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Simplified model of Switched Reluctance Motor for real-time calculations

Abstract. *The purpose of the paper is to show a new approach for three phase, 12/8 poles SRM modeling. It was assumed that the model need to fulfill a general principles of electromagnetic energy conversion, has simple structure for fast computation and has possibility to invert for control system synthesis. The main idea bases on the replacing the non-linear flux linkage function of two variables on the product of two nonlinear functions of one variable. The analytical basis and corresponding formulas are shown in details. Two different approaches for derivation of the electromagnetic torque generation formulas are presented. The motor model implemented in the Matlab/Simulink packet is described. Simulation results with the experimental waveforms are compared what confirmed agreement of model with real object. Comparison bases on the current waveforms of three phases in selected operation points. The paper presents the way of conversion of nonlinearity from 2D arrays to the analytical form of approximated functions with only 4 parameters. Presented in the article simple, nonlinear model of SRM gives an opportunity to use in real time control systems and is dedicated for model adaptive reference structures (MRAS). The MRAS structure can be evaluated for fault detection system (FDS) and fault tolerant control (FTC) purposes.*

Streszczenie. *W artykule przedstawiono oryginalne podejście do modelowania silnika reluktancyjnego przełączalnego. W przykładzie wzięto pod uwagę popularną konstrukcję trójpasmową o strukturze biegunów 12/8. Założono zgodność z zasadami przetwarzania energii elektromagnetycznej, prostą konstrukcję oraz zdolność do odwrócenia na potrzeby syntezy toru sterowania. Idea bazuje na zastąpieniu nieliniowej, sprzężonej funkcji dwóch zmiennych strumienia magnetycznego na wynik mnożenia dwóch funkcji jednej zmiennej. Przedstawiono podstawy analityczne idei oraz jej szczegółowe wyprowadzenia. Opisano zaimplementowany model w pakiecie Matlab/Simulink oraz wyniki badania jego zbieżności z eksperymentem w wybranych punktach pracy. Funkcje nieliniowe stablicowane w procesie wstępnej identyfikacji maszyny zostają zastąpione aproksymacjami analitycznymi z sumaryczną liczbą tylko czterech parametrów. Taki analityczny model daje się łatwo odwrócić, jego obliczenia są efektywne, a same parametry dają się stroić w procesie adaptacji do modelu rzeczywistego. Przedstawiona idea zastosowana w strukturze z modelem referencyjnym pozwala na rozwinięcie o blok funkcjonalny detekcji uszkodzeń (ang. Fault Detection System) oraz samego sterownia odpornego na awarię (ang. Fault Tolerant Control).**(Model matematyczny silnika reluktancyjnego przełączalnego przeznaczony do obliczeń w czasie rzeczywistym)***

Keywords: modeling, SRM, switched reluctance motor, electric drive, nonlinear approximation, real-time systems
Słowa kluczowe: modelowanie, SRM, silnik reluktancyjny przełączalny, napęd elektryczny, aproksymacja nieliniowa, systemy czasu rzeczywistego

Introduction

The aim of the study is to develop a model of switched reluctance motor (SRM) with simple structure which ensures satisfactory agreement with the real system. Simple and inversion-able motor model would allow further development and real-time testing of advanced control algorithms for fault-tolerant purposes (Fault Detection System – FDS and Fault Tolerant Control – FTC) in MRAS(Model Reference Adaptive System) structure. Furthermore, a fast computing model gives an opportunity to implement its inversion in simple embedded control systems e.g. to minimize torque fluctuations.

The way to present flux-linkage highly nonlinear function of two variables for SRM by two separately defined nonlinear expressions, each of one variable introduced in [1] is applied. Due to it one can obtain the model with only 4 parameters that can be easily converted to the lookup tables (and back) in embedded control system.

SRM modeling challenges

Nonlinearity modeling of switched reluctance motor is a problem discussed in many publications. Many presented modeling method are not applicable in the embedded system due to its complexity. The section focuses on a brief analysis of several papers in international context with comparison to the proposed model.

In [6] nonlinearity of flux-linkage generation in motor was estimated by several „gage curves" with lot of parameters what cause difficulty to adopt or even apply some approximations in different motor structure.

