Power control of a doubly fed induction machine via output feedback

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Abstract

A new output feedback control algorithm for a doubly fed induction machine (DFIM) is presented. The asymptotic regulation of active and reactive power is achieved by means of direct closed-loop control of active and reactive components of the stator current vector, presented in a line-voltage-oriented reference frame. To get the maximum generality of the solution, the usual assumption of negligible stator resistance is not made. A full-order DFIM model is used for the control algorithm development. The proposed control system is robust with respect to bounded machine parameter variations and errors on rotor position measurement. In the paper, it is also shown how the proposed current control algorithm can be modified in order to achieve asymptotic active current tracking and zero reactive current stabilization during steady state. An extension for the speed control objective and output EMF control during the excitation–synchronization stage are also presented. Simulation and experimental tests demonstrate high dynamic performance and robustness of the control algorithm for typical operating conditions. The proposed controller is suitable for both energy generation and electrical drive application with restricted speed variation range.

Keywords: Electric machines; Inverter drives; Electric energy generation; Speed control; Output feedback; Lyapunov-based control

1. Introduction

A vector-controlled doubly fed induction machine (DFIM) is an attractive solution for high-performance restricted speed-range electric drives and energy generation applications (Leonhard, 1995). In Fig. 1, the typical connection scheme of this machine is reported. The stator windings are directly connected to the line grid, while the rotor windings are supplied by a bi-directional power converter. This solution is suitable for all of the applications where limited speed variations around the synchronous speed are present. Since the power handled by the rotor side (slip power) is proportional to the slip, the energy conversion requires a rotor-side power converter which handles only a small fraction of the overall system power. Moreover, when the DFIM is used as a variable-speed drive, the slip power is regenerated during motor operating conditions by the rotor-side converter to the line grid, resulting in highly efficient energy conversion. Electric energy generation systems operating at variable speed have several advantages when compared with fixed-speed synchronous and induction generation. In generation systems driven by a diesel engine, the variable-speed operation depending on the generated power allows for a reduction of fuel consumption. In hydroelectric generation systems it increases the energy efficiency up to 10%. In wind energy generation systems the adjustment of the shaft speed as a function of the wind speed permits a higher energy capture by maximizing the turbine efficiency. Reduction of the torque ripple in the drive train due to torsional mode resonance can be
An important feature of the vector-controlled DFIM reported in Leonhard (1995) and Vas (1990) is the possibility to achieve decoupled control of the stator-side active and reactive power in both motor and generator applications. Moreover, if a suitably controlled AC/AC converter is used to supply the rotor side, the power components of the overall system can be controlled with low-current harmonic distortion in the stator and rotor sides.

The fundamentals of DFIM vector control are presented in Leonhard (1995). Different strategies were proposed to solve the DFIM control problem. The most important results are reported in Leonhard (1995), Pena, Clare, and Asher (1996a) and Hopfensperger, Atkinson, and Lakin (2000). All of them are based on the classical concept of field orientation (stator or air-gap flux) used as a torque–flux decoupling technique for induction motor control. Since in DFIM both stator and rotor currents are available from measurement, the flux vectors (stator, air-gap or rotor) can be computed using flux–current correlation equations. Consequently, the DFIM control problem is typically classified as a nonlinear state-feedback problem.

Under the assumption of rotor current-fed DFIM and negligible stator resistance, the torque and stator-side reactive power control problem is transferred to rotor current control if the rotor currents are defined in a flux-oriented reference frame. Torque (active power) or speed control objective, together with the stator-side reactive power regulation (stabilization) are typically considered.

The structure of a standard DFIM controller includes two-axis high-gain rotor current control loops with PI current controllers, implemented in a flux-oriented reference frame. Two rotor current references are used as scaled references for torque and reactive power. The solutions based on direct stator flux orientation, reported in Leonhard (1995), Yamamoto and Motoyoshi (1992), Pena et al. (1996a), Hopfensperger et al. (2000), Walczyńska (1991), rely on some simplifying assumptions. In particular, stator resistance is usually considered negligible. This hypothesis, which is typical for high-power DFIMs, leads to neglect also the stator flux poorly damped dynamics in the controller design since the stator flux vector is always assumed constant in quadrature with the line-voltage vector. As far as the authors know, no analytically proven full-order control algorithms based on the stator flux field orientation are available in literature. State-feedback linearization has been applied in Bogalecka and Kzreminski (1993) to solve the DFIM control problem. The assumption of current-fed rotor is used with an additional first-order filter in the control loop. Rotor position sensorless solutions have been considered in Xu and Cheng (1995), Hopfensperger et al. (2000) and Bogalecka (1993). The operation of a vector-controlled DFIM supplying an isolated load is reported in Pena, Clare, and Asher (1996b). The classical approach for DFIM vector control (Leonhard, 1995) requires measurements of stator, rotor currents and rotor position. In order to achieve synchronization with the line-voltage vector for soft connection to the line grid, the information about line voltages is also needed. Exact knowledge of induction machine inductances (including the saturation effect) is required to compute fluxes from currents. The necessity for high-precision rotor position measurement has been addressed in Xu and Cheng (1995). In Pena et al. (1996a, b), the authors use the integration of stator voltage equations in order to estimate stator fluxes. This solution requires particular adjustments to avoid open-loop-integration drift due to variations of stator resistance and measurements offset. However, it must be underlined that the compensation of such effect for DFIMs is less difficult than for typical induction motor drives. In fact, the stator flux components in a fixed a–b reference frame are sinusoidal with a frequency equal to that of the line grid and the stator resistances are very small in large DFIMs, usually adopted in industrial practice. Different approaches for the implementation of the stator flux-oriented reference frame are discussed in Hopfensperger et al. (2000).

In Peresada, Tilli, and Tonelli (1998), an alternative approach for the design of DFIM active–reactive power control is proposed. The controller development is based on implementation of a line-voltage vector-oriented reference frame. Since the line-voltage vector can easily be measured with negligible errors, this reference frame is DFIM parameter-independent in contrast to the field-oriented one. Moreover, information about line voltage is typically needed in order to perform the soft connection of the DFIM to the line grid during the preliminary excitation–synchronization stage. This full-order control algorithm ensures globally asymptotically stable torque tracking and stator-side unity-power-factor. It is demonstrated that conditions of stator flux field orientation and line-voltage vector orientation are equivalent if the stator-side power factor
is controlled at unity level. In Peresada, Tilli, and Tonielli (1999a), the approach of Peresada et al. (1998) is extended to the control problem of speed tracking and stator-side power factor stabilization of rotor current-fed DFIMs. A rotor speed/position sensor is used in the algorithms Peresada et al. (1998, 1999a), but no particular accuracy is required. The above control algorithms utilize the concept of indirect flux regulation (similarly to indirect field-oriented control of squirrel-cage induction motors).

Both direct and indirect stator flux field-oriented solutions are open loop with respect to the output variables, i.e. torque (active power) and reactive power. Robustness of the above control algorithms with respect to parameter variation and rotor position measurement are based on the natural stability properties of the DFIM electromagnetic subsystem. In order to improve robustness with respect to parameter variations and errors in rotor position measurements, the outer stator-side reactive power and active power loops are added in many publications. Nevertheless, no stability analysis is given for such solutions. In Peresada, Tilli, and Tonielli (1999b), a new full-order nonlinear control algorithm is proposed. It is shown that direct closed-loop control of active and reactive power guarantees global asymptotic regulation of output variables and internal stability, under the condition of measurable stator currents and voltages, as well as rotor position and speed. The concepts of cascaded architecture and field-orientation are not used to develop the controller.

