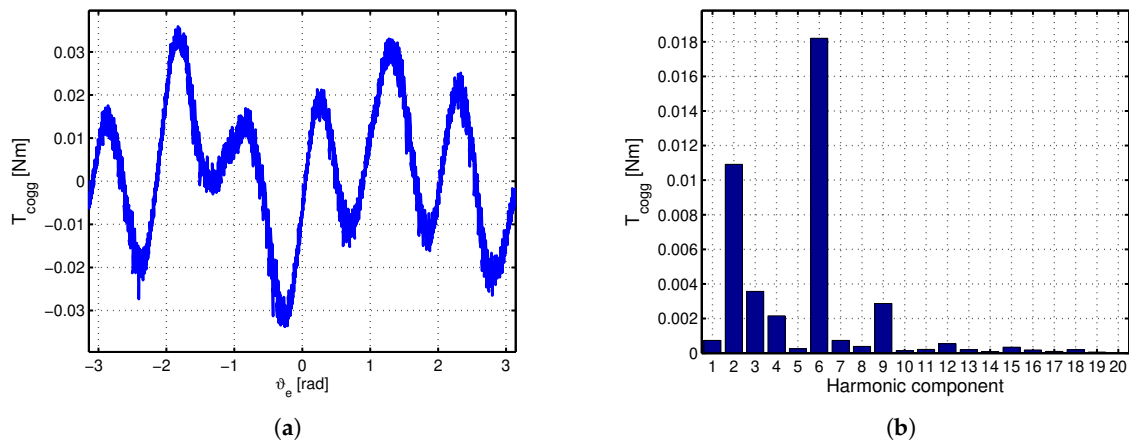


Figure 4. (a) measured cogging torque at no-load condition; (b) FFT analysis of measured cogging torque.

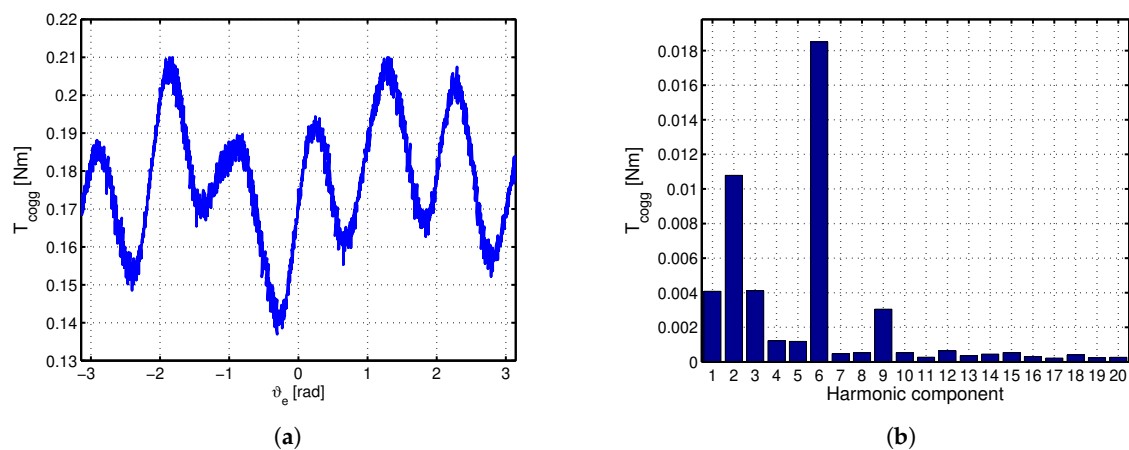


Figure 5. (a) Measured cogging torque at 45% of rated load (b) FFT analysis of measured cogging torque.

3. Experimental Setup

Figure 6 shows the experimental setup, which was used for all experiments to get the results shown below. Setup consists of three-phase PM motor ACT 57BLF02 (parameters in Table 1) with low-resolution incremental encoder QEDS-5886 (512 ppr), control board containing microcontroller NXP MPC5643L where the FOC has been implemented [32], NXP three-phase low voltage power inverter (Figure 7), 12 V power source, piezoelectric element to measure vibrations. Piezoelectric element is coupled with oscilloscope to measure its voltage output. Laptop with microphone and program Spectroid is used to measure acoustic noise. Real-time visualization of control variables (currents, speed, position, ...) handled by the control unit are displayed through the program NXP FreeMASTER.

Figure 8 shows variables and functions implemented in FOC calculation for MPC5643L using Automotive Math and Motor Control Library (AMMCLib). More details about implemented FOC are in [32].

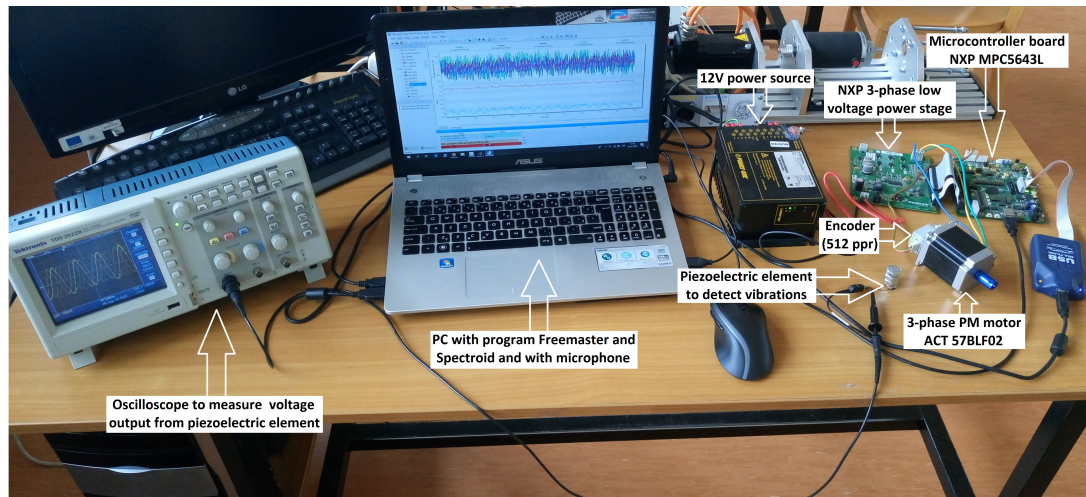


Figure 6. Used experimental setup.

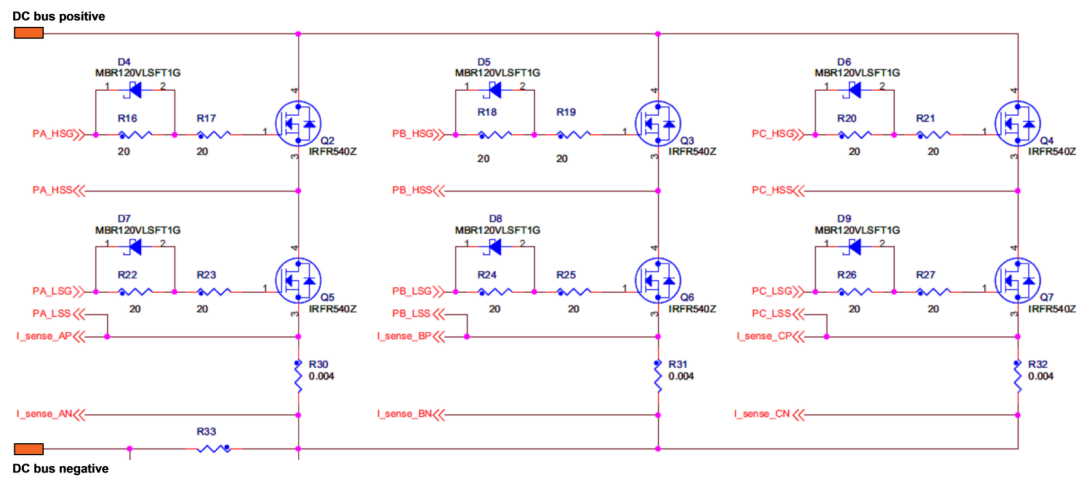


Figure 7. NXP three-phase low voltage power inverter with shunt resistors for current measurement [32].