In the [7] an approach is presented with mutual interaction effects between adjacent phases what is omitted in many other publications. Nevertheless, proposed model bases on FEM (Finite Elements Method) computation of parameters with complex analytical functions. This makes solution very hard to apply in embedded system. No experimental results are shown.

An article [8] is very interesting with solution close to the one that the paper presents. Simple relations of trigonometric functions so the exponential in flux-linkage relation with only five parameters are shown. The model is simple enough to implement in on-line self-tuning system. Still [8] bases on entangled function of two variables that can not be presented by simple lookup table.

Similar approach (nonlinear analytical function approximation) as in [8] is presented in the article [9]. Flux-linkage function is shown in the way of Fourier series expansion up to the 6th grade, what demands 6 parameters and off-line precise identification process with necessity of external, precise mechanism of rotor movement.

There are also publications that uses artificial neural networks (ANNs) to solve nonlinearity of the SRM modelling. Papers [13] and [10] are an examples. The results of phase current and flux-linkage estimations compared to the measured ones gives a prove for good model convergence, but the presented ANN structures – more distinguished themselves – requires much more computational power than solution presented in this paper.

The ANN approach in not the only available from methods based on artificial intelligence (AI) in the literature. In the [12] fuzzy model is introduced but with no details about fuzzy rule table and tuning possibilities (to obtain higher accuracy), there are also no experimental waveforms shown to verify the model.

It could be seen that simple model gives an opportunity to implement in FDS (fault detection system) what is presented in [11] and this application is also the purpose of the model proposed in this paper.

Model principles

As indicated in [1], the motor model should be very simple but in the same time as accurate as possible in comparison to the real system. This model should be lack of computational problems with possibility to give a „inversion" for control task and fulfill a general principles of electromagnetic energy conversion. In the presented SRM model some simplified assumption are made like: lack of interaction between the phases, lack of eddy currents and full symmetry of stator and rotor [2]. For such assumptions instantaneous electromagnetic torque T_e could be expressed as the sum of the torques of individual phases T_p, for three phase motor: $p = 1, 2, 3$. T_e and other model equations are determined from the general electromagnetic principles. If start from the phase voltage (v_p) equation:

$$v_p = i_p R_p + v_{pF}, \tag{1}$$

where: $i_p(A)$ – phase winding current, $R_p(\Omega)$ – phase winding resistance, $v_{pF}(V)$ – Faraday's induction law voltage,

$$v_{pF} = \frac{d\Psi_p}{dt}, \tag{2}$$

$\Psi_p(Wb)$ – phase winding magnetic flux linkage, $t(s)$ – time. In general case, where the flux linkage is a function of two variables: θ_p – motor rotor angle and i_p – phase current, equation (2) can be expanded to the form:

$$v_{pF} = \frac{d\Psi_p(i_p, \theta_p)}{dt}, \tag{3}$$

$$v_{pF} = \frac{\partial\Psi_p(i_p, \theta_p)}{\partial i_p}\frac{di_p}{dt} + \frac{\partial\Psi(i_p, \theta_p)}{\partial\theta_p}\frac{d\theta_p}{dt}. \tag{4}$$

Above equation is an entry point to solve electromagnetic torque value T_p dependence, what is shown in details in the next section.

Model equations

The simple, linear model of SRM bases on the assumption that the flux linkage is line dependent on the phase current:

$$\Psi_p(\theta_p) = L_p(\theta_p) i_p, \tag{5}$$

where: $L_p(H)$ – phase winding inductance. Equation (5) applied to (1) and (4) gives:

$$v_p = i_p R_p + L_p(\theta_p)\frac{di_p}{dt} + \frac{dL_p(\theta_p)}{d\theta_p} i_p \omega, \tag{6}$$

from where can be seen that expression:

$$\epsilon_p = \frac{dL_p(\theta_p)}{d\theta_p} i_p \omega \tag{7}$$

is an electromotive force (ϵ_p) generated in motor winding dependent from rotor speed $\omega = \frac{d\theta_p}{dt}$.

