A Sensorless Vector Control Using new BS_PCH Controller structure and SC MRAS Adaptive Speed Observer for Electric Vehicles

NGOC THUY PHAM 1*, THUAN DUC LE1
Dept. of Electrical Engineering Technology
Industrial University of Ho Chi Minh City
VIETNAM

Abstract: - In this paper, a new control structure are proposed to sensorless vector control the in-wheel motor drive system of Electric Vehicle (EV) to improve its performance and robustness. The design of the controller is based on Backstepping and Hamiltonian control combined with a improved stator current MRAS adaptive speed observer proposed to estimate the vehicle speed and it also can compensate for the uncertainties caused by the machine parameter variations, measurement errors, and load disturbances, improving dynamic performance and enhancing the robustness of the SPIM drive system, perfect tuning of the speed reference values, fast response of the motor current and torque, high accuracy of speed regulation. A global EV model is also evaluated based on the vehicle dynamics in this paper. The simulation results lead to the conclusion that the proposed system for the propulsion system of electric vehicle is feasible. The simulation results on a test vehicle propelled by two SPIM showed that the proposed control approach operates satisfactorily.

Key-Words: - Backstepping control, Port Controlled Hamiltonian control, stator current MRAS, neuron network, SPIM.

1 Introduction

In recent years, due to the effect of the greenhouse effect and the lack of fossil energy, a lot of research efforts have focused on developing high quality and efficient EV drive system such as a solution to energy and environmental issues. Electric motor drive is one of the main technologies in EV. An EV is driven by the electric motor, in addition to being environmentally friendly, it also has the advantage of creating a torque that responds quickly and accurately [1-4]. Electric drive system in EV includes: electric motors, converters and controllers. This system is designed to meet the speed and torque requirements set by the driver. For motors used in EVs, SPIM is one of the most competitive proposals for propulsion systems in EVs due to their outstanding advantages such as fault tolerance, reduced phase current ½ with the same source, pressure applied to motors compared to three-phase motors of the same capacity, high torque, low torque, vibration and noise reduction, high efficiency due to reduced rotor current, [5]...

In order to improve the performance of the drive system in the EV, many modern control solutions have been developed such as the field oriente control (FOC), direct torque control (DTC), and model predict control (MPC). However, both DTC and MPC need to analyze and select the vector voltage, the number of these voltage vectors increases exponentially as the number of stator phases increases. With its advantages, FOC Control Technology is the dominant control solution currently applied in the drive system for electric vehicle [6-7]. However, the conventional FOC strategies using PI control cannot provide satisfactory quality for high performance drives. In order to overcome these, recently nonlinear control methods have been developed to replace PID controllers such as [8-18]: Regulatory linear control, sliding mode control, Backstepping control, Fuzzy Logic control and neural network control, prediction control, passive control, Hamiltonia control,.v..v. However, these nonlinear control techniques are usually quite complex, demanding high computational effort, and requiring a precise mathematical model. They were difficult to obtain satisfactory control performance when using independently, especially in the cases applied to control the nonline systems. Therefore, the combination different control techniques to both simple and enhance the performance of SPIM drives, such as combined [19-21] to get a effective control system. In this paper, the author proposes a
combined control structure using BS controller for outer speed control, SPIM parameters are updated for controller to minimizing the effect of changing the parameters on the efficiency of the controller. The PCH controller is proposed for inner current controller to enhance the performance of the in-wheel motor drive system of Electric Vehicle. Beside, to reduce the parameter sensitivity, noise, cost, size, weight and increase the reliability of the system [22], a speed observer instead the mechanical sensors is proposed. In the proposed observer, the reference model uses the stator current components to free of pure integration problems and insensitive to motor parameter variations. In this scheme are, first: Adaptive model uses a two layer linear neural network trained online by a BPN algorithm. Second, the rotor flux is identified by the (VM) Voltage Model with the Rs value is estimated online to enhance the performance of the proposed observer, in addition, using VM will avoid the instability in the regenerating mode. The effectiveness of this proposed schemes for controlling the in-wheel motor drive system of Electric Vehicle is verified by simulation using MATLAB/Simulink.