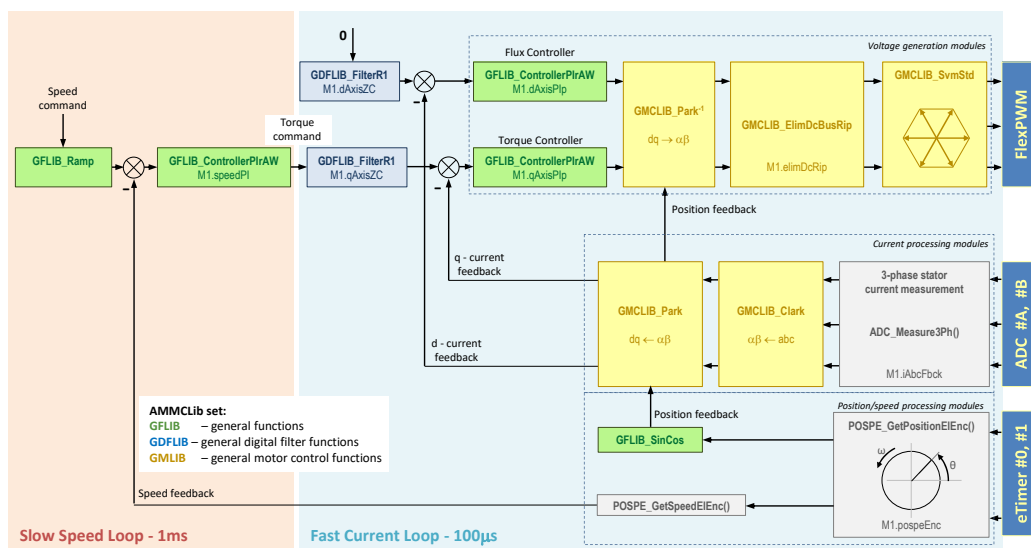


Figure 8. Variables and functions as implemented in FOC calculation for MPC5643L [32].

to insufficient reduction and the torque ripple, acoustic noise and vibrations. A suitable modification of current control structure depicted in Figure 9 is thus required to reduce torque ripple, acoustic noise and vibrations caused by 6th harmonic component of the cogging torque in the whole speed range. 2nd harmonic of the cogging torque is sufficiently compensated only by current PI controller.

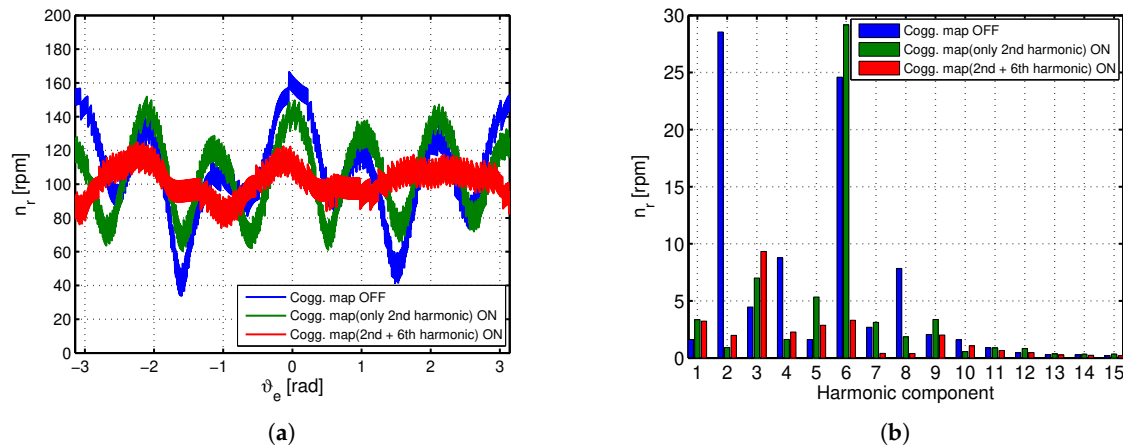


Figure 10. (a) speed ripple; (b) FFT of speed waveform, required $n_r = 100$ rpm.

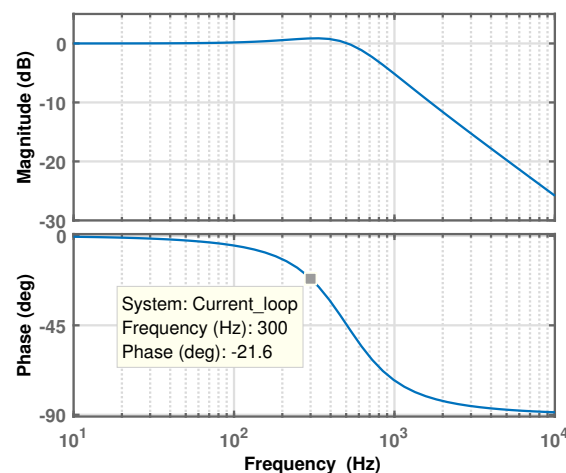


Figure 11. Bode diagram of closed current loop with PI controller.

4.2. Proposed Current Harmonics Controller

Precise dq currents trajectory tracking without amplitude attenuation or phase shift must be ensured for the effective compensation of the cogging torque and non-sinusoidal Back-EMF voltage. This can be a challenge because dq current controllers have to control signals with frequency a few times higher than fundamental frequency f_e . For compensation of the non-sinusoidal Back-EMF, it is mainly 6-multiply of f_e . For compensation of the cogging torque for used PM motor these are mainly 2 and 6-multiply of f_e , but challenging is mainly 6-multiply of f_e . In order to overcome the limits of standard PI controller based FOC, a modified current control structure is described here, including a Current Harmonic Controller (CHC), which is able to precisely control dq current harmonics for reducing influence of cogging torque or non-sinusoidal Back-EMF. The most effective approach for used PM machine in terms of performance of CHC vs. computational time of CHC is to extend PI controller with novel CHC structure only for 6th harmonic component.

Current PI controllers in FOC structure are extended by proposed CHC algorithm, which is placed parallel to dq currents PI controllers. Figure 12 shows extension of standard FOC, where CHC blocks

(CHC_{d6} and CHC_{q6}) in both current loops calculate required injected 6th harmonic d -axis or q -axis voltage. Injection of 6th dq voltages harmonic with correct phase and amplitude results in minimizing error between reference and feedback currents.

