Fault Tolerant Model Predictive Control of Three-Phase Permanent Magnet Synchronous Motors

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Abstract: - A new fault tolerant model predictive control (FTMPC) strategy is proposed for three-phase magnetically isotropic permanent magnet synchronous motor (PMSM) with complete loss of one phase (LOP) or loss of one leg (LOL) of the inverter. The dynamic model of PMSM with LOP or LOL is derived in *abc*-System. The principle of FTMPC is investigated, its predictive model for remaining two stator phase currents is established after LOP or LOL occurs, and the flux estimator based on current model is employed in order to calculate the stator flux & its corresponding torque. Extra-leg extra-switch inverter is used as power unit. The PI controller is put to use for regulating rotor speed and generating reference torque. Dynamic responses of healthy MPC and unhealthy FTMPC for PMSM systems are given to compare their performance via simulation and some analysis is presented. The simulation results show that the proposed FTMPC strategy not only allows for continuous and disturbance-free operation of the unhealthy PMSM with LOP or LOL but also preserves satisfactory torque and speed control. And then the effectiveness of the proposed schemes in this paper is demonstrated.

Key-Words: - Fault tolerant control; Model predictive control; Permanent magnet synchronous motor; Motor model; Flux estimator; Inverter

1 Introduction

The electrical drive systems in electric vehicle, aerospace as well as other important fields must be required to be high reliable and safe. Due to a variety of complex factors, potential failures are often inevitable. Once the electric drive system is out of order, if repairs and maintenance cannot be completed on the spot, this will result in the system to stop working, may cause great financial losses, and even result in enormous human and property losses. Therefore, there is an urgent need to research fault control for electrical motor.

One of the most common types of potential faults in electrical motor is the loss of one phase (LOP) of the motor, or alternatively, the loss of one leg of the inverter (LOL). If LOP or LOL happens suddenly, the corresponding phase is open-circuited, the drive system supply and load currents are significantly distorted and the load phase current in which the failure occurred has large zero periods resulting in a loss of torque control and in high pulsating unacceptably torques. Consequently, the drive system's operation has to be interrupted [1, 2]. So it is indispensable to solve the problem such that motor system can be controlled to be disturbance-free.

As for the aforementioned fault, nowadays there are two modulation techniques, one being based-on hardware techniques and the other based-on software (i.e. control method). By means of some different approaches such as using matrix converter structure[3,4], adding redundant switch [1,5-7], introducing phase-redundant topology [8-10], proposing cascaded two-level converter [11] as well as giving redundant converter[12,13], etc., the effective fault tolerant results have been achieved. However, these methods are less preferable in some applications because of their complicated hardware and high operation cost. Therefore the fault tolerant method using software with low-cost reconfiguration has been highly praising [14-17].

Over the past years, as for unhealthy electrical motor, making use of field oriented control strategy (FOC) [18], the performance of faulty electric drive systems can be maintained via controlling current [2,10,19-21]. Due to the inherent bandwidth of inner current loop, the dynamic response of FOC drive systems is limited. To improve the dynamic performance, direct torque control (DTC) has recently begun to be applied for unhealthy electrical motor [22-24].Compared to FOC, DTC directly manipulates the inverter's output voltage vector, hence eliminates the inherent delay caused by current loops, then features comparatively good dynamic response. Despite the above merits, switching-table-based DTC presents some unavoidable drawbacks [25], such as high torque and flux ripples, variable switching frequency along with acoustic noise.

Apart from FOC and DTC methods, model predictive torque control (MPTC) is an emerging control concept and is received significant attention from three-phase electrical drives community [26-32], which adopts the principles of model predictive control (MPC) [33]. Compared with state-of the art schemes, such as DTC and FOC, MPTC achieves a reduction of the torque and flux ripples [34]. Furthermore, switching losses can be reduced [35].It is the abovementioned advantages of MPTC that motivates this paper.

Permanent magnet synchronous motor (PMSM) drive is nowadays widely used in the industry applications due to their high efficiency and high power/torque density. For healthy three-phase PMSM inverter, Fig.1 is its topology. For unhealthy three-phase PMSM with LOP or LOL, there are mainly three solving schemes at present: the first is called the extra-leg split capacitor control strategy [2, 36], which adds a redundant switch to connect the source's neutral to the load's neutral. The second is known as split capacitor scheme for isolating the phase with a faulty switching device of motor drive system and connecting to the midpoint of DC link [37]. The disadvantage of aforementioned two reconfiguration topologies lies in that the maximum speed in the post-fault operation is half of its nominal value due to the applied voltage on the machine terminals is decreased to half of its original value. Then appears the third termed as extra-leg extra-switch (ELES) scheme shown in Fig.2. In the scheme, the added switch connects the motor neutral point to an extra inverter leg, which provides the current path during the fault operation.

Based-on the third scheme, employing voltage & current model flux estimator, [22-24] discussed fault tolerant DTC for PM AC motor with one phase open-circuit fault. However, MPC has never been applied to fault tolerant control for electrical motor. Using the merits of MPC, this paper investigates fault tolerant control for PMSM. It will be shown that utilizing the method developed in this paper, the reliability and satisfactory performance can be achieved for the drive system under LOP or LOL operating condition.

