

# Direct predictive speed control of permanent magnet synchronous motor fed by matrix converter

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## ABSTRACT

This paper proposes a non-cascade -single loop- Direct Speed Control algorithm for surface mounted Permanent Magnet Synchronous Motor (PMSM) fed by Matrix Converter. The proposed method uses Finite Control Set Model Predictive Control (FCS-MPC) to manipulate system speed and currents simultaneously. Also, for better performance of the predictive method, an observer designed to estimate mechanical torque and other uncertain parameters of the mechanical subsystem as a lumped disturbance. Simulation results using Matlab/Simulink demonstrate the performance of proposed algorithm.

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## 1. INTRODUCTION

In recent years Permanent Magnet Synchronous Motors (PMSM) are getting some real attention due to their high-power density, fast dynamic response, high efficiency and capability of working without gearbox. These characteristics makes PMSM a strong candidate for different applications like wind energy systems, electric vehicles and industrial equipment [1, 2]. Matrix Converter (MC) is a kind of converter that can connect an AC device directly to a three phase AC source by using 9 bidirectional switches [3]. Unlike conventional back to back capacitor-based converters, MCs achieve ac-ac conversion without energy storage link, which helps them to have more reliability, compact size and longer lifetime [3-6].

A typical PMSM drive using matrix converter with three-phase source, input filter and motor is represented in Figure 1. It can be seen that every input phase can connect to an output phase using a switch. To filter input currents and avoid voltage spikes during switching, MCs need an LC filter combined with parasitic resistor. More detail on MC and filter will be given in modeling section. There are different control algorithms for PMSMs but Field Oriented Control (FOC) and Direct Torque Control (DTC) are the most popular methods [7-8]. Interaction between control variables, windup problem, bandwidth limitation due to cascade structure and modulation stage, makes FOC a non-ideal controller choice [9]. DTC uses a lookup table based on switching states to control motor torque directly. Output voltage vectors are not always optimal, as a consequence DTC suffers from high torque and stator flux ripples [10, 11].

Simple treatment of constraints, Multivariable structure and good performance in the presence of nonlinearities make MPC one of the best choices for drive systems [12-13]. There are two main categories

for MPC, Continuous Control Set MPC (CCS-MPC) and Finite Control Set MPC (FCS-MPC) [9, 14-16]. CCS-MPC needs a modulator to generate gate signals, but FCS-MPC takes advantage of the fact that the number of possible switching states is limited, so it uses the model of the system to predict next state of it for each possible control action and best control action will be chosen by minimizing a cost function [17]. Using cost function helps FCS-MPC to directly control more than one objective at the same time.

Cascade structure is a common strategy for speed/position control of PMSMs, but comparing to current dynamic, mechanical dynamic is sluggish and this difference between time constants enforces the designer to consider longer prediction horizon for DPSC [18]. Moreover, elimination of outer loop controller not only brings stability issues but also decrease system performance and causes steady state error [19]. In [20-22] some methods proposed to control speed of a PMSM fed by a two-level back to back converter directly. This topology eliminates both outer linear controller and modulator.

In this paper we propose a direct predictive speed control for PMSM fed by matrix converter. A new cost function introduced by combining speed dynamic, current dynamic and system constraints to have high performance without cascade structure. To tackle stability issue and high current distortion, due to difference between mechanical and electrical time constants, a current reference designed based on sliding mode concept is added to cost function. This term guarantees outer loop stability and decrease current distortion. Also, to have lower steady state error and better performance, a disturbance observer designed to estimate load torque value and other uncertainties of the mechanical subsystem as a lumped disturbance.

The rest of this paper is organized as follows: In section 2 a model of the system and matrix converter is presented. In section 3 direct predictive speed control cost function described and in section 4 stability proof for current reference part and observer is provided. In section 5 simulation results are discussed to demonstrate the performance of the controller.

