Efficiency-Optimized Variable Structure Direct Torque Control for Synchronous Reluctance Motor Drives

Direct torque control (DTC) is known to produce fast response and robust control in ac adjustable-speed drives. However, in the steady-state operation, notable torque, flux and current pulsations occur. A new direct torque and flux control strategy based on variable-structure control and space vector pulse width modulation is proposed for synchronous reluctance motor drives considering iron losses. The DTC transient merits and robustness are preserved and the steady-state behavior is improved by reducing the torque and flux pulsations. In addition, efficiency optimization strategy is executed by adjusting the amplitude of stator reference flux continuously, independent of drive parameters. Finally, the validity and capability of the method are verified by simulation and experimental results.

Keywords: Flux Control, Robust Control, Synchronous Reluctance, Torque Control, Variable Structure Control.

1. INTRODUCTION

In the past two decades, as a number of benefits of the Synchronous Reluctance Motor (SynRM) over other types of ac machines stand out, the interest in this machine has increased [1]. A SynRM is advantaged on induction motor by the absence of rotor copper losses, on brushless motors by inexpensive rotor structure, and on switched reluctance motor by a much lower torque ripple and low noise. Compared to Surface type Permanent Magnet Synchronous Motor (SPMSM), it is capable of high-speed operation and for use in high-temperature environments.

As another topic in ac machine drives, the efficiency optimization has been developed in the case of adjustable speed drive systems. Since there are limitations in improving the efficiency through the machine design or alternation of mechanical structures, several papers have reported on using optimization techniques for the induction machine [2], [3]. In these control schemes, the efficiency optimization has been mainly focused on adjusting the flux level or balancing between the copper and iron losses. One main problem of such optimization schemes is that they are applicable only to the drives with slowly changing state because the rotor flux linkage of the induction motor has a relatively large time constant [3]. However, in the case of SynRM, the open circuit of rotor makes the flux linkage directly proportional to stator current and thus, without losing the fast dynamics, the efficiency optimization strategy for the SynRM can be applied to various adjustable speed drive applications.

The well-known direct torque control (DTC) strategy for ac motors control seems to be particularly useful for these ASDs, due to its robustness and functional simplicity [4-9]. Despite its simplicity, DTC is able to produce very fast torque and flux control and is robust with respect to motor parameters changes and to perturbations. However, in the steady-state operation, notable torque, flux and current pulsations, and acoustical noise occur. To overcome the above drawbacks, some researchers have tried to propose some different DTC space vector modulation (SVM) techniques or to improve switching state patterns [8],

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In [8] a model-based predictive DTC has been presented for SynRM with constant switching frequency and relative long sampling time. The model-based of [8] is difficult to implement in practice and needs a lot of computation time and storage. In addition in [8], no any conventional strategy related to SynRM has been implemented and the iron loss has been also ignored. Compared to [8], in this paper we introduce a new DTC-SVM scheme for SynRM in which the iron loss resistance is taken into account and can be applied for efficiency-optimized control strategy related to SynRM. In [9], a nonlinear method capable of high dynamic torque regulation and efficiency optimization of SynRM has been described based on input-output feedback linearization (IOFL) DTC-SVM. In [9], the SynRM efficiency is optimized by using Lagrange’s theorem [10]. Then, the efficiency optimization criterion and the motor torque are chosen as output variables. IOFL technique however, requires the full knowledge of the motor parameters with sufficient accuracy. In addition, two PI controllers employed in [9] that its coefficients should be properly adjusted.

Variable structure control (VSC) or sliding mode control (SMC) is an effective, high frequency switching control for nonlinear systems with uncertainties [10]. It features simple implementation, disturbance rejection, strong robustness and fast responses. Recently, several solutions that integrate the VSC and DTC principles (VS-DTC) within high performance ac drives have been proposed [11]-[13]. A VSC-DTC drive for induction machine [11, 12] was proposed with stator flux oriented. It employs a switching component and a linear PI regulator for torque and flux control respectively. The controllers were operated in the stator reference frame and it needs frame transformation from stationary reference to stator field orientation frame and vice versa, therefore has high computation time and computation storage. In order to solve some of the drawbacks of [11, 12], a VSC-DTC interior PMSM drive [12] was presented in stationary reference frame. Although the proposed method eliminates both PI regulators of flux and torque and reduces computation time, but that research work ignores iron losses and does not implement a certain torque control strategy.