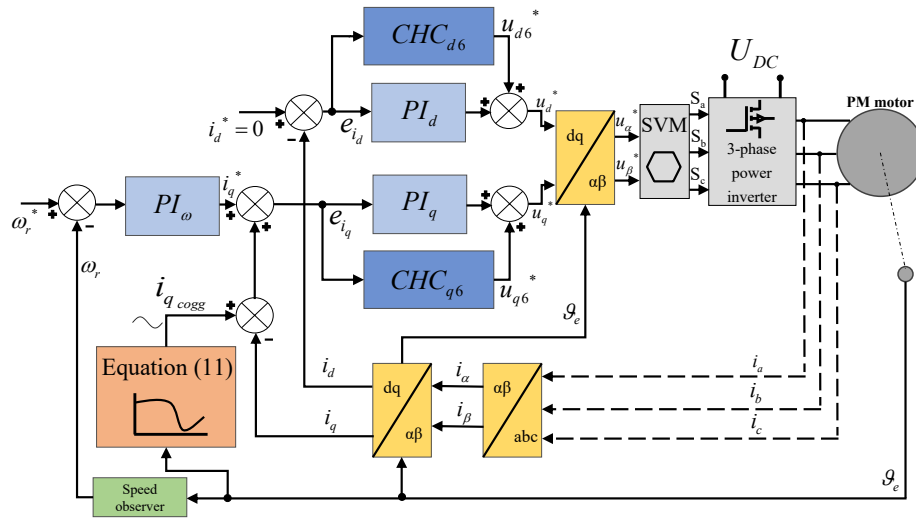


Figure 12. Proposed modification of FOC.

Shape of these injected voltages in dq frame is modified online, based on the error in dq currents to achieve precise control of 6th dq current harmonic. It results in correct dq currents harmonic, which counteracts the cogging torque and non-sinusoidal Back-EMF.

The input of this compensation algorithm is error of dq currents and ϑ_e . Amplitude x_{q6} and phase ϕ of compensation voltage in q -axis are adapted during operation, based on error in q -axis current. Compensation voltage is described as follows:

$$u_{q6}^* = x_{q6} \sin(6\vartheta_e + \phi) = x_{q6} [\sin(6\vartheta_e) \cos(\phi) + \sin(\phi) \cos(6\vartheta_e)] \quad (12)$$

Voltage u_{q6}^* is composed from cosine and sine components:

$$u_{q6}^* = x_{q6} \sin(6\vartheta_e + \phi) = a_{q6\sin} \sin(6\vartheta_e) + a_{q6\cos} \cos(6\vartheta_e) \quad (13)$$

where $a_{q6\sin}$ is amplitude coefficient of sine part and $a_{q6\cos}$ is amplitude coefficient of cosine part for q -axis. Variations in these amplitude coefficients result in modification of injected q voltage. Amplitude coefficients $a_{q6\sin}$ and $a_{q6\cos}$ are DC values in steady-state.

Integrators with constant gain k_{CHC} are used for calculation of these amplitude coefficients to reach minimal error of 6th harmonic in q -axis current:

$$\begin{aligned} a_{q6\sin} &= e_{q\sin} \frac{k_{CHC}}{s} \\ a_{q6\cos} &= e_{q\cos} \frac{k_{CHC}}{s} \end{aligned} \quad (14)$$

Integrator behaves as a low-pass filter with the time constant, which is proportional to $\frac{1}{k_{CHC}}$. In this paper, $k_{CHC} = 10$ is chosen. This means that 63.3% of convergence to steady-state values of amplitude coefficients is done in 0.1 s. Input errors into these integrators are calculated as follows:

$$\begin{aligned} e_{q\sin} &= e_{i_q} \sin(6\vartheta_e) \\ e_{q\cos} &= e_{i_q} \cos(6\vartheta_e) \end{aligned} \quad (15)$$

where e_{iq} describes actual error in q -axis current. After discrete transformation of Equations (14) by using Backward-Euler discrete substitution $s \approx \frac{z-1}{T_s z}$ amplitude coefficients in k -th step are calculated in microcontroller as follows:

$$\begin{aligned} a_{q6\sin}(k) &= a_{q6\sin}(k-1) + e_{q\sin}(k)k_{CHC}T_s \\ a_{q6\cos}(k) &= a_{q6\cos}(k-1) + e_{q\cos}(k)k_{CHC}T_s \end{aligned} \quad (16)$$

where k represents step number, $e_{q\sin}(k)$ represents input error in k -th step and T_s is a sampling time of current loop, $T_s = 0.0001$ s. Figure 13 shows block scheme of the algorithm for determination of q -axis compensation voltage harmonic. The same procedure applies to d -axis current harmonics controller.

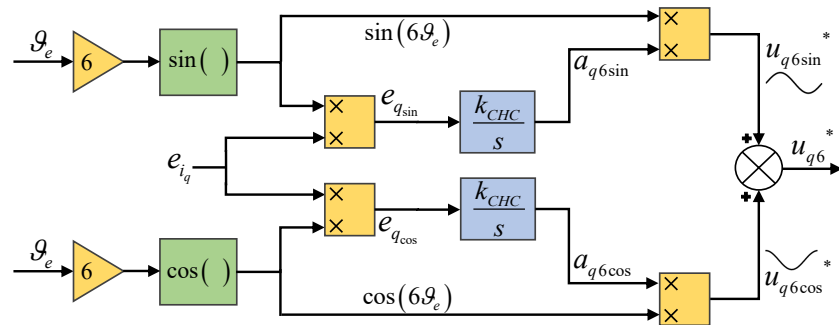


Figure 13. Current harmonic controller for calculation of compensation q -axis voltage.

In case that CHC is applied like in Figure 12 without applied cogging torque map added into required q -axis current, CHC rejects only the impact of non-sinusoidal Back-EMF and dead-time. Figure 14 shows abc currents and dq currents significantly distorted by harmonics of non-sinusoidal Back-EMF and by dead-time. The sixth harmonic component in dq currents is significantly reduced, when CHC is applied (CHC ON). This leads to significant reduction of dominant 5th and 7th harmonic components in abc currents. Figure 15 shows transient state of amplitude coefficients and continuous reduction of 6th harmonics in dq currents after turning on the CHC algorithm.

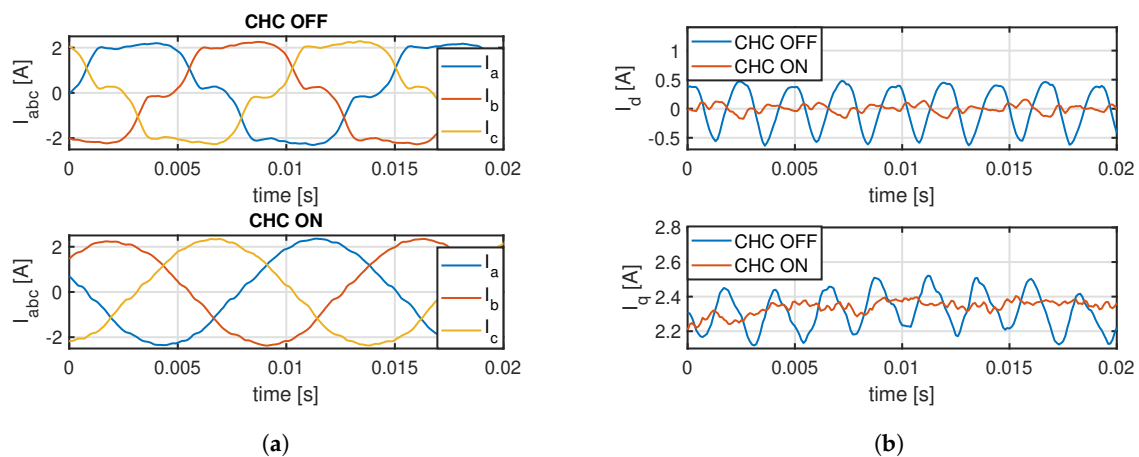


Figure 14. (a) abc (b) dq currents after application of CHC.