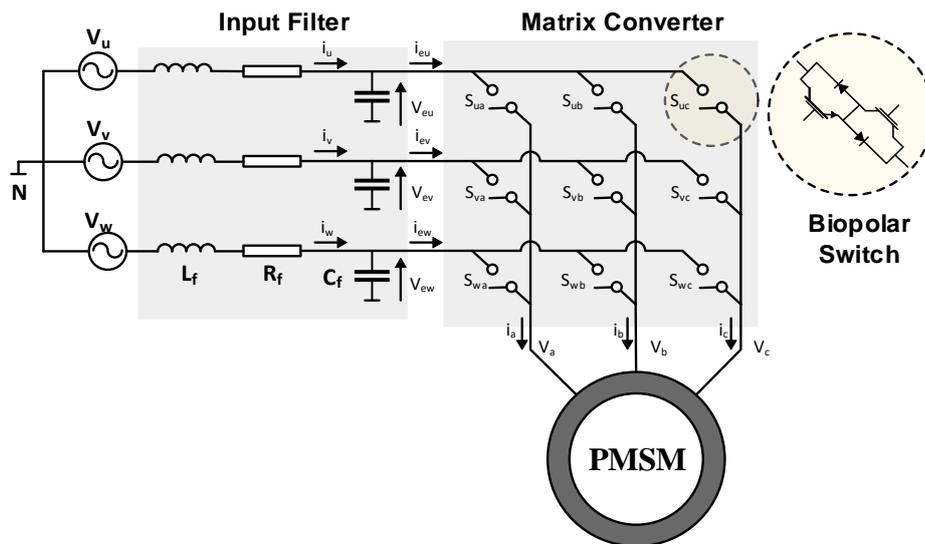


Figure 1. Schematic of a matrix converter drive for PMSM

## 2. SYSTEM MODEL

As it was mentioned, all MPC based control algorithms uses system model to predict next state, so system model has significant impact on this algorithm. In the following section the models for MC, PMSM and input filter are described.

### 2.1. Matrix converter model

Figure 1 shows a  $3 \times 3$  MC with 9 bidirectional switches. Each switch is composed of two power transistors and two parallel diodes. The MC is connected to the three-phase source through the input filter that its inductance is  $L_f$ , its capacitance is  $C_f$  and parasitic resistance is  $R_f$ . This filter will eliminate high frequency harmonics in input currents ( $i_u, i_v, i_w$ ) and avoid voltage spikes. The mathematical relationship between input and output voltage and current of MC are [23]:

$$\begin{bmatrix} v_a \\ v_b \\ v_c \end{bmatrix}_{v_o} = \underbrace{\begin{bmatrix} S_{ua} & S_{va} & S_{wa} \\ S_{ub} & S_{vb} & S_{wb} \\ S_{uc} & S_{vc} & S_{wc} \end{bmatrix}}_T \begin{bmatrix} v_{eu} \\ v_{ev} \\ v_{ew} \end{bmatrix}_{v_i}, \begin{bmatrix} i_{eu} \\ i_{ev} \\ i_{ew} \end{bmatrix}_{i_i} = \underbrace{\begin{bmatrix} S_{ua} & S_{ub} & S_{uc} \\ S_{va} & S_{vb} & S_{vc} \\ S_{wa} & S_{wb} & S_{wc} \end{bmatrix}}_{T^T} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix}_{i_o} \quad (1)$$

Where  $T$  and  $T^T$  are instantaneous transfer matrix and its transpose respectively and  $i = \{u, v, w\}$ ,  $j = \{a, b, c\}$ .  $S_{ij} = 1$  if the switch that connects  $i_{th}$  input to  $j_{th}$  output phase is ON and  $S_{ij} = 0$  if the switch is OFF.

Due to the inductive nature of the load, sudden interrupt in current would cause overvoltage that can destroy the component. Also, switches should not short circuit two input phases; because of this restriction, following equation should always be satisfied which reduces the number of possible switchings to 27.

$$S_{uy} + S_{vy} + S_{wy} = 1 \forall y \in \{a, b, c\} \quad (2)$$

## 2.2. Input filter model

Input filter shown in Figure 1 is just like an RLC circuit, and can be described by the following state space model [23]:

$$\dot{\mathbf{x}}(t) = \underbrace{\begin{bmatrix} 0 & \frac{1}{C_f} \\ -\frac{1}{L_f} & -\frac{R_f}{L_f} \end{bmatrix}}_{A_c} \mathbf{x}(t) + \underbrace{\begin{bmatrix} 0 & -1 \\ \frac{1}{L_f} & 0 \end{bmatrix}}_{B_c} \mathbf{u}(t) \quad (3)$$

Where:

$$\begin{aligned} \mathbf{x}(t) &= [\mathbf{v}_i(t) \quad \mathbf{i}_s(t)]^T, \mathbf{u}(t) = [\mathbf{v}_s(t) \quad \mathbf{i}_i(t)]^T \\ \mathbf{v}_s(t) &= \frac{2}{3(v_u + \mathbf{a}v_v + \mathbf{a}^2v_w)} \\ \mathbf{i}_s(t) &= \frac{2}{3(i_u + \mathbf{a}i_v + \mathbf{a}^2i_w)} \end{aligned} \quad (4)$$

Where  $\mathbf{a} = e^{j\frac{2\pi}{3}}$ .