Recognizing the VSC merits and the advantages of using the SVM, this paper presents a VSC-DTC solution for SynRM drive systems. The main purpose of this paper is to continue the research work described in [9] for the following gains

• To achieve the SynRM direct torque and stator flux control and to eliminate two PI controllers employed in [9]
• To implement efficiency-optimized strategy independent of drive parameters by adjusting the stator reference flux amplitude
• To use VSC for robust control of SynRM drive system instead of IOFL.

According to the authors little knowledge and search, this is the first time that the decoupled stator flux and torque control of SynRM is discussed based on VSC DTC-SVM. To confirm the validity of the presented approach, simulation results are carried out.

### 2. MODEL IN STATINARY REFERENCE FRAME INCLUDING IRON LOSSES

The circuit equations of SynRM including iron losses in the rotor d-q rotating coordinate are given as

\[
\begin{bmatrix}
    v_d \\
    v_q
\end{bmatrix} =
\begin{bmatrix}
    R_s + \Re p L_d & -\omega \Re L_q \\
    \omega \Re L_d & R_s + \Re p L_q
\end{bmatrix}
\begin{bmatrix}
    i_d^T \\
    i_q^T
\end{bmatrix}
\]

(1)
Where \( \mathfrak{g}_R = \left(1 + \frac{R_s}{R_i}\right) \), \( p = \frac{d}{dt} \) and \( R_s \) denotes stator resistance, \( L_d \) and \( L_q \) represent d and q axes inductances, respectively. Note that the resistance \( R_i \) represents the iron losses connected in parallel to both rotational and transient back EMF. Since \( R_i \) supplies an additional current path to each axis equivalent circuit, the torque depends not on measured terminal current, \( i_d^s \) and \( i_q^s \), but on \( i_d^T \) and \( i_q^T \). Relation between \( (i_d^T \) and \( i_q^T \)) and \( (i_d^s \) and \( i_q^s \)) can be obtained as

\[
\begin{bmatrix}
i_d^s \\
i_q^s
\end{bmatrix} = \begin{bmatrix}
\left(1 + \frac{L_d}{R_i}\right) & -\omega_\alpha \mathfrak{g}_L_q \\
\omega_\alpha \mathfrak{g}_L_d & \left(1 + \frac{L_q}{R_i}\right)
\end{bmatrix} \begin{bmatrix}
i_d^T \\
i_q^T
\end{bmatrix}
\]

Or

\[
\begin{bmatrix}
i_d^T \\
i_q^T
\end{bmatrix} = \frac{R_i}{\left(p + \frac{R}{L_d}\right)\left(p + \frac{R}{L_q}\right) + \omega_\alpha^2} \begin{bmatrix}
\frac{1}{L_q}\left(p + \frac{R}{L_q}\right) & \frac{\omega_\alpha}{L_q} \\
\frac{\omega_\alpha}{L_d} & \frac{1}{L_d}\left(p + \frac{R}{L_d}\right)
\end{bmatrix} \begin{bmatrix}
i_d^s \\
i_q^s
\end{bmatrix}
\]

Direct transformation of (1) to the stationary \( \alpha - \beta \) reference frame will make the model complicated containing both \( 2\theta_{re} \) and \( \theta_{re} \) (electrical rotor position) terms. The SynRM model can be re-written explicitly to retrieve the saliency based EMF [14] in the rotor reference frame.

