A Direct Stator Current Controller for a Doubly-Fed Induction Machine

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Abstract

In this paper, a new control scheme is presented for the doubly-fed induction machine with specific applications to renewable energy (wind farms in particular). The proposed control algorithm offers the advantages of proven stability and remarkable simplicity. In contrast with the classical vector control method, where the doubly-fed induction machine is represented in a stator-flux oriented frame, a model with orientation of the stator voltage is adopted. This allows for a decomposition of the active and reactive powers on the stator side and their regulation on the rotor side. A main contribution of the paper is the proof that a linear PI control with a skew-symmetric matrix regulating the stator currents ensures global stability for a feedback linearized doubly-fed induction machine, the specific condition that the PI gains must satisfy is derived as a simple inequality. The proof is based on a Hurwitz test for polynomials with complex coefficients that does not seem to have had prior application in control theory. The feedback linearization stage only uses the direct measurement of the rotor and stator currents and is thus easily implementable. Furthermore, it is also shown that a simpler PI control is also stable for a large range of control gains. Finally, the control system is tested in simulations.

I. INTRODUCTION

Doubly-fed induction machines (DFIM) have become very popular, especially in the field of renewable energy as hybrid engines or high performance storage systems [1][2][3][4] and for wind turbines [5][6][7]. The attractiveness of the DFIM stems primarily from its ability to handle large speed variations around the synchronous speed. Another advantage is that the power electronic equipment that controls the machine only has to handle a fraction of the total power, reducing losses (and the cost) of the power electronic converter.

In this paper, a typical connection of the DFIM is considered, as shown in Figure 1. In this case, the stator is directly connected to the power grid, while the machine is controlled through the rotor voltages. A back-to-back (B2B) converter, consisting of an AC-DC rectifier and a DC-AC inverter stage, is used for generating the rotor voltages.

For generation, the control goals for a DFIM are usually the active and reactive power delivered to the grid. For drive applications, the DFIM control is composed of an inner current control loop and an external (and considerably slower) mechanical loop (see in [8] a counter example with a unique control loop). This paper only focusses on the electrical loop, and assumes that, for driving applications, an outer mechanical loop (in terms of torque or speed, depending on the system) provides the active power reference for the electrical controller, and that the speed varies slower than the electrical variables.
Most DFIM controllers proposed in the literature are based on vector control and decoupling [9][6][10]. The methodology is derived from the description of the electrical part of the DFIM in a stator flux oriented reference frame which allows the decoupling of the active and reactive power of the stator side and their independent control through the rotor currents. To achieve the stator flux orientation, the flux angle must be estimated and several complicated operations implemented. An alternative control scheme that is widely used for the DFIM is direct torque control (DTC) [11], which uses switching tables depending on the rotor and stator fluxes, and assures a fast response of the system. However, rigorous stability proofs have not been reported in the literature for DTC. Other control schemes with rigorous stability and robustness analysis are the output feedback algorithm presented in [12], and the passivity-based controller proposed in [2].

This paper presents a control algorithm for the DFIM that is computationally simpler than other methods for which stability proofs are available. In contrast to vector control, where the DFIM is represented in a stator flux oriented frame, a model with orientation along the stator voltage vector is proposed. Assuming an infinite bus, control of the stator currents in this reference frame directly translates into the control of the active and reactive powers absorbed or delivered by the machine. The main contribution of this paper is the proof that a linear PI control of the stator currents ensures stability for a large range of PI gain values. Furthermore, global stability can be guaranteed if a feedback linearizing term is added. This feedthrough term only uses the direct measurement of the stator and rotor currents (both accessible for a DFIM), instead of the flux estimation required in the stator flux oriented methods.

The stability proofs are based on a little-known Hurwitz test for complex polynomials [13]. This method allows to significantly reduce the complexity of obtaining the stability conditions by reducing the 6th order characteristic polynomial with real coefficients to a cubic polynomial with complex coefficients. Interestingly, while a Routh-Hurwitz test for the 6th order polynomial of [14] was found to be intractable, application of its version with complex polynomials yields a simple stability test, requiring that a single quadratic inequality be satisfied by the PI gains.