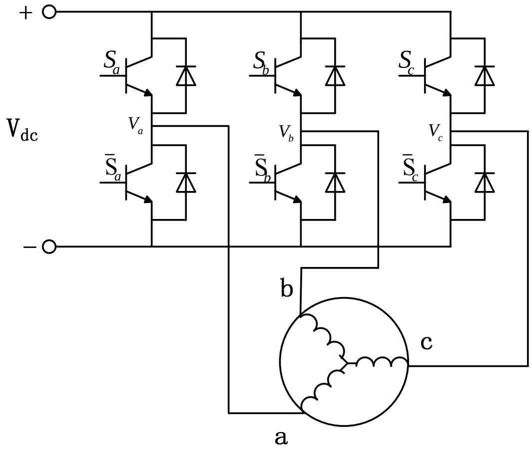


Fig. 1 Healthy three phase inverter

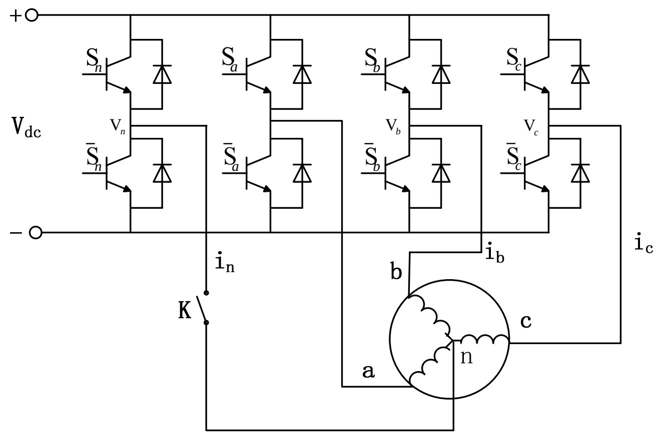


Fig.2 ELES three phase inverter

The structure of the paper is as follows: modeling of PMSM with LOP or LOL is established in section two. In section three, the principle of FTMPC is described and its corresponding predictive model for remaining two stator phase currents is established, and the flux & torque estimator based on current model is given. The numerical simulation results & analysis and conclusion are reported in section four and five, respectively.

2 Dynamic Model of PMSM with LOP or LOL in *abc*-System

In this paper, as for three-phase magnetically isotropic PMSM, schematic diagram of the motor-inverter is shown in Fig. 2. In the event that any one phase is open-circuited or any one transistor fails open in the inverter, then switch K is on. In the following analysis, suppose phase *a* or its corresponding one leg in the inverter is off. In this case, the current in phase *a* suddenly drops to zero. Like the process of building the model [38] for healthy motor, modeling of unhealthy motor includes three equations. They are flux linkage, voltage and torque equations as follows.

2.1 Stator flux linkage expression in *abc*-system

Suppose three-phase stator self-inductances L_a L_b and L_c are same, i.e. $L_a=L_b=L_c=L$ and three-phase stator mutual-inductances M_{ab}, M_{bc} and M_{ca} are same, i.e. $M_{ab}=M_{bc}=M_{ca}=M$ (neglecting stator self-inductance's and mutual-inductance's second harmonics). And i_b and i_c are stator phase currents. When phase *a* is off, stator flux-linkages Ψ_{sa}, Ψ_{sb} and Ψ_{sc} produced only by the stator currents in *abc*-system (that means three-phase stationary coordinate) are shown as in Fig. 3 and can be expressed as follows:

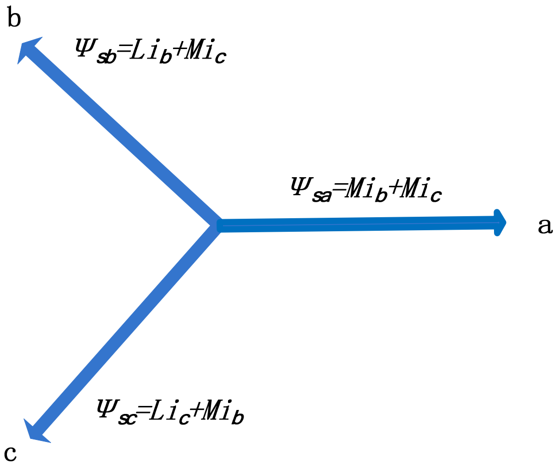


Fig. 3 Flux-linkages of three phases produced by stator currents in *abc*-system

$$\begin{bmatrix}\psi_{sa}\\ \psi_{sb}\\ \psi_{sc}\end{bmatrix}=\begin{bmatrix}M & M\\ L & M\\ M & L\end{bmatrix}\begin{bmatrix}i_b\\ i_c\end{bmatrix} \tag{1}$$

Considering the rotor magnet, the stator flux-linkage vector in *abc*-system can be expressed as follows:

$$\begin{bmatrix}\psi_a\\ \psi_b\\ \psi_c\end{bmatrix}=\begin{bmatrix}\psi_{sa}\\ \psi_{sb}\\ \psi_{sc}\end{bmatrix}+\begin{bmatrix}\psi_m\cos\theta_r\\ \psi_m\cos(\theta_r-\frac{2\pi}{3})\\ \psi_m\cos(\theta_r+\frac{2\pi}{3})\end{bmatrix} \tag{2}$$

Where Ψ_a,Ψ_b and Ψ_c are resultant of stator flux-linkages produced both by the stator currents and by the rotor magnetic along *a*-axis, *b*-axis and *c*-axis, respectively. θ_r and ψ_m are electrical angular rotor position with reference to phase *a* and permanent magnet flux, respectively.

2.2 Stator voltage equation in *abc*-system

Suppose that as soon as phase *a* is detected being off, switch K is on as shown in Fig.2. When phase *a* is off, the remaining stator phase voltage vector of PMSM is given by:

$$\begin{bmatrix}v_{bn}\\ v_{cn}\end{bmatrix}=\begin{bmatrix}R_b & 0\\ 0 & R_c\end{bmatrix}\begin{bmatrix}i_b\\ i_c\end{bmatrix}+\frac{d}{dt}\begin{bmatrix}\psi_b\\ \psi_c\end{bmatrix} \tag{3}$$

Where R_b and R_c are stator phase resistances, v_{bn} and v_{cn} stator phase voltages which are defined as follows

$$\begin{aligned}v_{bn}&=V_b-V_n\\ v_{cn}&=V_c-V_n\end{aligned} \tag{4}$$

V_b and V_c in (4) are shown in Fig. 2.

Substituting (2) into (3), the following expression can be obtained,

$$\begin{bmatrix}v_{bn}\\ v_{cn}\end{bmatrix}=\begin{bmatrix}R_b & 0\\ 0 & R_c\end{bmatrix}\begin{bmatrix}i_b\\ i_c\end{bmatrix}+\begin{bmatrix}L & M\\ M & L\end{bmatrix}\begin{bmatrix}\frac{di_b}{dt}\\ \frac{di_c}{dt}\end{bmatrix}-\begin{bmatrix}\psi_m\omega_r\sin(\theta_r-\frac{2\pi}{3})\\ \psi_m\omega_r\sin(\theta_r+\frac{2\pi}{3})\end{bmatrix} \tag{5}$$

where ω_r is rotor speed, the phase currents i_b, i_c and neutral line current i_n in Fig. 2 meet following mathematical relationship

$$i_n=i_b+i_c \tag{6}$$

Remark 1 : Suppose that motor parameters, such as stator phase resistances, inductances, etc., are time-invariant in this paper.