The aim of this paper is the generalization of the preliminary result given in Peresada et al. (1999b) for the case of torque and speed control. Rigorous stability analysis of the electromagnetic dynamics is given. Relying on this result, a new dynamic second-order speed controller with load compensation has been developed. This solution guarantees stator-side reactive power regulation and asymptotic speed regulation under the condition of constant load torque provided that DFIM physical limits are satisfied. In addition, a new closed-loop synchronization–excitation control algorithm is proposed in the paper which guarantees transient free connection of the DFIM to the line grid. The relative-degree equal to one of the DFIM between the stator currents (outputs) and rotor voltages (inputs) is exploited in order to achieve robustness properties with respect to parameter perturbations and particular attention is paid to the residual internal dynamics. No information on rotor currents is required to achieve the control objectives hence, in industrial plants, only a rough measurement of these variables is necessary for protection purposes. It is worth observing that the absence of rotor current feedback does not imply that these variables are ‘uncontrolled’, in fact, as enlightened in the Lyapunov-based controller development, global asymptotic stabilization of all state variables is guaranteed.

The proposed nonlinear output feedback controller demonstrates strong robustness properties with respect to stator and rotor resistances/inductances variation and to rotor position measurement errors. In addition, owing to the closed-loop structure with true-stator-current feedback, the controller compensates for non-idealities of the electric machine magnetic structure, delivering improved stator current waveforms. The controller is suitable both for drive application and electric energy generation (e.g. in alternative energy plants) including operation as an autonomous generator during the excitation–synchronization stage.

The paper is organized as follows. In Section 2, the DFIM model and the control problem statement are presented. In particular, the selection of the line-voltage-vector-oriented reference frame is deeply discussed, recalling also the active/reactive power control objective considered in Peresada et al. (1999b). Stator-side active and reactive current controllers are designed in Section 3. The extension of the active current control algorithm for speed control objective is given in Section 4. The results of simulation and experimental tests are presented in Section 5. As underlined in Section 3, a variant of the proposed stator current controller, which guarantees stator active current tracking and stabilization to zero of the stator reactive current during steady state, is reported in Appendix A; while, in Appendix B, the above-mentioned excitation–synchronization algorithm, which replicates the current controller structure of Section 3, is presented.

2. DFIM model and control objectives

Under the assumption of linear magnetic circuits and balanced operating conditions, the equivalent two-phase model of the symmetrical DFIM, represented in an arbitrary rotating \((d-q)\) reference frame is

\[
\dot{e} = \omega, \\
\dot{\omega} = \frac{1}{J} [T_q - T], \quad T_q = \mu (\psi_d i_q - \psi_q i_d),
\]

\[
i_d = -\gamma i_d + \omega_0 i_q + \alpha \beta \psi_d + \beta \omega \psi_q + \frac{1}{\sigma} u_d - \beta u_{2d},
\]

\[
i_q = -\omega_0 i_d - \gamma i_q - \beta \omega \psi_d + \alpha \beta \psi_q + \frac{1}{\sigma} u_q - \beta u_{2q},
\]

\[
\dot{\psi}_d = -\omega \psi_d + (\omega_0 - \omega) \psi_q + \alpha L_m i_d + u_{2d},
\]

\[
\dot{\psi}_q = -(\omega_0 - \omega) \psi_d - \omega \psi_q + \alpha L_m i_q + u_{2q},
\]

\[
i_{0d} = \omega_0,
\]

where \(i_d, i_q, \psi_d, \psi_q\) are the components of the stator current and rotor flux vectors, \(u_{2d}, u_{2q}\) are the components of the rotor voltage vector, while \(u_d, u_q\) are the components of the line voltage vector; \(\omega, J, \mu, \alpha, \beta, \sigma,\) and \(L_m\) are the electrical angular speed, inductance, and resistance of the DFIM, the ratio of the stator to the rotor inductances and electrical angular frequency of the line, the ratio of the stator to the rotor resistances, and the mutual inductance of the DFIM.
represent the line-voltage components (stator windings are directly connected to the line grid); \( e \) and \( \omega \) are rotor angular position and speed; \( T \) is the external torque applied to the mechanical system of the DFIM; \( T_g \) is the torque produced by the electrical machine; \( J \) is the total rotor inertia; \( \omega_v \) and \( \omega_h \) are angular position and speed of the \((d-g)\) reference frame with respect to the \(a\)-axis of the fixed stator reference frame \((a-b)\). Variables expressed in the \((d-g)\) reference frame are given by:

\[
\begin{align*}
   x_{1dq} &= e^{-j\omega_v t} x_{1ab} \\
   x_{2dq} &= e^{-j(\omega_h - \omega_v) t} x_{2av}
\end{align*}
\]

where \( x_{pq} \) stands for two-dimensional vectors in the generic \((p-q)\) reference frame; subscript ‘1’ indicates stator variables while subscript ‘2’ indicates rotor variables; \((u-v)\) indicates the rotor reference frame and \( \epsilon \) is the rotor angle (i.e. the angle between the \( u \)- and the \( a \)-axis).

Positive constants in (1), related to electrical and mechanical parameters of the DFIM, are defined as follows:

\[
\begin{align*}
   \sigma &= L_1 \left( 1 - \frac{L_m^2}{L_1 L_2} \right), \\
   \beta &= \frac{L_m}{\sigma L_2}, \\
   \mu &= \frac{3 L_m}{2 L_2}, \\
   \alpha &= \frac{R_2}{L_2}, \\
   \gamma &= \left( \frac{R_1}{\sigma} + \alpha L_m \beta \right)
\end{align*}
\]

where \( R_1, R_2, L_1, L_2 \) are stator/rotor resistances and inductances, respectively, \( L_m \) is magnetizing inductance. One pole pair is assumed without loss of generality.

Depending on the DFIM application, \( T \) in (1) has different meanings.

- When the DFIM is used as an electric generator, \( T \) is the torque produced by a controlled primary mover. Torque \( T_g \) produced by the DFIM is a perturbation for speed control system of the primary mover. Without loss of generality, it is assumed that the mechanical dynamics of the DFIM is properly stabilized by the primary mover speed controller.

- When the DFIM is used as a motor, \( T \) is the external load torque. Usually, in this condition, a speed control loop, acting on the torque \( T_g \), generated by the DFIM, is present. In wind power plants, the speed control objective can be formulated in order to adjust the turbine speed as a function of the wind speed.

The main control objective considered is the regulation of DFIM stator-side active and reactive powers (i.e. the active and reactive power exchanged between the line grid and the stator port of the DFIM). The stator active power control objective is significant for energy generation applications since, given a prime mover able to produce a given power at a given speed, it is possible to compute (Leonhard, 1995) the stator power which must be imposed to have a total DFIM active power equal to that available from the prime mover (keeping in mind the overall efficiency) at the considered speed. Alternatively, the control of stator reactive power is relevant since it can coincide with the control of the total reactive power delivered by the DFIM system if a vector-controlled rectifier is adopted at rotor side imposing the exchange of active power only with the line grid.

Active and reactive power at stator side are given by

\[
\begin{align*}
   P_a &= \frac{3}{2} (u_id_i + u_q i_q), \\
   P_r &= \frac{3}{2} (u_q i_d - u_i i_q).
\end{align*}
\]

The first step in the mathematical formulation of the control problem is the selection of the reference frame for control algorithm development. It is well known that vector control of an induction machine with the same specifications can be designed in any reference frame. The criteria adopted in this case for reference frame selection are:

- reliability of realization under the condition of plant parameter variation and measurement errors;
- simplicity and approximation avoidance in the formulation of control objectives.

