

Novel methodology for direct speed control of a permanent magnet synchronous motor with sensorless operation

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Abstract
A new formulation of the direct speed control (DSC) method for permanent magnet synchronous motor (PMSM) is proposed. The proposed DSC method employs a direct voltage control to synthesise the \( d \) and \( q \) axes stator voltage. In this method, no \( q \)-axis current loop regulation is required. Thus, a new concept of the dynamic limiter (DL) is proposed to limit the undesirable current rises. The mathematical analyses, frequency and time domain simulations, and experimental test results highlight the simplicity and effectiveness of the proposed DSC method. Furthermore, sensorless operation of the method can make it a reliable alternative for low-cost drive of PMSM.

1 | INTRODUCTION

The permanent magnet synchronous motor (PMSM) has high efficiency, high torque to inertia ratio, high power density, and low rotor losses [1, 2], which make it a preferable choice in industrial applications. In spite of the mentioned merits of a PMSM, the control method design is demanding [3]. A control method should be robust against changes to the motor parameters. In addition, it is desirable to have a simple structure.

Over the past few decades, several researches have been carried out regarding PMSM control methods. First, the field-oriented control (FOC) method was proposed in 1969 [4]. In this method, the stator current is resolved on the \( dq \)-axes. In order to regulate \( d \)-axis current (flux component) and \( q \)-axis current (torque component), two proportional-integral (PI) controllers are used. In addition, one PI controller is employed for the speed loop regulation [5]. The key element of the FOC method is the phase angle of the rotor flux-linkage. Generally, the measured/estimated speed of the PMSM is used to determine the position of the rotor flux-linkage [6].

Another main control method of the PMSM is the direct torque control (DTC) method [7]. In contrast to the FOC, the torque and stator flux-linkage are the controlled variables in the DTC. It requires two hysteresis band controllers for regulating the torque and stator flux-linkage, and one PI regulator to control the motor speed [8]. The DTC method benefits from a fast dynamic, simple structure, and position sensorless operation. However, it suffers from high torque ripple and variable switching frequency [7–9]. In order to tackle these problems, space vector modulation-based DTC (DTC-SVM) was introduced. Contrary to the conventional DTC, in the DTC-SVM method, two PI controllers are used to regulate the torque and flux loop. Thus, considering a PI controller for speed loop, the DTC-SVM incorporated three PI controllers [9, 10].

With the revolutionary advancement of digital signal processing devices, model predictive control (MPC) was presented, which had several benefits such as simple structure and fast dynamic [11]. The MPC methods can be divided into two groups based on the purpose of the control method. If the current (torque) is the controlled variable, it is called model predictive current control (MPCC), whereas model predictive direct speed control (MPDSC) refers to the method that directly controls the speed of the motor. In addition to the general advantages of the MPC methods, the cascade controller was eliminated in the MPDSC. However, both MPCC and MPDSC are complicated and needed huge calculations in every sample time [11–13].

Sliding mode controller (SMC) is another method that is robust to internal parameter variation [14]. Thus, it is utilised in a variety of applications. Artificial neural network and fuzzy controller are new categories of control methods, known as modern control methods. These methods show robust performance; however, they undergo heavy computational burden [15, 16].
The adaptive speed controller (ASC) was introduced to deal with parameter variation and non-linearity of PMSM. The presented methods are capable of rejecting the whole system (including motor, inverter, the environment, etc.) parameter variation. However, it is difficult to model the overall process dynamics with the mathematical formulation. As a result, the utilisation of ASC is burdensome and undesirable for industry [17, 18].

Having a cascade controller or being computationally complicated are the drawbacks of the aforementioned control methods. The drawback of the cascade controllers is that the inner loop dynamics (current control loop) should be significantly faster than the dynamic of the outer loop (speed control loop). In other words, the bandwidth of the inner loop must be 5–10 times higher than that of the outer loop. The advantages of high bandwidth are as follows: stiffer motor performance, error decrement, and transient response time improvement. Also, the main drawback of the high bandwidth is that the motor responds to higher frequency disturbances. Therefore, the power dissipation can be increased significantly. As a result, the temperature of the motor will increase.

In the high-power applications of AC motors which operate in the low-speed region, the switching frequency of the inverter is kept at low value. As already explained in the above paragraph, the current loop should have significantly higher bandwidth than the speed loop. However, due to the defect of high bandwidth, it is advisable to choose the lower one in low-speed–high-power applications. Otherwise, interference between the inner loop and outer loop is probable [19, 20].

In order to address the described problems, a novel direct speed control (DSC) method is proposed. In the proposed DSC, the \( q \)-axis PI controller is omitted. Thus, unlike the FOC and DTC methods, the proposed DSC has two regulation loops.

