Passivity-Based Adaptive Sliding Mode Speed Control of Switched Reluctance Motor Drive Considering Torque Ripple Reduction

M.M. Namazi Isfahani, S.M. Saghaian-Nejad, A. Rashidi and H. Abootorabi Zarchi

Abstract—This paper presents a high-performance nonlinear controller for Switched reluctance machine (SRM), acting as a speed regulation motor drive. We use a cascaded torque control structure for torque ripple minimization. In a cascaded control structure, accurate torque control requires inner current controller. Passivity-based control (PBC) approaches is proposed for current control purpose. By using the port-controlled Hamiltonian (PCH) systems theory, a full-order nonlinear controlled model is first developed. Then nonlinear passivity-based adaptive sliding mode control algorithm in the presence of external disturbances for the purpose of torque ripple reduction and characteristic improvement is presented. The proposed controller design is separated into the inner loop and the outer loop controller. In the inner loop, passivity-based control (PBC) is employed by using energy shaping techniques to produce the proper switching function. The outer loop control is employed by adaptive sliding controller to determine the appropriate Torque command. It can also overcome the inherent nonlinear characteristics of the system and make the whole system robust to uncertainties. The performance of the proposed controller algorithm has been demonstrated in simulation and also experimentally using a 4kW, four-phase, 8/6 pole SRM DSP-based drive system.

I. INTRODUCTION

In recent years, there is a growing concern about the use of switched reluctance motor. Robustness, high efficiency, low cost, high speed, simple structure, easy to maintain, high controllability, high torque to inertia ratio, simple power converter circuits with reduced number of switches and smaller dimension of the motor, are unique features of this machine. High torque ripple and acoustic noise are most disadvantages of the motor. The nonlinear characteristics include the nonlinear torque function of position and current and the magnetic saturation at certain operation regions [1].

Several control methods and schemes have been proposed to overcome these problems. For example, variable structure controller made the SRM drive system insensitive to parameter variations and load disturbance [2].

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Artificial neural network and fuzzy controller needs a lot of designer experience [3]. Nonlinear internal model control for SRM drive required very complicated computations and implementation of the system is very difficult [4].

Also, for the above mentioned control strategies of the SRM it is assumed that its parameters are known exactly or the unknown parameters can be identified by the adaptive technique. However the parameters of the SRM are not exactly known and always vary with current and position. Actually, control is difficult to implement owing to its complex algorithm when considering the structural information of SRM in design. Improving the applicability of the SRM on the basis of taking the structural characteristics into account is a significant step in designing the controller of the SRM.

In this paper, a cascaded torque control structure for its well-known advantages is used. In used cascaded control structure, first an adaptive sliding Mode speed controller is designed and then compared to a PI controller, finally to achieve high performance torque control, a nonlinear feedback current controller, which effectively uses the natural energy dissipation properties of the SRM, is proposed. Passivity-based control (PBC), introduced to define a controller design methodology which achieves stabilization by passivation [5]. The Control algorithm is first simulated through SIMULINK and then tested on a four-phase 8/6 pole 4kW oulton SR motor.

This paper is organized as follows. At first, in Section II, the SRM nonlinear modeling is presented. In Section III controller design is separated into the inner loop and the outer loop and a passivity-based adaptive sliding controller is designed based on combination of passivity-based current control and adaptive sliding mode technique. In Section V, experimental and simulation results confirm that the desired speed reference command is perfectly tracked and torque ripple reduced in spite of external load disturbances. Finally, conclusions are given in Section VI.