\[
\begin{bmatrix}
v_d \\
v_q
\end{bmatrix} = \begin{bmatrix}
R_s + \mathfrak{g}_L pL_d & -\omega_\alpha \mathfrak{g}_L qL_q \\
\omega_\alpha \mathfrak{g}_L L_d & R_s + \mathfrak{g}_L pL_d
\end{bmatrix} \begin{bmatrix}
i_d^T \\
i_q^T
\end{bmatrix} + \mathfrak{g} \begin{bmatrix}
L_d - L_q \end{bmatrix} \begin{bmatrix}
\omega_\alpha \left(\omega_\alpha i_d^T - p i_q^T\right)
\end{bmatrix}
\]

(4)

The (4) can be transformed into the stationary reference frame

\[
\begin{bmatrix}
v_\alpha \\
v_\beta
\end{bmatrix} = \begin{bmatrix}
R_s + \mathfrak{g}_L pL_d & \omega_\alpha \mathfrak{g}_L \left(L_d - L_q\right) \\
-\omega_\alpha \mathfrak{g}_L \left(L_d - L_q\right) & R_s + \mathfrak{g}_L pL_d
\end{bmatrix} \begin{bmatrix}
i_d^T \\
i_q^T
\end{bmatrix} + \mathfrak{g} \begin{bmatrix}
L_d - L_q \end{bmatrix} \begin{bmatrix}
\omega_\alpha \left(\omega_\alpha i_d^T - p i_q^T\right)
\end{bmatrix} \begin{bmatrix}
\sin \theta_{re} \\
\cos \theta_{re}
\end{bmatrix}
\]

(5)

where \( v_\alpha \), \( v_\beta \), \( i_\alpha \) and \( i_\beta \) are voltages and currents in stationary reference frame. The second term on the right side of (5) is defined as a salient EMF expressed by

\[
e = \begin{bmatrix}
e_\alpha \\
e_\beta
\end{bmatrix} = \mathfrak{g} \begin{bmatrix}
L_d - L_q \end{bmatrix} \begin{bmatrix}
\omega_\alpha \left(\omega_\alpha i_d^T - p i_q^T\right)
\end{bmatrix} \begin{bmatrix}
\sin \theta_{re} \\
\cos \theta_{re}
\end{bmatrix}
\]

(6)

Using (7), a new model of the SynRM can be described by
\[
\begin{align*}
\frac{d}{dt} i_{\alpha}^T &= \begin{bmatrix} -R_d/(\Re L_d) & -\omega_n (L_d - L_q)/L_d \\ \omega_n (L_d - L_q)/L_d & -R_r/(\Re L_d) \end{bmatrix} i_{\alpha}^T + \\
& \begin{bmatrix} -1/(\Re L_d) & 0 \\ 0 & -1/(\Re L_d) \end{bmatrix} \begin{bmatrix} e_{\alpha} \\ e_{\beta} \end{bmatrix} + \frac{1}{(\Re L_d)} \begin{bmatrix} v_{\alpha} \\ v_{\beta} \end{bmatrix} \\
\frac{d}{dt} \lambda_{\alpha} &= v_{\alpha} - R_s i_{\alpha}, \quad \frac{d}{dt} \lambda_{\beta} = v_{\beta} - R_s i_{\beta}
\end{align*}
\]

where \( v_{\alpha}, v_{\beta}, i_{\alpha} \) and \( i_{\beta} \) are terminal voltages and currents in stationary reference frame. \( \lambda_{\alpha} \) and \( \lambda_{\beta} \) are stator flux linkages and \( i_{\alpha}^T \) and \( i_{\beta}^T \) are torque producing currents.

\[ T = \frac{3P}{2} (\lambda_{\alpha} i_{\beta}^T - \lambda_{\beta} i_{\alpha}^T) \]

where \( T \) is the estimated torque and \( P \) is number of pole pairs.

\[ \lambda_s^2 = \lambda_{\alpha}^2 + \lambda_{\beta}^2 \]

where \( \lambda_s^2 \) is the square stator flux linkage norm.