Remark 2: Here it is necessary to mention of how to obtain phase inductance L and phase mutual inductance M.

Suppose mutual inductance M is one half of phase inductance L. Neglecting the second harmonics of stator self-inductance and mutual-inductance, we have,

$$L_d = L_q \approx L + M = L + \frac{1}{2}L = \frac{3}{2}L \tag{7}$$

Thus

$$L = \frac{2}{3}L_d \tag{8}$$

2.3 Electromagnetic torque equation

The electromagnetic torque equation of PMSM with LOP or LOL fault is as follows,

$$J\frac{\mathrm{d}\omega_r}{\mathrm{d}t} = T_e - T_l - B_m\omega_r - T_f \tag{9}$$

where J, T_e, T_l, B_m and T_f are respectively inertia of moment, electromagnetic torque, load torque, viscous friction coefficient and coulomb friction torque.

Combination of the above-given flux-linkage vector equation, phase voltage vector equation and electromagnetic torque equation is the model for PMSM with LOP or LOL.

3 Fault Tolerant MPC for PMSM with LOP or LOL

The objective of FTMPC for PMSM is that when LOP or LOL happens, the motor control system not only is still stable, but also its speed and torque can be controlled to meet given requirements. The system in Fig.4 shows block diagram for implementing FTMPC scheme by modifying conventional MPC strategy. Suppose that as soon as LOP or LOL happens, it can be detected and the system can be switched from conventional MPC to FTMPC scheme immediately.

Besides PMSM with LOP or LOL, FTMPC system mainly comprises of four components. They are FTMPC, Calculator for reference stator flux linkage, Power unit and PI controller, respectively. FTMPC includes three parts: Cost function minimization, Predictive model as well as Flux & torque estimators.

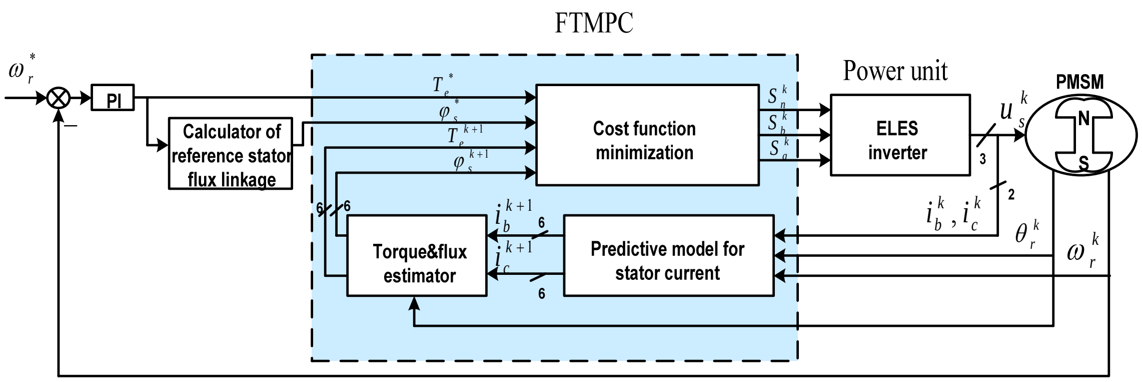


Fig. 4 FTMPC system for PMSM with LOP or LOL

3.1 FTMPC

3.1.1 Basic principle of FTMPC

***A.* Conventional MPC for healthy motor drives**

The basic idea of MPC is to predict the future behavior of the variables over a time frame (integer multiple of the sample time) based on the model of the system. MPC has several merits such as easy inclusion of nonlinearities and constraints [26]. In fact, MPC is an extension of DTC, as it replaces the look-up table of DTC with an online optimization process in the control of machine torque and flux. Different from the employments of hysteresis comparators and switching table in conventional DTC, the principle of vector selection in MPC is based on evaluating a defined cost function. The selected voltage vector from conventional switching table in DTC is not necessarily the best one in terms of reducing torque and flux ripples. There are limited discrete voltage vectors in the two-level inverter-fed PMSM drives, as a result, it is possible to evaluate the effects of each voltage vector and select the one minimizing the cost function.

For conventional MPC, the cost function for healthy motor drives is such chosen that both torque and flux at the end of the cycle is as close as possible to the reference value. Generally, the minimum value of cost function is defined as

$$\begin{aligned} &\min .g = \left|T_e^* - T_e^{k+1}\right| + k_1 \left\|\psi_s^*\right| - \left|\psi_s^{k+1}\right\| \\ &s.t. \quad u_s^k \in \{V_1, \ V_2, \ \cdots \ V_5, \ V_6\} \end{aligned} \tag{10}$$

Where T_e^* and Ψ_s^* are reference values for torque and stator flux, respectively. T_e^{k+1} and Ψ_s^{k+1} predictions for torque and stator flux at $(k+1)th$ instant, respectively. V_1, V_2, V_3, V_4, , V_5 and V_6 are six nonzero voltage space vectors and can be generated by healthy three phase inverter with respect to the different switches states. Voltage space vector u_s^k at kth instant is defined as

$$u_s^k = \frac{2}{3} V_{dc} \left[S_a^k + e^{i2\pi/3} S_b^k + \left(e^{i2\pi/3}\right)^2 S_c^k \right]$$

Where S_x^k ($x=a,b,c$) at kth instant is upper power switch state of one of three legs as shown in Fig. 4. $S_x^k=1$ or $S_x^k=0$ when upper power switch of one leg is on or off. k_1 is the weighting factor. The selection of k_1 is still an open problem for answer [26]. In this paper k_1 is selected to be T_n/Ψ_n in order to give torque and flux the same weight, where T_n and Ψ_n are the rated value for torque and stator flux, respectively.