Many approaches to DFIM control, (Leonhard, 1995), use the field-orientation concept based on the availability of stator and rotor current measurements. Stator, rotor or air-gap flux can be computed directly from current vectors to organize the flux-oriented reference frame. Nevertheless, one of the two current vectors should be transformed to rotor or stator reference frame using the rotor position angle in the coordinate transformation matrix (2). This requires an accurate position measurement (resolution of standard sensors is usually satisfactory, but even small misalignment could be very harmful). Moreover, stator-side power factor control implies variations of the flux modulus which produces variations of the magnetizing inductance (involved in flux computation) due to the saturation effect.

In order to reduce the effect of the above inaccuracies in the reference frame generation and in vector transformations, a line (stator) voltage vector reference frame \((d-g)\) has been adopted (the \(d\)-axis is aligned with the line-voltage vector). This reference frame is independent of machine parameters and position measurement accuracy. The space location of DFIM vectors in the line-voltage-vector-oriented reference frame is shown in Fig. 2.
Using the line-voltage vector reference frame, a simple and smooth connection of the stator windings to the line grid can be performed during the start-up procedure of the DFIM-based system.

Measured line (stator) voltages in two-phase presentation are equal to
\[ u_a = U \cos(\omega_0 t + \varphi_0), \quad u_b = U \sin(\omega_0 t + \varphi_0). \]
where \( U \) and \( \omega_0 \) are the line-voltage amplitude and the angular frequency, \( \varphi_0 \) is the initial angular position of the line-voltage vector.

The synchronous stator voltage-oriented reference frame is defined by setting in (1) and (2)
\[ \cos(\omega_0 t) = u_a / U, \quad \sin(\omega_0 t) = u_b / U, \quad \omega_0 = \omega_1. \]

Under this transformation \( u_d = U \) and \( u_q = 0 \) in the DFIM model (1). In addition, currents \( i_d \) and \( i_q \), in the line-voltage-oriented reference frame (see Fig. 2) represent the active and reactive components of the stator current vector. The expressions of active and reactive powers (3) can be presented as
\[ P_a = \frac{2}{3} U i_d, \quad P_r = -\frac{2}{3} U i_q. \]

For convenience, a positive power flow is defined in (6) when its direction is from the power source to the electric machine. Conditions \( i_q < 0 \) and \( i_q > 0 \) indicate a lagging and leading power factor, respectively; \( i_q = 0 \) is the condition of unity power factor. From (6), it follows that active–reactive power control objective is equivalent to active–reactive stator currents control. Let \( P^*_a \) and \( P^*_r \) be the references for the power components at stator side for the DFIM. Using (6), references for the components of the stator current, in the voltage-oriented reference frame, are given by
\[ i^*_d = \frac{2}{3} P^*_a / U, \quad i^*_q = -\frac{2}{3} P^*_r / U. \]

The control problem of the DFIM generator is formulated in terms of stator active–reactive current regulation as follows.

**Proposition.** Consider the DFIM model (1) under coordinate transformation (2), (5). Let us assume that:

A.1. The stator voltage amplitude and frequency are constants (stator windings are directly connected to the line grid).

A.2. References for active and reactive stator currents are constant and bounded, or represent ramp signals with bounded first derivative and bounded amplitude.

A.3. Under the assumption of a properly controlled primary mover, the rotor speed is time varying, measurable and bounded together with its first time derivative.

A.4. Stator currents and voltages as well as rotor position and speed are available from measurements.

Under these conditions a dynamic output feedback controller exists in the form
\[ \begin{pmatrix} u_{2d} \\ u_{2q} \end{pmatrix} = f \left( i^*_d, i^*_q, i^*_d, i^*_q, i_d, i_q, U, \omega_0, \omega, z \right), \]
\[ \dot{z} = \varphi \left( \omega_0, i^*_d, i^*_q, i_d, i_q \right), \]
which guarantees:

O.1. Asymptotic active–reactive stator current regulation independent of speed variations; i.e.
\[ \lim_{t \to \infty} (i_d) = 0, \quad \lim_{t \to \infty} (i_q) = 0, \]

O.2. Boundness of all internal signals and synchronization of the control actions with the stator voltage vector; and which has the following qualitative property:

O.3. Good robustness with respect to stator and rotor resistance variations and constant error in rotor position measurement due to position sensor misalignment.

The proof of O.1 and O.2 of the proposition is given by the controller design and the stability analysis presented in the next section. The validity of O.3 is based on the general stability arguments for the closed-loop systems having proportional-integral controllers and confirmed by simulation and experimental tests reported in Section 5. In particular, the choice of a line-voltage-oriented reference frame (as discussed before) and the adopted control algorithm structure (as discussed in the following paragraphs) play a key role in achieving good robustness properties.

In Section 4, it is shown how the stator active–reactive current control can be fitted to the speed control of the DFIM.

### 3. Design of the output feedback control algorithm for the DFIM

Before starting the design procedure let us make the following remark: the objective of active–reactive
current control must be achieved without violations of the physical constraints on the DFIM operation. These constraints require that machine fluxes are bounded and rotate synchronously with the stator voltage vector during steady state. Let us consider stator flux equations:

\[
\begin{align*}
\psi_{1d} &= L_{1d}i_d + L_mL_2^{-1}(\psi_d - L_mi_d), \\
\psi_{1q} &= L_{1q}i_q + L_mL_2^{-1}(\psi_q - L_mi_q). 
\end{align*}
\]

(10)

The stator flux dynamics derived from (1) with \(u_{1d} = U, u_{1q} = 0\) is equal to

\[
\begin{align*}
\dot{\psi}_{1d} &= \omega_0\psi_{1q} - R_1i_d + U, \\
\dot{\psi}_{1q} &= -\omega_0\psi_{1d} - R_1i_q. 
\end{align*}
\]

(11)

Achieving the condition of ideal stator current tracking, i.e. \(i_d \equiv \dot{i}_d, i_q \equiv \dot{i}_q\), makes the dynamics of the stator fluxes unobservable by the controller. This dynamics has oscillatory behavior and it is not suitable for practical application. Even if in a real system the perfect tracking condition is not achievable due to the limited power of the rotor converter and line impedance, problems would be encountered if a feedback linearizing strategy were applied. From this consideration, it is possible to conclude that the stator active–reactive current error dynamics cannot be independent of flux dynamics, restricting the controller development on the level of the regulation problem.

The design procedure is performed in two steps: a flux control is designed first, to achieve flux regulation, then the current control algorithm is developed.

Let us define the flux regulation errors as

\[
\begin{align*}
\psi_d &= \psi_d - \psi_d^*, \\
\psi_q &= \psi_q - \psi_q^*,
\end{align*}
\]

(12)

where flux references, \(\psi_d^*\) and \(\psi_q^*\), will be defined later according to stator current control objectives.

Using definition (12), the last two equations of the DFIM model (1) can be rewritten in ‘error form’ as

\[
\begin{align*}
\dot{\psi}_d &= -\gamma (\psi_d + \psi_d^*) + \omega_2 (\psi_q + \psi_q^*) \\
&\quad + \alpha L_m(\dot{i}_d + \dot{i}_d^*) + u_{2d} - \dot{\psi}_d^*, \\
\dot{\psi}_q &= -\gamma (\psi_q + \psi_q^*) - \omega_2 (\psi_d + \psi_d^*) \\
&\quad + \alpha L_m(\dot{i}_q + \dot{i}_q^*) + u_{2q} - \dot{\psi}_q^*,
\end{align*}
\]

(13)

where \(\omega_2 = \omega_0 - \omega\) is the slip angular frequency.