Mechanical power P_{me} consumed to generate electromagnetic torque T_p can be expressed as an difference between supplied electrical power P_{el} and power of the phase coil magnetic field plus resistive loses ($P_R = {i_p}^2 R_p$):

$$P_{me} = P_{el} - (P_{ma} + P_R). \tag{8}$$

If consider P_{el} from product of equation (6) and phase current (i_p), there is:

$$P_{me} = T_p \omega = T_p \frac{d\theta_p}{dt}, \tag{9}$$

and power of magnetic field P_{ma}:

$$P_{ma} = \frac{\int_{i_p} L(\theta_p) i_p di}{dt}, \tag{10}$$

$$P_{ma} = L_p(\theta_p) i_p \frac{di_p}{dt} + \frac{1}{2} i^2 \omega \frac{dL_p(\theta_p)}{d\theta_p}. \tag{11}$$

Equations (8),(9),(11) gives finally:

$$T_p = \frac{1}{2} i^2 \frac{dL_p(\theta_p)}{d\theta_p}. \tag{12}$$

Unfortunately, model based on formulas (6) and (12) provides a good agreement with experimental results only in the range of small currents [1]. To make more accurate model, the equation (5) could be written in the form:

$$\Psi_p(\theta_p) = L_p(\theta_p) sat(i_p), \tag{13}$$

where: $sat(i_p)$ – saturation function of the phase current (i_p).

The role of the $sat(i_p)$ from equation (13) is to take account of the saturation of the magnetic circuit. When (11) applied to the phase voltage equation (1) and (4) it can be written:

$$v_p = i_p R_p + \frac{d\Psi_p(i_p, \theta_p)}{dt}. \tag{14}$$

If assumed v,i,R,L related to the motor phase winding (subscript p), expression (14) may be expanded to the form:

$$v = iR + \frac{\partial(L(\theta)sat(i))}{\partial i}\frac{di}{dt} + \frac{\partial(L(\theta)sat(i))}{\partial\theta}\frac{d\theta}{dt}. \tag{15}$$

Introducing operator D as follows:

$$D(F(x)) = \frac{\partial F(x)}{\partial x} \tag{16}$$

equation (15) can be write in the form:

$$v = iR + L(\theta)D(sat(i))\frac{di}{dt} + D(L(\theta))sat(i)\omega, \tag{17}$$

where: D – differential operator of the dependent variable indicated in parentheses. There can be easily derived from (17) phase winding current i_p state equation:

$$\frac{di_p}{dt} = \frac{u_p - i_p R_p - D(L_p(\theta_p))sat(i_p)\omega}{L_p(\theta_p)D(sat(i_p))}. \tag{18}$$

To get a torque value based on nonlinear model equation (13), the start point is mechanical power equation (9). This power is also equal to the time derivative of the *coenergy* (E_C):

$$P_{me} = \frac{dE_c}{dt}, \tag{19}$$

Alignment of (9) and (19) results in:

$$T_p = \frac{dE_c}{d\theta_p}. \tag{20}$$

The E_C as a current integral of flux linkage from definition gives:

$$dE_C(\theta_p, i_p) = \int_{\tau=0}^{i_p} \Psi_p(\theta_p, \tau) d\tau, \tag{21}$$

$$dE_C(\theta_p, i_p) = \int_{\tau=0}^{i_p} (L_p(\theta_p) sat(\tau)) d\tau, \tag{22}$$

and finally:

$$T_p(\theta_p, i_p) = \frac{L_p(\theta_p)}{d\theta_p} \int_{\tau=0}^{i_p} (sat(\tau)) d\tau. \tag{23}$$