B. FTMPC for unhealthy motor drives

For FTMPC algorithm, in the evaluating cost function, the stator phase voltages (v_{bn}, v_{cn}) are directly employed instead of voltage space vector. Therefore, the minimum value of cost function for FTMPC of unhealthy motor drives should be modified as

$$\begin{aligned} &\min .g = \left|T_e^* - T_e^{k+1}\right| + k_1 \left\|\psi_s^*\right| - \left|\psi_s^{k+1}\right\| \\ &s.t. \quad u_{sn}^k \in \{V_{bn1\_cn1}, \ V_{bn2\_cn21}, \ \cdots, \\ &\qquad V_{bn5\_cn5}, \ V_{bn6\_cn6}\} \end{aligned} \tag{11}$$

Where V_{bni_cni} (i=1,2, $\cdots$, 5, 6), mapping to V_i.(i=1,2, $\cdots$, 5, 6), represents two stator phase voltages v_{bn}, v_{cn}. ELES three phase inverter in Fig.2 generates six nonzero voltage space vectors V_1, V_2, V_3, V_4, V_5 and V_6, the same as what healthy three phase inverter does.

For ELES three phase inverter, the mapping relationship between V_i and V_{bni_cni} is as shown in table 1.

Similar to conventional MPC, the inputs in FTMPC are the reference and predicted torque & flux. After evaluating the cost function in (9) for every possible group of v_{bn}, v_{cn}, one and only one group v_{bn}, v_{cn} is selected, so is one and only one voltage space vector accordingly.

Table 1 The mapping relationship between voltage space vector and stator phase voltages

	v_{bn}	v_{cn}
V_1	$-V_{dc}$	$-V_{dc}$
V_2	0	$-V_{dc}$
V_3	V_{dc}	0
V_4	V_{dc}	V_{dc}
V_5	0	V_{dc}
V_6	$-V_{dc}$	0

3.1.2 Predictive model for stator currents

Similar to MPC, the basic idea of FTMPC is also to predict the future behavior of the variables. The predictions for FTMPC are phase currents in *abc*-system instead of the currents in *dq*-system. In order to obtain the predictions, (5) can be rewritten as following,

$$\begin{aligned} \frac{\mathrm{d}i_b}{\mathrm{d}t} = \frac{1}{L^2 - M^2} \Big[&L v_{bn} - M v_{cn} + R_c M i_c - R_b L i_b + \\ &L \psi_m \omega_r \sin(\theta_r - \frac{2\pi}{3}) - M \psi_m \omega_r \sin(\theta_r + \frac{2\pi}{3}) \Big] \end{aligned} \tag{12}$$

$$\begin{aligned} \frac{\mathrm{d}i_c}{\mathrm{d}t} = \frac{1}{L^2 - M^2} \Big[&L v_{cn} - M v_{bn} + M R_b i_b - L R_c i_c + \\ &L \psi_m \omega_r \sin(\theta_r + \frac{2\pi}{3}) - M \psi_m \omega_r \sin(\theta_r - \frac{2\pi}{3}) \Big] \end{aligned} \tag{13}$$

(12) and (13) can be used to obtain prediction of the stator currents at the next sampling instant based on given stator voltages v_{bn}^k and v_{cn}^k as well as measured currents i_b^k and i_c^k at current sampling instant. The prediction can be expressed as

$$\begin{aligned} i_b^{k+1} = i_b^k + \frac{T_s}{L^2 - M^2} \Big[&L v_{bn}^k - M v_{cn}^k + R_c M i_c^k - R_b L i_b^k + \\ &L \psi_m^k \omega_r^k \sin(\theta_r^k - \frac{2\pi}{3}) - M \psi_m^k \omega_r^k \sin(\theta_r^k + \frac{2\pi}{3}) \Big] \end{aligned} \tag{14}$$

$$\begin{aligned} i_c^{k+1} = i_c^k + \frac{T_s}{L^2 - M^2} \Big[&L v_{cn}^k - M v_{bn}^k + M R_b i_b^k - L R_c i_c^k + \\ &L \psi_m^k \omega_r^k \sin(\theta_r^k + \frac{2\pi}{3}) - M \psi_m^k \omega_r^k \sin(\theta_r^k - \frac{2\pi}{3}) \Big] \end{aligned} \tag{15}$$

Where i_b^{k+1} and i_c^{k+1} are predicted values of stator currents for the next sampling period, T_s is the sampling period.

After obtaining i_b^{k+1} and i_c^{k+1}, both the torque and flux at the $(k+1)th$ instant can be estimated.

3.1.3 Torque & flux estimators

Generally, two flux estimators can be employed. One is voltage-based model, the other current-based one. The former involves integrator which is sensitive to not only DC offset but also initial value [39], and bigger DC offset along with improper initial value easily leads to the saturation problem, which consequently results in the whole system being unstable. Nevertheless, the latter is able to avoid the troublesome problem. The currents involved in the latter can be calculated from measuring phase currents. Therefore this paper concentrates on the latter. By current-based model, the flux estimator will be discussed in dq-system (that means two-phase rotary coordinate). Obviously, the estimator used for healthy PMSM cannot be directly applied to unhealthy one. So the modified flux estimator suitable for FTMPC is established as following.

A. Flux estimator in dq-system

In *dq*-system, the flux-linkage Ψ_d and Ψ_q can be expressed as following vector:

$$\begin{bmatrix} \psi_d^{k+1} \\ \psi_q^{k+1} \end{bmatrix} = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \begin{bmatrix} i_d^{k+1} \\ i_q^{k+1} \end{bmatrix} + \begin{bmatrix} \psi_m^{k+1} \\ 0 \end{bmatrix} \tag{16}$$

Where L_d and L_q are inductances in *dq*- system, i_d and i_q are currents in *dq*-system.

By Park and Clarke transformations, i_d and i_q can be yielded from phase current vector as follows

$$\begin{bmatrix} i_d^{k+1} \\ i_q^{k+1} \end{bmatrix} = \begin{bmatrix} \cos\theta_r^{k+1} & -\sin\theta_r^{k+1} \\ \sin\theta_r^{k+1} & \cos\theta_r^{k+1} \end{bmatrix} \cdot \begin{bmatrix} -1/2 & -1/2 \\ \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_b^{k+1} \\ i_c^{k+1} \end{bmatrix} \tag{17}$$

The magnitude of stator flux linkage ψ_s is

$$\psi_s^{k+1} = \sqrt{\left(\psi_d^{k+1}\right)^2 + \left(\psi_q^{k+1}\right)^2} \tag{18}$$

B. Torque estimator in dq-system

Electromagnetic torque developed in *dq*-system can be estimated s following

$$T_e^{k+1} = \frac{3}{2}\mathrm{p}\left[\psi_m^{k+1} i_q^{k+1} + \left(L_d - L_q\right) i_d^{k+1} i_q^{k+1}\right] \tag{19}$$

where p is number of pole pairs.