The method proposed in this paper is also applicable to other control problems with certain symmetry properties. In [15], the Hurwitz test was used to find analytic conditions for spontaneous self-excitation in induction generators. The same test was applied to the algorithm presented in [12], see [16].

II. BACKGROUND: THE COMPLEX HURWITZ TEST

The extension of the well-known Routh-Hurwitz criterion to polynomials with complex coefficients is an old result of the literature [13], possibly not well-known due to the lack of relevant applications. The main result presented in that paper is...
summarized by the following Theorem.

**Theorem 1:** The polynomial

\[ P(s) = s^n + \alpha_1 s^{n-1} + \alpha_2 s^{n-2} + \ldots + \alpha_n \]  

where \( \alpha_k = a_k + jb_k \) and \( k = 1, 2, \ldots, n \), has all its zeros in the half-plane \( \Re(s) < 0 \) if and only if the determinants, \( \Delta_1 \ldots \Delta_k \),

\[
\Delta_1 = a_1
\]

\[
\Delta_2 = \left| \begin{array}{ccc}
a_1 & a_3 & -b_2 \\
a_0 & a_2 & -b_1 \\
0 & b_2 & a_1 \\
\end{array} \right|
\]

\[
\Delta_3 = \left| \begin{array}{ccc}
a_1 & a_3 & 0 & -b_2 & 0 \\
a_0 & a_2 & 0 & -b_1 & -b_3 \\
0 & a_1 & 0 & 0 & -b_2 \\
0 & b_2 & 0 & a_1 & a_3 \\
0 & b_1 & b_3 & a_0 & a_2 \\
\end{array} \right|
\]

for \( k = 2, 3, \ldots, n \) and \( a_r = b_r = 0 \) for \( r > n \), are all positive.

**Proof:** See Theorem 3.2 of [13].

Based on the previous Theorem, the particular case of a cubic polynomial with complex coefficients can be derived.

**Lemma 1:** Assuming that \( a_0 \) is real and positive, the roots of a third-order polynomial with complex coefficients

\[ P(s) = a_0 s^3 + (a_1 + j b_1) s^2 + (a_2 + j b_2) s + a_3 + j b_3 \]

are in the open left-half plane if and only if \( \Delta_1 > 0, \Delta_2 > 0 \) and \( \Delta_3 > 0 \), where

\[
\Delta_1 = a_1
\]

\[
\Delta_2 = \left| \begin{array}{cc}
a_1 & a_3 \\
a_0 & a_2 \\
0 & b_2 \\
\end{array} \right|
\]

\[
\Delta_3 = \left| \begin{array}{ccc}
a_1 & a_3 & 0 \\
a_0 & a_2 & 0 \\
0 & a_1 & 0 \\
0 & b_2 & 0 \\
0 & b_1 & b_3 \\
\end{array} \right|
\]

We will show in Sections IV and V how the complex Hurwitz test can be used to prove stability in an elegant and simpler manner.
III. MODEL OF THE DOUBLY- FED INDUCTION MACHINE

The model comes from the three phase dynamical equations of a DFIM, assuming that the machine is symmetric (all windings are identical), the stator-rotor mutual inductances are sinusoidal functions of the rotor angle [17][18], and the three phase system is balanced. These assumptions enable the use of transformations, which greatly simplify the control problem.

The basic transformation (also known as Blondel–Park transformation) is widely used in the study of electric machines [18]. This mathematical transformation is used to decouple one of the (balanced) phases, to refer all variables to a common reference frame, and to obtain state-space models whose parameters are independent of the relative angle between rotor and stator.

