Direct torque control based on second order sliding mode controller for three-level inverter-fed permanent magnet synchronous motor: comparative study

Introduction. The permanent magnet synchronous motor (PMSM) has occupied a large area in the industry because of various benefits such as its simple structure, reduced moment of inertia, and quick dynamic response. Several control techniques have been introduced for the control of the PMSM. The direct torque control strategy associated to three-level clamped neutral point inverter has been proven its effectiveness to solve problems of ripples in both electromagnetic torque and stator flux with regard to its significant advantages in terms of fast torque response. Purpose. The use of a proportional integral speed controller in the direct torque control model results in a loss of decoupling with regard to parameter fluctuations (such as a change in stator resistance value induced by an increase in motor temperature), which is a significant drawback for this method at high running speeds. Methods. That is way a second order sliding mode controller based on the super twisting algorithm (STA) was implemented instead of PI controller to achieve a decoupled control with higher performance and to ensure stability while dealing with parameter changes and external disturbances. Results. The simulation results carried out using MATLAB/Simulink software show that the model of direct torque control based on a three-level inverter-fed permanent magnet synchronous motor drive has better performance for second order sliding mode controller than the proportional integral controller. Through the response characteristics we see greater performance in terms of response time and reference tracking without overshoots. Decoupling, stability, and alignment toward equilibrium are all guaranteed. References 9, table 2, figures 9.

Key words: permanent magnet synchronous motor, direct torque control, second order sliding mode controller.
to a reduction of the machine equations. The transformation best known by electrical technicians is that of Park, the representation is shown in Fig. 1.

![Fig. 1. Two-phase representation of PMSM](image)

The stator voltage equations are given by the following equations in the matrix form [3]:

$$
\begin{bmatrix}
V_{dq}
\end{bmatrix} = \begin{bmatrix}
R_s & L_d \\
L_q & R_s
\end{bmatrix} \begin{bmatrix}
i_d \\
i_q
\end{bmatrix} + \frac{d}{dt} \begin{bmatrix}
\phi_d \\
\phi_q
\end{bmatrix} + p \cdot \Omega \begin{bmatrix}
\phi_d^* \\
\phi_q^*
\end{bmatrix},
$$  \hspace{1cm} (1)

where

$$
\begin{bmatrix}
V_{dq} \\
R_s \\
i_d \\
i_q
\end{bmatrix} = \begin{bmatrix}
V_d \\
0 \\
i_d \\
i_q
\end{bmatrix},
$$  \hspace{1cm} (2)

$$
\begin{bmatrix}
L_d \\
R_s \\
i_d \\
i_q
\end{bmatrix} = \begin{bmatrix}
0 \\
L_q \\
i_d \\
i_q
\end{bmatrix},
$$  \hspace{1cm} (3)

$$
\begin{bmatrix}
\phi_d \\
\phi_q
\end{bmatrix} = \begin{bmatrix}
\phi_d^* \\
\phi_q^*
\end{bmatrix},
$$  \hspace{1cm} (4)

$$
\begin{bmatrix}
\phi_d \\
\phi_q
\end{bmatrix} = \begin{bmatrix}
\phi_d^* \\
\phi_q^*
\end{bmatrix},
$$  \hspace{1cm} (5)

We note also that:

$$
\begin{bmatrix}
\phi_d \\
\phi_d
\end{bmatrix} = \begin{bmatrix}
\phi_d^* \\
\phi_d^*
\end{bmatrix} = \begin{bmatrix}
\phi_d \\
\phi_d
\end{bmatrix}.
$$  \hspace{1cm} (6)

where $V_d$, $V_q$, $i_d$, $i_q$, $L_d$, $L_q$, $\phi_d$ and $\phi_q$ are the $dq$ components of the stator voltage, current, inductance and flux linkage, respectively; $R_s$ is the stator resistance; $\phi_f$ is the rotor flux linkage generated by the permanent magnets; $p$ is the pairs of poles.

The electromagnetic torque expression is given by:

$$
T_e = \frac{3}{2} \cdot p \cdot \left[ L_d - L_q \right] \cdot i_d \cdot i_q + \phi_f \cdot i_q,
$$  \hspace{1cm} (7)

In the case where the machine has non-salient poles ($L_d = L_q$), this equation (7) is simplified to:

$$
T_e = \frac{3}{2} \cdot p \cdot \phi_f \cdot i_q.
$$  \hspace{1cm} (8)

**DTC with three-level inverter.** The principle of the DTC is to maintain the stator flux within a specific range [3, 4]. This technique is based on the direct determination of the commands sequences applied to the switches of a three level inverter. This strategy is generally based in the use of hysteresis comparators whose role is to control the amplitudes of the stator flux and the electromagnetic torque. The synoptic of DTC control is shown in Fig. 2.

Once the 2 flux components are obtained, the electromagnetic torque can be estimated by the formula below:

$$
C_e = \frac{3}{2} \cdot p \cdot \left[ \phi_{q\alpha} I_{q\beta} - \phi_{q\beta} I_{q\alpha} \right].
$$  \hspace{1cm} (11)

Moreover, in order to obtain the sector, the rotor flux angle is determined by:

$$
\theta = \arctg \frac{\phi_{q\beta}}{\phi_{q\alpha}}.
$$  \hspace{1cm} (12)

This model is updated with 3 level hysteresis controller for the flux and 5 level for the torque in order to build the optimized switching table as illustrated in Table 2 that led to determine a vector between 27 state vectors to apply to a three level NPC inverter (Fig. 3) noting that those vectors are distributed on 12 sectors of the stator flux plane Fig. 4.