Substituting (17) into (19), torque can be estimated.

3.2 Power unit

The power unit adopts ELES inverter structure as shown in Fig.2. Then a voltage space vector V_i (i=1,2,…,6) is output of ELES inverter, which is controlled by the switching states S_i (i=1,2,…,6). Fig.5 shows the symmetrical layout of six nonzero voltage vectors V_1-V_6 and its corresponding six switching states S_1-S_6.

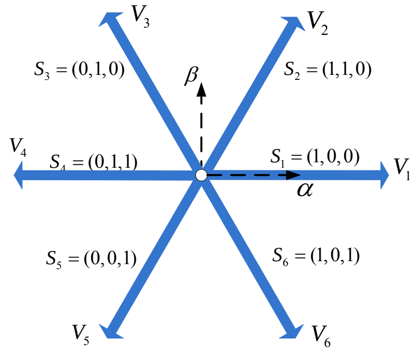


Fig.5 Layout of voltage space vectors and its corresponding switching states

3.3 PI controller

PI controller is used to regulate the rotor speed. The PI controller compares the reference speed with the motor real-time speed and generates a torque command (or reference value), i.e., T_e^*. Properly selecting proportional & integral parameter could keep the speed and torque ripple small.

3.4 Calculator for reference stator flux linkage

In accordance with reference torque T_e^*, the desired reference stator flux linkage Ψ_s^* as shown in Fig. 4 can be obtained by means of the following algorithm which is called MTPA (maximum torque per ampere),

$$\psi_s^* = \sqrt{\left(T_e^* L_q \Big/ \frac{3}{2} p\psi_m\right)^2 + \left(i_d L_d + \psi_m\right)^2} \tag{20}$$

Where i_d^* is assumed to be zero, i.e.,

$$i_d^* = 0$$

As a result, the corresponding reference stator flux linkage Ψ_s^* is written as

$$\psi_s^* = \sqrt{\left(T_e^* L_q \Big/ \frac{3}{2} p\psi_m\right)^2 + \psi_m^2} \tag{21}$$

4 Simulation Results of FTMPTC for PMSM with LOP or LOL

4.1 PMSM parameters and other control parameters selection

We take a PMSM as an example to validate the effective of proposed FTMPC scheme, parameters of which are given in Table 2 [22-24]. The computer model of FTMPC system for unhealthy PMSM is established. In order to comparatively investigate the performance of healthy MPC and unhealthy FTMPC motors, the computer model of conventional MPC system for healthy PMSM is also built. These models are based on MATLAB/SIMULINK/SIMSCAPE platform.

Table 2 Parameters of PMSM

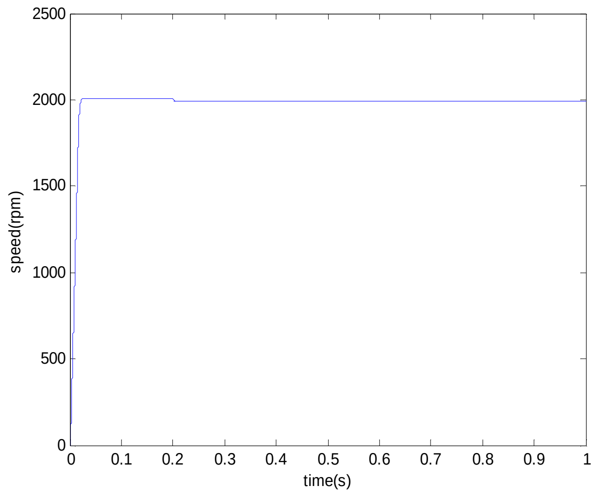
Symbol	Quantity	Value
R_a, R_b, R_c	Phase resistance	0.466Ω
L_d, L_q	*dq*-coordinate inductance	3.19mH
ψ_m	Rotor magnetic flux	92.8mWb
p	Number of pole pairs	1
V_{dc}	DC bus voltage	70V
ω_n	Rated speed	3000rpm
T_n	Rated torque	0.3Nm
J	Moment of inertia	0.0002Kg.m^2
B_m	Viscous friction coefficient	0
T_f	Coulomb friction torque	0

The reference speed ω_r^* is set to 2000 rpm and the external load of 0.3Nm is applied at t=0.2s. The sampling period is 100 μ s, the value k_1 is selected to be 60. The proportional & integral parameters for MPC and FTMPC systems are same. They are selected as follows.

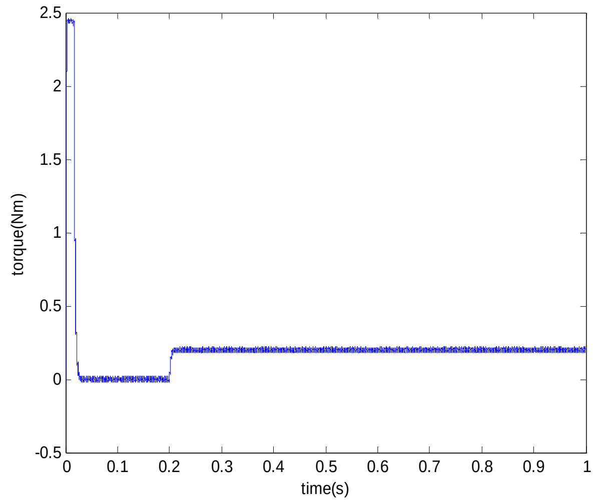
$$K_p = 0.01, K_i = 0.005$$

4.2 Simulation and its analysis

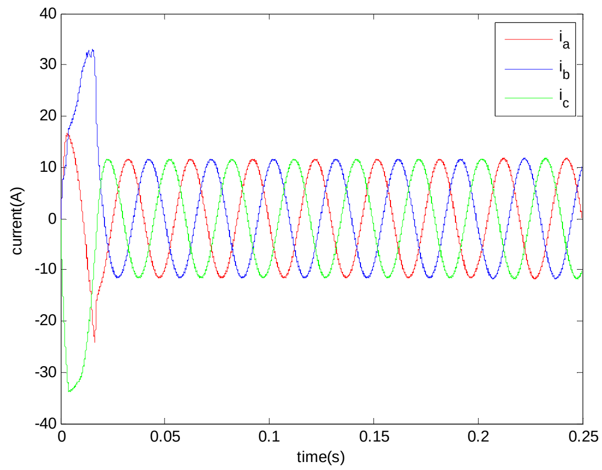
The numerical simulation results are given from Fig. 6 to Fig.8 in terms of torque, rotor speed, stator currents, trajectory of stator flux linkage, stator flux linkages, etc.