Elimination of the \( q \)-axis current control loop makes \( i_{sq} \) prone to undesirable overshoots. Therefore, a current-limiting method should be used. In [20, 21], cost functions were introduced, which could limit the current magnitude during the control process. The current limitation was one of the objectives of the mentioned cost functions. Although the methods in [20, 21] yield acceptable results, they are complicated and burdensome. In [22], the dynamic limiter (DL) was presented. The input of the DL is a dimensionless quantity, which is compared with a constant value. Then, according to the value of the mentioned quantity, a constant factor is multiplied by \((\omega * - \omega)\). The obtained signal is the input of the PI controller in the speed regulation loop. The explicit advantage of DL is its simplicity. In comparison to [22], several works are carried out here. The method in [22] was not experimentally verified. However, the complete experimental results are added here. Moreover, the operation of the control method in [22] is not sensorless. In addition, a novel concept of the DL is proposed which shows faster dynamic than that of [22].

According to the various references, the methods of the motor positioning determination can be classified into two categories: position detection using mechanical sensors and position detection based on the electrical sensors (sensorless methods) [23–25]. The mechanical sensors are robust against faults. On the other hand, the electrical sensors (current and voltage) are sensitive to faults. Also, utilising several electrical sensors for speed and position estimation reduces the reliability of the control system. However, since the mechanical sensors cost more than the electric ones, much effort is concentrated on the sensorless operation of the PMSM. Therefore, the proposed method is free from mechanical sensors. The primary goal is to propose a simple (from the viewpoints of design and mathematical formulation), low cost, and sensorless control method for PMSM.

In Section 2, the dynamic model of a PMSM is formulated. Section 3 outlines the proposed control method. Thorough mathematical analyses and explanations about the proposed method are presented. In addition, the new methodology of the DL is proposed. In Section 5, the simulation and experimental results are illustrated. The comparative simulation results consist of frequency domain analysis and time domain analysis for both proposed and FOC methods. Then, in Section 5, a comparison is made between the proposed DSC method and the conventional and modern control methods. Finally, the conclusion is given in Section 6.

## 2 | Dynamic Model of PMSM

The voltage-current model of PM in the synchronous reference frame can be written as follows [22]:

\[
\begin{align*}
  u_{sd} &= R_i s_i d + \frac{d\phi_{sd}}{dt} - \omega \phi_{sq} \\
  u_{sq} &= R_i s_i q + \frac{d\phi_{sq}}{dt} + \omega \phi_{sd}
\end{align*}
\]  

(1)

(2)

In addition, the flux-linkage-current model of PMSM is expressed by the following matrix:

\[
\begin{align*}
  \phi_{sd} &= L_s i_{sd} + \phi_{pm} \\
  \phi_{sq} &= L_s i_{sq}
\end{align*}
\]  

(3)

(4)

In Equations (1) to (4), \( u_{sd} \) and \( u_{sq} \) are \( d \)- and \( q \)-axes components of the stator voltage, respectively. \( i_{sd} \) and \( i_{sq} \) are \( d \)- and \( q \)-axes components of the stator current, respectively. \( \phi_{sd} \) and \( \phi_{sq} \) are \( d \)- and \( q \)-axes components of the stator flux-linkage, respectively. \( L_s, L_{sd}, L_{sq}, R_i \) are \( d \)-axis stator inductance, \( q \)-axis stator inductance, and stator resistance, respectively. \( \phi_{pm} \) is the flux-linkage of the permanent magnet of rotor. Also, \( \omega_e \) is the synchronous speed of the rotor. The electromagnetic torque of the PMSM is:

\[
T_e = 0.75P[(L_{sd} - L_{sq})i_{sd}i_{sq} + \phi_{pm}i_{sq}]
\]  

(5)
where, $T_e$ and $P$ are the electromagnetic torque and pole numbers, respectively. Finally, the dynamic equation of the mechanical motion is:

$$T_e - T_L = J \frac{d\omega_{mehc}}{dt} + B\omega_{mehc}$$  \hspace{1cm} (6)

where, $T_L$, $J$, $\omega_{mehc}$, and $B$ are load torque, rotor inertia, rotor mechanical speed, and friction coefficient, respectively.

3 | THE PROPOSED CONTROL METHOD

In [26], the concept of active flux was introduced. For PMSM, $\varphi_a$, which is defined as active flux, can be written as follows:

$$\varphi_a = (L_{s,d} - L_{s,q})i_{s,d} + \varphi_{pm}$$  \hspace{1cm} (7)

Considering Equations (1)–(4) and (7), the stator voltage equations are rewritten as:

$$u_{s,d} = R_i i_{s,d} + L_{s,d} \frac{di_{s,d}}{dt} - \omega_e L_{s,q} i_{s,q}$$  \hspace{1cm} (8)

$$u_{s,q} = R_i i_{s,q} + L_{s,q} \frac{di_{s,q}}{dt} + \omega_e [\varphi_a + L_{s,q} i_{s,d}]$$  \hspace{1cm} (9)

According to Equations (5) and (7), the electromagnetic torque expression can be written as follows:

$$T_e = 0.75P \varphi_a i_{s,q}$$  \hspace{1cm} (10)

If $L_{s,q} = L_{s,d}$ is satisfied, the PMSM will be the surface-mounted PM (SPMSM). Therefore, Equations (1)–(10) can be simplified. However, it only affects $\varphi_a$, and the overall control procedure remains unchanged. The block diagram of the proposed DSC method is shown in Figure 1. The objective is to tackle the pre-defined desired value of $\omega_m^*$ by the motor speed $\omega_m$. This is achieved by direct and sensorless control of the speed of PMSM. The stator voltages and currents (of two phases) are merely the measured quantities for the proposed DSC. The analogue-to-digital converter (ADC) is utilised to convert analogue measurements into digital ones. $i_{s,d}^*$ is the reference value of the $d$-axis current component. According to the types of PMSM (surface mounted/interior mounted) and the operational region (constant torque/constant power), various values can be assigned to $i_{s,d}^*$. Moreover, $\omega_m^*$ is the commanded speed value.