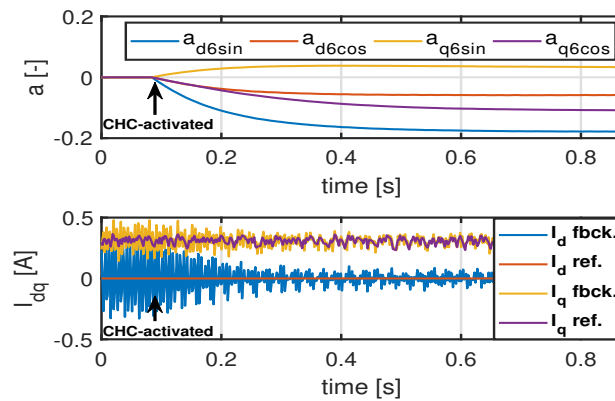


Figure 15. Transient state of amplitude coefficients and dq currents after turning on CHC.

5. Results of Cogging Torque Reduction

The described method to reduce acoustic noise, vibrations and torque ripple connected with the 2nd and 6th harmonic of the cogging torque was applied and verified on an experimental setup, which is in Figure 6 with motor ACT 57BLF02.

5.1. Torque Ripple Reduction

The effectiveness of the proposed approach as shown in Figure 12 has been experimentally validated on a 125 W PMSM drive. Significant reduction of torque ripple is achieved compared with application of only a cogging torque map (without using CHC) or without any application of cogging torque map. Figure 16 shows this comparison at $f_e = 66.7$ Hz. CHC significantly reduces torque ripple. Figure 17 shows ratio between torque ripple T_{ripple} and rated torque T_N for wider speed range. Torque ripple T_{ripple} is a maximal amplitude of ripple in mechanical torque.

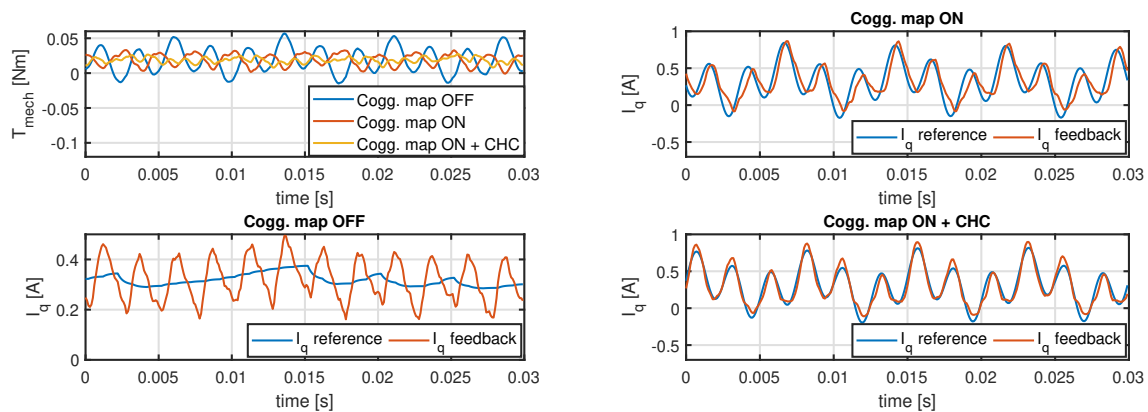


Figure 16. Torque ripple at $n_r = 1000$ rpm and $f_e = 66.7$ Hz.

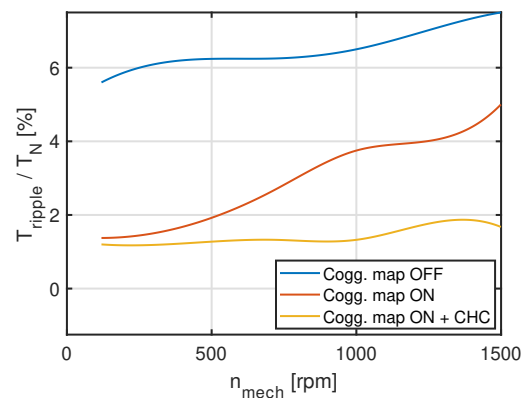


Figure 17. Peak value of torque ripple vs. mechanical speed.

5.2. Acoustic Noise and Vibrations Reduction

The Spectroid program was used to evaluate the produced acoustic noise, which performs FFT analysis from the signal from the microphone and visualizes the measured data as a spectrogram. Motor was located at a distance 70 cm from the PC microphone. The lower part of Figures 18, 20, 22, 23, 25 and 27 shows the sound FFT analysis over time as a spectrogram (FFT analysis is performed every 20 ms during the time range of the spectrograph, which is 9 s). The time axis is placed on the right side of the spectrograms. Color determines noise level in *dB* based on the color-scale placed on the left side of the spectrograms. The upper part of the figures with the yellow curve shows the last FFT analysis of measured noise.

Figures 18 and 19 show motor performance during start-up. The motor is controlled by the standard FOC algorithm. Speed ω_e started from 0 to $640 \frac{\text{rad}}{\text{s}}$ ($f_e = 102 \text{ Hz}$; mechanical speed $n = 1529 \text{ rpm}$). The white curve in all spectrogram figures below indicates fundamental frequency f_e curve. The full spectrum of acoustic noise is described by grid of curves ($2f_e, 3f_e, \dots$) which are an integer multiple of f_e . Figure 18 shows that the most visible noise curves are $2f_e$ and $6f_e$.

Figures 20 and 21 show the start-up sequence with applied cogging torque map (2nd + 6th harmonic) and CHC algorithm for 6th harmonic. Figure 20 shows that significant reduction of acoustic noise connected with 2 and 6-multiply of f_e ($2f_e$ and $6f_e$ curves) was achieved. In addition, $4f_e$ and $8f_e$ noise curves were greatly reduced.