Introducing integral operator S of the dependent variable indicated in parentheses:

$$S(F(x)) = \int_{\tau=0}^{x} (F(\tau))d\tau, \tag{24}$$

in above equation, one can obtain:

$$T_p(\theta_p, i_p) = D(L_p(\theta_p))S(sat(i_p)), \tag{25}$$

where: S – integral operator of the dependent variable indicated in parentheses. Easily can be proved that above equation with additional form:

$$satT(i_p) = \sqrt{2S(sat(i_p))}, \tag{26}$$

gives the electromagnetic torque formula $T_p(\theta_p, i_p)$ more similar to the linear model:

$$T_p(\theta_p, i_p) = \frac{1}{2}D(L_p(\theta_p))satT(i_p)^2. \tag{27}$$

As mentioned before, overall electromagnetic torque T_e generated by motor is the sum of all torques generated by phases separately:

$$T_e(\theta_r, i_p) = \sum_{p=1}^{3} T_p(\theta_p, i_p), \tag{28}$$

where: θ_r – an absolute rotor position in relation to θ_p as follows:

$$\theta_p = \theta_r - \pi(p-1)/12. \tag{29}$$

Nonlinear functions

Nonlinear functions: $sat(i_p)$ and $L_p(\theta_p)$ and its derivatives so integrals are implemented in the form of tables in the functional, limited ranges (as period of the $L_p(\theta_p)$ which is $\pi/8$ for 12/8 topology and 0-5 A for $sat(i_p)$ function). It is easy to keep this ranges while operating because of control signal limitations. Look-up tables simplifies model computation and makes it as quick as possible what is important in practical MCU (micro-controller unit) implementation. The function $sat(i_p)$ shape is similar to the one proposed in ([1]) based on finite elements method (FEM) computation. Function $sat(i_p)$ shape is shown on the figures: 1 and 2.

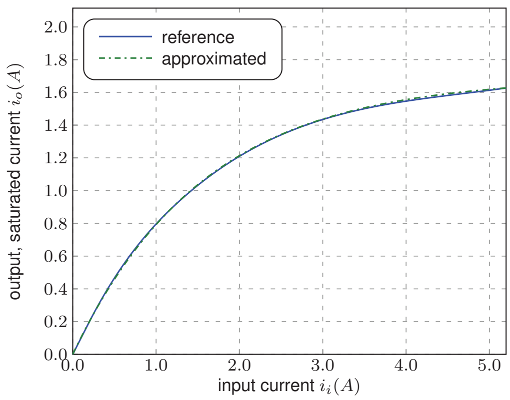

Figure 1. Approximation results of the $sat(i_p)$ expression to the exponential function

Phase inductance $L_p(\theta_p)$ table is fulfilled by direct off-line measurement (result shown in figure 3, so the its derivative of the angle - $dL_p(\theta_p)$ in the figure 4). The important thing is that nonlinear functions $sat(i_p)$ and $L_p(\theta_p)$ can be also approximated by sinus and exponential functions respectively. It must be pointed out that sine and exponential

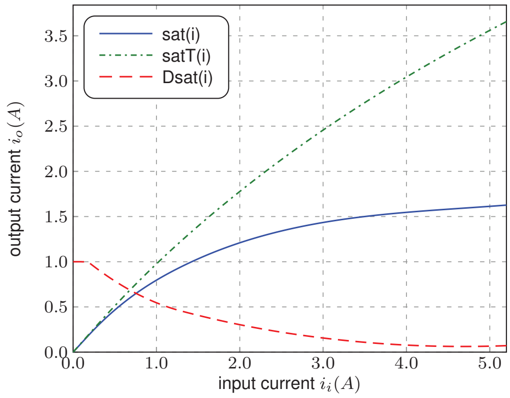

Figure 2. Function $sat(i)$ and its derivatives: $Dsat(i)$, $satT(i)$ plots

function have the advantages that the analytical form of the least squares approximation method is relatively simple. Implication of this is possibility of parametrization of model nonlinear functions and adaptation of these parameters to the real object in the off or on-line process. Moreover, analytical form allows to extend the range of the initial array (based on direct measurement/computation) and gives flexibility in determining the resolution of the table. After periodic model parameters adaptation, the analytical form can be easily converted to a table for on-line computing in control process. There are two functions with four parameters to approximate

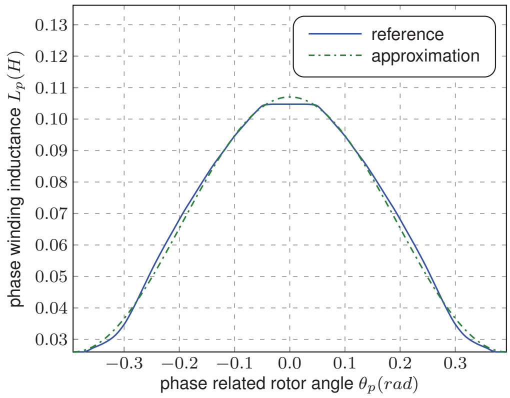

Figure 3. Phase inductance $L_p(\theta_p)$ plot taken from direct measure compared with its approximation to the cos function

from table values.