The paper is organized into five sections, in section 2, the basic theory of the model of the SPIM and the SPIM drive are presented. Section 3 introduces BS_PCH controller and proposed observer. Simulation and discuss are presented in Section 4. Finally, the concluding is provided in Section 5.

2 Model of SPIM and electric propulsion system in EV

2.1 Model of SPIM

The SPIM drive system scheme is provided in Fig.1. The original six-dimensional space of the machine is transformed into three two-dimensional orthogonal subspaces in the stationary reference frame (D-Q), (x - y) and (zl -z2) based on the Vector Space Decomposition (VSD) technique [9]. This transformation is obtained by means of 6x6 transformation matrix (Eq.1).

\[
T_6 = \begin{bmatrix}
1 & -\frac{1}{2} & -\frac{1}{2} & -\frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 0 \\
0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & \frac{1}{2} & \frac{1}{2} & 0 \\
0 & \frac{1}{2} & \frac{1}{2} & -\frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 0 \\
0 & \frac{1}{2} & \frac{1}{2} & -\frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 0 \\
1 & 0 & 0 & 0 & 0 & 0
\end{bmatrix}
\]

The electrical matrix equations in the stationary reference frame for the stator and the rotor may be written as (Eq.2)

\[
\begin{bmatrix}
V_s \\
I_r
\end{bmatrix} = \begin{bmatrix}
R_s & L_s & 0 & 0 & 0 & 0 \\
0 & 0 & R_r & L_r & 0 & 0
\end{bmatrix} \begin{bmatrix}
I_s \\
I_r
\end{bmatrix} + \begin{bmatrix}
0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0
\end{bmatrix} + \begin{bmatrix}
P & 0 & 0 & 0 & 0 & 0 \\
0 & P & 0 & 0 & 0 & 0
\end{bmatrix} \begin{bmatrix}
I_s \\
I_r
\end{bmatrix}
\]

where: [V], [I], [R], [L] and [Lm] are voltage, current, resistant, self and mutual inductance vectors, respectively. P is differential operator. Subscript r and s related to the rotor and stator resistance respectively. Since the rotor is squirrel cage, [Vr] is equal to zero. The electromechanical energy conversion only takes place in the DQ subsystem:

\[
\begin{bmatrix}
V_{m_r} \\
I_{m_r}
\end{bmatrix} = \begin{bmatrix}
R_r & PL_r & 0 & 0 & 0 & 0 \\
0 & 0 & R_r & PL_r & 0 & 0
\end{bmatrix} \begin{bmatrix}
I_{m_r} \\
I_{m_r}
\end{bmatrix} + \begin{bmatrix}
0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0
\end{bmatrix} + \begin{bmatrix}
P & 0 & 0 & 0 & 0 & 0 \\
0 & P & 0 & 0 & 0 & 0
\end{bmatrix} \begin{bmatrix}
I_{m_r} \\
I_{m_r}
\end{bmatrix}
\]

where  is the rotor angular position referred to the stator as shown in Fig. 1. To control for SPIM drives in EV, a transformation matrix must be used to represent the stationary reference fame (α-β) in the dynamic reference (d-q). This matrix is given:

\[
T_2 = \begin{bmatrix}
\cos(\delta) & -\sin(\delta) \\
\sin(\delta) & \cos(\delta)
\end{bmatrix}
\]