To use  $\mathbf{i}_s$  in FCS-MPC, a discrete model for input filter is needed. Considering a sample time  $T_{sp}$  (5) describes this model [24].

$$\begin{aligned} \mathbf{x}[k+1] &= A_k \mathbf{x}[k] + B_k \mathbf{u}[k] \\ A_k &= e^{A_c T_{sp}}, B_k = \int_0^{T_{sp}} e^{A_c(T_{sp}-\tau)} B_c d\tau \end{aligned} \quad (5)$$

To predict  $\mathbf{i}_s[k+1]$  one can use equation (5) or  $A_k, B_k$  and equation (6).

$$\begin{aligned} \mathbf{i}_s[k+1] &= A_k(2,1)\mathbf{v}_i[k] + A_k(2,2)\mathbf{i}_s[k] \\ &+ B_k(2,1)\mathbf{v}_s[k] + B_k(2,2)\mathbf{i}_i[k] \end{aligned} \quad (6)$$

Using equation (6), a term with relation to  $\mathbf{i}_s$  can be added to objective function and control using FCS-MPC.

## 2.3. PMSM model

The dynamic model for a PMSM in the  $dq$  reference frame can be described as follows:

$$\begin{aligned} \frac{di_d}{dt} &= \frac{1}{L_d} (v_d - R_s i_d + \omega_e L_q i_q) \\ \frac{di_q}{dt} &= \frac{1}{L_q} (v_q - R_s i_q - \omega_e L_d i_d - \omega_e \psi_{mg}) \\ \frac{d\omega_e}{dt} &= \left( \frac{3}{2J_m} Z_p^2 \psi_{mg} i_q - \frac{B_v}{J_m} \omega_e + d \right) \end{aligned} \quad (7)$$

Whereas  $R_s$  is the stator resistance,  $L_d = L_q = L_s$  are d and q axis inductances of the motor,  $Z_p, J_m, B_v, \psi_{mg}$  are number of pole pairs, motor inertia, friction coefficient and flux linkage of the permanent

magnet motor respectively and  $d$  is a lumped sum disturbance for load torque and other uncertainties in mechanical subsystem  $d = f(T_L, \Delta J_m, \Delta B_v)$ . The value of  $d$  has important impact on controller performance, so it will be estimated using an observer in section 5. Also  $v_d$  and  $v_q$  are voltage vectors in  $dq$  reference frame that are related to switching voltages as Figure 2.

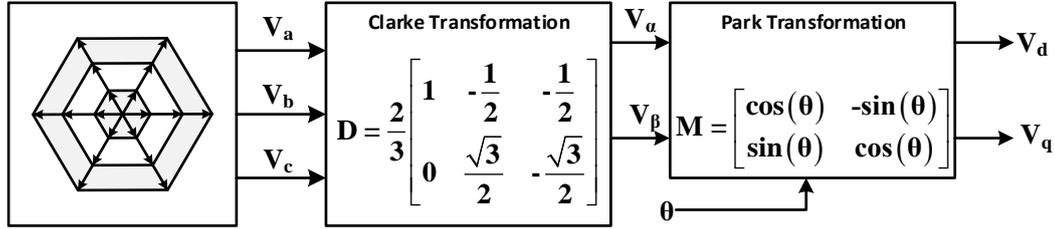


Figure 1. Park-Clarke transformation for dq axis voltages

To predict future values of current and speed, a discrete model of PMSM is needed for FCS-MPC. To reduce calculation burden, a simple discretization method with good performance should be selected. Here the model discretized using forward Euler method [23] as follow, is used, whereas  $T_{sp}$  is sampling time.

$$\begin{aligned} i_d[k+1] &= \left(1 - \frac{T_{sp}R_s}{L_s}\right) i_d[k] + \frac{T_{sp}}{L_s} v_d[k] + T_{sp}\omega_e[k]i_q[k] \\ i_q[k+1] &= \left(1 - \frac{T_{sp}R_s}{L_s}\right) i_q[k] + \frac{T_{sp}}{L_s} v_q[k] - T_{sp}\omega_e[k]i_d[k] - \frac{T_{sp}\psi_{mg}}{L_s}\omega_e[k] \\ \omega_e[k+1] &= \left(1 - \frac{T_{sp}B_v}{J_m}\right) \omega_e[k] + T_{sp}\left(\frac{3}{2J_m}Z_p^2\psi_{mg}i_q[k] + d\right) \end{aligned} \quad (8)$$

### 3. COST FUNCTION SELECTION

FCS-MPC takes advantage of the inherent discrete form of the power converters. The control objectives will be predicted for finite number of acceptable switching states of the power converter. The predicted variables will be compared with their reference values through a cost function. The switching state that minimizes the cost function will be applied to the converter to exert the voltage vector to the load in the next control interval, therefore the presence of a modulator is not required [16].