Similarly to [1] or [2], a transformation to a synchronous frame rotating at the constant frequency of the stator voltage of the grid is proposed. This yields the electrical equations

\[ \dot{\lambda}_s = - (\omega_s L_s J_2 + R_s I_2) i_s - \omega_s L_{sr} J_2 i_r + v_s \] (6)

\[ \dot{\lambda}_r = - (\omega - \omega) L_{sr} J_2 i_s - (\omega_s - \omega) L_r J_2 R_r I_2 i_r + v_r \] (7)

where \( \lambda_s, \lambda_r \in \mathbb{R}^2 \) are the stator and rotor fluxes, \( i_s, i_r \in \mathbb{R}^2 \) are the stator and rotor currents, \( v_s = \text{col}(v_{sd}, v_{sq}) \), is the stator voltage, \( v_r \in \mathbb{R}^2 \) is the rotor voltage and the control input, \( \omega \) is the mechanical speed, and \( \omega_s \) is the stator frequency. \( R_s, R_r \) are the stator and rotor resistances, \( L_s, L_r \) and \( L_{sr} \) are the stator and rotor self-inductances and mutual inductance, with \( L_s L_r > L_{sr}^2 \), and the matrices \( J_2 \) and \( I_2 \) are defined as

\[
J_2 = \begin{pmatrix} 0 & -1 \\ 1 & 0 \end{pmatrix}, \quad I_2 = \begin{pmatrix} 1 & 0 \\ 0 & 1 \end{pmatrix}.
\]

Linking fluxes, \( \lambda = \text{col}(\lambda_s, \lambda_r) \), and currents, \( i = \text{col}(i_s, i_r) \), are related by

\[
\lambda = L i = \begin{pmatrix} L_s I_2 & L_{sr} I_2 \\ L_{sr} I_2 & L_r I_2 \end{pmatrix} i.
\] (8)

The mechanical dynamics is given by

\[ J \dot{\omega} = \tau_e - B_r \omega - \tau_L \] (9)

where the electrical torque is

\[ \tau_e = L_{sr} i_s^\top J_2 i_r \] (10)

and \( J \) is the inertia, \( B_r \) is the friction coefficient, and \( \tau_L \) is an external constant torque.

Following standard convention, all electrical (two-dimensional vector) signals are partitioned into their so-called \( d \) and \( q \) components. For instance, the stator current is decomposed as \( i_s = \text{col}(i_{sd}, i_{sq}) \). The use of the stator-oriented synchronous frame, \( v_s = \text{col}(V_s, 0) \) (with \( V_s \) the amplitude of the three-phase stator voltage), allows us to express the stator active and reactive powers in terms of \( i_{sd} \) and \( i_{sq} \), respectively, so that,

\[ P_s = v_s^\top i_s \] (11)

\[ Q_s = v_s^\top J_2 i_s \] (12)
simplify to

\[ P_s = V_s i_{sd} \] (13)
\[ Q_s = V_s i_{sq} \] (14)

In particular, assigning a desired value \( i_{sq}^* \) allows one to regulate the power factor of the stator side of the machine, while \( i_{sd}^* \) can be used to control the active power (delivered or consumed) by the DFIM. In a drive application, \( i_{sd}^* \) is fixed as a desired value to achieve the target torque in the presence of voltage constraints. This paper concentrates only on the problem of robust regulation of \( i_s \) to its desired value.

**IV. FEEDBACK LINEARIZING CURRENT CONTROLLER**

The proposed control scheme is presented in Figure 2. It is composed of the current control block and the well-know Blondel-Park transformation to recover the dq-measurements of the currents and the stator voltages, and its inverse to generate the three-phase rotor voltages from the computed \( v_r \) values (in the dq-framework).

The proposed algorithm is simpler than the classic stator-flux oriented control [6], where the stator flux estimation (or reconstruction) is required for the reference frame orientation. Furthermore, the stator voltage oriented framework allows us to directly use the stator currents \( i_s \) to regulate the active and reactive stator power, see equations (13) and (14). This fact also simplifies the algorithm compared with the standard approach.