The PMSM speed error ($\omega_m^* - \dot{\omega}_m$) and the $d$-axis current error ($i_{s,d}^* - i_{s,d}$) are inputs of the PI controllers. Outputs of the mentioned controllers are $u_{s,d}^*$ and $u_{s,q}^*$, respectively. The proposed method uses $u_{s,q}^*$, $u_{s,q}^*$, and $u_{s,e}$, which are the result of inverse park transformation, to fed to the SVM block [27]. Then, the required duty cycles of the inverter are generated.

According to Figure 1, usage simplicity, elimination of the cascade controller, sensorless operation, and $q$-axis current loop (PI controller) elimination are the advantages of the proposed DSC.

The proposed DSC method consists of three main parts, which are explained in the following subsections.

3.1 | $d$-Axis control loop

The objective of this control loop is to regulate the $d$-axis component of the stator current. Figure 2 shows the detailed $d$-axis control loop of the proposed DSC method. The $d$-axis model of PMSM is derived from Equation (8). $\tau_d$ is the time constant of the $d$-axis equation, $k_d^*$ and $k_e^*$ are the coefficients of the PI controller of the $d$-axis. In addition, the inverter is modelled as a first-order transfer function, in which $k_{inv}$ and $\tau_{inv}$ are the gain of the mentioned transfer function and the time constant of the inverter, respectively. According to Equation (8), the non-linear term can be eliminated by the input–output linearisation technique. Given that $\tau_{inv}$ is almost equal to a few tens of microseconds, the effect of the inverter model can be neglected in the open-loop transfer function of the $d$-axis current loop. As a result, the controller design becomes simple. Finally, the PI controller can be chosen for regulating the $d$-axis control loop.

3.2 | $q$-Axis control loop

The primary goal of the $q$-axis loop is to apply direct control to the speed of PMSM. This is achieved using a single control
loop. In other words, the \( q \)-axis current regulation loop, which existed in the conventional methods, is omitted. Consequently, the \( q \)-axis voltage of the stator is controlled directly for speed regulation purposes. The secondary goal for the \( q \)-axis loop is control of \( q \)-axis current component, which will be clarified at the end of this subsection. Figure 3 is a detailed illustration of the control of the \( q \)-axis loop. In other words, the \( q \)-axis current regulation loop is neglected.

Similar to the \( d \)-axis control loop, the non-linearity of Equation (9) is compensated using the input–output linearisation method. In addition, considering the previous subsection explanations, the impact of the inverter model on the open-loop transfer function of the \( q \)-axis loop is neglected.

In the following, the effects of \( i_{s,q} \)-axis elimination are assessed. Considering \( \omega _{\text{mech}} = (2/\pi) \omega _e \), substituting Equations (6) and (10) into Equation (9) yields:

\[
\frac{1.5(P/2) \varphi _q}{JL_s} u_{s,q} = \left[ 1 + \frac{1}{\tau_q} \right] \frac{d^2 \omega_e}{dt^2} + \left[ \frac{1}{\tau_m} + \frac{1}{P/2} \right] \frac{d \omega_e}{dt} + \left[ \frac{1}{\tau_q \tau_m} \right] \omega_e + \left( \frac{1}{\tau_q} \right) T_L + \left( \frac{1}{J} \right) \frac{d T_L}{dt}
\]

(11)

According to Equation (11), \( u_{s,q} \) is obtained by summation of two groups of voltages. The first group is dependent on \( \frac{d \omega_e}{dt} \), \( \frac{d \omega_e}{dt} \), and \( \omega_e \), which is named \( u_{s,q}^{\omega} \). The second group, \( u_{s,q}^{\tau} \), is dependent on \( \frac{d T_L}{dt} \) and \( T_L \). According to the previous explanations, the non-linear part of \( u_{s,q} \) is omitted.

During operation in steady state and transient state, \( u_{s,q} \) can be expressed as follows:

1. Steady-state operation:

2. Transient state operation:

\[
\frac{1.5(P/2) \varphi _q}{JL_s} u_{s,q} = \omega_e + \left( \frac{1}{\tau_q} \right) T_L + \left( \frac{1}{J} \right) \frac{d T_L}{dt}
\]

(12)

The second-order characteristic equation of (12) is as below:

\[
s^2 + \left( \frac{1}{\tau_q} + \frac{1}{\tau_m} \right) s + \frac{1}{\tau_q \tau_m} = 0
\]

(13)

Equation (13) has two poles on the real axis, which are obtained as follows:

If the PMSM is rotating with a pre-defined speed without load torque, \( u_{s,q} = u_{s,q}^{\omega} (\omega_e) \).