According to the FCS-MPC scheme cost function design is a key point. By proper selection of cost function, FCS-MPC is able to control multiple objectives at the same time. However, proper selection of items and weightings is a challenging task. In this section items of cost function are discussed.

#### 3.1. PMSM cost term

The main goal of the controller is to control the speed of PMSM so that the error between predicted speed and measured speed could be a candidate for speed cost.

$$e_\omega = (\omega_e^* - \omega_e^p)^2 \quad (9)$$

Where  $\omega_e^p$  is predicted value and  $\omega_e^*$  is the reference value of speed. In contrast to electrical subsystem, mechanical subsystem dynamic is sluggish. So for better performance MPC needs a longer horizon to decrease current and torque distortions when speed error becomes small [18]. We add a current term to speed cost function to not only eliminate longer horizon necessity, but also take care of outer loop stability and current distortion. So the tracking part of cost function should be as follow:

$$C_T = \lambda_\omega(\omega_e^*[k] - \omega_e[k+1])^2 + \lambda_{i_q}(i_q^*[k] - i_q[k+1]) \quad (10)$$

$$i_q^* = \frac{1}{a}(\dot{\omega}_e^* + b\omega_e - \hat{d} + ke_\omega + k_{sw}sign(e_\omega)) \quad (11)$$

Where  $\hat{d}$  is the estimation of disturbance  $d$  and  $a$ ,  $b$  and  $sign(e_\omega)$  are as follows:

$$a = \frac{3}{2J_m} Z_p^2 \psi_{mg}, b = \frac{B_v}{J_m} \&sign(e_\omega) = \begin{cases} +1 & e_\omega > 0 \\ -1 & e_\omega < 0 \end{cases} \quad (12)$$

Stability proof for  $i_q^*$  in presence of  $\hat{d}$  is discussed in section 4.

### 3.2. Constraints cost term

There are two terms that should be considered, one for current limitation and another one for Maximum Torque Per Ampere (MTPA) criteria [24]:

$$C_L = \begin{cases} \lambda_L (\sqrt{i_d^2 + i_q^2} - I_{\max}) & \sqrt{i_d^2 + i_q^2} > I_{\max} \\ 0 & o.w \end{cases}$$

$$C_{Z_i} = \lambda_{id} \left( i_d + \frac{L_d - L_q}{\lambda} (i_d^2 - i_q^2) \right)^2 \xrightarrow{L_d = L_q} C_{Z_i} = \lambda_{id} i_d^2 \quad (13)$$

### 3.3. Input filter cost term

When we are controlling speed of PMSM using direct MC, control of input current is a very challenging task even by FCS-MPC, but if one needs to control input current as primary control objective, like unit power factor in generative mode, a reactive power term should be added to the cost function. Using equation (6) for grid current, reactive power would be calculated using following equation:

$$Q^p = v_{s\beta} i_{s\alpha}^p - v_{s\alpha} i_{s\beta}^p \quad (14)$$

Where  $\alpha$  and  $\beta$  are real and imaginary part of  $\mathbf{v}_s$  and  $\mathbf{i}_s$  in equation (4). So by adding  $C_{Z_i}$  and reactive term to single term we will have a term called zero term that means this part of cost function should go to zero.

$$C_Z = \lambda_{id} (i_d)^2 + \lambda_Q (Q^p)^2 \quad (15)$$

As in this work, our primary task control is speed of PMSM,  $C_Z$  does not have reactive power term. By combining  $C_T$  as tracking term,  $C_L$  as limitation term and  $C_Z$  as zero term the cost function will be ready for FCS-MPC.

$$g = C_T + C_Z + C_L \quad (16)$$

It is worth mentioning that finding optimal weighting factors when there are constraints in cost function is not a straight task. For this work we used multiple simulations to find weighting factors with good performance of the system.