The control algorithm can be also used as the inner-current loop for drive applications. Fixing \( i_{sq} \) for the reactive power compensation, a suitable \( i_{sd} \) can be obtained from a mechanical outer-loop for the torque (using (10)), or speed regulation (from the mechanical dynamics (9)).

The transformation of the three phase (stator and rotor) currents to the synchronous-reference (aligned to the stator voltages) is achieved with the rotation matrices

\[
K(\theta, \delta) = \begin{pmatrix}
e^{J_2 \delta} & O_2 \\
O_2 & e^{J_2(\delta-\theta)}
\end{pmatrix}
\] (15)
where $\dot{\theta} = \omega_s$, $\theta$ is the rotor position ($\dot{\theta} = \omega$), $O_2$ is a $2 \times 2$ matrix of zeros, and
\[
e^{J_2 \eta} = \begin{pmatrix} \cos(\eta) & -\sin(\eta) \\ \sin(\eta) & \cos(\eta) \end{pmatrix},
\] (16)

Notice that, since $\delta$ is defined by the stator voltages, and can be easily obtained via
\[
\delta = \arctan \left( \frac{v_{\alpha}}{v_{\beta}} \right).
\] (17)

This part of the scheme is easier to implement than the stator-flux oriented control, which depends on the stator flux estimation.

A. Feedback linearizing current controller

The proposed controller consists of a feedback linearization stage
\[
v_r = (\omega_s - \omega)L_{sr}J_2i_s + [(\omega_s - \omega)L_rJ_2 + R_rI_2]i_r + u
\] (18)

and a PI action
\[
u = -k_pJ_2(i_s - i_s^*) - k_IJ_2 \int (i_s - i_s^*)dt
\] (19)

with the scalar proportional and integral gains $k_p$ and $k_I$, respectively.

**Remark 1:** The first two terms in (18) exactly cancel the first two terms in (6), feedback linearizing the system and transforming the rotor equation in $\dot{\lambda}_r = u$.

**Remark 2:** In contrast to the standard practice, as mentioned in [1], the PI controller (19), is defined with the skew-symmetric matrix $J_2$. This fact turns out to be critical for the stability analysis. As explained in [19], this controller was obtained by applying passivity-based nonlinear control techniques.

Substituting (18) in (7), a linear closed-loop system is obtained, and applying the Laplace transform
\[
A(s) \begin{pmatrix} I_s(s) \\ I_r(s) \\ V_r(s) \end{pmatrix} = \begin{pmatrix} V_s(s) \\ O_2 \\ (k_ps + k_IJ_2)I_s^*(s) \end{pmatrix}
\] (20)

where
\[
A(s) = \begin{pmatrix} (L_s + R_s)I_2 + \omega_sL_sJ_2 & L_{sr}sI_2 + \omega_sL_{sr}J_2 & O_2 \\ L_{sr}sI_2 & L_rI_2 & -I_2 \\ (k_ps + k_IJ_2)I_2 & O_2 & sI_2 \end{pmatrix},
\] (21)

and $I_s(s) = \text{col}(I_{sd}(s), I_{sq}(s))$, $I_r(s) = \text{col}(I_{rd}(s), I_{rq}(s))$, $V_s(s) = \text{col}(V_{sd}(s), V_{sq}(s))$, $V_r(s) = \text{col}(V_{rd}(s), V_{rq}(s))$, and $(\cdot)^*$ refers to the desired values.