Constructing the flux control algorithm as

\[
\begin{align*}
u_{2d} &= \omega\psi_d^* - \omega_2\psi_q^* - \alpha L_m\dot{i}_d^* + \psi_d^* + v_d, \\
u_{2q} &= \omega\psi_q^* + \omega_2\psi_d^* - \alpha L_m\dot{i}_q^* + \psi_q^* + v_q,
\end{align*}
\]

(14)

the flux error dynamics becomes

\[
\begin{align*}
\dot{\psi}_d &= -\alpha\dot{\psi}_d + \omega_2\dot{\psi}_q + \alpha L_m\dot{i}_d + v_d, \\
\dot{\psi}_q &= -\alpha\dot{\psi}_q - \omega_2\dot{\psi}_d + \alpha L_m\dot{i}_q + v_q,
\end{align*}
\]

(15)

where \(v_d, v_q\) will be defined later.

Applying the control algorithm (14), the current error dynamics from the first two equations of (1) can be rewritten as

\[
\begin{align*}
\dot{i}_d &= -\gamma\dot{i}_d + \omega_0\dot{i}_q + \omega_2\dot{\psi}_d + \alpha L_m\dot{i}_d + \psi_d^* + \frac{1}{\sigma}U + \beta\omega_0\psi_q^* + v_d, \\
\dot{i}_q &= -\gamma\dot{i}_q - \omega_0\dot{i}_d + \omega_2\dot{\psi}_q + \alpha L_m\dot{i}_q + \psi_q^* - \frac{1}{\sigma}\dot{\psi}_d^* - \beta\omega_0\psi_d^*. 
\end{align*}
\]

(16)

From Eq. (16) it follows that a reasonable choice of the flux references dynamics is given by the following differential equations:

\[
\begin{align*}
\dot{i}_d^* &= \frac{1}{\beta} \left( \beta \omega_0\psi_d^* + \frac{1}{\sigma}U - \frac{\alpha L_m\dot{i}_d^*}{\sigma} + \omega_0\dot{i}_q^* - \dot{i}_q^* \right), \\
\dot{i}_q^* &= \frac{1}{\beta} \left( \beta \omega_0\psi_q^* - \frac{\alpha L_m\dot{i}_q^*}{\sigma} - \omega_0\dot{i}_d^* - \dot{i}_d^* \right). 
\end{align*}
\]

(17)

Substituting (17) into (16) the resulting current–flux error dynamics becomes

\[
\begin{align*}
\dot{\psi}_d &= -\alpha\dot{\psi}_d + \omega_2\dot{\psi}_q + \alpha L_m\dot{i}_d + v_d, \\
\dot{\psi}_q &= -\alpha\dot{\psi}_q - \omega_2\dot{\psi}_d + \alpha L_m\dot{i}_q + v_q.
\end{align*}
\]

(18)

According to (17), the flux references are given by a linear time-invariant differential equation which is not really implementable, since it is stable but not asymptotically (i.e. it is a non-autonomous harmonic oscillator).

A particular solution of (17), where oscillating terms are avoided by means of a suitable selection of the initial condition, is given by

\[
\left[
\begin{array}{c}
\psi_d^* \\
\psi_q^*
\end{array}
\right] = -\frac{1}{\sigma\beta} \left[
\begin{array}{c}
1 \\
0
\end{array}
\right] \left[
\begin{array}{c}
U \\
0
\end{array}
\right] + \left[
\begin{array}{c}
\sigma I - \frac{\alpha L_m}{\sigma} \\
0
\end{array}
\right] \left[
\begin{array}{c}
\dot{i}_d^* \\
\dot{i}_q^*
\end{array}
\right] - R_1 \sum_{k=1}^{\infty} \left[ \left( \frac{1}{\sigma\omega_0} \right)^{k+1} \frac{d^k}{dk} \left[ \dot{i}_d^* \right] \right].
\]

(19)

From (19), it follows that for arbitrary trajectories of current reference all of the time derivatives together with their initial conditions should be known. The following development is based on the assumption that both
current reference signals have bounded first time-
derivative with all of the higher-order ones equal to zero. Then the flux references are
\[
\psi_d^* = \frac{1}{\beta \omega_0} \left( \frac{R_1}{\sigma} i_d^* - \omega_0 i_q^* - \frac{1}{\sigma} U - \frac{R_1}{\omega_0} i_q^* \right),
\]
\[
\psi_q^* = \frac{1}{\beta \omega_0} \left( \frac{R_1}{\sigma} i_q^* - \omega_0 i_d^* - \frac{R_1}{\omega_0} i_d^* \right)
\]
where \( \tilde{q}^* \equiv \tilde{q}^* \equiv 0 \), according to assumption A.2 on current references.

To simplify stability analysis, it is worth performing the following linear time-invariant coordinate transformation
\[
\begin{bmatrix}
\hat{i}_d \\
\hat{i}_q \\
\hat{z}_d \\
\hat{z}_q
\end{bmatrix} =
\begin{bmatrix}
1 & 0 & 0 & 0 \\
0 & 1 & 0 & 0 \\
1 & 0 & \beta & 0 \\
0 & 1 & 0 & \beta
\end{bmatrix}
\begin{bmatrix}
i_d \\
i_q \\
\psi_d \\
\psi_q
\end{bmatrix}.
\]

In the new coordinates, system (18) becomes
\[
\begin{align*}
\dot{\hat{i}}_d &= -(\gamma + \sigma) \dot{\hat{i}}_d + \omega_0 \hat{i}_q + \alpha \hat{z}_d + \omega_0 z_q - \beta vd,
\dot{\hat{i}}_q &= -(\gamma + \sigma) \dot{\hat{i}}_q - \omega_0 \hat{i}_d + \alpha \hat{z}_q - \omega_0 z_d - \beta vq,
\dot{\hat{z}}_d &= \omega_0 z_q - \frac{R_1}{\sigma} \hat{i}_d,
\dot{\hat{z}}_q &= -\omega_0 \hat{z}_d - \frac{R_1}{\sigma} \hat{i}_q,
\end{align*}
\]
(21)

Setting \( v_d = v_q = 0 \) in (22) an open-loop control algorithm is obtained (no current measurement is needed), based on the natural passivity properties of the DFIM. To check stability of the equilibrium point \( \mathbf{x}_1 = (\hat{i}_d, \hat{i}_q, \hat{z}_d, \hat{z}_q)^T = 0 \) consider the following Lyapunov function
\[
V_1 = \frac{1}{2} \mathbf{x}_1^T \mathbf{P}_1 \mathbf{x}_1
\]
(23)
with \( \mathbf{P}_1 = \mathbf{P}_1^T > 0 \) equal to
\[
\mathbf{P}_1 =
\begin{bmatrix}
1 & 0 & -1 & 0 \\
0 & 1 & 0 & -1 \\
-1 & 0 & \gamma_1 & 0 \\
0 & 1 & 0 & \gamma_1
\end{bmatrix},
\]
(24)
where \( \gamma_1 > 1 \) in order to guarantee that \( \mathbf{P}_1 > 0 \).

Selecting \( \gamma_1 = 1 + \sigma/R_i(2x + \alpha L_m \beta) \), which verifies the previous inequality owing to positiveness of the DFIM parameters, the derivative of \( V_1 \) along the system trajectories is equal to
\[
\dot{V}_1 = -\alpha \left( 1 + L_m \beta \right) (\hat{i}_d^2 + \hat{i}_q^2) + (\hat{z}_d^2 + \hat{z}_q^2)
\]
\[
< -\alpha \| \mathbf{x}_1 \|_2^2
\]
(25)
where \( \| \cdot \|_2 \) indicates the Euclidean norm.

From (23) and (25), according to standard Lyapunov stability arguments, it can be concluded that the equilibrium point \( \mathbf{x}_1 = 0 \) is globally exponentially stable.