## 4. DISTURBANCE OBSERVER AND CURRENT REFERENCE STABILITY

Disturbances and uncertainties have high impact on FCS-MPC performance so an accurate disturbance observer can improve proficiency of the system. In this section a disturbance observer designed to estimate load torque value and other uncertainties in mechanical subsystem as a lumped sum disturbance. Let the motor currents and speed be measurable and consider mechanical part of (7) as follow:

$$\dot{\omega}_e = a i_q - b \omega_e + d, a = \frac{3}{2J_m} Z_p^2 \psi_{mg}, b = \frac{B_v}{J_m}$$

$$d = \dot{\omega}_e + b \omega_e - a i_q \quad (17)$$

The design procedure of the disturbance observer is as follow:  
Define the disturbance dynamic as:

$$\dot{\hat{d}} = K(d - \hat{d}) \quad (18)$$

Where  $K$  is a positive constant. To have globally stable estimation without steady state error, an auxiliary variable is defined like the one introduced in [25]:

$$z = \hat{d} - K\omega_e \quad (19)$$

By differentiating (19) and replacing (17) and (18) to new equation we have:

$$\begin{aligned} \dot{z} &= \dot{\hat{d}} - K\dot{\omega}_e = K(d - \hat{d}) - K\dot{\omega}_e \\ &= K(\dot{\omega}_e + b\omega_e - ai_q - \dot{\omega}_e) - K\dot{\hat{d}} \\ &= K(b\omega_e - ai_q) - K\dot{\hat{d}} \end{aligned} \quad (20)$$

So disturbance observer can be designed using equation (21):

$$\begin{cases} \dot{z} = K(b\omega_e - ai_q) - K\dot{\hat{d}} \\ \hat{d} = z + K\omega \end{cases} \quad (21)$$

Theorem 1. Using the observer designed with (21) and  $i_q$  from (11), the speed error, defined as  $\tilde{\omega}_e = \omega_e^* - \omega_e$ , and the error of disturbance, defined as  $\tilde{d} = d - \hat{d}$ , will converge to zero asymptotically. Proof. Consider the following Lyapunov function candidate:

$$V = \frac{1}{2}(\tilde{\omega}^2 + \tilde{d}^2) \quad (22)$$

The derivation of  $V$  respect to time is:

$$\dot{V} = \tilde{\omega}\dot{\tilde{\omega}} + \tilde{d}\dot{\tilde{d}} \quad (23)$$

By substituting (17), (18) and (21) into (23), it can be written that:

$$\begin{aligned} \dot{V} &= \tilde{\omega}(\dot{\omega}_e^* - ai_q + b\omega_e - d) + \tilde{d}(\dot{d} - \dot{\hat{d}}) \\ &= \tilde{\omega}(\dot{\omega}_e^* - ai_q + b\omega_e - d) + \tilde{d}(\dot{d} - \dot{z} - K\dot{\omega}_e) \end{aligned} \quad (24)$$

Using  $i_q$  from (11),  $\dot{z}$  from (20) and considering this fact that in practical engineering we always consider  $\dot{d} = 0$  [26] we have:

$$\begin{aligned} \dot{V} &= \tilde{\omega}(-k\tilde{\omega} - k_{sw}\text{sign}(\tilde{\omega}) - (d - \hat{d})) \\ &+ \tilde{d}(-K(b\omega_e - ai_q) + K\dot{\hat{d}} - K\dot{\omega}_e) \\ &= -k\tilde{\omega}^2 - k_{sw}|\tilde{\omega}| - \tilde{d} \\ &+ \tilde{d}\left(-K(\underbrace{\dot{\omega}_e + b\dot{\omega}_e - ai_q}_d) + K\dot{\hat{d}}\right) \\ &= -k\tilde{\omega}^2 - k_{sw}|\tilde{\omega}| - \tilde{d}\tilde{\omega} - K\tilde{d}^2 \\ &\leq -k\tilde{\omega}^2 - K\tilde{d}^2 - k_{sw}|\tilde{\omega}| + |\tilde{d}||\tilde{\omega}| \end{aligned} \quad (25)$$

Using observer error dynamic  $\dot{\tilde{d}} = -K\tilde{d}$ , we know that the observer is asymptotically stable and we can assume  $|\tilde{d}| \leq D$  for sure. So, the following inequality for derivation of Lyapunov function is satisfied:

$$\begin{aligned} \dot{V} &\leq -k\tilde{\omega}^2 - K\tilde{d}^2 - k_{sw}|\tilde{\omega}| + D|\tilde{\omega}| \\ &= -k\tilde{\omega}^2 - K\tilde{d}^2 - (k_{sw} - D)|\tilde{\omega}| \end{aligned} \quad (26)$$