Stability of (20) is determined by the 6th order characteristic polynomial $\det A(s)$, whose stability was analyzed in [14] but without reaching complete analytic conditions for stability. Interestingly, this sixth order characteristic polynomial with real coefficients can be reduced into a 3rd order polynomial with complex coefficients for which complete analysis is possible.
Defining $\mathcal{I}_s(s) = I_{sd}(s) + jI_{sq}(s)$, $\mathcal{I}_r(s) = I_{rd}(s) + jI_{rq}(s)$, $\mathcal{V}_s(s) = V_{sd}(s) + jV_{sq}(s)$, $\mathcal{V}_r(s) = V_{rd}(s) + jV_{rq}(s)$, (20) can be rewritten as

$$A(s) \begin{pmatrix} \mathcal{I}_s(s) \\ \mathcal{I}_r(s) \\ \mathcal{V}_r(s) \end{pmatrix} = \begin{pmatrix} \mathcal{V}_s(s) \\ 0 \\ j(kps + kl)\mathcal{I}_s(s) \end{pmatrix}$$  \hspace{1cm} (22)

where

$$A(s) = \begin{pmatrix} L_ss + R_s + j\omega_sL_s & L_{sr}s + j\omega_sL_{sr} & 0 \\ L_{sr}s & L_rs & -1 \\ j(kps + kl) & 0 & s \end{pmatrix}.$$  \hspace{1cm} (23)

Notice that the polynomial $\det A(s)$ has 6 roots, which must be either real or appear as complex pairs, while $\det A(s)$ has 3 roots that can lie anywhere in the complex plane. The complex polynomial has the form

$$\det A(s) = a_0s^3 + (a_1 + jb_1)s^2 + (a_2 + jb_2)s + a_3 + jb_3$$  \hspace{1cm} (24)

where

$$a_0 = \mu$$  \hspace{1cm} (25)

$$a_1 = L_rR_s$$  \hspace{1cm} (26)

$$b_1 = \omega_s\mu - kpL_{sr}$$  \hspace{1cm} (27)

$$a_2 = kp\omega_sL_{sr}$$  \hspace{1cm} (28)

$$b_2 = -k_1L_{sr}$$  \hspace{1cm} (29)

$$a_3 = k_1\omega_sL_{sr}$$  \hspace{1cm} (30)

$$b_3 = 0$$  \hspace{1cm} (31)

where $\mu = L_sL_r - L_{sr}^2 > 0$ ($L_sL_r > L_{sr}^2$ in all electrical machines).

Now the stability of the closed loop system (6)-(7) with (18), can be analyzed with Lemma 1, where $a_0 = \mu > 0$ is fulfilled. Computing conditions (3)-(5) yields

$$\Delta_1 = L_rR_s$$  \hspace{1cm} (32)

$$\Delta_2 = L_{sr}(kp\omega_sL_{sr}^2 + kp\omega_sL_{sr}L_rR_s - 2k_1\omega_s\mu L_rR_s - k_1^2\mu L_{sr}$$  \hspace{1cm} (33)

$$\Delta_3 = k_1\omega_s^3L_{sr}^2L_rR_s(k_1^2L_{sr}L_rR_s - kp\omega_sL_{sr}L_r - k_1\omega_s\mu^2).$$  \hspace{1cm} (34)

The first condition is automatically fulfilled because $L_r, R_s > 0$. It can be shown (see Appendix) that one must have $k_1 > 0$, and the third condition is more restrictive than the second one. Consequently, the Hurwitz conditions reduce to $\Delta_3 > 0$. Notice that the stability condition does not depend on the mechanical speed. This stability condition has as asymptote the result presented in [14], which implies a generalization of the previous work. Figure 3 represents both conditions for the machine.
Fig. 3. Stability boundary for the proposed control law (18) as determined by $\Delta_2$ and $\Delta_3$. The stable region is on the right side of the curve.

parameters used in Section VI.

**Proposition 1:** Consider the DFIM system (6)-(7) in closed-loop with the control law (18). If

$$0 < k_I < \frac{k_p^2 L_{sr} L_r R_s}{\mu (\mu \omega_s + k_p L_{sr})}$$  \hspace{1cm} (35)$$

the closed-loop system is stable.

