UDC 621.3.013 https://doi.org/10.20998/2074-272X.2022.5.02

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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 insure 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 with second order sliding mode speed 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 convergence 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.**

Вступ*. Синхронний двигун з постійними магнітами (СДПМ) зайняв велике місце в промисловості завдяки різним перевагам, таким як його проста конструкція, зменшений момент інерції та швидкий динамічний відгук. Для керування СДПМ було введено декілька методів керування. Стратегія прямого управління крутним моментом, пов'язана з трирівневим інвертором з фіксованою нейтральною точкою, довела свою ефективність для вирішення проблем пульсацій як електромагнітного крутного моменту, так і магнітного потоку статора, враховуючи його значні переваги з точки зору швидкої реакції крутного моменту.* ***Мета****. Використання пропорційно-інтегрального регулятора швидкості моделі прямого управління крутним моментом призводить до втрати розв'язки по відношенню до коливань параметрів (таких як зміна значення опору статора, викликане підвищенням температури двигуна), що є істотним недоліком для цієї моделі на високих робочих швидкостях.* ***Методи****. Таким чином, замість ПІ-регулятора був реалізований ковзний регулятор другого порядку, заснований на алгоритмі суперскручування (STA), для досягнення розв'язаного управління з більш високою продуктивністю та забезпечення стабільності при роботі зі змінами параметрів та зовнішніми збуреннями.* ***Результати*** *моделювання, виконаного з використанням програмного забезпечення MATLAB/Simulink, показують, що модель прямого керування крутним моментом, заснована на трирівневому приводі синхронного двигуна з постійними магнітами з інверторним живленням, має кращу ефективність з регулятором швидкості з ковзним режимом другого порядку, ніж пропорційний інтегральний контролер. Завдяки характеристикам відгуку бачимо більш високу ефективність з точки зору часу відгуку та відстеження посилань без перерегулювання. Розв'язка, стабільність та збіжність до рівноваги гарантовані..* Бібл. 9, табл. 2, рис. 9.

Ключові слова: **синхронний двигун з постійними магнітами, пряме керування крутним моментом, контролер режиму ковзання другого порядку.**

Introduction. In recent years, thanks to the rapid development of power electronics technology, permanent magnet synchronous motor (PMSM) has taken a wide space in the industry due to many advantages such as higher energy efficiency and higher torque to weight ratio. In the term of its control, several techniques have been presented in the literature; such as voltage/frequency control, which is limited to low-performance uses, and field oriented control, which has a complex process and is highly sensitive to parameter changes. nevertheless, the direct torque control (DTC) has been the most attractive control due to its notable advantages regarding its simplicity and its rapid torque response [1], This command need an external loop to control the speed either by the classic PID controller [8] or through other methods of control such as: fuzzy logic control [9], sliding mode [2], as well as the second order sliding mode controller (SOSMC) [2] which is designed to achieve high performance for systems with parametric variation and to ensure closed loop stability.

The purpose of this work is to compare between the PI and the second order sliding mode speed controllers in a direct torque control for three-level neutral point multilevel inverter-fed permanent magnet synchronous motor drive system.

In this paper, the mathematical model of PMSM has been introduced. In order to obtain a better performance for the PMSM, the technique of DTC associated to three-level neutral point clamped (NPC) multilevel inverters with a PI controller has been used. However, this solution has a major drawback represented by the speed's sensitivity to load variation. In order to remedy this problem, the DTC with the sliding mode controller (SMC) was generally used to improve its robustness and its insensitivity to parameter variations and external disturbances. However, the presence of the chattering is a major drawback for conventional SMC controllers. In order to minimize the phenomena, a SOSMC was used. For the purpose of evaluating and testing the different control techniques proposed for the PMSM fed by three-level NPC inverters drive system, a simulation study was carried out. The comparison study based on the obtained simulation results confirms the efficiency of the second order sliding mode controller over the PI controller under different operating conditions. [2].

Model and equations of the PMSM. The two-phase model of the PMSM is carried out by a transformation of the real three-phase reference into a fictitious two-phase reference, which is in fact only a change of base on the physical quantities (tensions, fluxes and currents), it leads 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.