If we choose a new variable as  $\bar{k}_{sw} = k_{sw} - D$  and  $k_{sw} > D$ :

$$\begin{aligned} \dot{V} &\leq -k\tilde{\omega}^2 - K\tilde{d}^2 - \bar{k}_{sw}|\tilde{\omega}| \\ \dot{V} &\leq -\bar{K}\left(\frac{1}{2}\tilde{\omega}^2 + \frac{1}{2}\tilde{d}^2\right) = -\bar{K}V \end{aligned} \quad (27)$$

Where  $\bar{K} = 2 \min\{k, K\}$ .

Using the lemma at [26] the solution of  $\dot{V} \leq -\bar{K}V$  is:

$$V(t) \leq e^{-\bar{K}(t-t_0)}V(t_0) \tag{28}$$

So the sliding mode surface converge to zero exponentially and by converging  $V$  to zero, error of speed and disturbance estimation goes to zero and the proof is completed.

### 5. SIMULATION AND RESULTS

To prove the effectiveness of the proposed model, a PMSM fed by MC is simulated using MATLAB/Simulink. Parameters of the PMSM are listed in Table 1. Also Figure 3 shows the flowchart of proposed method based on FCS-MPC and  $V(n)$  is the  $n^{\text{th}}$  vector of MC voltage in  $abc$ ,  $dq$  or  $\alpha\beta$  reference frame when it is needed.

To show the effectiveness of the proposed method, two other different speed control schemes are used in simulation. First scheme is direct predictive speed control (PSC) without current term that introduced in [18] for back to back converter and other scheme is a control system using PI with anti-windup structure as outer loop controller and FCS-MPC as inner loop current controller.

**Remark:** One of the drawbacks of FCS-MPC is the time of optimization procedure. Evaluation of cost function for 27 different available voltages is time consuming. Also, in some cases more than 27 calculations are required. The method in [18] used three step horizons for lower torque oscillation and current distortion, that it means three times prediction for each voltage vector are required, so the method needs  $27^3$  calculations. If one needs to use lower number of voltage vectors in each sampling interval, there is some methods for back to back convertors [27] and matrix converters [28]. In proposed method we used just one step to predict future values of variables and need 27 calculation of cost function however we could use method in [28] to have even lower calculation burden.

Table 1. Parameters of simulation system

Parameter	Value	Unit	Parameter	Value	Unit
Stator resistance $R_s$	1	$\Omega$	Viscous damping $B_v$	0.0093	Nm.s
Stator inductance $L_s, L_{\sigma}, L_{\sigma q}$	3.2	mH	Filter resistance $R_f$	$\Omega$	1
Number of pole pairs $Z_p$	4		Filter inductance $L_f$	mH	4
Flux linkage $\psi_{mg}$	0.126	web	Grid voltage/Freq	V/Hz	100/50
Moment of inertia $J_m$	0.126	gr/m <sup>2</sup>	$I_{max}$	A	7

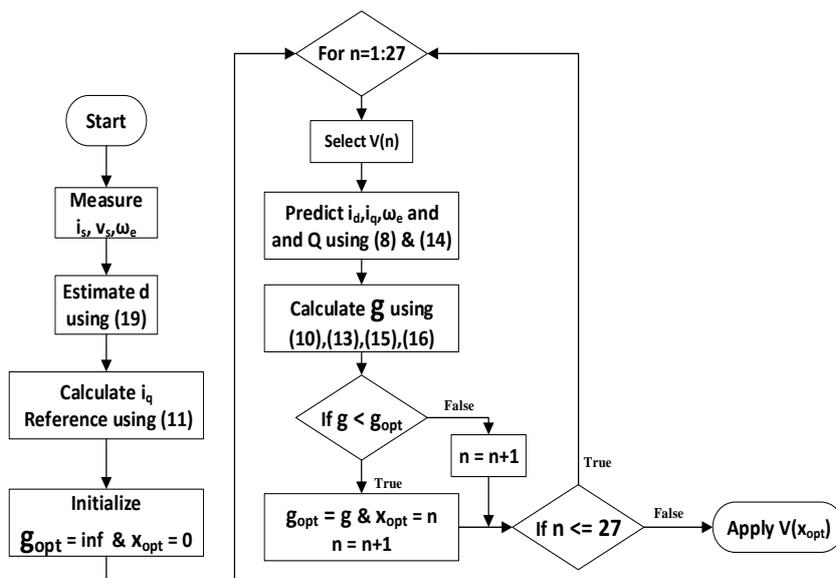


Figure 2. Flowchart of FCS-MPC

Figure 4 illustrates reference values of speed and load torque during simulation. Also estimated value of load torque using disturbance observer is shown and it can be seen that the observer has very good performance during step changes of load torque and steady state error is zero.