**Proof:** See Appendix.

**B. Effect of unknown parameters**

As seen from (18), the feedback linearization term requires the knowledge of $R_r$, $L_s$ and $L_{sr}$, which are in general uncertain and time-varying parameters. In particular, due to thermal effects, the value of $R_r$ is highly varying. In order to evaluate the effect of a possibly incorrect estimation of $R_r$, let us assume that one uses in (18) an estimated value $\hat{R}_r$, which can differ from the actual value of the rotor resistance. Then, (23) slightly modifies to

$$A_r(s) = \begin{pmatrix} L_s s + R_s + j \omega_s L_s & L_{sr}s + j \omega_s L_{sr} & 0 \\ L_{sr}s & L_r S + \hat{R}_r & -1 \\ j (k_p s + k_I) & 0 & s \end{pmatrix},$$  \hspace{1cm} (36)$$

where $\hat{R}_r = R_r - \hat{R}_r$.

Applying again Lemma 1 to the new matrix $A_r(s)$, it is possible to obtain a set of conditions given by $\Delta_1, \Delta_2, \Delta_3 > 0$. The determinant of $A_r(s)$ still has the form of (24), with the same parameters except the following three:

$$a_1 = L_r R_s + L_s \hat{R}_r$$  \hspace{1cm} (37)$$

$$a_2 = k_p \omega_s L_{sr} + R_s \hat{R}_r$$  \hspace{1cm} (38)$$

$$b_2 = -k_I L_{sr} + \omega_s L_s \hat{R}_r$$  \hspace{1cm} (39)$$
First, note that $a_1 > 0$, and consequently $\Delta_1 > 0$, if and only if $\hat{R}_r < R_r + \frac{L_r}{L_r^*} R_s$, which warns against an overestimation of $R_r$. Conditions $\Delta_2 > 0$ and $\Delta_3 > 0$ are quite more complicated,

$$\Delta_2 = c_1 k_1^2 + c_2 k_1 k_1 + c_3 k_1 + c_4 k_p + c_5$$  \hspace{1cm} (40)

$$\Delta_3 = d_1 k_1^3 + d_2 k_1^2 + d_3 k_1 + d_4 k_1 k_1 + d_5 k_1$$  \hspace{1cm} (41)

where

$$c_1 = -\mu L_{sr}^2$$  \hspace{1cm} (42)

$$c_2 = L_{sr}^2 (L_r R_s + L_s \hat{R}_r)$$  \hspace{1cm} (43)

$$c_3 = -2 \omega_s \mu L_{sr} L_r R_s$$  \hspace{1cm} (44)

$$c_4 = \omega_s L_{sr} L_r R_s (L_r R_s + L_s \hat{R}_r)$$  \hspace{1cm} (45)

$$c_5 = R_s \hat{R}_r \left( (L_r R_s + L_s \hat{R}_r)^2 + \omega_s^2 \mu L_s L_r \right)$$  \hspace{1cm} (46)

and

$$d_1 = -\omega_s \mu L_{sr}^3 R_s \hat{R}_r$$  \hspace{1cm} (47)

$$d_2 = \omega_s L_{sr}^3 R_s \left( \hat{R}_r (L_r R_s + L_s \hat{R}_r) - \omega_s^2 \mu L_r \right)$$  \hspace{1cm} (48)

$$d_3 = -\omega_s^2 \mu L_{sr}^2 R_s (L_s \hat{R}_r^2 + \omega_s^2 \mu L_r + 3 L_r R_s \hat{R}_r)$$  \hspace{1cm} (49)

$$d_4 = \omega_s^3 L_{sr}^2 L_r R_s (L_r R_s + L_s \hat{R}_r)$$  \hspace{1cm} (50)

$$d_5 = \omega_s^2 L_{sr}^2 R_s \hat{R}_r^2 (\omega_s^2 \mu L_s L_r + (L_r R_s + L_s \hat{R}_r)^2 + L_r R_s (L_r R_s + L_s \hat{R}_r))$$  \hspace{1cm} (51)

$$d_6 = \omega_s L_{sr} R_s \hat{R}_r \left( (\omega_s^2 \mu L_s L_r + (L_r R_s + L_s \hat{R}_r)^2 \right)$$  \hspace{1cm} (52)

At this point, in order to simplify the stability conditions, it is possible to compute the asymptotes of (40) and (41), which turn out to be the same and given by

$$k_1 = \frac{L_r R_s + L_s \hat{R}_r}{\mu} k_p - \frac{\omega_s L_s R_s}{L_{sr}}.$$  \hspace{1cm} (53)

Notice that for $\hat{R}_r = 0$, this corresponds to the asymptote of (35). Also, the stability condition presented in [14] is recovered.