If the load torque is applied on the PMSM, while it is rotating with the pre-defined speed value, \( u_{s,q} = u_{s,q}^{\omega} (\omega_e) + u_{s,q}^{\tau} (T_L) \).

2. Transient state operation:

○ If the load torque is applied on PMSM, while it is rotating with the pre-defined speed value,

\[
u_{s,q} = u_{s,q}^{\omega} \left( \omega_e + \frac{d \omega_e}{dt} \right) + u_{s,q}^{\tau} (T_L).
\]

○ If the reference value of speed changes during the rotation (or start-up) of PMSM with pre-defined speed value,

\[
u_{s,q} = u_{s,q}^{\omega} \left( \omega_e + \frac{d \omega_e}{dt} \right).
\]

Considering the above explanation and the structure of the conventional method, it can be understood that the \( q \)-axis current control loop is responsible for producing the desired load torque. Moreover, the outer speed loop will generate the required \( u_{s,q}^{\omega} \).

In contrast to the conventional method, in the proposed DSC method, \( u_{s,q}^{\omega} \) is equal to \( u_{s,q}^{\tau} \). This is because applying/removing the load torque causes changes in the speed value. As a result, \( u_{s,q}^{\omega} \) can supply the required \( u_{s,q}^{\tau} \) value. Therefore, the electromagnetic torque, which is proportional to the load torque and the shaft speed, reaches the reference values if the stator current and voltage stay within safe regions.

Applying the above details into Equation (11), it is rewritten as follows:

\[
\frac{1.5(P/2) \varphi _q}{JL_s} u_{s,q} = \frac{d^2 \omega_e}{dt^2} + \left( \frac{1}{\tau_q} + \frac{1}{\tau_m} \right) \frac{d \omega_e}{dt} + \frac{1}{\tau_q \tau_m} \omega_e
\]

(13)

\[
s^2 + \left( \frac{1}{\tau_q} + \frac{1}{\tau_m} \right) s + \frac{1}{\tau_q \tau_m} = 0
\]
\[ s_1 = \frac{1}{\tau_m} \]
\[ s_2 = -\frac{1}{\tau_q} \]

where, \( s_1 \) and \( s_2 \) are mechanical and electrical poles, respectively. According to the design procedure and structure of typical PMSM, \(|s_1| < |s_2|\). Therefore, the mechanical dynamic is slower than the electrical dynamic. As a result, the impact of the electrical pole on the response of the system in Equation (12) can be ignored. According to the simplification procedure of Equation (12), the open-loop transfer function of the system is written as:

\[
\frac{\Omega_m}{U_{s/q}} = 0.75P\varphi_e \frac{1}{\varphi_e s + \tau_m s^2} \]

where, \( \Omega_m \) and \( U_{s/q} \) are the Laplace transformation of \( \omega_m \) and \( u_{s/q} \), respectively. Consequently, a PI controller is used in the proposed DSC method for speed regulation.

The main drawback of the cascade controller in conventional control methods, which incorporates two PI controllers in the \( q \)-axis, was the interference between an inner and outer loop in the low-speed–high-power applications. In the proposed DSC methods, according to Figure 3 and previous explanations, a single loop is utilised for speed regulation purposes. In other words, the interference between the inner and outer loops is not of concern for the proposed DSC method.

In spite of the mentioned merits, the proposed method is not complete. Due to the elimination of the \( q \)-axis current control loop, \( i_{i,q} \) can experience large overshoots/undershoots during applying/removing the load torque. Thus, a limiting block should be considered in the \( q \)-loop. The DL is a simple current limiter. The input of the DL is an index named ‘\( a \)’, which is a dimensionless quantity that depends on the values of \( \omega^*_m \) and \( \omega_m \). According to the value of ‘\( a \)’ (0 \( \leq a \leq 1 \)), a constant value is multiplied by \( (\omega^*_m - \omega_m) \) as the output of the DL. The mentioned procedure is completed by comparing the value of ‘\( a \)’ with the predefined ranges [22].

The new concept of the DL in [22] is proposed. The proposed DL has a stepped line characteristic, which results in two advantages.

First, the slope of variation in the DL is not constant. That is, when the difference between the reference speed and shaft speed of PMSM is high, the average slope is gentle, and vice versa. The second benefit is the simple table-based structure of the DL in a way that the DL operation can be completed by a simple comparison and evaluation of the mentioned table. This makes the design process considerably simple. The novelty of the proposed DL is about the input parameter. In contrast to the definition in [22], \( a = 0.5 \left( \frac{|\omega^*_m - \omega_m|}{|\omega^*_m + \omega_m|/2} \right) \) is defined as an input variable of the DL. In the nominator and denominator, the absolute mark is utilised. As a result, when the reference speed is increased or decreased, the current-limiting ability of the proposed controller is maintained. Also, in the denominator, instead of \( \omega^*_m \), \( (\omega^*_m + \omega_m)/2 \) is used. According to Figure 4, ‘\( a \)’ of the proposed DL shows a faster dynamic than ‘\( a' \) of [22]. Therefore, in addition to a restricted current response, faster dynamic is achievable.