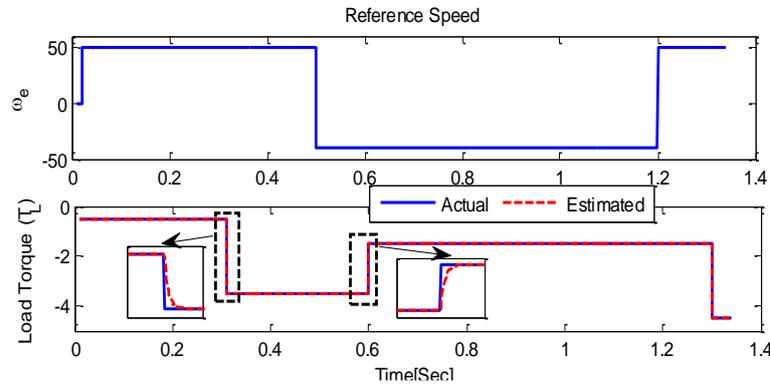


Figure 3: Reference speed, actual and estimated value of load torque

Figure 5 shows reference speed and PMSM speed using 3 different control schemes. It can be seen that two-method based on predictive control have faster dynamic in contrast to cascade structure with PI control. It is worth to mention that PI controller design is optimal and higher gains for higher speeds causes more overshoot in system. Figure 6 illustrates zoomed areas marked with green rectangle in Figure 5 for more details. Subplot (a) shows very low overshoot for proposed method in comparison to other methods even PSC and subplot (b) support this opinion. Subplot (c) shows speed variation during load torque step change and it can be seen that proposed method does not cause oscillation and track the reference speed as fast as possible. Finally, subplot (d) shows a steady state performance and it demonstrates that proposed method has lower steady state error and Table 2 contains Mean Square Error (MSE) values for this part of simulation just for comparison.

During speed transient, current performance is an important issue. With higher speed error weighting in tracking function (10), speed dynamic will be faster but it would cause more current distortion. Current variation during simulation is illustrated in Figure 7 for  $i_q$ . It can be seen that non-cascade structure schemes have more similar behavior and cascade controller with PI have different response.

For more details Figure 8 shows zoomed areas marked with green rectangle in Figure 7. It is noticeable that all control schemes exert current limitation and PI controller have lower current value in transient mode. Subplot (c) shows step down in q axis current for PSC and proposed method during speed change. It is clear that proposed method has lower undershoot with no oscillation and subplot (d) illustrate this fact during step up change.

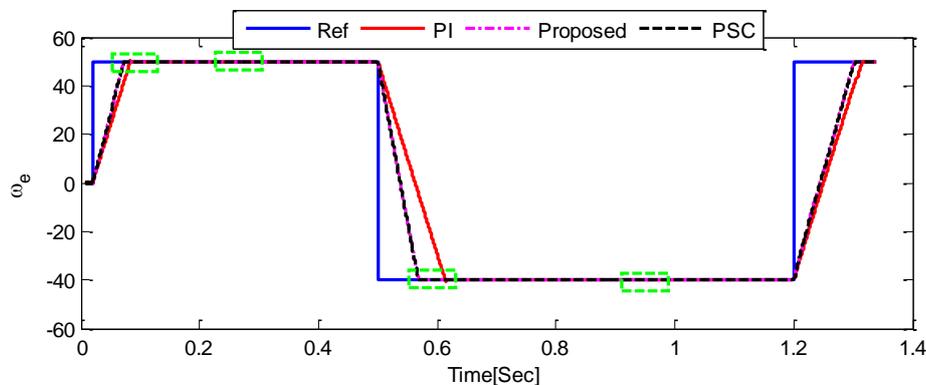


Figure 4: Reference speed and PMSM speed using three different control methods, Predictive Speed control (PSC) [18] (dash, black), PI control as outer loop (solid, red), proposed method (dash-dot, magenta)