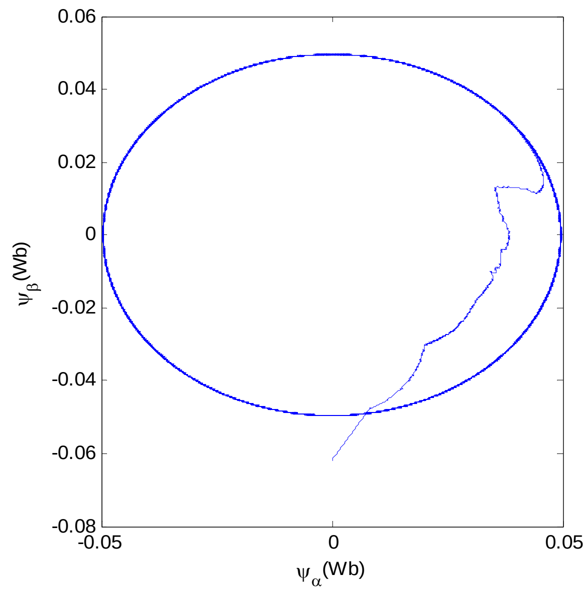
(a) Rotor speed response



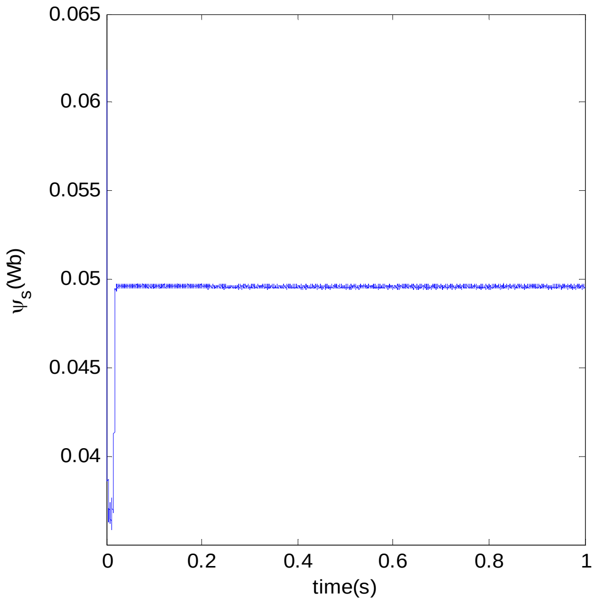
(b) Torque response



(c) Stator currents i_a,i_b and i_c

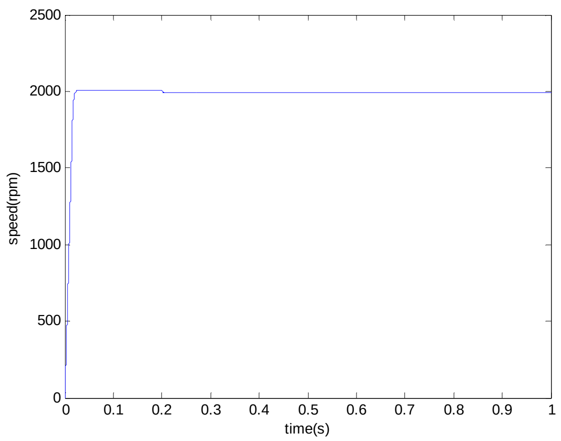


(d) Trajectory of stator flux linkage

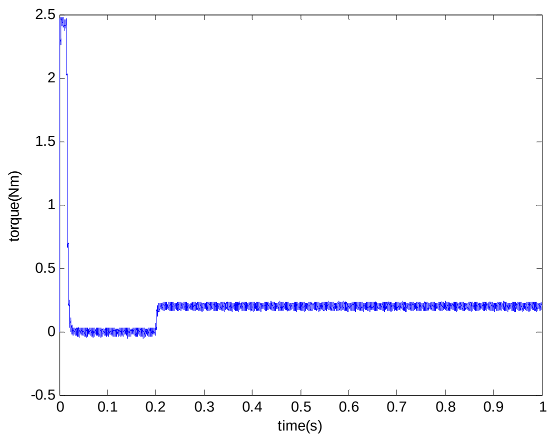


(e) Stator flux linkage Ψ_s

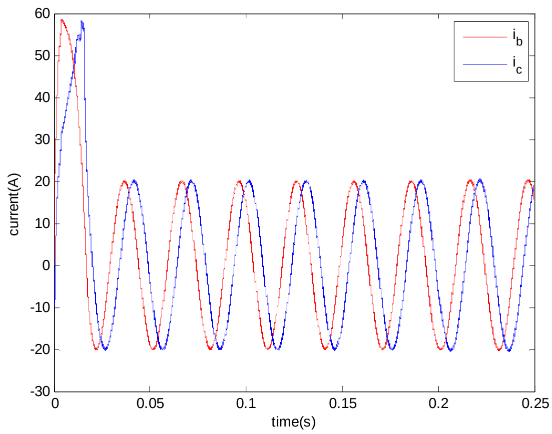
Fig.6 Dynamic responses of conventional MPC for healthy PMSM