Figure. 1. A SPIM drive general diagram

2.2 Model of electric propulsion system

In the FOC control case for SPIM, we have: \(\psi_{s0}=0\), \(\psi_{s\alpha}=\psi_{s\beta}\). The equation of the SPIM drive in the rotary reference coordinates can be written as follows

\[
\begin{align*}
L_r \frac{di_{sd}}{dt} &= -ai_{sd} + L_s \omega i_{sd} + bR_{r}\psi_{rd} + cu_{sd} \\
L_r \frac{di_{sq}}{dt} &= -ai_{sq} + L_s \omega i_{sq} + bR_{r}\psi_{rd} + cu_{sq} \\
\frac{d\omega}{dt} &= \frac{3}{2} \frac{\delta \psi_{rd} (\psi_{rd} i_{sd}) - T_f}{J} - B\omega_r \\
\frac{d\psi_{rd}}{dt} &= \frac{I_{m_r}}{T_f} \frac{1}{T_f} - \frac{1}{T_f} \psi_{rd}
\end{align*}
\]
The electromagnetic torque and the slip frequency can be expressed in dq reference frame:

$$T_e = \frac{3}{2} n_p \frac{L_m}{L_r} \psi_m \alpha d i_q$$  \quad (6)

$$\omega_d = \frac{L_m}{L_r} \psi_m \alpha d i_q$$  \quad (7)

We also have:

$$T_e = T_k + B \omega_r + j \frac{d \omega_r}{dt}$$  \quad (8)

$$\sigma = 1 - \frac{\dot{\psi}_r}{\dot{\psi}_s} \Rightarrow \dot{\psi}_r = \sigma \dot{\psi}_s$$

$$\frac{\sigma}{\alpha} = \frac{\dot{\psi}_r}{\psi_s}$$

$$b = \frac{\dot{\psi}_r}{\psi_s}$$

$$c = \frac{1}{\sigma}$$

$$r_r = \frac{L_r}{R_r}$$

The power required to drive a vehicle at a speed $v$ has to compensate for counteracting forces

$$P = vF_w = v(m_r g + m g \sin f + \frac{1}{2} rAC_d v^2 + m \frac{dv}{dt} + F_L)$$

when using the field-oriented control scheme, the electromagnetic torque $T_e$ can be expressed simply as

$$T_e = K_i i_q$$

where $K_i$ is the torque coefficient; $i_q$ is the q-axis torque current command.

From the vehicle dynamic model shown in (1)–(3), the required torque of a single in-wheel motor $T_m$ can be shown as

$$\frac{T_m}{n} = T_{motor} = K_i i_q$$

where $i_q$ is the required torque current command; $n$ is the number of in-wheel motor drive. The torque equation of an in-wheel motor is as follows

$$T_e = J \frac{d \omega_r}{dt} + B \omega_r + T_{motor}$$

$$= J \frac{d \omega_r}{dt} + B \omega_r + \frac{T_m}{n}$$

where $J$ is the inertia of the in-wheel motor using SPIM including the tyre set; $B$ is the viscous coefficient. Moreover, assuming the system parameter variations of the SPIM are absent, then the overall vehicle dynamic model of EV system including the torque equation of the PMSM can be written as

$$T_e = \left( J + m \frac{r^2}{G^2} \right) \frac{d \omega_r}{dt} + B \omega_r + T_{load} \left( \mu_r g + m g \sin \phi + F_L \right)$$

$$= \left( J + m \frac{r^2}{G^2} \right) \frac{d \omega_r}{dt} + B \omega_r + T_L$$

where $T_L$ is the load torque including the effects of rolling resistance, air drag, hill climbing and external disturbances, and the overbar symbol represents the system parameter in the nominal condition.
3 The proposed sensorless vector control scheme for SPIM drives in EV propulsion system.