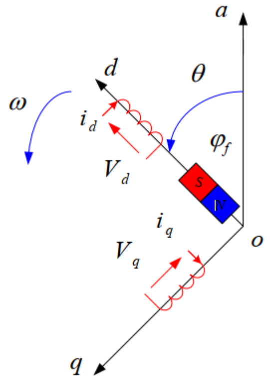

Fig. 1. Two-phase representation of PMSM

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

$$[V_{dq}]=[R_s]\cdot[i_{dq}]+\frac{d}{dt}[\varphi_{dq}]+p\cdot\Omega[\varphi^*_{dq}], \quad (1)$$

where

$$[V_{dq}]=\begin{bmatrix}V_d\\V_q\end{bmatrix}, \quad (2)$$

$$[R_s]=\begin{bmatrix}R_s & 0\\0 & R_s\end{bmatrix}, \quad (3)$$

$$[i_{dq}]=\begin{bmatrix}i_d\\i_q\end{bmatrix}, \quad (4)$$

$$[\varphi_{dq}]=\begin{bmatrix}\varphi_d\\\varphi_q\end{bmatrix},\quad \varphi^*_{dq}=\begin{bmatrix}-\varphi_q\\\varphi_d\end{bmatrix}. \quad (5)$$

We note also that:

$$\begin{bmatrix}\varphi_d\\\varphi_q\end{bmatrix}=\begin{bmatrix}L_d & 0\\0 & L_q\end{bmatrix}\cdot\begin{bmatrix}i_d\\i_q\end{bmatrix}+\begin{bmatrix}\varphi_f\\0\end{bmatrix}. \quad (6)$$

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

The electromagnetic torque expression is given by:

$$T_e=(3/2)\cdot p\cdot\left[(L_d-L_q)\cdot i_d\cdot i_q+\varphi_f\cdot i_q\right]. \quad (7)$$

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

$$T_e=(3/2)\cdot p\cdot\varphi_f\cdot i_q\,. \quad (8)$$

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.

The stator flux equation is expressed as:

$$\overline{\phi_s}=\hat{\phi}_{s\alpha}+j\hat{\phi}_{s\beta}\,, \quad (9)$$

where

$$\begin{cases}\hat{\phi}_{s\alpha}=\int_0^t (V_{s\alpha}-R_s\,I_{s\alpha})\,\mathrm{d}t;\\ \hat{\phi}_{s\beta}=\int_0^t (V_{s\beta}-R_s\,I_{s\beta})\,\mathrm{d}t,\end{cases} \quad (10)$$

where $V_{s\alpha}$, $V_{s\beta}$, $i_{s\alpha}$, $i_{s\beta}$, are the $\alpha\beta$ components of the stator voltage and current, respectively.

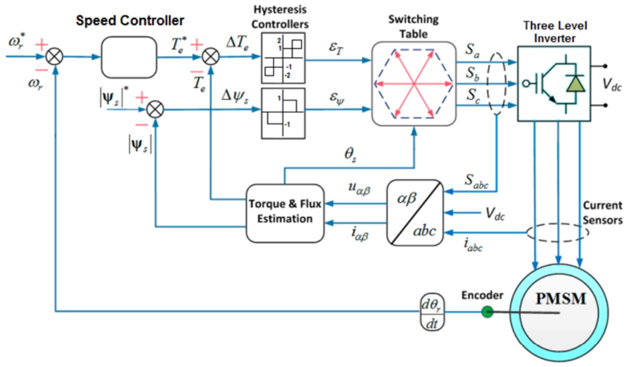

Fig. 2. The synoptic of DTC control

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

$$\hat{C}_e=\frac{3}{2}p\left[\hat{\phi}_{S\alpha}\,I_{S\beta}-\hat{\phi}_{S\beta}\,I_{S\alpha}\right]. \quad (11)$$

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

$$\theta=\mathrm{arctg}\frac{\hat{\phi}_{S\beta}}{\hat{\phi}_{S\alpha}}\,. \quad (12)$$

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.