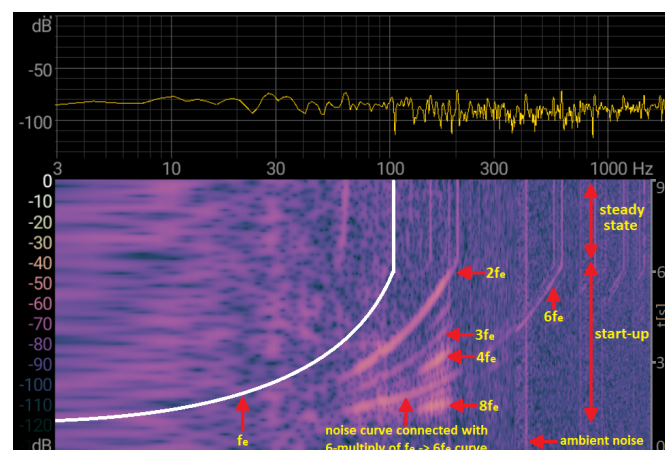


Figure 18. Sound spectrogram measured during start-up (ω_e from 0 to $640 \frac{\text{rad}}{\text{s}}$). Cogging map is not applied. Speed ramp $300 \frac{\text{rpm}}{\text{s}}$ and no-load condition.

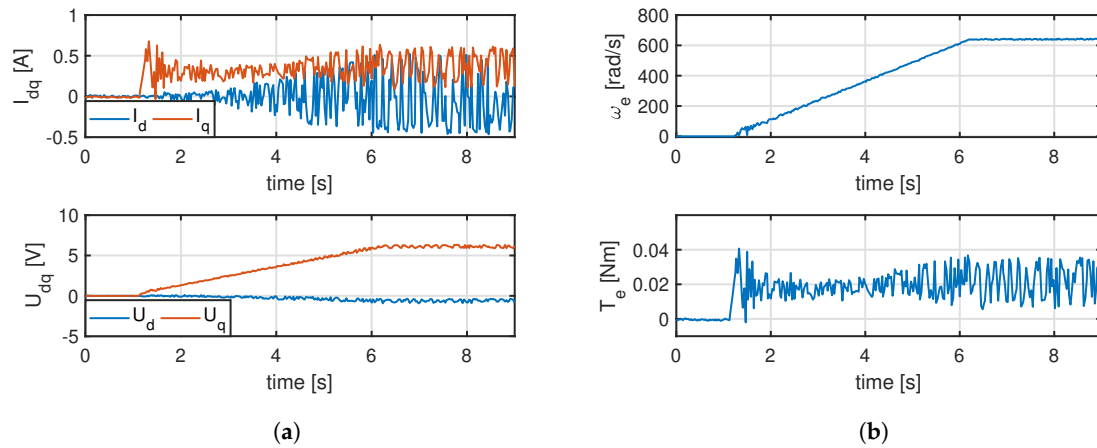


Figure 19. (a) dq currents and voltages; (b) ω_e and T_e during start-up sequence of Figure 18.

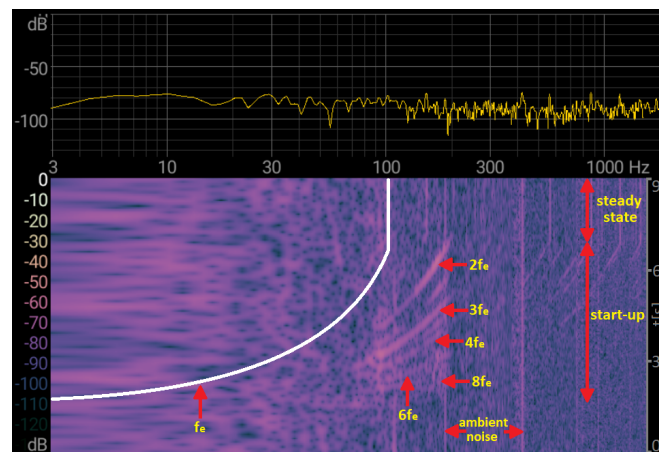


Figure 20. Sound spectrogram measured during start-up (ω_e from 0 to $640 \frac{\text{rad}}{\text{s}}$). Cogging torque map (2nd + 6th harmonic) + CHC are applied. Speed ramp $300 \frac{\text{rpm}}{\text{s}}$ and no-load condition.

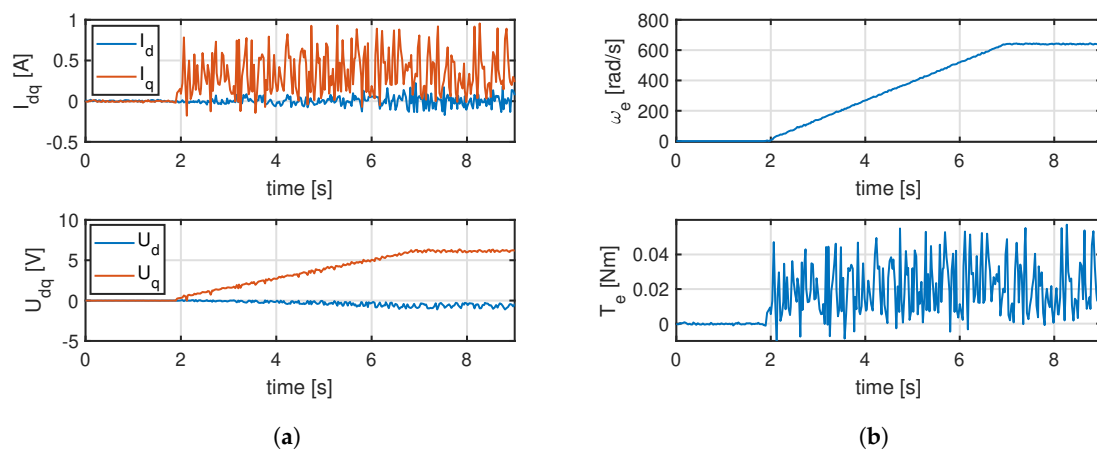


Figure 21. (a) dq currents and voltages; (b) ω_e and T_e during start-up sequence of Figure 20.

Figure 22 shows acoustic noise at steady-state case at $\omega_e = 300 \frac{\text{rad}}{\text{s}}$. Green arrows indicate adding 2nd and 6th harmonic of cogging torque map into q -axis current and turning on CHC algorithm.

After that, acoustic noise connected with 2, 4 and 6-multiply of f_e are significantly reduced (at least -10 dB).

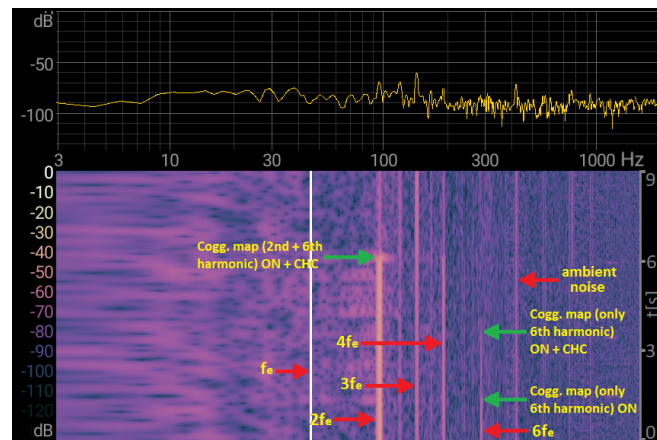


Figure 22. Sound FFT analysis at steady-state $\omega_e = 300 \frac{\text{rad}}{\text{s}}$ with applied cogging torque map and CHC algorithm. No-load condition.