### 3.3 Speed estimation block

Among various speed estimator/observer candidates, the back-electromotive force (EMF)-based method is selected for estimation in low-speed to high-speed operations [28]. The advantage of this method is its simplicity. However, the back-EMF method undergoes sensitivity to the motor parameter correctness and the low switching frequency of the inverter during low-speed operation [29].

The principle of the estimation method relies on the existence of the back-EMF only on the \( q \)-axis of the reference frame. The synchronous reference frame is based on the knowledge of the rotor position \( \theta_r \). Inside the proposed DSC, this position is unknown. Therefore, to drive the PMSM, another fictitious frame \( d' \) and \( q' \)-axes are used with assuming the position \( \theta_r \). Figure 5a shows the position of the \( dq \)-axes and \( d'q' \)-axes towards each other. According to Figure 5a, the transformation matrix between \( dq \)-axes and \( d'q' \)-axes is as below:

\[
\begin{bmatrix}
  x_d' \\
  x_q'
\end{bmatrix} =\begin{bmatrix}
  \cos \Delta \theta & \sin \Delta \theta \\
  -\sin \Delta \theta & \cos \Delta \theta
\end{bmatrix} \begin{bmatrix}
  x_d \\
  x_q
\end{bmatrix}
\]

Considering Equations (17), (8), and (9), the voltage-current model of the PMSM in \( d'q' \)-axes frame could be expressed as follows:

\[
u_{is,d} = R_i i_{is,d} + L_{s,d} \frac{di_{is,d}}{dt} - \omega_L L_{s,q} i_{is,q} + \omega_e \varphi_e \sin(\Delta \theta)
\]

\[
u_{is,q} = R_i i_{is,q} + L_{s,q} \frac{di_{is,q}}{dt} + \omega_L L_{s,d} i_{is,d} + \omega_e \varphi_e \cos(\Delta \theta)
\]

\[
\Delta \theta = \tan^{-1} \left( \frac{u_{i,s,d} - R_i i_{is,d} - L_{s,d} \frac{di_{is,d}}{dt} + \omega_L L_{s,q} i_{is,q}}{u_{i,s,q} - R_i i_{is,q} - L_{s,q} \frac{di_{is,q}}{dt} - \omega_L L_{s,d} i_{is,d}} \right)
\]

**Figure 4** Comparison of ‘\( a \)’ of the proposed dynamic limiter and ‘\( a' \) of [22]
The speed estimation algorithm is achieved by compensating the position error \( \Delta \theta = \theta_e - \theta' \) in Equation (19) between the real dq-axes and fictitious d'q'-axes. The compensation is completed by utilizing a PI controller. The detailed block diagram of the speed estimator is shown in Figure 5.

In order to decrease the speed estimation error during low-speed operation, two methods are chosen. Firstly, to compensate for the error caused by imprecise motor parameters estimation, a trial and error method is used. In other words, according to the initial value of the parameters, the motor is controlled and based on the respective outputs, the controller parameters are changed to obtain the best possible responses. Secondly, to compensate for the error caused by low switching frequency, the estimator controller bandwidth is adjusted so that the effect of high-frequency signals on the speed estimation becomes as low as possible.

4 | EXPERIMENTAL AND SIMULATION VALIDATIONS

According to Figure 1, the simulation of the proposed method is applied on the PMSM in MATLAB/SIMULINK environment. Furthermore, a laboratory prototype is implemented to validate the effectiveness of the proposed DSC. A DSP controller board is used for this purpose. The general view of the experimental setup is shown in Figure 6. According to Figure 6, the prototype consists of the following parts: (a) a SPMSM (with parameters given in Table 1) (b) a 4600 \( \mu F \) and 600 V capacitor as DC link, (c) a tapped transformer with rectifying diode, (d) a small transformer for gate drive supply, (e) electrical sensors, (f) inverter switches, (g) a DSP board, and (h) an oscilloscope. To supply the input of the inverter, the tap of multi-tap transformer is chosen to obtain almost 550 V on the DC link capacitor. The switching frequency and sampling frequency are set to 5 and 10 kHz, respectively.

The gate drive pulses generated by the DSP board are fed to the gate driver of the inverter using an op-amp circuit that shifts the voltage level of the DSP from 3.3 to 15 V. The stator current is sensed using the ACS712 current sensor and is sent to the controller board (DSP) with the help of an analogue-digital converter.