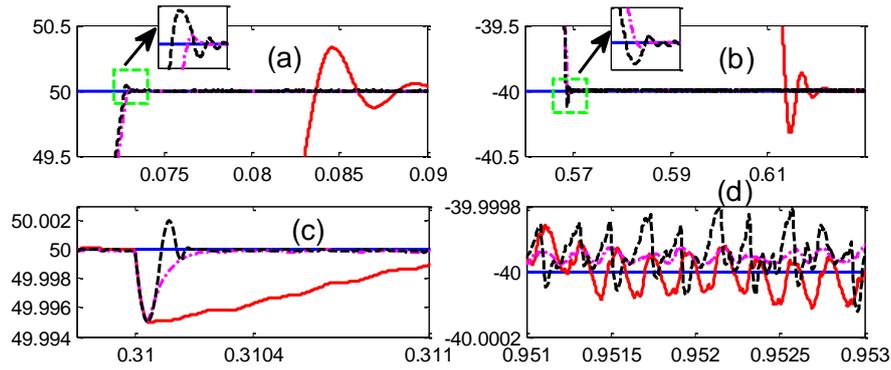


Figure 5. Zoom view of Figure 5 marked with green rectangle

Table 2. Mean square error of PMSM speed for  $t = 0.951s$  to  $t = 0.953$

	Proposed method	PSC method	PI method
MSE	3.4553e-09	2.6177e-06	8.7708e-07

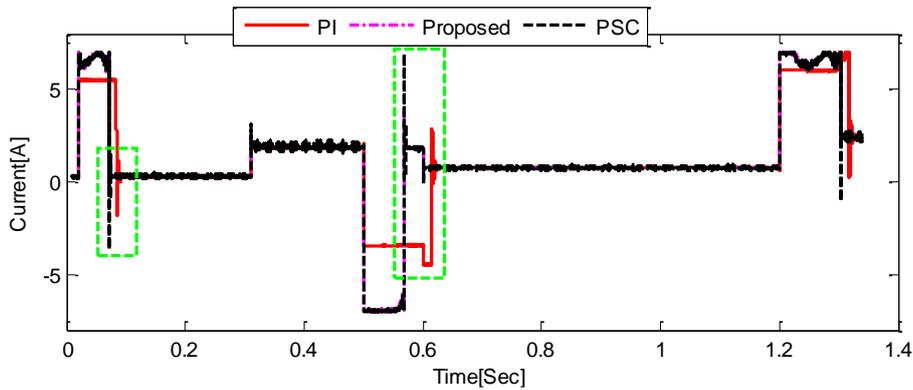


Figure 6. Current transient during simulation time using three different control method, Predictive Speed control (PSC) [18] (dash, black), PI control as outer loop (solid, red), Proposed method (dash-dot, magenta)

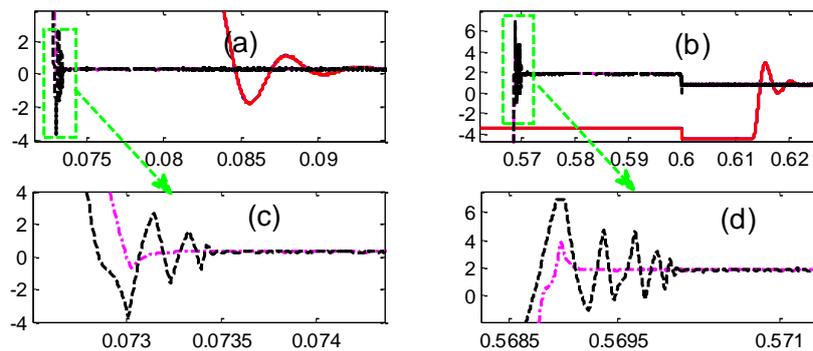


Figure 7: Zoomed view of Figure 7 marked with green rectangle in Figure 7

### 6. CONCLUSION

In this paper a direct predictive speed control for PMSM fed by matrix converter is introduced. A new speed tracking cost function is designed, by combining speed and current dynamics with system constraints to have high performance with low oscillation in speed without cascade structure. To resolve

stability issue and eliminate high current distortion due to difference between mechanical and electrical time constants, a current reference constructed based on sliding mode concept is been added to tracking term of cost function. This term guarantees outer loop stability and using simulation results it can be seen that this term can decrease current distortion. Also, to have lower steady state error and better performance a disturbance observer is designed to estimate load torque value and other uncertainties of the mechanical subsystem as a lumped disturbance. This observer permits elimination of load torque sensor and increases reliability of the system.

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