If start with the general rule for least square method function approximation:

$$\sum_{i=1}^{n}[F(x_i, p_1, ...p_k) - y_i]^2 = min, \tag{30}$$

where: n – number of sample points; i – sample point number; $F()$ – approximating function; x_i – function argument; p_l – function parameter l-numbered; y_i – measured value at i-sample; and exponential approximation function in the form:

$$F_i(x_i, \gamma, \epsilon) = \gamma(1 - e^{\epsilon x_i}), \tag{31}$$

where parameters are: γ, ϵ, it is easy to prove that those parameters can be computed from equations:

$$\epsilon = \frac{n\sum_{i=1}^{n}(x_i ln(y_i)) - \sum_i^n(x_i)\sum_i^n(ln(y_i))}{n\sum_{i=1}^{n}x_i^2 - (\sum_{i=1}^{n}x_i)^2}, \tag{32}$$

$$\gamma = exp(\frac{\sum_{i=1}^{n}(ln(y_i)) - B\sum_{i=1}^{n}(x_i)}{n}). \tag{33}$$

The results of the exponential approximation of $sat(i_p)$ function and its derivatives are shown on the figure 1 and figure 2. Computed values applied to figure 1 plot values: $\gamma = 1,68$; $\epsilon = -0,65$.

For cosine approximation function:

$$F_i(x_i, \alpha) = \alpha[cos(8x_i) + 1], \quad (34)$$

parameter α can be computed from:

$$\alpha = \frac{\sum_i y_i cos(8x_i)}{\sum_i cos^2(8x_i)}. \quad (35)$$

The equation (34) as an $L_D(\theta_p)$ – dynamic inductance approximation (omitted constant minimal phase inductance – β) plus equation (35) related to the motor phase winding inductance $L_p(\theta_p)$ gives:

$$L_p(\theta_p) = L_D(\theta_p) + \beta = \alpha[cos(8\theta_p) + 1] + \beta, \quad (36)$$

where: $\theta_p \in (-\frac{\pi}{8}; \frac{\pi}{8})(rad)$; $\alpha(H)$ – dynamic inductance L_D amplitude; $\beta(H)$ – minimal measured phase inductance;

As could be seen in the figure 4, some measurements errors in the phase inductance from rotor angle gives much noise in its derivative function, so the equation (36) is method for analytical smoothing that reduce measurement errors. Basing on the FFT (Fast Fourier Transform) analysis, the second the highest harmonics amplitude of the sourced inductance signal is less than 2% of the main one, so another harmonics in the study are omitted without consequences for the required accuracy of the model. The proposed approximation of cosines function gives good results what illustrates waveforms from figure 3.

The results of the cosine approximation of $L_p(\theta_p)$ function and its derivatives are shown on the figure 3 and figure 4. Computed values applied to figure 3 and 4 waveforms generation: $\alpha = 0,041$; $\beta = 0,026$.

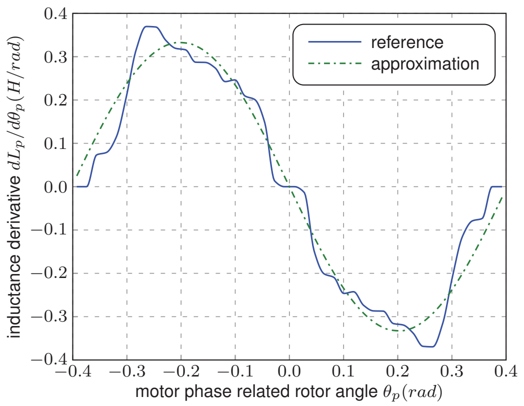

Figure 4. Derivatives of the phase inductance $D(L_p(\theta_p))$ from measure and of the cos approximation