II. NONLINEAR CONSTRUCTION AND MODELING OF SRM

Owing to the doubly salient construction, the SRM presents a highly nonlinear load to the current controller, thus the design of a high-performance current controller for an SRM drive is a challenge. For SRM drives, the mutual coupling between phases is usually neglected for low-speed applications, so the phase currents can be controlled independently. The unipolar converters is used as the power converter for SR motor drives, while this converter topology provides the most flexible and effective control to the current waveforms of SRM. The voltage equation for one phase of an SRM is
\[ v_k = r_k i_k + \frac{d\lambda_k(\theta, i_k)}{dt} = r_k i_k + L_k(\theta, i_k) \frac{di_k}{dt} + \frac{\partial \lambda_k(\theta, i_k)}{\partial \theta} \omega \]  

Where \( u_k \) is the phase voltage, \( i_k \) is the phase current, \( r_k \) is the phase resistance, \( k = a, b, c, d \) is the active phase, \( \theta \) is the rotor position and \( \lambda_k(\theta, i_k) \) is the flux linkage, \( L_k(\theta, i_k) = \frac{\partial \lambda_k(\theta, i_k)}{\partial i_k} \) is the incremental inductance, \( \omega = d\theta/dt \) is the motor speed and \( e = \omega \partial \lambda_k(\theta, i_k)/\partial \theta \) is the back EMF. Using numerical methods, the incremental inductance and back-EMF characteristics of the SRM can be obtained from the measured flux linkage data. Both incremental inductance and back-EMF coefficient are nonlinear functions of rotor position and current.

III. PASSIVITY-BASED ADAPTIVE SLIDING MODE CONTROL ALGORITHM

In this section, we present a Passivity-Based Adaptive Sliding Mode Control method for designing SRM controller. The overall controller block diagram, shown in Fig. 1 is separated into the inner loop and the outer loop controller. Based on the assumption that stator current \( i_k \) as well as rotor speed \( \omega \) are available for measurement, the controller design procedures can be divided into three steps. The first step is to design an adaptive sliding speed controller for speed tracking of the overall system.

The final step is passivity-based current control of the electrical subsystem by injecting a nonlinear electrical damping term and a set of reference current vectors \( i_k^* \) are found out to achieve current tracking. The sake of the outer loop is to generate the appropriate Torque command fed for the inner loop. Finally, passivity-based inner loop current controller will produce the switching functions.

A. Adaptive Sliding Mode Outer Loop Speed Controller

The first step is designing the outer-loop speed controller to determine the appropriate reference torque. An adaptive sliding mode control scheme is proposed for the speed control. The conventional adaptive sliding mode control estimates the unknown uncertainty upper boundary with sign function which causes the chattering phenomenon. Therefore, in [6] a novel method has been proposed to reduce chattering by using exponential function wht adaptive approach instead of sign function. For speed control design purposes, the dynamic model of the SRM can be written as:

\[ \frac{d\omega}{dt} = \frac{1}{J} \left[ T_e - T_L - B\omega \right] = \beta \]

\[ \frac{d\alpha}{dt} = \frac{1}{J} \left[ \sum_{k=a}^{d} \left( \frac{\partial T_e}{\partial i_k} \right) \left( \frac{\partial \lambda_k}{\partial i_k} \right) - v_k - R_k i_k - \frac{\partial \lambda_k}{\partial \theta} \omega \right] \]  

\[ + \frac{1}{J} \left[ \omega \sum_{k=a}^{d} \left( \frac{\partial T_e}{\partial \theta} \right) - T_u - B\alpha \right] \]  

Where \( T_u = \bar{T}_L \) , Let \( x = (x_1, x_2)^T = (\beta, \omega)^T \). Also let the output of the SRM be \( y = \omega \). Thus, the model of the SRM system can be written in compact form as \( \dot{y} = f(x) + g(x)u \), where \( f \) and \( g \) are determined from (2). We consider \( T_u(t) \in \Omega_u = \{ T_u(t) : \left| T_u(t) \right| \leq \alpha \} \) as an external disturbance and control law as:

\[ u = -\frac{1}{g} \left[ Ke + \left( f - \omega^* \right)^2 e + \delta_1 \exp(-\sigma_1 t) \right] \]

\[ + \frac{1}{g} \left[ \alpha \frac{e}{\left| e \right| + \left( \delta_2 / \rho \right) \exp(-\sigma_2 t)} \right] \]  

Where \( K > 0 \), is the state feedback gain and \( \alpha \) denotes the estimated value of the unknown parameter \( \alpha \). The constants \( \delta_1 \) and \( \sigma_1 \) are (small) positive constants specified by the designer to avoid discontinuity and control chattering. The control law (3) with adaptation mechanism \( \dot{\alpha} = \gamma \dot{\alpha} \), where \( \gamma > 0 \) is adaptation gain, guarantees that all the closed loop signals are bounded and tracking error \( e = \omega - \omega^* \) is asymptotically converged to zero. The smaller value of \( \delta_1 \), give less smoothness to the control law.