![Fig. 2. The synoptic of DTC control](image)

![Fig. 3. One leg of 3 level inverter layout](image)

![Fig. 4. 12 sectors with switching vectors](image)
In 3-level NPC inverter there are 3 switching state for each leg (S1, S2); (S1, S1'); (S1', S2') which result in 3 voltages levels respectively $V_{dc}$, $V_{dc}/2$, 0. Consequence of these switching possibilities, 27 state vectors will be obtained as shown in Table 1 [5, 6].

### Table 1

| Distribution of the 3-level inverter voltage vectors into 4 groups |
|---------------------------------------------------------------|
| **Zero state vectors**                                      |
| \( V_1 \), \( V_2 \), \( V_{15} \) | \((0, 0, 0)(1, 1, 1)(-1, -1, -1)\) |
| **Short Vectors**                                           |
| \( V_2 \), \( V_3 \), \( V_{14} \), \( V_{15} \) | \((1, 0, 0)(1, 1, 0)(0, 1, 1)(0, 0, 1)\) |
| **Long vectors**                                            |
| \( V_{16} \), \( V_{17} \), \( V_{18} \), \( V_{19} \), \( V_{20} \) | \((1, -1, -1)(1, 1, -1)(-1, 1, 1)(1, -1, 1)\) |
| **Medium vectors**                                          |
| \( V_{22} \), \( V_{23} \), \( V_{24} \), \( V_{25} \), \( V_{26} \), \( V_{27} \) | \((0, -1, 0)(0, 0, 0)(0, 1, 0)(0, -1, 1)(1, -1, 0)\) |

Digital simulation using MATLAB/Simulink has been used to test the techniques described in this paper. In this simulation, the frequency is 50 Hz, stator resistance is 2.3 $\Omega$, inductance \( L_d = L_q = 7.6 \, \text{mH} \), moment of inertia is 0.032 kg-m², permanent flux is 0.4 T and number of poles is 4. The PMSM starts with a constant reference speed equal to 100 rad/s. At \( t = 0.2 \, \text{s} \) the rotor speed decreases to 80 rad/s. At \( t = 0.5 \, \text{s} \) a reverse of rotation to –100 rad/s was performed finally at \( t = 0.7 \, \text{s} \), a nominal load torque \( T_L = 5 \, \text{N-m} \) was applied, then removed at \( t = 0.9 \, \text{s} \).

### Second order sliding mode controllers (SOSMC)

The conventional sliding mode controller is known by the chattering phenomena, to address this issue, we proposed to extend the basic SMC model to a second derivative of the sliding surface (SOSMC), with the objective of minimizing the chattering band [2]. The equivalent component remains the same, the switching component become:

\[
T_n = K_\Omega \cdot \text{sign}(S_\Omega),
\]

where \( K_\Omega \) and \( K_\Omega \) are the positive constants.

### Simulation results

The controller parameters \( K_p, K_i \) become as:

\[
K_p = 2 \cdot \rho^2 \cdot J;
K_i = 2 \cdot \rho \cdot J \cdot f_r,
\]

where \( \rho \) represents the module of the real part and imaginary part of the 2 poles.

### Conventional sliding mode controller

The sliding mode controller has been built to control the speed to ensure good tracking, accurate response and insensitivity to changes in drive system [2, 7]. The sliding surface has been selected as:

\[
S_\Omega = \Omega_{ref} - \Omega.
\]

In addition, the electromechanical equation of the motor is expressed by:

\[
\frac{d\Omega}{dt} + f_r \cdot \Omega = T_e - T_L.
\]

By considering (17), the derivative of (16) becomes:

\[
\dot{S}_\Omega = \Omega_{ref} - \frac{1}{J}(T_e - T_L - f_r \cdot \Omega).
\]

The reference control variable is written such as:

\[
T_{eq} = T_{ref} - T_n,
\]

where \( T_{eq} \) and \( T_n \) are the equivalent and switching components of the control variable, respectively.

In the sliding mode \( S = 0 \), the equivalent component is determined by:

\[
T_{eq} = J \cdot \dot{\Omega}_{ref} + T_L + f_r \cdot \Omega.
\]

Moreover, the switching component is written by

\[
T_n = K_\Omega \cdot \text{sign}(S_\Omega),
\]

The functional diagram of speed controller is shown in Fig. 5.

![Fig. 5. Speed control loop](image)

Adopting the pole placement method considering that the closed loop speed transfer function is given by:

\[
F_C = \frac{\omega}{\omega_{ref}} = \frac{K_p}{J \cdot P^2 + (f_r + K_p) P + K_i}.
\]

The controller parameters \( K_p, K_i \) become as:

\[
\begin{align*}
K_p &= 2 \cdot \rho^2 \cdot J; \\
K_i &= 2 \cdot \rho \cdot J \cdot f_r,
\end{align*}
\]

where \( \rho \) represents the module of the real part and imaginary part of the 2 poles.
The return to the equilibrium point was 0.02 s less than the proportional integral controller. In the other hand, the dynamic performance and steady-state accuracy of the PI controller were not very satisfactory since the motor speed sensitivity to load disturbance and motor variations was very high (1 rad/s) then it was improved and reduced to 0.1 rad/s with the sliding mode controller.

**Conflict of interest.** The authors declare that they have no conflicts of interest.

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**Conclusions.** In this paper a comparative study of proportional integral and second order sliding mode controller in a direct torque control system based on a 3-level neutral point clamped inverter-fed permanent magnet synchronous motor drive has been presented.

Simulation results prove that second order sliding mode controller provide better tracking performances than the proportional integral in terms of rise time and overshoot as well as less sensitivity of motor speed to load disturbance.

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