### 3. DIRECT TORQUE CONTROL PRINCIPLES

The DTC for ac drives is a strategy exclusively based on stator voltage control. The consecutive voltage vectors applied to the motor are directly selected on the basis of torque and flux errors. In this way, fast response and robust torque and flux control are obtained, without intermediate current control. The classic DTC uses bang-bang torque and flux controllers, without decoupling [15]. A simple switching logic (switching table) employs the output signals of these controllers to select the most appropriate voltage vector, i.e., the one which rapidly reduces the torque and flux errors. Due to the fact that the voltage vector is maintained for the whole duration of the control period, the classic approach causes large torque, flux and current ripple, accompanied by acoustical noise. The switching frequency of the power devices is variable and uncontrollable. One way to decrease the ripple is to significantly shorten the duration of the control period. This requires powerful/expensive digital signal processors (DSPs), and the switching frequency remains variable.

Another approach is to increase the control resolution. The torque ripple was reduced, in [11], when the control period was divided into three subintervals and a sequence of three voltage vectors was applied in this time. This refinement can be further continued, the control resolution will continue to increase and the ripple will decrease.

A suitable solution of increasing control resolution and reducing torque and flux ripples is the use of a variable structure controller with the DTC SynRM drive. This new control approach is described and evaluated in the next section.

### 4. VARIABLE STRUCTURE CONTROL SCHEME

The variable structure control strategy is based on the design of discontinuous control signal that drives the system states toward special manifolds in the state-space. These manifolds are chosen in a way that the system will have the desired behavior as the state converges to them. In this paper, based on standard DTC concept, a variable structure
controller for direct torque and stator flux control is exploited. The overall block diagram of the SynRM drive is shown in Fig. 1. The outer loop contains a speed PI controller which produces the reference torque command for the torque controller.

The inner loop includes variable structure DTC (VS-DTC) which calculates the most appropriate stator voltage vectors to drive the torque and flux to track their references. Space vector modulation provides a solution for high resolution control and constant inverter switching frequency.

The control strategy has more advantages over the DTC. In DTC scheme, the voltage vector selected by the switching table might be correct at the sampling instant, but not proper during the whole sampling interval. A consequence is the unwanted chattering of torque and flux, unless the sampling period is small enough. Another problem of the classic DTC is that large and small errors in torque and flux are not distinguished. In other words, the switching table based DTC always applies full voltage vectors both during transients and steady states regardless of large and small errors in torque and flux. The proposed control scheme operating in the stationary frame predicts the most appropriate voltage vector which is adaptive to error amplitude, i.e., small errors will be compensated by small voltage vectors and not driven by full voltage vectors. The switching vector is generated by symmetric SVM module. More voltage vectors are applied during each sampling interval which is divided into more sub-intervals. This increases control resolution and reduces torque and flux ripples greatly.

Fig. 1: Block Diagram of the proposed Variable Structure Direct Torque Controlled SynRM Drive system

5. TORQUE AND FLUX CONTROLLER

5.1 Sliding Surfaces

The control objectives are to track desired torque and flux trajectories. So the sliding surface is set as \( S = \begin{bmatrix} S_1 & S_2 \end{bmatrix}^T \). The switching surfaces are considered in the integral forms by

\[
S_1 = e_T(t) + K_T \int_0^t e_T(\tau) d\tau - e_T(0) \tag{11}
\]

\[
S_2 = e_\lambda(t) + K_\lambda \int_0^t e_\lambda(\tau) d\tau - e_\lambda(0) \tag{12}
\]

Where \( e_T = T_{e, \text{ref}} - T_e \) and \( e_\lambda = \lambda_{s, \text{ref}}^2 - \lambda_s^2 \) are the errors between the references and the estimated values of torque and square of stator flux. \( K_T \) and \( K_\lambda \) are positive control
gains and are chosen for the desired system dynamics. The manifold $S_1 = 0$ represents the torque regulation, and $S_2 = 0$ represents the tracking of square of flux magnitude. When the system states have reached the sliding manifold and stay on the surface, then we have $S_1 = S_2 = \xi_1 = \xi_2 = 0$. From (11) and (12), derivatives of $S$ are equal to zero, which gives
\[
\frac{d}{dt} e_T = -K_T e_T \tag{13}
\]
\[
\frac{d}{dt} e_\lambda = -K_\lambda e_\lambda \tag{14}
\]