3.1. The proposed BS controller for outer speed control and rotor flux loops

As we known, the speed response depends on the inertia of the machine. Therefore, if a fast acting controller is used into the outer speed of vector control of SPIM drives, it can always provide $i_{sq}$ reference current in excess the allowed limit and then the controllers will default work at the critical value. In other words, the output of the speed and rotor loop always work on saturated modes at the limit values, the working time in these modes depend on the sudden change in reference speed and the time constant due to inertia mechanics of the machine. These make the system lost control whenever the controller reaches saturation. To overcome this, the BS controller is proposed, it is quite good from a mechanical mechanic point of view. The stability and performance of the subsystems is studied by Lyapunov theory. Therefore, at each step of the design, a virtual command is created to ensure the convergence of subsystems.

$$\varepsilon_{\omega} = (\omega_{r} - \omega_{r}) + k_{\omega} \int_{0}^{t} (\omega_{r} - \omega_{r}) dt$$

$$\varepsilon_{\psi} = (\psi_{rd} - \psi_{rd}) + k_{\psi} \int_{0}^{t} (\psi_{rd} - \psi_{rd}) dt$$

The error dynamical equations are

$$\dot{\varepsilon}_{\omega} = \omega_{r} - \frac{3}{2} P \sigma_{L} i_{sq} \psi_{rd} + \frac{T_{i}}{J} + B \omega_{r} + k_{\omega} (\omega_{r} - \omega_{r})$$

$$\dot{\varepsilon}_{\psi} = \psi_{rd} + \frac{L_{m}}{\tau_{r}} i_{sd} + \frac{1}{\tau_{r}} \psi_{rd} + k_{\psi} (\psi_{rd} - \psi_{rd})$$

To obtain the virtual controller of speed and rotor flux loop, the following Lyapunov function candidate is considered:

$$V(\varepsilon_{\omega}, \varepsilon_{\psi}) = \frac{1}{2} (\varepsilon_{\omega}^{2} + \varepsilon_{\psi}^{2})$$

Differentiating $V$:

$$\frac{dV(\varepsilon_{\omega}, \varepsilon_{\psi})}{dt} = \varepsilon_{\omega} \frac{d\varepsilon_{\omega}}{dt} + \varepsilon_{\psi} \frac{d\varepsilon_{\psi}}{dt}$$

$$= \varepsilon_{\omega} \left[ \frac{1}{2} \frac{d\omega}{dt} - \frac{3}{2} \sigma_{L} i_{sq} \psi_{rd} + \frac{T_{i}}{J} + B \omega_{r} + k_{\omega} (\omega_{r} - \omega_{r}) \right]$$

$$+ \varepsilon_{\psi} \left[ \frac{1}{\tau_{r}} \psi_{rd} + \frac{L_{m}}{\tau_{r}} i_{sd} + \frac{1}{\tau_{r}} \psi_{rd} + k_{\psi} (\psi_{rd} - \psi_{rd}) \right]$$

where: $k_{\omega}$, $k_{\psi}$ are positive design constants that determine the closed-loop dynamics. To $V' < 0$, the stabilizing virtual controls are chosen

$$i_{sq}^{*} = \frac{1}{k_{ij} \psi_{rd}} \left[ k_{ij} \omega_{r} + \frac{1}{T_{i}} \frac{\partial \omega_{r}}{\partial t} + B \omega_{r} + k_{\omega} (\omega_{r} - \omega_{r}) \right]$$

$$i_{sd}^{*} = \frac{1}{L_{m}} \left[ k_{ij} \omega_{r} + \frac{1}{T_{i}} \frac{\partial \omega_{r}}{\partial t} + B \omega_{r} + k_{\omega} (\omega_{r} - \omega_{r}) \right]$$

We obtain:

$$\frac{dV(\varepsilon_{\omega}, \varepsilon_{\psi})}{dt} = -k_{ij} \omega_{r}^{2} - k_{\psi} \psi_{rd}^{2} < 0$$