The motor speed is obtained using a speed estimator, which has been explained in the previous section. The speed reference and its estimation are monitored using a computer system and the current responses are shown by an oscilloscope. The output range of the current sensor is 0–5 V. Considering that the output of the sensor is fed to the DSP, an intermediate circuit is utilised to reduce the voltage to 3.3 V. Thus, in the stator current waveforms, 1 V corresponds to 1.52 A. In order to control the PMSM in the constant torque region, \( i_s^* \) is set to zero. Also, in order to achieve a fair performance comparison between the FOC and the proposed DSC method, the PI coefficients are selected based on the pole assignment method. Then, the obtained coefficients are modified to achieve two goals. Firstly, the most rapid settling time for speed

Table 1: Permanent magnet synchronous motor ratings and parameters

| Ratings and Parameters       | Value  |
|------------------------------|--------|
| Rated power                  | 400 W  |
| Line voltage                 | 380 V  |
| Rated current (rms)          | 2.5 A  |
| Rated speed                  | 3000 rpm|
| Pole                         | 8      |
| Stator inductance            | 5.9 mH |
| Stator resistance            | 0.9 \( \Omega \) |
| Permanent magnet flux        | 0.0933 Wb|
| Rotor inertia                | 0.62 \( \times 10^{-4} \) kg.m\(^2\)/s |
| Viscous coefficient          | 3.183 \( \times 10^{-4} \) kg.m\(^2\)/s |
response is obtained. Secondly, the current response of the motor does not exceed two times the nominal current. The reason behind the comparison between the FOC method and the proposed DSC method relies on the fact that the \( d \)-axis control loop of the proposed DSC method is similar to that of the FOC method. Various simulation and experimental results are obtained, which are given in the following paragraphs. This part starts with a frequency domain analysis and is followed by time domain simulation and experimental results. With the nominal parameters in Table 1, the open-loop Bode plot of the \( q \)-axis in the proposed DSC method is shown in Figure 7. The DL in the proposed method has a non-linear characteristic. Thus, using [22], the impact of the different DL factors (\( a = 0.9, 0.8, \ldots \)) on the magnitude of the frequency response is plotted in Figure 7. Without the DL (when \( a < 0.3 \)) the crossover frequency of the system is 1150 rad/s. When the DL restriction is applied, the crossover frequency is reduced until it reaches 342 rad/s (for \( a \geq 0.9 \)). Since the DL only has an impact on the Bode magnitude diagram, the Bode phase plot is constant for the mentioned cases. Consequently, the crossover frequency reduction results in an increment of the phase margin. According to Figure 7, the phase margin varies from 7.3\(^\circ\) to 22.8\(^\circ\) and the gain margin is infinite, theoretically. According to the achieved value for gain margin and phase margin, the proposed DSC method with the DL has stable performance.

Among the 'a' values, which are indicated in Figure 7, \( 'a: \text{average}' \) is not the case in [22]. Considering the 'a', which is defined here, and the conditions in [22], the operational range for each level of the flowchart in [22] can be achieved. For instance, \( a \geq 0.9 \) and \( a \geq 0.8 \) indicate \( \dot{\omega} \leq 0.053\omega^* \) and \( \dot{\omega} \leq 0.11\omega^* \), respectively. Also, the remaining conditions in [22] can be converted to the inequalities between \( \dot{\omega} \) and \( \dot{\omega}^* \). Finally, according to the per-unit outputs of the DL and the respective conditions (which are considered as a weighing factor), the average per-unit output of the DL is calculated as follows:

\[
[0.1 \times (0.053 - 0) + 0.2 \times (0.11 - 0.053) + \ldots] = 0.75 \quad (21)
\]

Using \( 0.75(\omega^* - \dot{\omega}) \) as the average output of the DL allows a fair comparison between the proposed method and the FOC method.

Figure 8 illustrates the Bode characteristic of the closed-loop plant for the proposed DSC method (when \( a < 0.3 \), \( a \geq 0.9 \), and \( a: \text{average} \)) and the FOC method, wherein the output is the mechanical speed and the input is the reference speed.

According to Figure 8, at low frequencies, the output tracks the reference signal. However, when \( a \geq 0.9 \) is satisfied, the mentioned condition lasts until the frequency of 71 rad/s. In addition, there are two common measures of the performance evaluation in the closed-loop Bode plot. The first is the closed-loop bandwidth \( \omega_{\text{bw}} \), which is defined as being the point where the Bode magnitude plot is 3 dB below 0 dB. Considering Figure 8, when the DL is in the restriction mode, \( (a \geq 0.9) \), \( \omega_{\text{bw}} \) has the lowest (362 rad/s) value. On the other hand, when the output of the DL equals \( \omega^* - \dot{\omega} (a < 0.3) \), the highest (1460 rad/s) value of \( \omega_{\text{bw}} \) is achieved, which is higher than that of the FOC method (1240 rad/s). Also, \( \omega_{\text{bw}} \) in the average performance of the DL is 1185 rad/s.