Figures 23–27 show proper functionality of the proposed approach even at approximately 50% of nominal load. Noise curves in frequency under f_e (e.g., $0.5f_e$) are caused by non-ideal coupling with load.

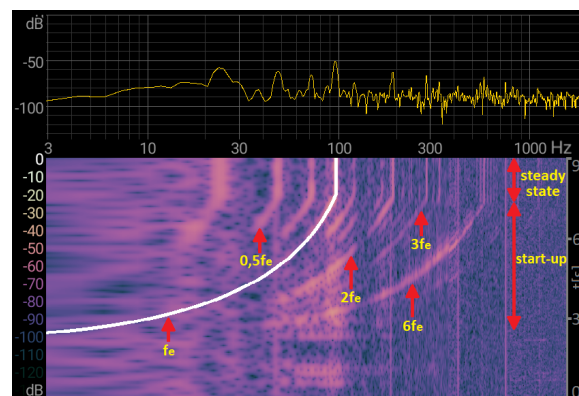


Figure 23. Sound spectrogram measured during start-up (ω_e from 0 to $600 \frac{\text{rad}}{\text{s}}$). Cogging map is not applied. Speed ramp $300 \frac{\text{rpm}}{\text{s}}$ and approx. 50% of rated load.

Figure 28 shows measurement of acoustic noise and vibrations for $2f_e$ curve and $6f_e$ curve vs. their own frequency for these three cases: 1. cogging torque map OFF (blue and yellow), 2. only cogging torque map (2nd and 6th harmonic) ON (red and purple), 3. cogging torque map ON + CHC (green). Measurement of acoustic noise and vibrations was performed in steady state for a more speed points. Application of only cogging torque map reduces acoustic noise and vibrations significantly only at lower speed. At higher speeds, the reduction of noise and vibrations is negligible. Figure 28 shows that noise and vibrations waveforms excited by 2nd harmonic of the cogging torque are the same as waveforms excited by 6th harmonic of the cogging torque. Figure 29 shows acoustic noise and vibrations for $2f_e$ curve and $6f_e$ curve vs. mechanical speed. Results shown in Figures 28 and 29 and in Tables 2 and 3 are measured during no-load condition.

For evaluation of produced vibrations, a piezoelectric element was used, which generates voltage based on detected vibrations. Output voltage was measured and processed on an oscilloscope. The piezoelectric element was placed on a desk in distance of 10 cm from the motor.

Figures 30–32 shows acoustic noise and vibrations at steady-state case at $\omega_e = 300 \frac{\text{rad}}{\text{s}}$ and no-load condition. Figure 30 shows measurements when adding 2nd and 6th harmonic of cogging torque map

into q current is not applied and CHC is turned off. Figure 31 shows measurements when adding only 2nd harmonic of cogging torque map into q current, while CHC is turned off. Figure 32 shows measurements when 2nd and 6th harmonic of cogging torque profile is added into q current and CHC is turned on. Acoustic noise and vibrations connected with 2, 4, 8 and 6-multiply of f_e are significantly reduced (at least -10 dB) in Figure 32.

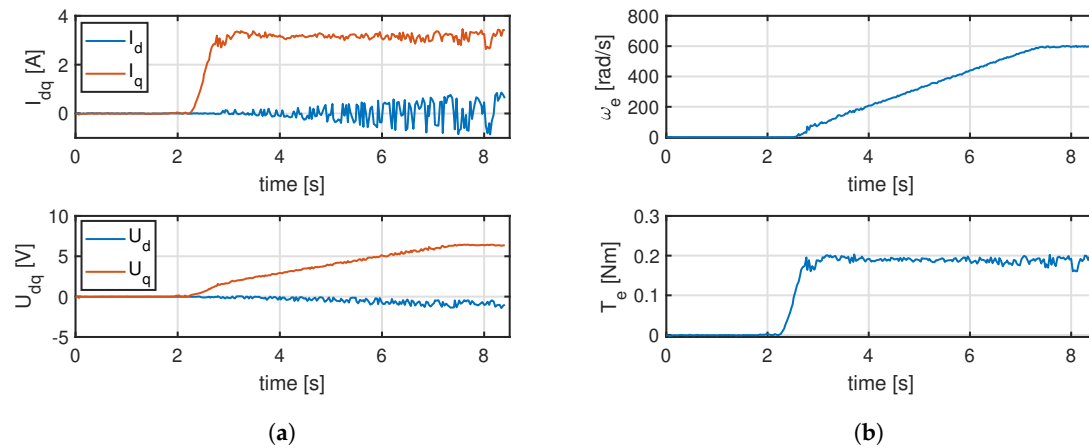


Figure 24. (a) dq currents and voltages; (b) ω_e and T_e during start-up sequence of Figure 23.

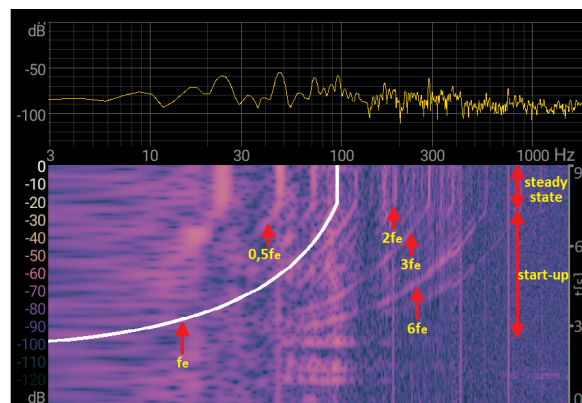


Figure 25. Sound spectrogram measured during start-up (ω_e from 0 to 600 $\frac{\text{rad}}{\text{s}}$). Cogging torque map (2nd + 6th harmonic) + CHC are applied. Speed ramp 300 $\frac{\text{rpm}}{\text{s}}$ and approx. 50% of rated load.

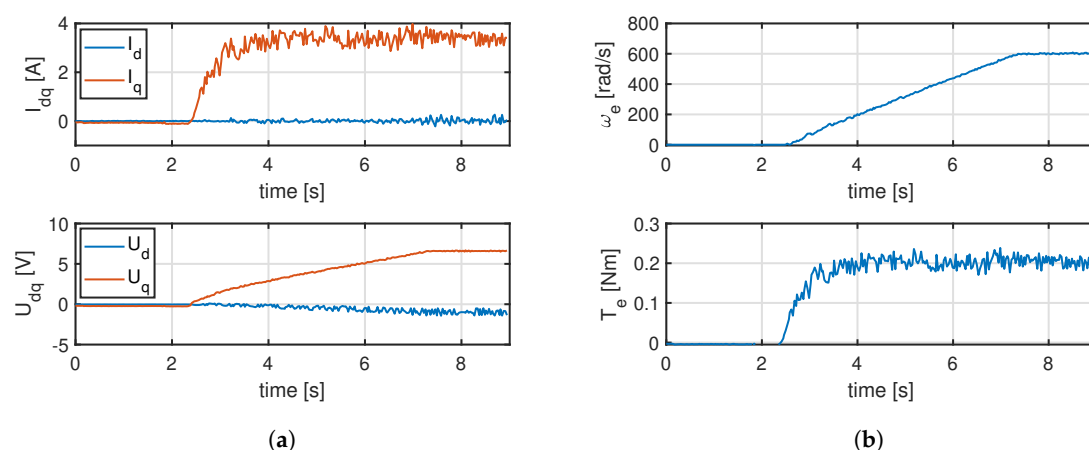


Figure 26. (a) dq currents and voltages; (b) ω_e and T_e during start-up sequence of Figure 25.