3.2 The inner current loop controllers using PCH

In contrast, unlike the speed controllers, requiring with the inner current loop are fast response and highly robust and stable. To meet these control criteria, internal current controller is proposed using PCH control. This controller can effectively compensate for load disturbance in the system so the proposed method is more robust, stability and faster dynamics response. A PCH system with dissipation is a representation of the form:

$$\frac{dx}{dt} = [J(x) - R(x)] \frac{dH}{dx}(x) + g(x)u$$

$$y = g^{T}(x) \frac{dH}{dx}(x)$$

where $R_{x} = R_{x}^{T} > 0$ represents the dissipation. The interconnection structure is captured in matrix $g(x)$ and the skew symmetric matrix $J_{x} = -J_{x}^{T}$, $H(x)$ is the total stored energy function of the system. We define the state vector, input vector and output vector are as follows, respectively

$$x = [x_{1} \ x_{2}]^{T} = [L_{s} i_{sd} \ L_{s} i_{sq}]^{T}$$

$$u = [u_{1} \ u_{2}]^{T} = [bR \psi_{rd} + cu_{sd} \ -bo \psi_{rd} + cu_{sq}]^{T}$$

$$y = [i_{sd} \ i_{sq}]^{T}$$

The Hamiltonian function of the system is given by
\[ H(x) = \frac{1}{2} x^T D^{-1} x = \frac{1}{2} \left( x_1^2 + x_2^2 \right) = \frac{1}{2} \left( L_{s}i_{sd}^2 + L_{r}i_{rd}^2 \right) \] (23)

Suppose we wish to asymptotically stabilize the system (23) around a desired equilibrium \( x_0 \), a closed-loop energy function \( H_d(x) \) is assigned to the system which has a strict minimum at \( x_0 \) (that is, \( H_d(x) > H_d(x_0) \) for all \( x \neq x_0 \) in a neighborhood of \( x_0 \)). The feedback stabilization theory of PCH system is given as follows [17]. Given \( J(x), R(x), H(x), g(x) \) and the desired equilibrium \( x_0 \). Assume we can find a feedback control \( u = \alpha(x) \), \( R(x) \), \( g(x) \) and a vector function \( K(x) \) satisfying:

\[
\frac{dK}{dx}(x) = \left[ \frac{dK}{dx}(x) \right]^T \; ; \; K(x_0) = -\frac{dH_d}{dx}(x_0);
\]

\[
\frac{dK}{dx}(x_0) = \frac{d^2H_d}{dx^2}(x_0) > 0
\]

(24)

The closed-loop system:

\[
\frac{dx}{dt} = \left[ J_d(x) - R_d(x) \right] \frac{dH_d}{dx}(x)
\]

will be a PCH system with dissipation.

\[
K(x) = \frac{dH_d}{dx}; \quad H_d(x) = H_d(x) - H(x)
\]

(25)

where \( H_d \) is the energy added to the system and \( x_0 \) will be a stable equilibrium of the closed-loop system. The expected Hamiltonian energy storage function is defined as

\[
H_d(x) = H(Ax) = H(x - x_0)
\]

(26)

\[ J_d(x) = J(x) + J_d(x) = -J_d^T(x) \]

(27)

\[ R_d(x) = R(x) + R_d(x) = R_d(x) > 0 \]

(28)

\[
J_a(x) = \begin{bmatrix} 0 & J_1 \\ -J_1 & 0 \end{bmatrix}; \quad R_a(x) = \begin{bmatrix} - \eta & 0 \\ 0 & - \eta \end{bmatrix}
\]

(29)

where, \( J_1, r1 \) and \( r2 \) are undetermined interconnect and damping parameters. According to equations (25)-(30), the controller of the current inner loop of the motor is

\[
\begin{align*}
\hat{u}_a^d &= \sigma \left[ a_{id}i_{id} + r_1 \left( \hat{i}_{id}^* - i_{id} \right) - J_1 \left( \hat{i}_{iq}^* - i_{iq} \right) - L_s \omega_{r}i_{aq} \right] \\
\hat{u}_a^* &= \sigma \left[ b_{id}\hat{q}_{id} + r_2 \left( \hat{i}_{id}^* - i_{id} \right) + J_1 \left( \hat{i}_{iq}^* - i_{iq} \right) + L_s \omega_{r}i_{aq} \right]
\end{align*}
\]

(30)

### 3.3 Structure of the NN_SC_MRAS Speed Observer

In this scheme, the measured stator current components are also used as the reference model of the MRAS observer to avoid the use of a pure integrator and reduce influence of motor parameter variation as in [25]. An Artificial Neural Network (ANN) can reproduce the equation (31), where \( w_1, w_2, w_3, w_4 \) are the weights of the neural networks defined as (31);
3.3.2.2 Stator Resistance Online Estimation Algorithm.