According to Figure 8 and the above paragraphs, \( \omega_{\text{bw}} \) of the proposed DSC method varies from 362 rad/s to 1460 rad/s. As a result, during speed variation, the bandwidth frequency of the proposed method increases simultaneously with an increment of the motor speed. The second point for performance evaluation is the resonant peak. According to Figure 8, the resonant peak of the proposed DSC method varies from 6.07 dB (for \( a \geq 0.9 \)) to 15.51 dB (for \( a = 0.9 \)). The resonant peak for the FOC method is 16.18 dB, which is higher than that of the proposed method. It is worthy of note that, when the difference between \( \omega^* \) and \( \dot{\omega} \) is high \( (a \geq 0.9) \), the current response is prone to the largest overshoots. However, thanks to the inherent characteristics of the proposed method, these overshoots are restricted by the DL. Although the frequency response of the system gives useful information, it cannot be utilised as the only factor for evaluation of the system performance. This is due to the fact that the system has various non-linearities, which cannot be considered in the presumed model. Therefore, the time domain analysis will be given in the following.

The functionality and reliability of the proposed DSC method are verified by comparing the simulation and experimental results of the DSC and the FOC methods in low-speed operation, medium-speed operation, and high-speed operation. In the experimental speed response, the red and blue graphs are the estimated speed and reference speed, respectively. The PI controller coefficients of the proposed DSC method are as follows: \( k_p^d = 5.04 \), \( k_i^d = 4864.25 \), \( k_p^w = 0.05 \), and \( k_i^w = 20.37 \). Also, the PI controller coefficients of the FOC method are as follows: \( k_p^d = 2.51 \), \( k_i^d = 3450.54 \), \( k_p^w = 1.87 \), \( k_i^w = 822.08 \), \( k_p^\omega = 0.26 \), and \( k_i^\omega = 3.45 \).

To evaluate the performance of the proposed DSC in the low-speed operation, a step reference command is generated for speed. Figure 9a–d show comparative results of the
simulation in the low-speed operation mode. Also, Figure 10a–d illustrate the experimental comparative results corresponding to Figure 9. A speed reference of 180 rpm (6% of nominal value) is applied at $t = 0.1\text{s}$. Then, at $t = 0.5\text{s}$, 25% of the nominal load torque (0.325 N.m) is applied on the shaft of the motor. From Figure 10a,c, the proposed DSC method has a faster speed response than the FOC method. The proposed method and the FOC method almost have settling times of 0.27 and 0.48s, respectively. Also, the undershoot and overshoot of the speed response are very similar (40 and 22 rpm, respectively). The speed oscillation occurred in the proposed method. This is due to the low inertia of the motor. The speed estimation errors in the mentioned figures are around 10 rpm and exist due to the weakness of the estimation method. Figures 10b,d show the current response in the loading moment. When the load is applied, the current rise in the FOC method is obvious, and is limited by the DL in the proposed method. Figure 11 shows the ability of the proposed DSC and FOC methods during medium-speed operation. Speed references of 450 rpm (15% of nominal value) and 1080 rpm (36% of nominal value) are applied at $t = 1.5\text{s}$ and $t = 3\text{s}$, respectively. Also, 50% of the nominal load torque is applied on the shaft of the motor. Figure 11a–d present the simulation results of the mentioned methods. Also, Figure 12a–c and Figure 13a–c illustrate the experimental results of the proposed DSC and FOC methods, respectively (corresponding to Figure 11). According to the simulation and experimental results of the FOC method, when the step change is introduced to the reference speed, the overshoot occurs in the speed response. However, in these moments, the smooth speed response of the DSC method is obvious. The settling times of the proposed method and the FOC method are around 0.26 and 0.37s, respectively. Therefore, the proposed method has faster dynamic during the load change. Figures 12b–c and 13b–c show the current response during the full time range of the experiment and during the second step change of the speed for the proposed DSC and FOC methods, respectively. The effective performance of the DL can be assessed by Figures 12b–c and 13b–c. The maximum current rise in the FOC method is 2.13 A, whereas that of the proposed method is

**FIGURE 8** The Bode diagram of the closed-loop system for the proposed direct speed control (DSC) method and the field oriented control (FOC) method!
1.36 A. The only defect in the performance of the proposed method is the estimated speed oscillation during the load change. The speed estimation error is around 14 rpm (for both methods), which is initiated by the inverter non-linearity mostly.

In order to evaluate the performance of the mentioned methods in the nominal speed operation, the speed of the motor is first ramped to the rated value of 3000 rpm in 1 s. After 1 s, the nominal torque is applied to the shaft of the motor and removed at $t = 3\, s$. Finally, the speed decreases to 0 rpm in 1 s. Figures 14a–d illustrate the results of the simulation in the high-speed operation mode. Also, Figures 15a–d show the experimental comparative results corresponding to Figure 14.

According to Figure 15a, the nominal speed performance of the proposed method can be measured from three viewpoints. During the ramp-up and ramp-down times, the speed estimation error occurs. This is a defect of the DL utilisation. However, after 0.2 s, the estimation error reduces considerably. In the steady state, the speed estimation overcomes the parameter uncertainty and inverter non-linearity. The current waveforms in the experimental results indicate the moment that the load is applied to the motor shaft. In

**FIGURE 10** Experimental results of the proposed and field oriented control (FOC) methods during low-speed operation ($25\%T_L$). (a) speed response of direct speed control (DSC), (b) current waveform of DSC, (c) speed response of FOC, (d) current waveform of FOC

**FIGURE 11** Simulation results of the proposed and field oriented control (FOC) methods during medium-speed operation ($100\%T_L$). (a) speed response of direct speed control (DSC), (b) current waveform of DSC, (c) speed response of FOC, (d) current waveform of FOC
the steady state, the estimation errors for both methods are 6–12 rpm. The settling times of the proposed method and the FOC method are 0.35 and 0.5s, respectively.