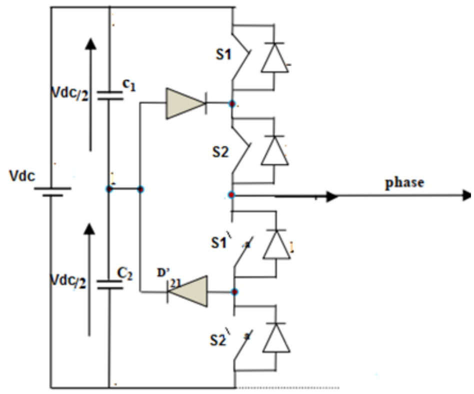

Fig. 3. One leg of 3 level inverter layout

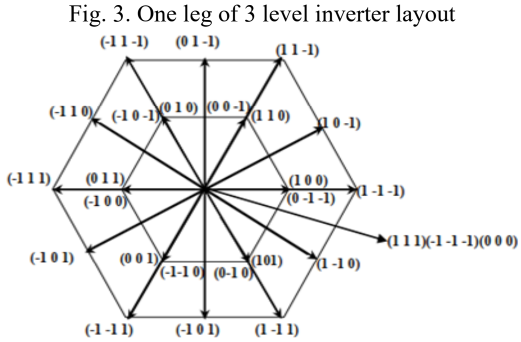

Fig. 4. 12 sectors with switching vectors

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_8 V_{15} (0, 0, 0) (1, 1, 1) (–1, –1, –1)
Short Vectors	V_2 V_3 V_4 V_5 V_6 V_7 (1, 0, 0) (1, 1, 0) (0, 1, 0) (0, 1, 1) (0, 0, 1) (1, 0, 1) V_8 V_9 V_{10} V_{11} V_{12} V_{13} (0, –1, –1) (0, 0, –1) (–1, 0, –1) (–1, 0, 0) (–1, –1, 0) (0, –1, 0)
Long vectors	V_{16} V_{17} V_{18} V_{19} V_{20} V_{21} (1, –1, –1) (1, 1, –1) (–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} (1, 0, –1) (0, 1, –1) (–1, 1, 0) (–1, 0, 1) (0, –1, 1) (1, –1, 0)

Table 2
DTC modified switching table

Φ	T_e	Stator flux sectors											
		1	2	3	4	5	6	7	8	9	10	11	12
+1	+2	22	17	23	18	24	19	25	20	26	21	27	16
	+1	22	3	23	4	24	5	25	6	26	7	27	2
	0	1	8	15	1	8	15	1	8	15	1	8	15
	–1	27	2	22	3	23	4	24	5	25	6	26	7
	–2	27	16	22	17	23	18	24	19	25	20	26	21
0	+2	23	18	24	19	25	20	26	21	27	16	22	17
	+1	23	4	24	5	25	6	26	7	27	2	22	3
	0	1	8	15	1	8	15	1	8	15	1	8	15
	–1	26	7	27	2	22	3	23	4	24	5	25	6
	–2	26	21	27	16	22	17	23	18	24	19	25	20
–1	+2	18	24	19	25	20	26	21	27	16	22	17	23
	+1	4	24	5	25	6	26	7	27	2	22	3	23
	0	1	8	15	1	8	15	1	8	15	1	8	15
	–1	6	26	7	27	2	22	3	23	4	24	5	25
	–2	20	26	21	27	16	22	17	23	18	24	19	25

PI speed controller. The PI controller determines the reference torque in order to maintain the corresponding speed [2]. The speed dynamics is given by the following mechanical equation:

$$\Omega = \frac{T_e - T_L}{JP + f_r}, \tag{13}$$

where T_e, T_L, J, f_r, P are the motor torque, load torque, moment of inertia, viscous friction factor and Laplace operator, respectively.

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

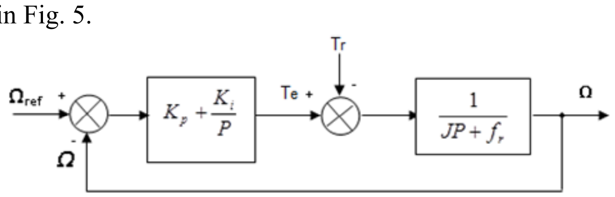

Fig. 5. Speed control loop

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

$$F_C = \frac{\Omega}{\Omega_{ref}} = \frac{K_p \cdot \left(P + \frac{K_i}{K_p}\right)}{J \cdot P^2 + \left(f_r + K_p\right) \cdot P + K_i}. \tag{14}$$

The controller parameters K_P, K_i become as:

$$\begin{cases} K_i = 2 \cdot \rho^2 \cdot J; \\ K_p = 2 \cdot \rho \cdot J - f_r, \end{cases} \tag{15}$$

where ρ 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:

$$S_\Omega = \Omega_{ref} - \Omega. \tag{16}$$

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

$$J \cdot \frac{d\Omega}{dt} + f_r \cdot \Omega = T_e - T_L. \tag{17}$$

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

$$\dot{S}_\Omega = \dot{\Omega}_{ref} - \frac{1}{J} \cdot \left(T_e - T_L - f_r \cdot \Omega\right). \tag{18}$$

The reference control variable is written such as:

$$T_{ref} = T_{eq} - T_n, \tag{19}$$

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

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

$$T_{eq} = J \cdot \dot{\Omega}_{ref} + T_L + f_r \cdot \Omega. \tag{20}$$

Moreover, the switching component is written by

$$T_n = K_\Omega \cdot \operatorname{sign}(S_\Omega), \tag{21}$$

where K_Ω is the positive coefficient.

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:

$$T_n = K_{\Omega 1}\sqrt{\left|S_\Omega\right|}\operatorname{sign}(S_\Omega) + K_{\Omega 2}\int \operatorname{sign}(S_\Omega)dt, \tag{23}$$

where $K_{\Omega 1}$ and $K_{\Omega 2}$ are the positive constants.