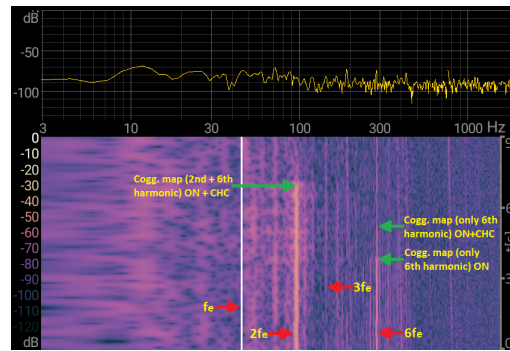


Figure 27. Sound FFT analysis at steady-state $\omega_e = 300 \frac{\text{rad}}{\text{s}}$ with applied cogging torque map and CHC algorithm. Approx. 50% of rated load.

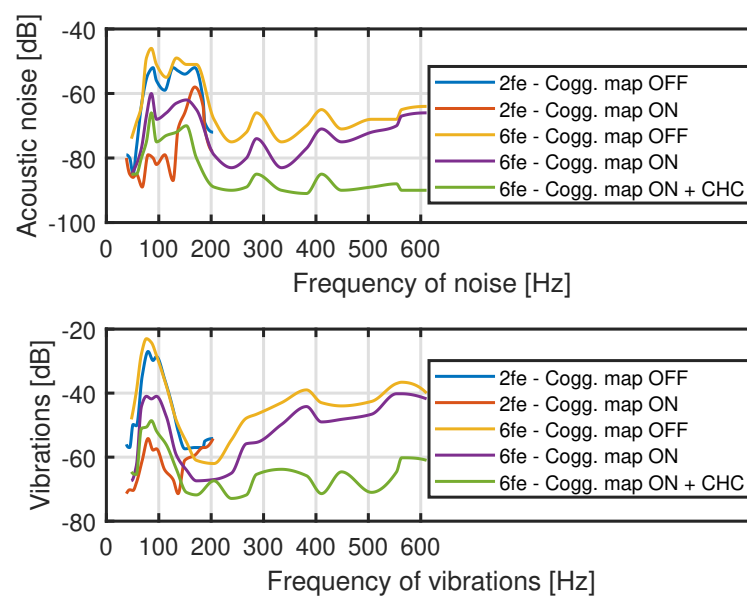


Figure 28. Acoustic noise and vibrations ($2f_e$ curve and $6f_e$ curve) vs. their own frequency.

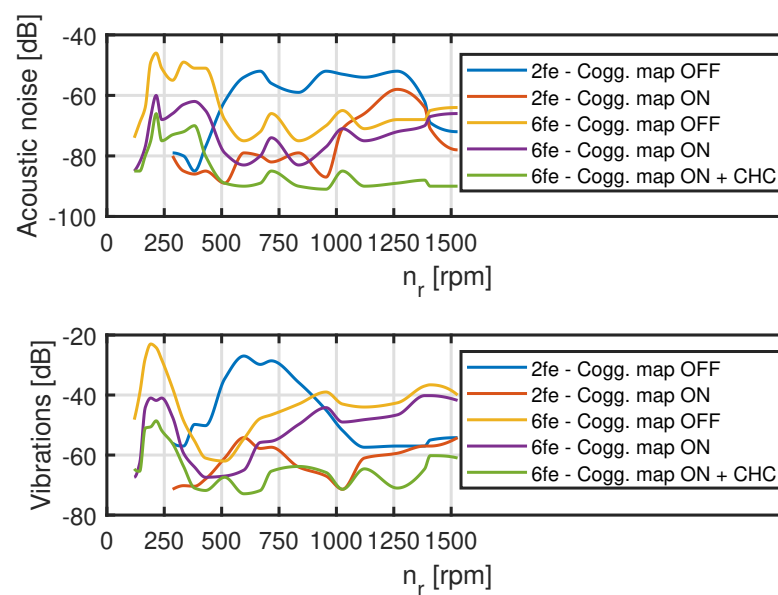


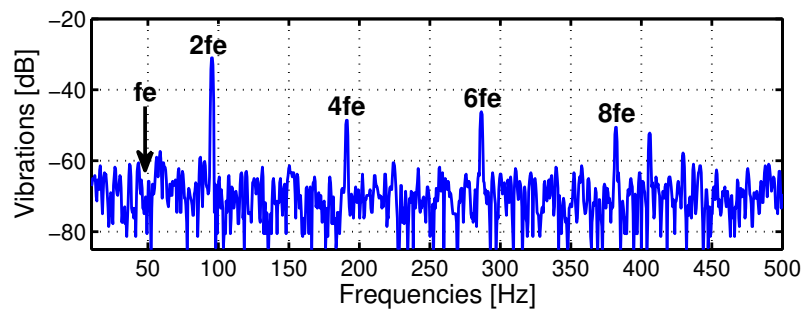
Figure 29. Acoustic noise and vibrations ($2f_e$ curve and $6f_e$ curve) vs. mechanical speed n_r .

Table 2. Acoustic noise and vibrations for $2f_e$ curve vs. mechanical speed n_r .

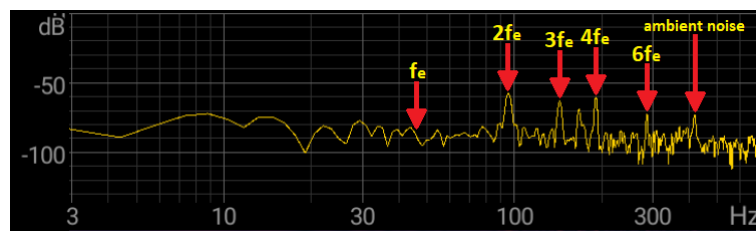
n_r [rpm]	Vibrations [dB]		Acoustic Noise [dB]	
	Cogg. map OFF	Cogg. map ON	Cogg. map OFF	Cogg. map ON
286	−56	−71.4	−79	−80
334	−57	−70.2	−80	−85
382	−49.8	−70.4	−85	−86
430	−50.2	−68	−77	−85
513	−34.6	−61	−62	−89
597	−27	−54.2	−54	−79
668	−29.8	−57.8	−52	−80
716	−28.6	−57.4	−56	−82
836	−35.8	−64	−59	−79
955	−45	−67	−52	−87
1027	−51.8	−71.4	−53	−71
1122	−57.4	−61	−54	−66
1265	−57	−59.4	−52	−58
1385	−57	−57	−62	−64
1409	−55	−57	−69	−71
1528	−54.2	−54.2	−72	−78

Table 3. Acoustic noise for $6f_e$ curve vs. mechanical speed n_r .