By defining the control signals \( v_d \) and \( v_q \) in (22) as
\[
v_d = \frac{1}{\beta} k_1 \tilde{i}_d, \quad v_q = \frac{1}{\beta} k_1 \tilde{i}_q,
\]
(26)
where \( k_i > 0 \) is the proportional gain of the current controllers, a closed-loop proportional current control is obtained. Previous stability analysis, based on (23) and (25) is still valid with\[
\gamma_2 = 1 + \frac{\alpha}{R_i} (2x + \alpha L_m \beta + k_i) > 1,
\]
\[
\dot{V}_1 = -\alpha \left( 1 + L_m \beta \right) (\hat{i}_d^2 + \hat{i}_q^2) + (\hat{z}_d^2 + \hat{z}_q^2)
\]
\[
< -\alpha \| \mathbf{x}_1 \|_2^2
\]
(27)

The proposed nonlinear active–reactive current controller, given by (14), (20) and (26) has a proportional action based on reliably measured current feedback signals. Since the linear time-varying dynamics of the DFIM (22) has relative degree equal to one and it is exponentially stable for any \( k_i > 0 \), the robustness properties of current regulation can be improved by increasing the current controller gain \( k_i \).

In order to compensate, during steady-state conditions, for constant perturbation generated by DFIM parameter variations and error in rotor position measurement, the following two-dimensional proportional-integral current controller is designed
\[
v_d = \frac{1}{\beta} (k_i \tilde{i}_d + \lambda \tilde{i}_q - y_d), \quad \dot{y}_d = -k_i \tilde{i}_d - \beta R_1 \tilde{i}_q,
\]
\[
v_q = \frac{1}{\beta} (k_i \tilde{i}_q - \lambda \tilde{i}_d - y_q), \quad \dot{y}_q = -k_i \tilde{i}_q + \beta R_1 \tilde{i}_d,
\]
(28)
where \( k_i > 0 \) is the integral gain of the current controllers and \( \lambda = k_i \omega_0 \) is the ‘cross gain’.

The resulting flux–current dynamics, with the PI controllers (28), is the following:
\[
\dot{\hat{i}}_d = -(\gamma + \sigma + k_i) \hat{i}_d + (\omega_2 - \lambda) \hat{i}_q + \alpha \hat{z}_d + \omega_0 z_q + y_d,
\]
\[
\dot{\hat{i}}_q = -(\gamma + \sigma + k_i) \hat{i}_q - (\omega_2 - \lambda) \hat{i}_d + \alpha \hat{z}_q - \omega_0 z_d + y_q,
\]
\[
\dot{\hat{z}}_d = \omega_0 z_q - \frac{R_1}{\sigma} \hat{i}_d,
\]
\[
\dot{\hat{z}}_q = -\omega_0 \hat{z}_d - \frac{R_1}{\sigma} \hat{i}_q,
\]
(29)

To investigate stability conditions of the equilibrium point \( \mathbf{x} = 0 \), where \( \mathbf{x} = (\hat{i}_d, \hat{i}_q, \hat{z}_d, \hat{z}_q, y_d, y_q)^T \), let us consider the non-negative function
\[
V = \frac{1}{2} \mathbf{x}_1^T \mathbf{P}_1 \mathbf{x}_1
\]
(30)
with \( P = P^T > 0 \) equal to

\[
P = \begin{bmatrix}
0 & \omega_0^{-1} \\
0 & -\omega_0^{-1} \\
-\omega_0^{-1} & 0 \\
\omega_0^{-1} & 0 & 0 & \gamma_2 \\
0 & 0 & \gamma_2 & 0
\end{bmatrix},
\]

where

\[
\gamma_1 = (1 + e) + \gamma_{11}, \quad e > 0, \quad \gamma_{11} > 0, \quad \gamma_2 > 0,
\]

\[
\gamma_{11}\gamma_2 > \frac{1}{\omega_0^2}
\]

to guarantee that \( P > 0 \).

The time derivative of \( V \) along the trajectories of (29) is negative semidefinite and equal to

\[
\dot{V} = \dot{V}_1 \leq 0
\]

with

\[
\gamma_1 = \left[ 1 + \frac{\sigma}{R_1} (2\alpha + \alpha L_{m0} \beta + k_i) + \frac{1}{\gamma_2 \omega_0^2} \right].
\]

Note that conditions (32) and (34) give freedom in the selection of the current controller parameters \( k_i \) and \( k_{ii} \); in fact for each \( k_i > 0 \) and \( k_{ii} > 0 \), it is possible to find \( \gamma_1 \) and \( \gamma_2 \) in order to satisfy the above conditions.

From (30) and (33), it follows that \( V \) is bounded and \( \lim_{t \to \infty} V(t) \) exists and is bounded. Moreover, since \( V \) is radially unbounded, \( \mathbf{x} \) is bounded too and, according to assumption A.3, also \( \dot{x} \) will be bounded. That means \( \dot{V} \) is bounded and then, according to Barbalat’s Lemma (see Khalil, 1994), \( \dot{V} \to 0 \) hence

\[
\lim_{t \to \infty} (\dot{\mathbf{i}}_d, \dot{\mathbf{i}}_q, \mathbf{z}_d, \mathbf{z}_q)^T = \mathbf{0}.
\]

Global current and flux regulation is achieved with bounded internal signals. In order to show that integral terms \( y_d, y_q \) converge to zero in ideal conditions (i.e. without perturbations in model (1)), let us consider the dynamics \( \dot{\mathbf{i}}_d, \dot{\mathbf{i}}_q, \mathbf{z}_d, \mathbf{z}_q \) given by the first four equations in (29). Since \( \dot{x} \) is bounded and, by assumptions A.2, A.3, \( \dot{\mathbf{i}}_d = \dot{\mathbf{i}}_q = 0 \) and \( \mathbf{z}_2 = -\dot{\omega} \) is also bounded, it follows that \( \dot{\mathbf{i}}_d, \dot{\mathbf{i}}_q, \mathbf{z}_d, \mathbf{z}_q \) are bounded. Hence, according to Barbalat’s Lemma

\[
\lim_{t \to \infty} (\dot{\mathbf{i}}_d, \dot{\mathbf{i}}_q, \mathbf{z}_d, \mathbf{z}_q)^T = \mathbf{0}.
\]

Combining (35) and (36), it follows from (29) that

\[
\lim_{t \to \infty} (y_d, y_q)^T = \mathbf{0}.
\]

The control objectives O.1 and O.2 are globally achieved according to stability analysis. The complete equations of the nonlinear current controller are given by

\[
u_{zd} = \omega_0 \dot{\psi}_d^* - \omega_2 \dot{\psi}_q^* - \omega_2 \mathbf{L}_{m} \dot{i}_d + \dot{\psi}_d^* + \frac{1}{\beta} (k_i \dot{i}_d + \lambda \dot{\gamma}_d - y_d),
\]

\[
u_{zq} = \omega_0 \dot{\psi}_q^* + \omega_2 \dot{\psi}_d^* - \omega_2 \mathbf{L}_{m} \dot{i}_q + \dot{\psi}_q^* + \frac{1}{\beta} (k_i \dot{i}_q - \lambda \dot{\gamma}_d - y_q),
\]

\[
\psi^*_d = \frac{1}{\beta \omega_0} \left( \frac{R_1}{\sigma} \dot{i}_d - \omega_0 \dot{i}_q^* \right) - \frac{1}{\sigma} U - \frac{R_1}{\sigma \omega_0} \dot{i}_q^* + \frac{1}{\beta \omega_0} \left( \frac{R_1}{\sigma} \dot{i}_q - \omega_0 \dot{i}_d^* \right) - \frac{1}{\sigma} U - \frac{R_1}{\sigma \omega_0} \dot{i}_d^* - \frac{1}{\sigma} U
\]

\[
\psi^*_q = \frac{1}{\beta \omega_0} \left( \frac{R_1}{\sigma} \dot{i}_d - \omega_0 \dot{i}_q^* \right) - \frac{1}{\sigma} U - \frac{R_1}{\sigma \omega_0} \dot{i}_q^* + \frac{1}{\beta \omega_0} \left( \frac{R_1}{\sigma} \dot{i}_q - \omega_0 \dot{i}_d^* \right) - \frac{1}{\sigma} U
\]

The block diagram of the controller (38) is shown in Fig. 3. In Appendix A, it is shown as to how the controller can be modified in order to achieve active current tracking and stabilization of zero reactive current during steady state. In Appendix B, a particular excitation–synchronization algorithm is presented to perform smooth connection to the line grid of a DFIM system used to generate electric energy. That algorithm replicates the current controller structure shown in the present paragraph in order to simplify the commutation between the two regulators after the connection of the generator to the line grid.