The weight adjustment has to be proportional to the negative of the error gradient to obtain a minimum squared error between actual and estimated stator current. 

\[
\frac{\partial E}{\partial w_i} = \frac{\partial E}{\partial \hat{i}_s(k)} \frac{\partial \hat{i}_s(k)}{\partial w_i} 
\]

where:

\[
\frac{\partial E}{\partial \hat{i}_s(k)} = \frac{1}{2} \frac{\partial}{\partial \hat{i}_s(k)} [i_s(k) - \hat{i}_s(k)]^2 = -\epsilon_i^T(k) 
\]

\[
\frac{\partial \hat{i}_s(k)}{\partial w_i} = \left[ \left( \psi_{sQ}(k - 1) - \hat{\psi}_{sQ}(k - 1) \right) \right]^T 
\]

Substituting (15) and (16) into (14):

\[
-\frac{\partial E}{\partial w_i} = \epsilon_{sO}^T(k) \psi_{sQ}(k - 1) - \epsilon_{sQ}^T(k) \psi_{sO}(k - 1) 
\]

The weight adjustment law of the NN_VM_SC_MRAS observer can be derived based on backpropagation network (BPN) algorithm and can be written:

\[
\Delta \omega_s(k) = -\eta \frac{\partial E}{\partial \hat{\omega}_s} 
\]

\[
= -\eta \left[ \epsilon_{sO}^T(k) \psi_{sO}(k - 1) - \epsilon_{sQ}^T(k) \psi_{sQ}(k - 1) \right] 
\]

where \( \eta \) is a positive constant called the learning rate. Hence the weight update equation:

\[
\omega_s(k) = \omega_s(k - 1) + \Delta \omega_s(k) + \alpha \Delta \omega_s(k) 
\]

where \( \alpha \) is a positive constant called the momentum constant. Due to simple structure of the NN stator current observer, the weight adjustment can be performed online and the stator resistance can be estimated from the weight w1 as:

\[
\hat{R}_s(k+1) = \frac{\sigma L_s}{T_s} \left[ 1 - \frac{T_s L_s}{\sigma L_s L_m} \right] - \omega_s(k+1) 
\]

3.3.2 Rotor Flux and Stator Resistance Online Estimation Algorithm.

3.3.2.1 Rotor Flux identifier

In this proposed observer, the rotor flux components are identified based on VM equation (42) to generate the estimated rotor flux value what are provided for the adaptive model of the proposed observer.

\[
\begin{align*}
\frac{dv_{sO}}{dt} &= \frac{x_s}{x_m} (v_{sO} - R_{sO} i_{sO} - \sigma T_n x_i p_{sO}) \\
\frac{dv_{sQ}}{dt} &= \frac{x_s}{x_m} (v_{sQ} - R_{sQ} i_{sQ} - \sigma T_n x_i p_{sQ})
\end{align*}
\]

3.3.2.2 Stator Resistance Estimator

From (32), (33) to minimize squared error between actual and estimated stator current the weight adjustment has to be proportional to the negative of the error gradient with respect to the weight. It can be written as:

\[
w_s(k) = w_s(k - 1) + \Delta w_s(k) + \gamma \Delta w_s(k) 
\]

where

\[
\Delta w_s(k) = -\beta \frac{\partial E}{\partial w_s} 
\]

\[
= -\beta \left[ \epsilon_{sO}(k) i_{sO}(k - 1) + \epsilon_{sQ}(k) i_{sQ}(k - 1) \right] 
\]