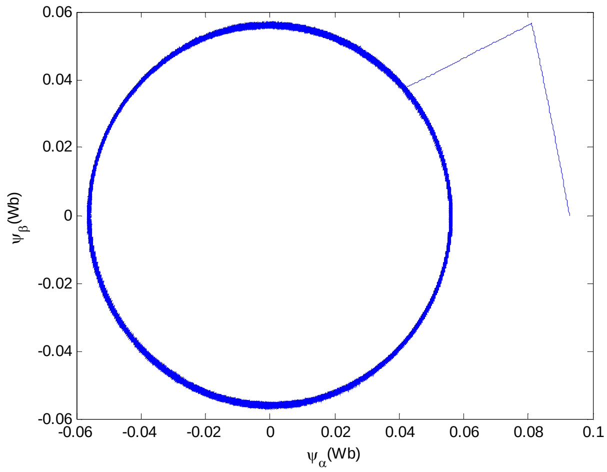
(a) Rotor speed response



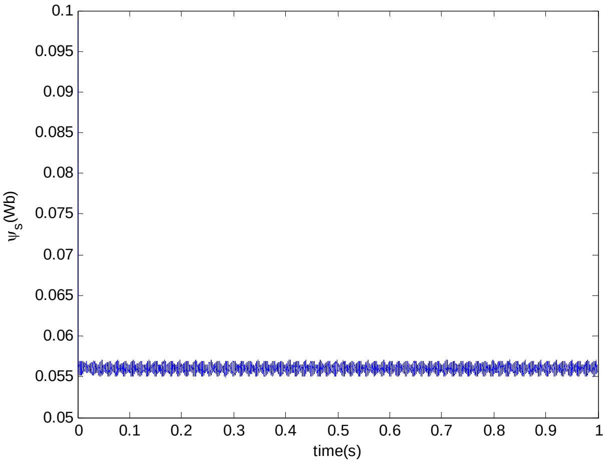
(b) Torque response



(c) Stator currents i_b and i_c

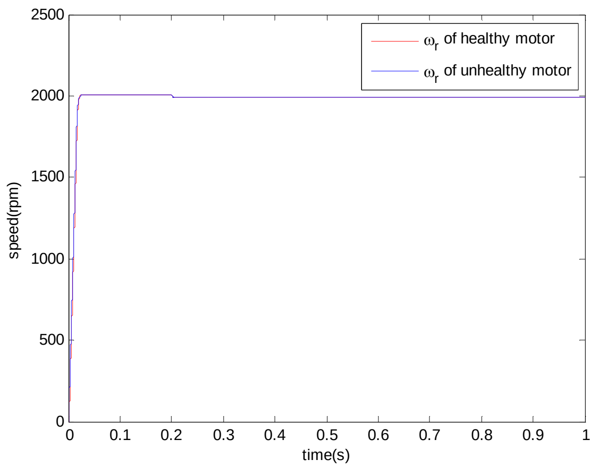


(d) Trajectory of stator flux linkage

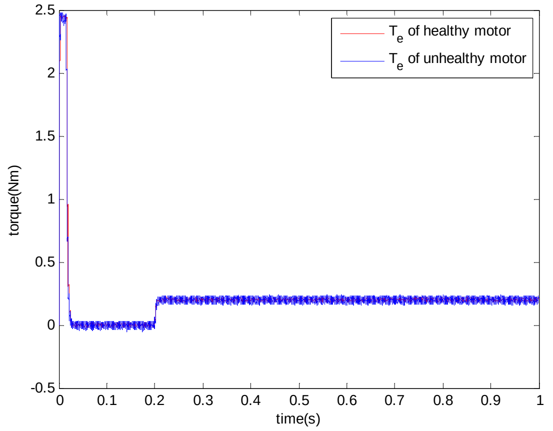


(e) Stator flux linkage Ψ_s

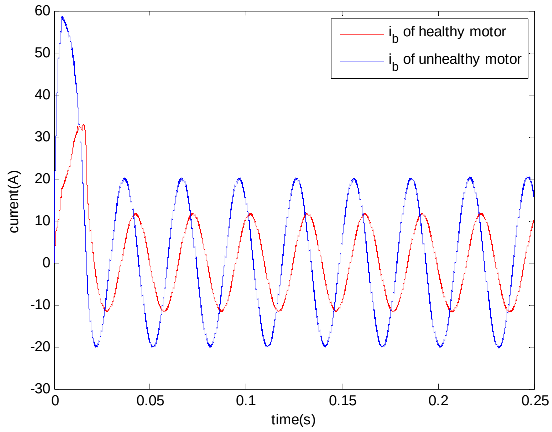
Fig. 7 Dynamic responses of FTMPC for unhealthy PMSM



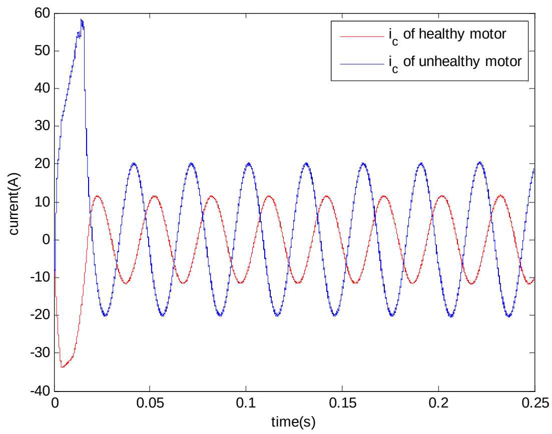
(a) Speed ω_r of healthy and unhealthy motor



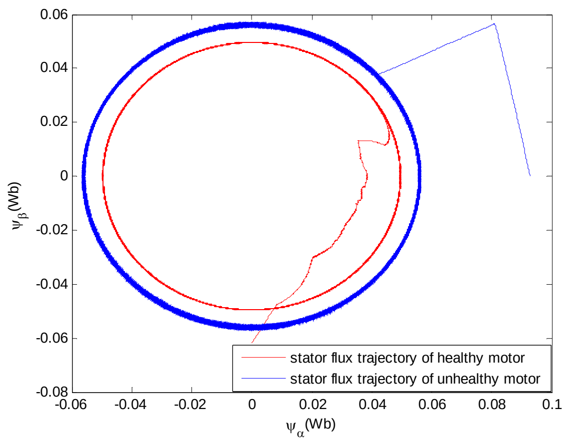
(b) Torque T_e of healthy and unhealthy motor



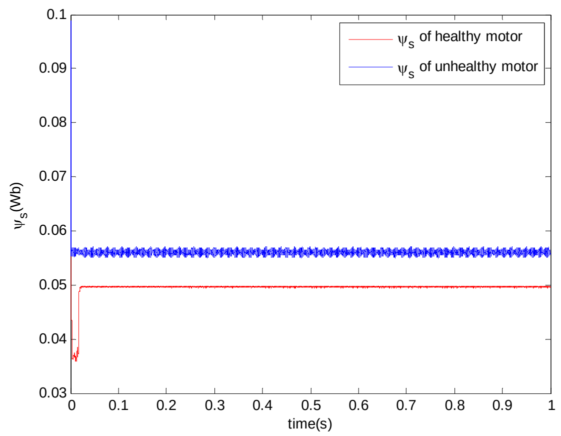
(c) Current i_b of healthy and unhealthy motor