According to the obtained simulation and experimental results (Figures 9–15), it can be seen that the proposed method has a faster speed response than the FOC method during the load change. Moreover, a smooth speed response for the proposed method is achievable. Meanwhile, this advantage is the result of the DL utilisation. The other impact of DL utilisation is the restricted current responses, which are attainable in the proposed method. Also, the superiority of the FOC method over the proposed method is as follows: the FOC method has a smooth speed response at the load change moment. Also, during step change of the reference speed, the estimated speed has a faster response.

In order to clarify the estimation error in the previously explained scenarios, Figures 16 and 17 are given to show the errors in the proposed method and the FOC method, respectively. According to the previous paragraphs, during low-speed and medium-speed operations the estimation error occurs due to the motor parameter uncertainty and inverter non-linearity, whereas the mentioned error is reduced mostly during the nominal speed operation. According to Figures 16 and 17, the steady-state estimation error of the proposed method is similar to that of the FOC method.

In order to verify the performance of the proposed method in the very low-speed operation, the 30 rpm step reference speed signal (1% nominal value) is generated at \( t = 0.1 \)s and is given to the controller. Figure 18 shows the comparative simulation result of the mentioned case. Because the back-EMF speed estimation method is prone to various errors and noise during the very low-speed operation, only the simulation result is shown here. Considering Figure 18, although the proposed DSC method and the FOC method have similar performance in the steady state, the FOC method has a faster response in the speed step change.

Finally, the steady-state operation of the proposed method is checked during the nominal speed and nominal load torque operation. Figures 19a–d present the speed response and current waveform of the proposed DSC method and the FOC method.

5 | COMPARISON

A novel, simple, and effective control method for PMSM has been proposed, and the comparison between the proposed DSC method and other well-known control methods has been accomplished from three points of view. The measured computational time of the proposed method and the FOC method are 37\( \mu \)s and 36\( \mu \)s, respectively. Therefore, the FOC method has a lower computational burden. However, the computational time of the proposed method is considerably lower than that of the model predictive methods and adaptive controller-based methods [13, 30, 31].
The proposed method utilises the general voltage-current and current-flux equations of the motor (the same as the conventional method). However, the model predictive methods and the adaptive controller-based methods use complex mathematical equations, which make the mentioned system complicated. Indeed, the adaptive controller-based methods are more complicated than the others [11–13, 32]. Moreover, unlike the conventional methods, the proposed DSC method uses a direct speed (voltage) control procedure. Therefore, the structure of the proposed controller is simple. In the modern control methods, the MPDSC has a simplicity in the structure. Table 2 summarises all of these comparisons.

Considering Sections 3 and 4, and Table 2, the advantages of the proposed DSC method are as listed here:

1. The cascade structure of the speed loop, which is necessary in the conventional methods, is omitted. The cascade structure has a problem with high power application, where
the switching frequency is low. Furthermore, careful attention should be paid to the coordinated design of the inner and outer control loops.
2. Considering the reduction in the number of PI controllers, the overall controller design process becomes simpler than the FOC method.
3. Mathematical formulation of the control method is as simple as the conventional methods.
4. It has a simple structure with acceptable responses.
response of the stator current remains within the permissible range, a novel concept of the DL has been suggested. As a result, the current increment during the load applying, removal, and start-up process is effectively restricted. However, the DL has non-linear characteristics and reduces the closed-loop bandwidth of the proposed controller. The proposed method has been verified by mathematical analyses. Also, in addition to the frequency domain and time domain simulations, an experimental prototype has been implemented (in the proposed DSC and FOC methods). The results of the analysis, simulation, and experimental tests prove that the proposed DSC method has a simple structure, effective performance, and accurate responses. Finally, a brief comparison is made between the proposed method and the conventional and modern control methods.

6 | CONCLUSION

A novel direct speed control method of PMSM with sensorless operation has been proposed. This method has reached the DSC with direct voltage control and without q-axis current loop regulation. Therefore, the proposed DSC method has a more simple structure than the conventional control method, which uses a cascaded control strategy. To ensure that the

| TABLE 2 | Comparison of the proposed DSC method and other well-known methods |
|----------|-------------------------------------------------|
| Method   | Simple Structure | Mathematical Formulation | Cascade Controller |
| FOC and DTC [11–13] | Yes | Low | Yes |
| MPCC [11, 30] | Yes | High | Yes |
| MPDSC [11, 13, 30, 31] | Yes | High | No |
| ASC [17, 18, 32] | No | High | Yes |
| Proposed DSC | Yes | Low | No |

Abbreviations: ASC, adaptive speed controller; DSC, direct speed control; DTC, direct torque control; FOC, field oriented control; MPCC, model predictive current control; MPDSC, model predictive direct speed control.

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