4. Design of the speed control algorithm for the DFIM

In this section, the speed regulation—load torque compensation control algorithm is designed on the basis of the inner current control loops developed in Section 3. The property of global asymptotic stability of the DFIM electrical subsystem is used for the design of the desired dynamics of the mechanical subsystem. First, the active–reactive currents control problem is transferred to the torque–reactive current control problem, then an outer speed control loop is designed with a dynamic speed controller, generating the torque reference command for the inner torque control loop.

Consider the DFIM torque Eq. (1)

\[
T_g = \mu (\dot{\psi}_d \dot{i}_q - \dot{\psi}_q \dot{i}_d) = \mu (\dot{\psi}_d \dot{\psi}_q^* - \dot{\psi}_q \dot{\psi}_d^*)
\]

\[
+ \mu (\dot{\psi}_d \dot{i}_q - \dot{\psi}_q \dot{i}_d) + \mu (\dot{\psi}_d \dot{i}_q - \dot{\psi}_q \dot{i}_d) + \mu (\dot{\psi}_d \dot{i}_q - \dot{\psi}_q \dot{i}_d) = T_g^* + T_g,
\]
where torque reference is defined as
\[ T_g^* = \mu (\psi_d^* i_q^* - \psi_q^* i_d^*). \]  
(41)

Substituting flux references given by (20), Eq. (41) becomes
\[
T_g^* = \frac{\mu}{\beta \omega_0} \left[ -\frac{R_1}{\sigma} (i_d^* + i_q^*) + \frac{U}{\sigma} i_d^* - \frac{R_1}{\sigma} \frac{1}{\omega_0} \frac{U}{\sigma} i_d^* \right] \times \left( \frac{i_d^*}{C_0} - \frac{i_q^*}{C_1} \right) = T_{g1}^* + T_{g2}^*.
\]  
(42)

where
\[
T_{g1}^* = \frac{\mu}{\beta \omega_0} \left[ -\frac{R_1}{\sigma} \left( \frac{1}{C_0} i_d^* + \frac{1}{C_1} i_q^* \right) + \frac{U}{\sigma} i_d^* \right],
\]
\[
T_{g2}^* = -\frac{\mu}{\beta \omega_0} \frac{R_1}{\sigma} \left[ \left( \frac{i_d^*}{C_0} - \frac{i_q^*}{C_1} \right) \right].
\]  
(43)

From the expression for \( T_{g1}^* \), the active current reference is given by
\[
i_d^* = \frac{U}{\sigma} - \frac{Q^1/2}{2 R_1/\sigma},
\]
\[
Q = \left[ \left( \frac{U}{\sigma} \right)^2 - 4 \frac{R_1}{\sigma} \left( \frac{R_1}{\sigma} i_d^* + \frac{T_{g1}^*}{\beta \omega_0} \right) \frac{\beta \omega_0}{\mu} \right].
\]  
(44)

Eq. (42) establishes the steady-state relation between active current reference and torque command. On inspecting (40)–(42), it can be seen that (42) reflects the power balance condition in the stator side of the DFIM, which establishes the solvability of (42) with \( Q > 0 \). Note that if \( i_d^* = 0 \) then \( T_{g2}^* = 0 \). The dynamic torque component \( T_{g2}^* \) can be further expressed using (44) as
\[
T_{g2}^* = \frac{\mu}{\beta \omega_0} \frac{R_1}{\sigma} \left( \frac{i_d^*}{C_0} - \frac{2}{Q^{1/2}} \frac{R_1}{\sigma} \right) \frac{1}{Q^{1/2}} - \frac{1}{\omega_0} \frac{1}{Q^{1/2}} \frac{R_1}{\sigma} \frac{i_d^*}{C_1},
\]  
(45)

For practical operating conditions of the DFIM with bounded \( i_d^* \), \( T_{g1}^* \), \( T_{g2}^* \) and small value of \( R_1 \), from (44), the result is: \( Q^{1/2} \approx U/\sigma \). Under this condition with \( i_d^* \) properly bounded, the dynamic torque component \( T_{g2}^* \) is small enough to be neglected.

Using definition (40) together with condition \( T_{g2}^* = 0 \), the speed dynamics of the DFIM, given by the first equation in (1) can be written as
\[
\dot{\omega} = \frac{1}{J} (T_{g1}^* + \dot{T}_g - T),
\]  
(46)

where \( T \) is the load torque.

The DFIM speed control problem under stator-side reactive power control is formulated as follows. Assume that the speed reference \( \omega^* \) is constant and it can assume values in a restricted slip speed range around motor synchronous speed \( \omega_0 \). Load torque \( T \) is assumed to be bounded unknown and constant. Reference for reactive current component \( i_q^* \) is formed as a ramp signal with a
properly bounded first-order time derivative. Under this condition, it is necessary to design the speed control algorithm forming the torque reference command \( T_{g1}^* \) which guarantees asymptotic speed regulation when \( i_q^* \) is constant, i.e.

\[
\lim_{t \to \infty} \delta \omega = 0, \quad \tilde{\omega} = \omega - \omega^*.
\] (47)

Before starting the speed controller development, it is important to note that current control algorithm (38) requires knowledge of the first-order time derivative of \( i_d^* \), which implies from (44) that \( T_{g1}^* \) should be known. To satisfy this condition, the following dynamic speed controller is defined

\[
T_{g1}^* = \xi, \quad \dot{\xi} = -\frac{1}{\tau} \xi + \frac{1}{\tau} J (\tilde{\omega} - \dot{\tilde{\omega}}),
\] (48)

\[
\dot{\tilde{\omega}} = -k_{out} \delta \omega,
\] (49)

where \((k_{out}, k_{in}) > 0\) are the proportional and integral gains of the speed PI controller with \( \dot{\tilde{\omega}} \) defined as estimation of the constant quantity \( T/J \), while \( \tau \) is the time constant of the first-order filter. Substituting (48) in (46), the speed error dynamics becomes

\[
\dot{\delta \omega} = -k_{out} \delta \omega, \quad \dot{\tilde{\omega}} = \frac{1}{\tau} \xi + \frac{1}{\tau} J (\tilde{\omega} - \dot{\tilde{\omega}}),
\] (49)

To show that the speed control objective is achieved with \( T_y = 0 \), the change of coordinates

\[
\eta = \frac{1}{J} (\xi - T)
\]

is considered. In addition, the load torque estimation error is defined as

\[
\tilde{T} = \frac{T}{J} - \tilde{T}
\]

with \( \delta \omega = 0 \), since \( T/J \) is constant. Hence system (49) becomes

\[
\dot{\tilde{\omega}} = -k_{out} \delta \omega, \quad \dot{\tilde{\omega}} = \eta + \frac{1}{\tau} \tilde{T}_g,
\]

\[
\dot{\eta} = -\frac{1}{\tau} \eta - \frac{1}{\tau} k_{out} \delta \omega - \frac{1}{\tau} \tilde{T}.
\] (50)

The dynamics of linear time-invariant system (50) can be specified by the selection of the three tuning parameters \( k_{out}, k_{in}, \tau \) in order to guarantee asymptotic stability and the desired transient performance.