Finally in this step a proportional-integral (PI) controller is designed here to compare with the nonlinear controller. The parameters of the PI controller are determined by pole placement. By assumption of \( \xi = 1.25, \omega_h = 0.7 \text{ rad/sec} \), the PI controller parameters obtained as \( K_p = 0.466, K_i = 7.47 \).

B. PCH Modeling of SRM

Energy conserving dynamical systems with independent storage elements can be described by the form of port controlled Hamiltonian (PCH) systems as follows [7]

\[ \begin{align*}
\dot{x} &= [J(x) - R(x)] \frac{\partial H(x)}{\partial x} + G(x)u \\
y &= h(x) = G^T(x) \frac{\partial H(x)}{\partial x}
\end{align*} \]  

Where \( G(x) \) denotes the input matrix, \( u \) and \( y \) denote the system input and output. The interconnection structure is captured in the matrix \( J(x) \) with \( J^T(x) = -J(x) \).

The dissipation effects are captured by the matrix \( R(x) \) which is a semi-positive definite or positive definite matrix.
The total energy of the system is defined by the positive definite function \( H(x) \). The port controlled Hamiltonian systems with dissipation (4) satisfy the following power-balance equation

\[
\frac{dH(x)}{dt} = u^T y - \frac{\partial^T H(x)}{\partial x} R(x) \frac{\partial H(x)}{\partial x}
\]

The port-controlled Hamiltonian system is passive if the Hamiltonian \( H(x) \) is bounded from below [8]. To simplify controller design, a complex system can be decomposed into subsystems by using some techniques of order reduction, such as multi-time-scale analysis. Consider two subsystems with negative feedback interconnection by the power port as shown in Fig. 2. After the interconnection, the closed-loop represented as

\[
\begin{bmatrix}
\dot{x}_1 \\
\dot{x}_2
\end{bmatrix} = \begin{bmatrix}
J_1 - R_1 & 0 \\
0 & J_2 - R_2
\end{bmatrix}
\begin{bmatrix}
\partial H_1 \\
\partial x_1 \\
\partial H_2 \\
\partial x_2
\end{bmatrix}
\]

\[
+ \begin{bmatrix}
G_1(u_1 - p y_2) \\
G_2(u_2 + p^T y_1)
\end{bmatrix}
\]

For the closed-loop model, the energy function is the total energy of the two subsystems; \( H(x) = H_1(x_1) + H_2(x_2) \) therefore, Using (5) and suppose that \( y_1^T p y_2 = y_2 y_1 \), the derivative of this energy function along time is given by

\[
\frac{dH}{dt} = -\frac{\partial^T H_1}{\partial x_1} R_1 \frac{\partial H_1}{\partial x_1} - \frac{\partial^T H_2}{\partial x_2} R_2 \frac{\partial H_2}{\partial x_2} + u_1^T y_1 + u_2^T y_2
\]

Stability of closed-loop system guaranteed supposing both of the two subsystems are stable.

In [9] it is proved that the complete model of a SRM can be decomposed as the feedback interconnection of the following two passive linked subsystems (electrical and mechanical passive subsystem) as shown in Fig. 2.

\[
\begin{align*}
\text{electrical subsystem:} & & \begin{bmatrix}
v_k \\
-\theta
\end{bmatrix} & \rightarrow \begin{bmatrix}
i_k \\
T_e
\end{bmatrix} \\
\text{mechanical subsystem:} & & (T_e - T_L) & \rightarrow \theta
\end{align*}
\]