The above equations ensure the errors ($e_T$ and $e_\lambda$) converge to zero. Since (11) and (12) are zeros from the beginning, complete robustness is achieved during the whole transients and the system will converge asymptotically to the origin with time constants of $1/K_T$ and $1/K_\lambda$. The convergence rates of error dynamics are determined by $K_T$ and $K_\lambda$. Then, the design task is reduced to enforcing sliding mode in the manifolds, $S_1 = 0$ and $S_2 = 0$ with discontinuous stator voltage space vectors.

5.2 Variable Structure control Law

The task is to design a VSC law to drive the state trajectory to the intersection of the above switching surfaces. In this effort, a VS controller is designed to generate the stator voltage command for SVM modulator. The motion projections of the system (7)-(10) on $S$ subspace are derived by differentiating the $S$ vector
\[
\xi_1 = \xi + K_T e_T = (\xi_{e,ref} - \xi_e) + K_T (T_{e,ref} - T_e) \tag{15}
\]
\[
\xi_2 = \xi + K_\lambda e_\lambda = (\xi_{s,ref} - \xi_s) + K_\lambda (\lambda^2_{s,ref} - \lambda^2_s) \tag{16}
\]

Substituting for $T_e$, $\lambda_s$ and their derivatives using (9) and (10) leads to
\[
\xi = F + E U \tag{17}
\]
where calculation for derivatives of $E$ and $F$ is shown below
\[
E = \begin{bmatrix}
-\frac{3P}{2} \left( i_{\beta} - \frac{\lambda_\beta}{\Re L_d} \right) & -\frac{3P}{2} \left( \frac{\lambda_\alpha}{\Re L_d} - i_{\alpha}^T \right) \\
-2\lambda_\alpha & -2\lambda_\alpha
\end{bmatrix} \tag{18}
\]
\[
F_1 = -\frac{3P}{2} \lambda_\alpha \left( \frac{\omega_e (L_d - L_q)}{L_d} i_{\alpha}^T - \frac{R_s}{\Re L_d} i_{\beta} - \frac{1}{\Re L_d} e_{\beta} \right) \tag{19}
\]
\[
F_2 = K_\lambda e_{\lambda} + 2\lambda_\alpha R_s i_{\alpha}^T + 2\lambda_\beta R_s i_{\beta}^T + K_T e_T \tag{20}
\]
and
\[
U = \begin{bmatrix} v_{\alpha} & v_{\beta} \end{bmatrix}.
\]

In VSC, a Lyapunov approach is used for deriving conditions on the control law that will drive the state orbit to the equilibrium manifold. The quadratic Layapunov function of the form (21) is selected
The time derivative of \( V \) on the state trajectories of (17) is given as
\[
\dot{V} = \frac{1}{2} (\mathbf{S}^T \mathbf{S}) \geq 0
\]  
(21)

\[
\dot{V} = \frac{1}{2} (\mathbf{S}^T \mathbf{S} + \mathbf{S}^T \dot{\mathbf{S}}) = \mathbf{S}^T \dot{\mathbf{S}} = \mathbf{S}^T (\mathbf{F} + \mathbf{E} \mathbf{U})
\]  
(22)

The switching control law must be chosen so that the time derivative of \( V \) is negative definite for \( S \neq 0 \). We may select
\[
\mathbf{U} = \begin{bmatrix} v_u \\ v_\beta \end{bmatrix} = -\mathbf{E}^{-1} \begin{bmatrix} \mu_1 & 0 \\ 0 & \mu_2 \end{bmatrix} \begin{bmatrix} \text{sign}(S_1) \\ \text{sign}(S_2) \end{bmatrix}
\]  
(23)