Drive structure

It is necessary to mention that the switched reluctance motor model verification needs to be as a part of the all drive system while this kind of motor could not be driven directly from the power grid – cannot work stand alone. SRM requires dedicated power inverter and control system with rotor position monitoring. As shown in figure 5 (general structure of the drive system), INV (inverter) generates voltage extortion based on PWM (pulse width modulation) signals with delay vary mainly on PWM frequency. IS (phase current sensor) and WS (position/speed sensor) are models of devices by use of 1st order lag system (figure 4). The time constants of

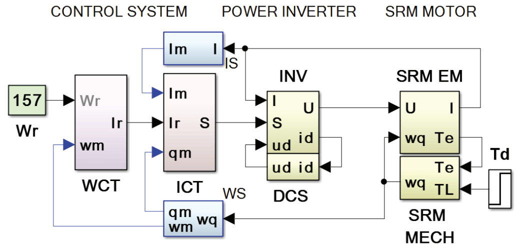

Figure 5. General structure of the SRM drive model

that inertia blocks comes from physical parameters of sensors and digital measurement methods, so the analog-to-digital converters implemented in the real embedded system. ICT is the current controller while WCT are speed controller based on the structure and parameters coming from MCU implementation [4], ICT and WCT are simple PI controllers with appropriate output limitations (specified in the nominal rates and model parameters summary below). The main part of the model - electromagnetic subsystem of the SRM - which is also the article main concern was shown in figure 6. The figure presents (in general) Simulink implementation of the equations (18) and (28) with nonlinear blocks called $Dsat$, sat, $satT$, $L5(Q)$ and $dL5(Q)$ which represents the functions: $D(sat(i_p))$, $sat(i_p)$, $satT(i_p)$, $L_p(\theta_p)$, $dL_p(\theta_p)/d\theta_p$ respectively. An additional part to the system – the position signal quantization was shown in the bottom-left part of the figure 6. Mechanic part of the SRM model was shown in the figure 7 (described as *SRM MECH* in the figure 5). It was implemented as an simple block with static and viscous friction.

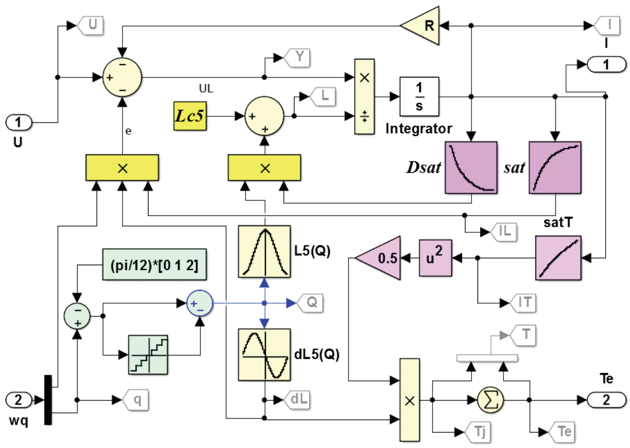

Figure 6. Electromagnetic part of the switched reluctance motor model

Turn-on and turn-off angles for phase excitation could have any values and the phase excitation period could be less than $\pi/12$ as opposite to the solution presented in other articles. There is also linear current controller modeled while in [1] is used hysteresis one.

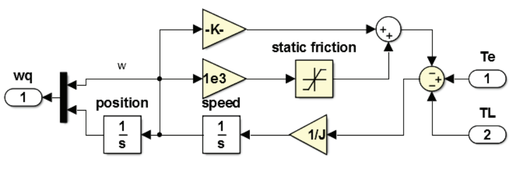

Figure 7. Mechanical part of the SRM model

The model parameters comes from the real system as follows: motor phase resistance: $R = 6,98\ \Omega$, rotor inertia: $J = 35 \cdot 10^{-6}\ kgm^2$, voltage supply: $U = 162\ V$, voltage drop across the power inverter: $U_d = 2\ V$, inverter PWM frequency $f = 20\ kHz$, PWM resolution: $9bit$, phase excitation turn-on angle: $\theta_{on} = -15,0deg$, phase excitation turn-off angle: $\theta_{off} = -2,0deg$, speed regulator output limit: $i_{max} = 5\ A$, minimum measured phase inductance: $L_p(\theta_p)_{min} = 23\ mH$, maximum measured phase inductance: $L_p(\theta_p)_{max} = 106\ mH$.