For the interconnected system, take

\[
x_1 = [x_1, x_2, x_3, x_4]^T = [i_a, i_b, i_c, i_d]^T
\]

as the state vector, the electrical subsystem can be shown as

\[
\begin{bmatrix}
D_1 \dot{x}_1 = [J_1(x) - R_1] x_1 + g_1 (v_k - p G^T x_2) + \xi_1 \\
i_k = G_1^T x_1
\end{bmatrix}
\]

Where:

\[
D_1 = \text{diag} (L_a, L_b, L_c, L_d), \quad R_1 = \text{diag} (R_a, R_b, R_c, R_d) \\
G_1 = \text{diag} (1, 1, 1, 1), \quad u_1^T = [u_a, u_b, u_c, u_d], \quad J_1 = 0
\]

\[
\xi_1 = \left[-\frac{1}{2} \frac{dL_i}{d\theta} i_k \omega, -\frac{1}{2} \frac{dL_i}{d\theta} i_k \right]^T, \quad p = \text{diag} \left(\frac{1}{2} \frac{dL_i}{d\theta} i_k\right)
\]

Here \( G_1 \) denotes the input matrix, \( u_i \) the control vector and \( \xi_1 \) the disturbance.
Thus, we can define desired storage energy function

\[ D_1 \dot{x}_1 + R_i x_1 = \left( v_k - p G_2^T x_2 \right) + \xi_l \] (10)

The objective of PBC design is to guarantee tracking of signal \( i_k \) toward to its desired state \( i_k^* \) [10]. This method proposes to make a copy of (10), where the state \( x_1 \) is replaced by the desired state \( x_1^* \) and the injection damping term is added, that is a desired system

\[ D_1 \dot{x}_1^* + R_i x_1^* - k \ddot{x}_1 = \left( v_k - p G_2^T x_2 \right) + \xi_l \] (11)

Where the injection damping term is \( k = \text{diag} \{ k_1, k_2, k_3, k_4 \} \), referred as damping injection matrix, is a positive definite diagonal matrix and \( \ddot{x}_1 = x_1 - x_1^* \). Subtracting (11) from (10) yields the following error dynamic model

\[ D_1 \dot{x}_1 + R_i \ddot{x}_1 + k \dddot{x}_1 + p G_2^T \ddot{x}_2 = 0 \] (12)

Thus, we can define desired storage energy function

\[ H_d = 0.5 \dot{x}_1^T D_1 \dot{x}_1 \] as a Lyapunov function. The desired equilibrium point is realized if the \( H \) has its minimum at the equilibrium point. Asymptotic stability is proved using LaSalle’s principle for the closed-loop system (12) where

\[ \frac{d}{dt} H_d = -\dot{x}_1^T \left( R_i + k \right) \dot{x}_1 \] (13)

Following LaSalle’s theorem, the time derivative of storage energy function is forced to be negative semidefinite, i.e. \( H_d \leq 0 \), being equal to zero only for the equilibrium points. Therefore (13) is equal to zero only when \( \ddot{x}_1 = 0 \). The Lyapunov stability and convergence can be proven that (10) is passive and input-output stable. Finally, the above method proceeds to restrict \( x_1 = x_1^* \) and solve from (11), which yields the switching functions as

\[ v_k = \dot{r}_k \dot{i}_k^* + L_k \frac{d i_k^*}{dt} + \frac{1}{2} \frac{d L_k}{d \theta} i_k \omega + \frac{1}{2} \frac{d L_k}{d \theta} i_k \omega^* - k (i_k - i_k^*) \] (14)

The reference current is calculated using the desired torque

\[ i_k^* = \sqrt{\frac{2 T_\text{e}^*}{d L_k (\theta, i_k^*)/d \theta}} \] (15)

IV. SIMULATION RESULTS

The proposed controllers are simulated using the SIMULINK software. The model takes magnetic saturation into account. The drive system simulations are used for comparison purposes to investigate the performance of the proposed PBC approach at different load conditions. Also, a DSP-based drive system using a four-phase 8/6, 4KW oultion SRM which has the nonlinear static flux linkage and torque characteristics is used to test the speed controller experimentally. From the imposed pole locations, the gains of the current PI controller are computed and the damping parameter values of passivity-based controller have been obtained by using a trial-and-error procedure.