5.3 Proof of the Stability

For stability to the switching surfaces, it is sufficient to have \( \dot{J} \leq 0 \). By setting control gains, stability can be achieved provided the following condition is satisfied

If \( \mu_1 > \left| F_1 \right| \) and \( \mu_2 > \left| F_2 \right| \) then
\[
\dot{J} = \mathbf{S}^T \dot{\mathbf{S}} = \mathbf{S}^T \left( \mathbf{F} - \begin{bmatrix} \mu_1 & 0 \\ 0 & \mu_2 \end{bmatrix} \begin{bmatrix} \text{sign}(S_1) \\ \text{sign}(S_2) \end{bmatrix} \right) < 0
\]  
(24)

The time derivative of Lyapunov function \( \dot{J} \) is negative definite, then the system becomes asymptotically stable.

The VS controller developed above assures the tracking of the torque and stator flux of the SynRM drive. However, the fast switching may generate unexpected chattering. To alleviate the problem, the discontinuous part of the controller is smoothed out by introducing a boundary layer around the switching surface, that is
\[
\text{Sat} (S_i/\phi_i) = \begin{cases} 
1 & \text{if } S_i > \phi_i \\
S_i/\phi_i & \text{if } |S_i| < \phi_i \\
-1 & \text{if } S_i < -\phi_i 
\end{cases}
\]  
(25)

where \( \phi_i > 0 \), is the width of the boundary layer and \( i = 1, 2 \).

Where \( \mu_1 \) and \( \mu_2 \) are positive control gains.

6. EFFICIENCY-OPTIMIZED CONTROL OF SYNRM

In this paper, a well-known control strategy relating to SynRM (efficiency-optimized) is considered under the constraint of constant torque production. In efficiency-optimized scheme the minimization of input active power is selected as objective function.

6.1 Model-based Method

The maximum efficiency method can be obtained by minimizing the expression for the power into the machine under the constraint of constant power output. The input power \( P_{in} \) of the SynRM is obtained as
\[
P_{in} = \frac{3}{2} (v_d i_d^s + v_q i_q^s)
\]  
(26)

Using (1)-(7) in steady-state condition, \( P_{in} \) can be expressed as
\[
P_m = R_s (i_d^{T^2} + i_q^{T^2}) + \frac{1}{R_i} (1 + \frac{R_s}{R_i}) (L_d^2 i_d^{T^2} + L_q^2 i_q^{T^2}) \omega_c^2 \\
+ \frac{2R}{R_i} (L_d - L_q) i_d^{T^2} i_q^{T^2} \omega_c
\]  

(27)

The minimum input power can be found by taking the derivative of (27) with respect to \( i_d^T \) and \( i_q^T \). In this condition, the following expression for the torque current components is obtained [16].

\[
i_q^T = \pm i_d^T \cdot \frac{(R_s + \frac{1}{R_i} (1 + \frac{R_s}{R_i}) L_q^2 \omega_c^2)}{(R_s + \frac{1}{R_i} (1 + \frac{R_s}{R_i}) L_d^2 \omega_c^2)}
\]  

(28)

6.2 Search Method

The flux linkage of SynRM is directly proportional to stator current and thus, without losing the fast dynamics, the efficiency optimization strategy for the SynRM can be applied to various drive applications by adjusting the flux level [17]. If (26) is regarded as objective function under the constraint of constant motor torque production, this function can be minimized by adjusting the stator flux value in consecutive steps. Then, the magnitude of filtered input power is obtained and compared with previous value. This procedure is continued until when the motor input power is being decreased. The required stator flux magnitude can be finally detected which is almost equal to the value obtained from model-based method. This method is quite independent of drive parameters.