Simulation results. 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 Ω, inductance L_d = L_q = 7.6 mH, moment of inertia is 0.032 kg·m^2, 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 s the rotor speed decreases to 80 rad/s. At t = 0.5 s a reverse of rotation to –100 rad/s was performed finally at t = 0.7 s, a nominal load torque T_L = 5 N·m was applied, then removed at t = 0.9 s.

In the instant t = 0.7 s when applying the load, a speed drop from –100 rad/s to –101 rad/s was noticed with the PI controller unlike the SOSMC model where the speed remains in an excellent range maintaining its reference (Fig 6, 7). Also an overshoot of the speed is observed when decreasing the speed to 80 rad/s (t = 0.2 s), contrary to the SOSMC results where the speed keep tracking the reference without overshooting (Fig. 6, 7).

As can be seen, compared with model of PI-DTC, the model of SOSMC-DTC has a better dynamic response for both speed and torque indicating that SOSMC controller was less sensitive to the load disturbance.

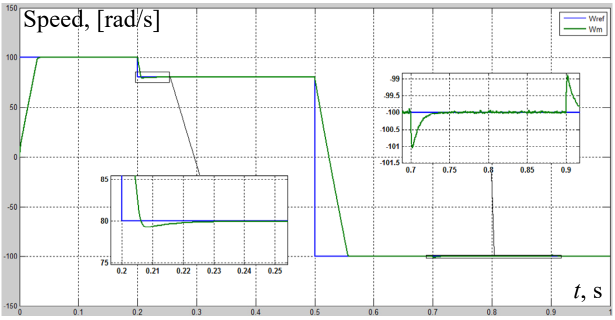

Fig. 6. Rotor speed response with PI controller

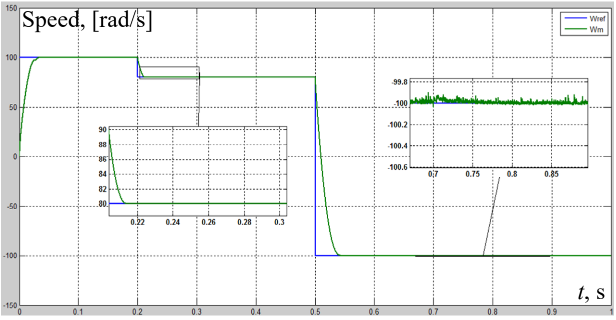

Fig. 7. Rotor speed response with SOSMC

When the rotation was reversed, the motor torque required longer time to reach equilibrium (0.06 s), this time was clearly reduced in SOSMC model (Fig. 8, 9).

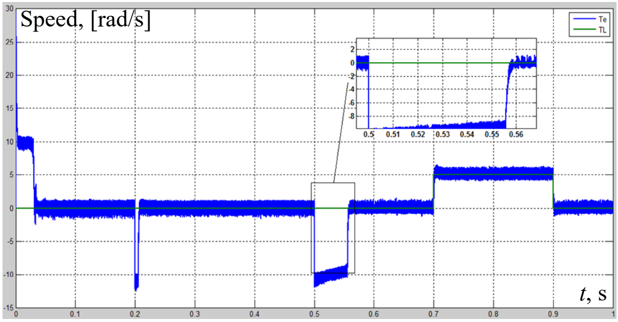

Fig. 8. Motor torque response with PI controller

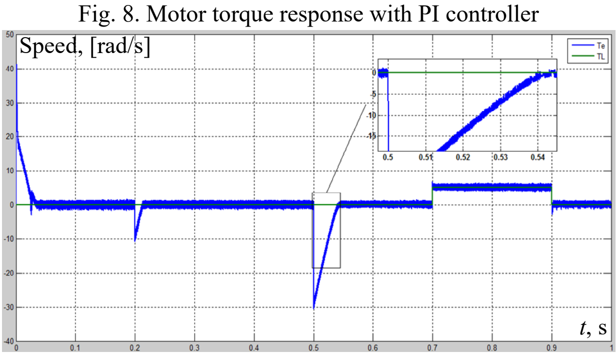

Fig. 9. Motor torque response with SOSMC

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. 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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Received 25.03.2022
Accepted 28.06.2022
Published 07.09.2022

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How to cite this article:
Guezi A., Bendaikha A., Dendouga A. Direct torque control based on second order sliding mode controller for three-level inverter-fed permanent magnet synchronous motor: comparative study. *Electrical Engineering & Electromechanics*, 2022, no. 5, pp. 10-13. doi: https://doi.org/10.20998/2074-272X.2022.5.02