Simulation and experimental results

Model verification bases on comparing simulation and experimental phase current waveforms at steady states at different speeds and load torques. The SRM used for experiment is applied in washer machine and it is quite popular unit. Detailed construction parameters can be obtained from [5]. It may be noted from figure 8 and figure 9 that the

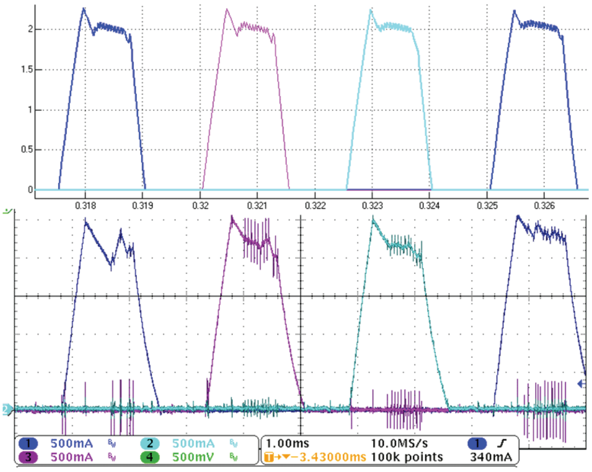

Figure 8. Phase currents waveforms comparison: simulation (top) and experiment (bottom) results at speed 1000 rpm, load torque 0,15 Nm; X(time) scale: 1 ms/DIV; Y(value) scale: 500 mA/DIV

proposed model in terms of phase current values, its time dependences and waveforms in general gives good convergence with the experimental results. Important in the present case is not only a model of the motor but also the model parameters of entire control system.

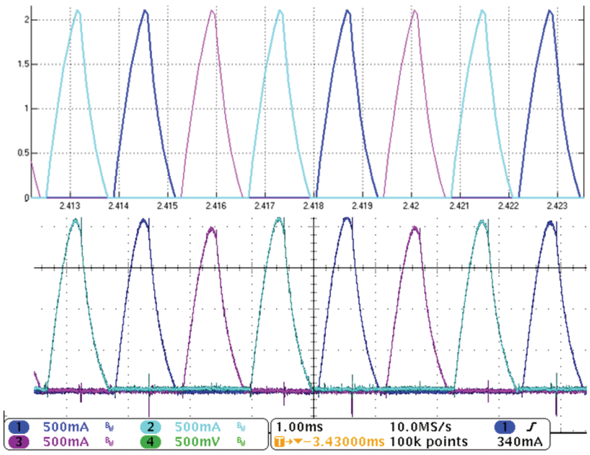

Figure 9. Phase currents waveforms comparison: simulation (top) and experiment (bottom) results at speed 2500 rpm, load torque 0,17 Nm; X(time) scale: 1 ms/DIV; Y(value) scale: 500 mA/DIV

Conclusions

A new model of the three phase, 12/8 poles switched reluctance motor was shown for the real-time computing purposes. The presented in the paper model of switched reluctance motor gave a good similarity in comparison with experimental results. All the complete drive system elements implementations were shown: electromagnetic and mechanics part of the motor, controllers, sensors and power inverter. The SRM model equations evaluation were presented. A cosines function was used for nonlinear inductance function smoothing purposes with good results. Used nonlinear functions approximation were shown with its parameters computing methods.

Such a model, evaluated for 12/8 topology due to its simplicity can be applied in real time systems on microprocessor or FPGA platform. Nowadays HDL generators which automates HIL (Hardware In the Loop) simulations (based on Matlab/Simulink) requires simple computational blocks as an elements of the model. If applicable for FPGA implementation, the results of the presented paper would be used for advanced and/or fault tolerant control (FTC), so fault detection systems (FDS) as an subject of further research and experimental implementation.

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