7. SIMULATION RESULTS

Some simulations are performed to show that the proposed VS-DTC is able to operate with reduced torque ripple, without compromising the fast dynamic response and robustness of torque and flux control, of the classic DTC [15]. A step by step computer program was developed to model the drive system control of Fig.1. The VS-DTC and classic DTC strategies are simulated for a 0.5 Hp four-pole three-phase SynRM drive (see Appendix A), at 5 kHz sampling frequency, and representative results are illustrated in Fig. 2. In both cases, at the instant 1 sec, the torque command is changed from -1 Nm to +1 Nm, with the stator flux maintained at the rated level, while the motor was running at 200 rpm. The motor torque magnitude in the VS-DTC drive is shown in Fig. 2(a), while Fig. 2(b) shows similar quantity in the classic DTC drive. The proposed controller is superior to classic DTC control, as it has low ripple, and is equally robust with respect to torque transients.
8. EXPERIMENTAL SETUP AND RESULTS

An overall block diagram of the proposed drive system is shown in Fig. 3(a). In order to evaluate the performance of the actual system, a PC-based prototype system was built and tested. A photo of laboratory setup is shown in Fig. 3(b) and consists of the following sections: A 0.5 Hp three-phase SynRM and a 1.1 kW dc generator which is used as its load and a three-phase voltage source.

The steady-state performance of the proposed VS-DTC for the SynRM drive is shown in Fig. 4 with the rotor speed 400 rpm and 1 Nm load torque. The chattering free torque and flux dynamics are illustrated in Fig. 5 when the torque command reverses between ±1 Nm. The corresponding trajectories of estimated rotor speed, flux and phase current are also shown in Fig. 5. The torque and flux are smooth enough, and the stator flux locus is almost like a circle. Fig. 6 shows the experimental results of Efficiency-optimized drive performance under the search algorithm. The motor is started under a low load condition in response to a medium step speed command as shown in Fig. 6(a). The original stator flux command is set to a maximum value to provide fast dynamics. At about \( t = 1.0 \) second, a steady state speed is detected. Then, a direction test determines a decreasing direction for \( \lambda^* \). Subsequently, the adjustment of \( \lambda^* \) is started towards its optimal value as shown in Fig. 6(b). After only about one second, the motor input power reaches its minimum value as shown in Fig. 6(c). This is about a 50% reduction in the input power. Then, a triangular mode of operation is initiated to test the robustness. By applying a speed command change at \( t = 7s \) promptly, \( \lambda^* \) returns to its original value, \( \lambda^{\text{Max}} \), and a new steady state speed is reached after a desirable transient period. The search algorithm becomes active again, and a minimum input power is obtained at the new operating point.
Fig. 3. Hardware implementation of the proposed controller, a) Laboratory implementation block diagram, b) Experimental setup.

Fig. 4: Experimental results of the steady-state performance of the proposed VS-DTC when the motor is running at 1400 Rpm with 1Nm Load, a) Rotor speed, b) The reference and estimated motor torque, c) Stator flux value
Fig. 5: Experimental results of SynRM Drive System under Proposed BELBIC-DTC, when Torque Reverses, a) Electromagnetic torque, b) Rotor speed, c) Phase current, d) Stator flux, in synchronous-frame, f) Stator flux $D_s$-$Q_s$ locus.
Fig. 6: Experimental results of Efficiency-optimized drive performance under VS-DTC. a) Speed response, b) Reference and estimated stator fluxes, c) Motor input power.

9. CONCLUSION

In this paper, a novel direct torque and flux control strategy based on variable-structure control and space vector pulse width modulation was proposed for synchronous reluctance motor drives considering iron losses. Combining the principles of the Sliding Mode, DTC, and space-vector modulation yields a simple but robust drive system. In particular, the sliding mode control contributes to robustness of the drive, the DTC results in a fast dynamic response, and the SVM improves the torque, flux and current steady-state waveforms by ripple reduction. In addition, based on the proposed control approach, for a given load torque and a given rotor speed, the optimized-efficiency related to SynRM has been adapted by tuning the amplitude of stator flux. Finally, the effectiveness of the proposed method has been verified by the simulation and experimental results.

APPENDIX A

Specifications and Parameters of three-phase SynRM

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
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REFERENCES