The composite speed current error dynamics is given by (50), with \( \tilde{T}_g \) defined in (40), and (29) with an additional perturbation arising from the non-compensating dynamics in (20) and from the second derivative of \( i_d^* \) according to (44). This interconnection term is a small gain feedback between the two subsystems due to the scaling factor proportional to \( R_1 Q^{-1/2}/\omega_0 \) which is much smaller than 1. Hence the mechanical subsystem given
by (50) and the electrical one given by (29) are almost decoupled. This structure of the error dynamics ensures that the composite system is practically asymptotically stable, i.e.

$$\lim_{t \to \infty} (\tilde{i}_d, \tilde{i}_q, z_d, z_q, y_d, y_q, T_L, \tilde{\omega}, \eta)^T = 0. \quad (51)$$

**Remark.** If stator resistance is negligible, the proposed solution can be simplified by setting its value to zero in the control algorithm computation.

A block diagram representing the proposed speed controller is given in Fig. 4.

Fig. 6. Stator current and rotor flux references.

Fig. 7. Transient during active–reactive current regulation with no parameter errors.
5. Simulation and experimental tests

Simulation and experimental tests have been performed using a small (5 kW) wound–rotor induction machine whose rated data are reported in Appendix C. The first set of simulations is reported to demonstrate the performance during active–reactive current regulation of the DFIM generator.

1. The initial time interval 0–0.95 s is used to start the primary mover, to perform the excitation and synchronization of the DFIM with the line-voltage vector and to connect the stator windings to the line grid. The trajectory of the primary mover speed (125 rad/s in no-load condition) and the sequence of the DFIM operation during excitation–synchronization preliminary stage is reported in Fig. 5.

2. At time $t = 0.95$ s the active current reference trajectory is applied, starting from zero initial value and reaching 90% of the rated value (corresponding to 45 Nm of produced torque with reactive current equal to zero). From Fig. 5, it can be noted that the primary mover speed reduces to 115 rad/s (no integral action is adopted in the primary mover speed controller). The DFIM generator still operates super-synchronously and usually, in this condition, the rotor port delivers active power (Leonhard, 1995), nevertheless, in the proposed example, the rotor port absorbs power owing to the relevant resistive losses.

3. At time $t = 1.3$ s, the reference trajectory for reactive component of the stator current is applied.

Both current and flux references computed using (20) are shown in Figs. 6a and b, respectively. The controller gains during all of the tests are set at $k_i = 200$; $k_{ii} = 10000$. The transients, reported in Fig. 7, demonstrate the dynamic performance of the proposed controller during active–reactive current regulation. The current errors are negligibly small during the above test; the soft almost transient-free connection to the line grid is achieved with the excitation algorithm given in Appendix B.

The second set of simulation results shows the dynamic behavior of the DFIM-based electrical drive. The following operating conditions have been considered:

1. At $t = 0$ s, the unloaded wound rotor induction motor is directly connected to the line grid. An additional start up resistor equal to $4 R_{2N}$ is inserted into the rotor circuit in order to damp the current transient.

2. At $t = 1$ s, the closed-loop control from rotor side is applied with the following speed controller parameters: $k_o = 80$; $k_{oi} = 3200$; $\tau = 0.005$ s. The speed reference and the load torque profile equal to the rated value and applied at $t = 2$ s, are shown in Fig. 8, together with the enable command for the speed controller.

![Fig. 8. Sequence of operation, speed reference and load torque profile for speed control application of the DFIM.](image-url)

![Fig. 9. Stator current and rotor flux references in speed control application of the DFIM.](image-url)
3. At $t = 4\, s$, a non-zero reference for the reactive component of the stator current is imposed. The stator current references as well as the flux reference trajectories, computed according to (20), are shown in Fig. 9.

Note that the speed, current and flux references and the related errors are not significant in the interval $(0–1)\, s$, since the speed control is disabled. In Figs. 10 and 11, the behavior of mechanical and electrical variables is reported. The speed tracking error is less than $0.5\, \text{rad/s}$ after the enable transient at $t = 1\, s$. The stator current tracking error is almost negligible. Some short transients arise only when a relevant change in the reference speed derivative occurs.

It is worth observing that no theoretical constraint prevents the speed control starting from zero speed. The impossibility of controlling low speed derives from voltage limits on the rotor inverter given by the usual sizing of this device in this application (Leonhard, 1995).

In Figs. 12–14 experimental results are reported. The proposed controller has been discretized using a simple backward-derivative method and implemented on a DSP-based control board (TMS320C32) with a sampling time of 200 $\mu\text{s}$. The PWM frequency of the inverter adopted to feed the rotor side of the DFIM is 5 kHz. A single-phase diode rectifier has been used to provide the DC supply to the inverter, hence the nominal DC–link voltage was 310 V. This hardware solution prevents a bi-directional exchange of power with the line grid at rotor side; hence it is not suitable for industrial plants, but it can be used to test the proposed controller anyway. The resolution of the encoder mounted on the rotor shaft is 1024 ppr. An electric drive based on a two-pole-pair induction motor with $V/\omega$ control was used as primary mover. The reference speed was set to 125 rad/s. The aim of this test is to verify the performance of the proposed active–reactive current controller. In Fig. 12, the behavior of the electric variables is reported, the references for both stator currents and rotor fluxes are the same as in Figs. 5 and 6, except for the initial interval from 0 to 0.5 s. In fact, in the experimental tests the excitation–synchronization stage is not considered. The initial condition of the results reported in Fig. 12 is characterized by stator windings connected to the three-phase line grid and stator currents regulated to zero by the proposed controller. Some transients can be noted in the stator current errors when the reference is variable, zero errors are guaranteed in steady-state conditions. A higher level of stator current transient errors as compared to the simulation test can be observed from Figs. 12 and 7. The reason for such behavior are an encoder misalignment of 0.1 rad (due to inaccuracy in the rotor magnetic axis definition) and strong saturation of the DFIM magnetic system for required flux levels (Fig. 6b) with a variation of the mutual inductance greater than 1.5 times. However, a good stator–current tracking is achieved confirming the robustness of the proposed solution.

In Fig. 13, the real stator phase voltage and current in a fixed stator reference frame are reported when the reference for the reactive current reference is set to zero, while the active current one is equal to 10 A. The result of the proposed solution (Fig. 13b) is compared with the stator current obtained with the open-loop stator flux...
field-oriented control algorithm proposed in Peresada et al. (1998) under the same conditions (Fig. 13a). In Fig. 14, the normalized harmonic content of the current waveforms of the previous pictures is considered. A relevant reduction of the stator current distortion can be noted passing from Peresada et al. (1998) to the proposed solution. This is an important feature of the proposed closed-loop control of the stator currents. In fact, owing to the direct feedback of the stator current errors on the rotor voltages, the proposed controller acts in order to compensate for non-idealities of the induction machine electromagnetic circuit.