where \( \gamma \) is a positive constant called the momentum constant, \( \beta \) is a positive constant called the learning rate. Due to simple structure of the NN stator current observer, the weight adjustment can be performed online and the stator resistance can be estimated from the weight w1 as:

\[
\hat{R}_s(k+1) = \frac{\sigma L_s}{T_s} \left[ 1 - \frac{T_s L_s}{\sigma L_s L_m} \right] - \omega_s(k+1) 
\]
standard test driving cycle with data points representing the vehicle reference speed over time. It is characterized by low vehicle speed (maximum of 50 km / h) and it is suitable electric vehicle testing in urban areas. (Fig 7b).

+ Test 3: This survey is conducted with a sample cycle similar to the survey 2 but based on [27, Fig 8] with the extended survey time with 5 test cycles. as in [27]. The test performed on EVs based on ECE-15 test cycle. The goal of this test is to evaluate the performance and dynamic response of the sensorless vector control using BS_PCH in combination with proposed observer when applying for the propulsion system in EV for 5 consecutive cycles to prove the control quality, stability, robustness of the control and estimation strategies. The test cycle is the urban ECE-15 cycle [27]. From the simulation results show in Fig 5, it is easy to see that the speed responses in both tests 1 and 2 in case 1 is very good, the speed error is nearly zero in the stable mode. BS_PCH controller can provide very good dynamic response during acceleration and deceleration of EV.

Test 3 performed on EVs based on ECE-15 test cycle. The goal of this test is to evaluate the performance and dynamic response of the sensorless vector control using BS_PCH in combination with proposed observer when applying for the propulsion system in EV for 5 consecutive cycles to prove the control quality, stability, robustness of the proposed control and estimation strategies.

Observing the simulation results shown in Fig. 6, we see that the NN_SC_MRAS speed observer works correctly and stably, the estimated speed is almost exactly equal to the reference and measured speed, in both the established and transient mode, even at zero and low speed regions.

![Fig. 7 Speed, stator current, torque response and estimation error:](image)

- a. First test cycle with reference speed: 0-270-135-270 rpm [26, Fig. 5 abc, ghi]).
- b. Second test cycle: Modified ECE-15 (Fig. 8) [26, Fig. 2].

![Fig. 8 Speed, stator current, torque response, estimated error in the third test cycle based on [27].](image)
Case 2:
Tests verify and evaluate the performance of the proposed SPIM drive using in EV based on recommended benchmark tests in [28, Fig.6] also make with describe the road during the test EV operation as shown in Fig. 9.

The survey results shown in Fig. 10 verified the effectiveness of the proposed drive for the wheel using SPIM. This proposed scheme provide fast, accurate speed and torque responses, control and estimation quality are very good during both stability and dynamic modes. it is suitable the requirements of the propulsion system of EV.

Fig. 9 Describe the road during the test EV operation[28]

Fig. 10 Speed, estimation error, stator current, torque response in during the EV test operation based on test in [140, Fig.6].

5 Conclusion
This paper are proposed a sensorless vector control scheme for the in-wheel motor drive system of Electric Vehicle (EV) using a new controllerstructure combined with a stator current MRAS based on adaptive speed observer using neuron network. The simulation results are performed to prove the successful development of the proposed sensorless vector control scheme of SPIM drives for the propulsion system of electric vehicle. The speed of SPIM proving for EV follows accurately the reference speed set commands, instantaneous torque response. The proposed scheme meets the requirements of the propulsion system of EV, even at zero and low speed regions. In fact, electric vehicle performance is greatly influenced by the quality of the vehicle's control system, electric motor,… the application of modern control techniques to improve and enhance the performance of electric vehicles have high practical significance.

References


Creative Commons Attribution License 4.0 (Attribution 4.0 International , CC BY 4.0)

This article is published under the terms of the Creative Commons Attribution License 4.0
https://creativecommons.org/licenses/by/4.0/deed.en _US