(d) Current i_c of healthy and unhealthy motor



(e) Trajectory of stator flux linkage for healthy and unhealthy motor



(f) Stator flux linkage Ψ_s of healthy and unhealthy motor

Fig.8 Comparison between MPC for healthy PMSM and FTMPC for unhealthy PMSM

Fig. 6 presents dynamic responses of conventional MPC for healthy PMSM and Fig. 7 shows corresponding dynamic responses of FTMPC for unhealthy PMSM. Fig. 8 illustrates the performance comparison between conventional MPC for healthy and FTMPC for unhealthy PMSM. Ψ_α, Ψ_β as shown in Figs. 6, 7 and 8 are stator flux-linkages in αβ-system. Appendix gives its detailed solving procedure.

Analyzing these simulations, the following consequence could be obtained:

- Comparing Fig.6 and 7, it can be seen that the proposed FTMPC system is still stable and allows for continuous operation when one phase of PMSM is open-circuited.
- As far as the dynamic responses of rotor speed and torque are concerned, comparing red with blue (represent healthy and unhealthy PMSM, respectively) curves in Fig.8 (a) and 8(b) demonstrates that the performance of proposed FTMPC strategy is satisfactory and meets given requirements.
- Figs. 8(c),8(d) and Figs. 8(e),8(f) show phase currents and stator flux relationship between MPC for healthy and FTMPC for unhealthy, respectively. It can be clearly seen that in order to maintain the same torque as its previous value after LOP or LOL occurs, phase *b* and phase *c* currents are regulated to dramatically increase and accordingly stator flux is increased simultaneously, which is consistent with the theoretical analysis [2].
- Although the torque T_e and stator flux Ψ_s ripple of the unsymmetrical two-phase machine are larger than that of symmetrical three-phase machine as shown in Fig. 8, FTMPC strategy can manage to get constant torque and circular stator flux, and then make two-phase machine to complete normal operation before shutting down for repair.

To sum up, proposed FTMPC strategy for unhealthy PMSM with LOP or LOL can make electrical drive system to tolerate fault and therefore are effective and correct.

Remark: When LOP or LOL happens, FTMPC can be used to replace MPC. In order to guarantee smoothness transition from MPC to FTMPC, the control transition strategy should be studied or else a braking torque will occur, which may result in mechanical damage[40-42]. In addition, the cost function in (11) assumes that all calculations and judgment are implemented at *kth* instant and the selected vector will be applied immediately. However, in practical digital implementation, the assumption is not true and the applied voltage vector is not applied until the $(k+1)th$ instant, which results in one step delay. To eliminate such adverse effect, the delay compensation should taken action [43]. The above-mentioned problems are our next work to be done.

5 Conclusion

This paper has puts forward a new FTMPC strategy. The model of three-phase PMSM with LOP or LOL is first built. Then based on conventional MPC, the cost function of FTMPC is modified. In order to realize predictive control, the predictive model for stator phase currents is derived, and the flux estimator based-current is built. The motor control system uses ELES inverter as power unit in the post-fault operation. Numerical simulation is performed to check the feasibility of the new FTMPC strategy. The results show that the proposed FTMPC system is still stable and allows for continuous operation when one phase of PMSM is open-circuited. In comparison with the conventional MPC strategy for healthy motor system, the proposed FTMPC strategy for unhealthy motor system preserves satisfactory torque & speed control and could meet given requirements and therefore is effective and correct. Taking real motor parameters variation (like stator resistance, etc.) into consideration, we will study FTMPC strategy with adaptive observer in the next work.

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Appendix

In αβ-system, the flux-linkages $\Psi_{s\alpha}$ and $\Psi_{s\beta}$, which is produced only by the stator current, can be expressed as following vector

$$\begin{bmatrix} \psi_{s\alpha}^{k+1} \\ \psi_{s\beta}^{k+1} \end{bmatrix} = \begin{bmatrix} \frac{1}{2}(M-L) & \frac{1}{2}(M-L) \\ \frac{\sqrt{3}}{2}(L-M) & -\frac{\sqrt{3}}{2}(L-M) \end{bmatrix} \begin{bmatrix} i_b^{k+1} \\ i_c^{k+1} \end{bmatrix} \tag{1}$$

Since stator currents i_α and i_β in αβ-system can be expressed as following vector:

$$\begin{bmatrix} i_\alpha^{k+1} \\ i_\beta^{k+1} \end{bmatrix} = \begin{bmatrix} -\frac{1}{3} & -\frac{1}{3} \\ \frac{\sqrt{3}}{3} & -\frac{\sqrt{3}}{3} \end{bmatrix} \begin{bmatrix} i_b^{k+1} \\ i_c^{k+1} \end{bmatrix} \tag{2}$$

Taking (2) into account, (1) can be rewritten as

$$\begin{bmatrix} \psi_{s\alpha}^{k+1} \\ \psi_{s\beta}^{k+1} \end{bmatrix} = \begin{bmatrix} L-M & 0 \\ 0 & L-M \end{bmatrix} \begin{bmatrix} i_\alpha^{k+1} \\ i_\beta^{k+1} \end{bmatrix} \tag{3}$$

Considering the rotor magnet, the stator flux-linkages Ψ_α and Ψ_β in αβ-system can be expressed as following vector

$$\begin{bmatrix} \psi_{\alpha}^{k+1} \\ \psi_{\beta}^{k+1} \end{bmatrix} = \begin{bmatrix} \psi_{s\alpha}^{k+1} \\ \psi_{s\beta}^{k+1} \end{bmatrix} + \begin{bmatrix} \frac{2}{3}\psi_m^{k+1}\cos\theta_r^{k+1} \\ \frac{2}{3}\psi_m^{k+1}\sin\theta_r^{k+1} \end{bmatrix} \tag{4}$$

In addition, the magnitude of stator flux linkage ψ_s can be given by following formula.

$$\psi_s^{k+1} = \sqrt{\left(\psi_\alpha^{k+1}\right)^2 + \left(\psi_\beta^{k+1}\right)^2} \tag{5}$$