First by using simulation and experimental results, two speed controllers without applying passivity-based current controller in motor drive are compared. The desired rotor speed is set to 200 RPM and the external load torque is suddenly changed at \( t = 0.05 \) second from 5 \( Nm \) to 10 \( Nm \). The speed response and produced torque obtained for SRM adaptive sliding and PI speed control are shown in Fig. 3. As it can be compared higher tracking performance of reference speed and much lower torque ripple is achieved from adaptive sliding mode case. For better comparison, the torque ripple values of adaptive sliding and PI speed controllers are given in Table 1. The torque ripple is defined as [11]

\[ \text{Torque Ripple} = \frac{T_{\text{inst}}(\text{max}) - T_{\text{inst}}(\text{min})}{T_{\text{avg}}} \times 100 \% \] (16)

Since a torque sensor was not available, for the electromagnetic torque, only simulation results comparing the methods are presented. The proposed adaptive sliding speed controller has been tested experimentally and compare to conventional PI controller. The work presented here employs a conventional digital-control platform. It is based on the eZdsp F2812 board as a suitable platform for implementing motor controllers. This board is built around the TMS320F2812 digital signal processor (DSP). This platform is compatible with SIMULINK and includes four dual pulse-width-modulation (PWM) channels (8 channels total), 4 analog-to-digital converters (ADCs), and a speed-encoder input. The processor is a 32-bit DSP with fixed-point arithmetic; thus, discrete and fixed-point math blocks from Simulink can be used to program it. The complete experimental hardware used for evaluating the 8/6 SRM drive is shown in Fig. 4a. The conventional asymmetric converter used for our four-phase SR drive circuit is shown in Fig. 4b. The main power converter is mounted on the workbench, as shown in the Fig. 4. Also, Fig. 5 compare the typical one phase current of SR motor obtained for a constant conduction angle without applying controller for PI and Adaptive Sliding speed controllers.

The experimental results of SRM speed control with a rotor-speed magnitude of 200 RPM and are shown in Fig. 6. From waveforms of the Fig. 6, the excellent speed command
tracking is seen in the case of adaptive sliding speed controller, also the low ripple current associated with this controller operation can be observed. Finally to improve the drive control performance the passivity-based current controller is applied. Figs. 7 and 8 show that performance of electromagnetic torque and response of the rotor speed has been increased to a value corresponding to the load. The phase current using PBC is shown in Fig 9.

![Graph](image)

(a)

(b)

Figure 3. Simulation results (SRM speed control without using passivity-based current controller). (a) Motor and load torques. (b) Motor speed.

<table>
<thead>
<tr>
<th>Control strategy</th>
<th>Load Torque 10 Nm</th>
<th>Load Torque 5 Nm</th>
</tr>
</thead>
<tbody>
<tr>
<td>PI</td>
<td>23%</td>
<td>38%</td>
</tr>
<tr>
<td>Adaptive Sliding</td>
<td>1.9%</td>
<td>4.5%</td>
</tr>
</tbody>
</table>

TABLE I. TORQUE RIPPLE COMPARISON

![Graph](image)

(a)

(b)

Figure 4. (a) SRM test setup used for experimentation. (b) Four phase asymmetric converter for SRM.

![Graph](image)

Figure 5. Phase current in 5 Nm Load (scale: phase current= 3A/div) for PI (left) and low adaptive sliding (right).
A nonlinear controller has been presented for a four-phase SRM drive based on the composite passivity-based adaptive sliding technique. Complete model of a SRM possesses two-time-scale characteristics and decomposed as the feedback interconnection of the two electrical and mechanical passive linked subsystems. Hence, by using cascaded torque control structure, the proposed PBC algorithm is designed. Because of tackling the machine physical structure characteristics into account, it can overcome the inherent nonlinear characteristics of the system and is robust to system uncertainties and bounded disturbances. The simulation and experimental results show the proposed controller has improved dynamic performance of rotor speed and torque, also produces lower torque ripple for SRM drives.

REFERENCES