n_r [rpm]	Acoustic Noise [dB]		
	Cogg. map OFF	Cogg. map ON	Cogg. map ON + CHC
119	−74	−85	−85
143	−69	−82	−85
167	−64	−77	−80
191	−49	−67	−75
215	−46	−60	−66
239	−51	−68	−75
286	−55	−66	−73
334	−49	−63	−72
382	−51	−62	−70
430	−51	−65	−80
513	−68	−79	−89
597	−75	−83	−90
668	−72	−80	−89
716	−66	−74	−85
836	−75	−83	−90
955	−70	−77	−91
1027	−65	−71	−85
1122	−71	−75	−90
1265	−68	−72	−89
1385	−68	−70	−88
1409	−65	−67	−90
1528	−64	−66	−90

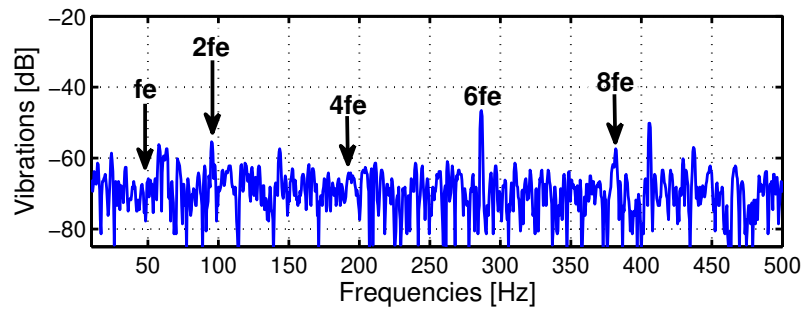


(a)

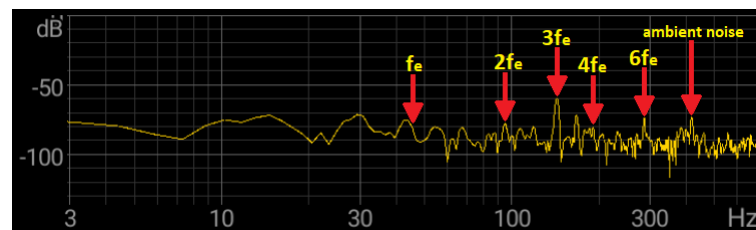


(b)

Figure 30. Cogging torque map OFF, CHC OFF, $\omega_e = 300 \frac{\text{rad}}{\text{s}}$: (a) vibrations FFT; (b) acoustic noise FFT.



(a)



(b)

Figure 31. Only 2nd harmonic of cogging torque map applied, CHC OFF, $\omega_e = 300 \frac{\text{rad}}{\text{s}}$: (a) vibrations FFT; (b) acoustic noise FFT.

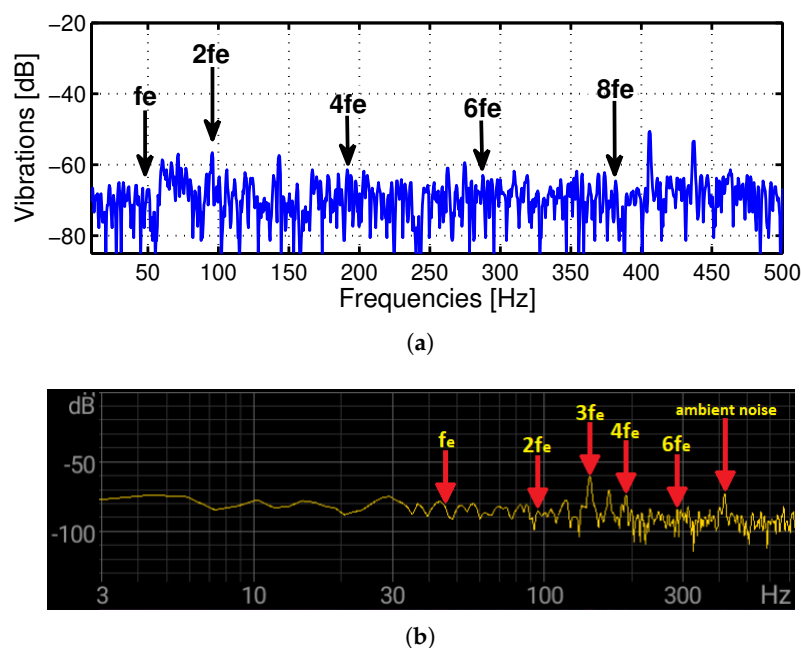


Figure 32. 2nd and 6th harmonic of cogging torque map applied, CHC ON, $\omega_e = 300 \frac{\text{rad}}{\text{s}}$: (a) vibrations FFT; (b) acoustic noise FFT.

6. Conclusions

This work proposes a new control approach to increase the performance of electric drives with PM motor in terms of vibrations and acoustic noise reduction by modification of FOC structure. Dominant harmonic components (2nd and 6th) of cogging torque profile are superimposed into q -axis current reference to reduce acoustic noise, vibrations, torque ripple and speed ripple caused by cogging torque.

The Novel Current Harmonic Controller (CHC) algorithm was presented. CHC is simple and easy to use algorithm for control of dq current harmonics in order to reduce acoustic noise, vibrations and torque ripple in whole speed range. It can be used not only for precise tracking of dq current references, but also for suppression of undesirable low-order current harmonics caused by non-sinusoidal Back-EMF and dead-time. Very precise tracking of q -axis current reference, which has six times higher frequency than stator electrical frequency f_e and contains suitably added cogging torque profile to suppress its impact on performance of the PM motor was presented in this paper.

Achieved results show that proposed approach significantly improves performance of low-cost PM motor in terms of lower acoustic noise, vibrations and torque ripple, which are highly influenced by poor construction, which often leads to significant cogging torque.

In future work, we would like to remove limitations of proposed approach and evaluate application of CHC in field oriented control for other machines (multi-phase motor, induction motor, synchronous reluctance motor) in order to improve their performance in terms of torque ripple, acoustic noise and vibrations. We would like to deeply investigate strong saturation of the stator as a limitation in terms of correct identification and thus reduction of cogging torque, despite of measured cogging torque map under no-load was almost the same as under 45% load. Although low-cost encoder with low-resolution (512 ppr) was used during measurement, which can be considered also as a limitation. Therefore, we would like to test functionality of the proposed approach even with less ppr or in sensorless operation.

Author Contributions: Conceptualization, methodology, investigation, software, validation, writing—original draft preparation, M.S. (Martin Sumega), formal analysis, resources, writing—review and editing, supervision, P.R., M.S. (Marek Stulrajter). All authors have read and agreed to the published version of the manuscript.

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Nomenclature

θ_r	rotor mechanical angular position
f_e	stator electrical frequency
ω_r	mechanical angular velocity
T_e	electromagnetic torque
*	required value
θ_e	rotor electrical angular position
ω_e	electrical angular velocity
n_r	rotor mechanical speed expressed in rpm
T_{cogg}	cogging torque
S_a, S_b, S_c	inverter switching signals

Abbreviations

The following abbreviations are used in this manuscript:

PM	Permanent Magnet
FOC	Field Oriented Control
CHC	Current Harmonics Controller
Back-EMF	Back Electromotive Force (voltage)
BLDC	Brushless Direct Current
PMSM	Permanent Magnet Synchronous Motor
PI	Proportional Integral
LCM	Least Common Multiple
SVM	Space Vector Modulation
FFT	Fast Fourier Transform

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