6. Conclusions

The new direct active–reactive power controller for the DFIM provides global asymptotic regulation in the presence of induction machine parameter variations and rotor position measurement errors. In addition, it delivers an improved stator current waveform compensating for non-idealities of the induction machine electromagnetic circuit. The simulation and experimental tests confirm the high dynamic performance and robustness of the proposed controller. Two extensions of the active–reactive power control algorithm are presented. The first one allows the mechanical speed control; while the second one is suitable to control the autonomous DFIM-based generator during the excitation–synchronization stage in order to achieve transient-free connection to the line grid. The proposed controller is suitable for both energy generation and electrical drive applications, where restricted variations of the speed around the synchronous velocity are present.

Appendix A. Variation of the proposed controller: the active current tracking and unity power factor stabilization problem

In this section, a variant of the proposed solution is presented to address a slightly different control problem, given by the following objectives: (a) active-current component tracking (i.e. \( \lim_{t \to \infty} i_d = 0 \)) and (b) stabilization at zero of reactive current during steady state (i.e. \( \lim_{t \to \infty} i_q = 0 \) if \( i_d \) is constant). The active current reference is assumed to be bounded together with its first and second time derivatives.
In order to solve this problem the following choice is adopted

\[ \psi_d^* = -\frac{1}{\beta} \dot{\psi}_d, \quad i_q^* = -\frac{1}{\omega_0} i_d'. \]  

(A.1)

Hence, redefining the control algorithm (14) as

\[ u_{2d} = 2\psi_d^* - \omega_2 \psi_q^* - \omega L_m i_d + v_d, \]

\[ \dot{i}_d = \frac{1}{L_m} (u_{2d} - \omega_2 \psi_q - \omega L_m i_d + v_d), \]

\[ \dot{i}_q = -\frac{1}{L_m} (u - \omega \psi_d - \omega L m i_d + v_d), \]
where $E_d$, $E_q$ are stator EMF components, generated by rotor excitation.

The control objective during the excitation–synchronization stage is to design rotor control voltages in such a way that the stator EMF vector is equal to the line-voltage vector, i.e.

$$\lim_{t \to \infty} E_d = -U, \quad \lim_{t \to \infty} E_q = 0. \quad (B.2)$$

Under this condition, the connection of the DFIM to line grid is transient-free. The open-loop excitation control algorithm can easily be designed using (B.1) as is shown in (Peresada et al., 1998). Otherwise, in this section, in order to provide fast and robust excitation, the result of Section 3 is adopted to design the closed-loop excitation control algorithm. For this purpose, the following filtered EMF signals $x_d$ and $x_q$ (given by a first-order two-dimensional filter) are defined:

$$\dot{x}_d = -k x_d + \omega_0 x_q + E_d,$$

$$\dot{x}_q = -k x_q - \omega_0 x_d + E_q, \quad (B.3)$$

where $\tau_0 = k^{-1}$ is time constant of the filter.

The reference for output filter variables are defined as a steady-state solution of (B.3) under condition (B.2)

$$x^*_d = -\frac{k U}{k^2 + \omega_0^2}, \quad x^*_q = \frac{\omega_0 U}{k^2 + \omega_0^2}. \quad (B.4)$$

Combining (B.1) and (B.3), the resulting dynamic subsystem has a similar structure to the model of the DFIM. Following the same conceptual line, the following definitions are derived

- **control voltages**

  $$u_{2d} = x\psi^*_d - \omega_2 \psi^*_q + v_d,$$

- **flux references**

  $$\psi^*_d = \frac{L_m}{L_2} \frac{1}{\omega_0} (-k x^*_d + \omega_0 x^*_d),$$

  $$\psi^*_q = \frac{L_m}{L_2} \frac{1}{\omega_0} (k x^*_d - \omega_0 x^*_q), \quad (B.6)$$

proportional-integral EMF controller given by (28) with $\dot{i}_d$, $\dot{i}_q$ replaced by $\dot{x}_d$, $\dot{x}_q$ where

$$\dot{x}_d = x_d - x^*_d, \quad \dot{x}_q = x_q - x^*_q. \quad (B.7)$$

Straightforward computation leads to the dynamics of the output filter variables in the form of (29).

Following the same line of stability analysis as presented in Section 3, it follows that condition (B.2) is globally achieved.

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**Appendix B. Excitation—synchronization control algorithm**

In this section, the DFIM with open stator circuits and operating as an autonomous generator is considered. This kind of working condition is interesting since it is necessary to perform transient-free connection of the stator windings to the line grid. The dynamics of the machine in the above conditions can be derived from (1) with $i_d = i_d = i_q = i_q = 0$ and it results as follows

$$E_d = \frac{L_m}{L_2} (x\psi^*_d + \omega \psi^*_q - u_{2d}),$$

$$E_q = \frac{L_m}{L_2} (x\psi^*_q - \omega \psi^*_d - u_{2q}),$$

$$\dot{x}_d = -x \psi^*_d + \omega_2 \psi^*_q + u_{2d},$$

$$\dot{x}_q = -x \psi^*_q - \omega_2 \psi^*_d + u_{2q}, \quad (B.1)$$


Appendix C. Rated data of the DFIM used for simulations and experiments

Nameplate data and parameters of the adopted DFIM are reported in Table 1.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Nominal voltage</td>
<td>$380 \text{V}_{\text{RMS}}$ (Y-connected)</td>
</tr>
<tr>
<td>Nominal power</td>
<td>5 kW</td>
</tr>
<tr>
<td>Nominal speed</td>
<td>100 rad/s</td>
</tr>
<tr>
<td>Stator resistance ($R_1$)</td>
<td>0.95$\Omega$</td>
</tr>
<tr>
<td>Rotor resistance ($R_2$)</td>
<td>1.8$\Omega$</td>
</tr>
<tr>
<td>Stator inductance ($L_1$)</td>
<td>0.094 H</td>
</tr>
<tr>
<td>Nominal frequency</td>
<td>50 Hz</td>
</tr>
<tr>
<td>Nominal torque</td>
<td>50 Nm</td>
</tr>
<tr>
<td>Pole pairs</td>
<td>3</td>
</tr>
<tr>
<td>Rotor inductance ($L_2$)</td>
<td>0.088 H</td>
</tr>
<tr>
<td>Magnetizing inductance ($M$)</td>
<td>0.082 H</td>
</tr>
<tr>
<td>Rotor inertia ($J$)</td>
<td>0.1 kgm$^{-2}$</td>
</tr>
</tbody>
</table>

Table 2

Nominal values of the electric variables of the adopted DFIM, under the conditions of zero reactive power at stator side, nominal speed and nominal torque

<table>
<thead>
<tr>
<th>Variable</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$i_d$</td>
<td>11.7 A</td>
</tr>
<tr>
<td>$\psi_d$</td>
<td>$-0.22$ Wb</td>
</tr>
<tr>
<td>$i_{2d}$</td>
<td>$-13.4$ A</td>
</tr>
<tr>
<td>$u_{2d}$</td>
<td>$-9.7$ V</td>
</tr>
<tr>
<td>$i_q$</td>
<td>0 A</td>
</tr>
<tr>
<td>$\psi_q$</td>
<td>$-1.02$ Wb</td>
</tr>
<tr>
<td>$i_{2q}$</td>
<td>$-11.6$ A</td>
</tr>
<tr>
<td>$u_{2q}$</td>
<td>$-24$ V</td>
</tr>
</tbody>
</table>

Appendix C. Rated data of the DFIM used for simulations and experiments

Nameplate data and parameters of the DFIM adopted in simulations and experiments are reported in Table 1.

Nominal values of $i_d$, $\psi_d$, $i_{2d}$, $i_{2q}$, $u_{2d}$, $u_{2q}$ for the considered DFIM working as a motor in conditions of nominal speed, nominal torque and zero reactive power imposed at stator side ($i_q = 0$) are reported in Table 2.

References