Figure 4, shows a numerical example of the stability region for different values of $\hat{R}_r$. With the parameters of the machine described in Section VI, and considering a 10% error in the $R_r$ estimation, $\Delta_3 > 0$ is more restrictive than $\Delta_2 > 0$. In Figure 4, the $\Delta_3 = 0$ for different errors in the rotor resistance estimation are shown.

This numerical example shows that a positive error, i.e. $\hat{R}_r > 0$, in the estimation of $\hat{R}_r$ is preferred. In other words, overestimating the rotor resistance implies decreasing the stability region and, as pointed out before, can even destabilize the system. Moreover, for small values of $\hat{R}_r$, a small stable region appears for $k_p < 0$. This can be easily seen from the asymptote (53), where $\hat{R}_r < 0$ implies a lower slope. This fact suggests taking $\hat{R}_r = 0$ to have a larger stability region, and also simplify
the control law. Then, asymptote (53) yields,

$$k_I = \frac{L_r R_s + L_x R_r}{\mu} k_P - \frac{\omega_s L_{tr} R_s}{L_{sr}}. \quad (54)$$

V. DIRECT PI STATOR CURRENT CONTROLLER

The control law introduced in the previous section guarantees stability for a large range of the PI parameter values. However, in order to implement the control algorithm, it is necessary to know some machine parameters and both the stator and the rotor currents. In this section, a simplification of the proposed controller (18)-(19) is analyzed, that only keeps the PI action, i.e.,

$$v_r = -k_{P} J_{2}(i_s - i_s^*) - k_{I} J_{2} \int (i_s - i_s^*) dt. \quad (55)$$

For this scheme, shown in Figure 5, the rotor currents are not required, and only the stator currents need to be measured. At this point, for the stability analysis, a constant mechanical speed is assumed. Using the same idea as before, the closed-loop dynamics can be written as (22) with $\mathcal{A}(s)$ becoming
The polynomial $\det A_{PI}(s)$ still has 3 roots, and the Hurwitz test described in Lemma 1 can be used. The determinant of (56) has the same form as (24) with the following coefficients

\begin{align*}
a_0 &= \mu \\
a_1 &= L_r R_s + L_s R_r \\
b_1 &= \mu(2\omega_s - \omega) - k_p L_{sr} \\
a_2 &= R_s R_r - \omega_s(\omega_s - \omega)\mu + k_p\omega_s L_{sr} \\
b_2 &= (\omega_s - \omega)L_r R_s + \omega_s L_s R_r - k_1 L_{sr} \\
a_3 &= k_1 \omega_s L_{sr} \\
b_3 &= 0.
\end{align*}

The stability of the closed loop system (6)-(7) with (55) can be analyzed by computing $\Delta_1$, $\Delta_2$, $\Delta_3$. $\Delta_1 = a_1$ always satisfies $\Delta_1 > 0$, while $\Delta_2$ and $\Delta_3$ are in the same form as (40) and (41), respectively, where the coefficients now take the following values

\begin{align*}
c_1 &= -\mu L_{sr}^2 \\
c_2 &= L_{sr}^2 (L_r R_s + L_s R_r) \\
c_3 &= \mu L_{sr}(\omega(L_r R_r - L_r R_s) - \omega_s(L_r R_s + L_s R_r)) \\
c_4 &= \omega L_{sr} L_r R_s(L_r R_s + L_s R_r) \\
c_5 &= R_s R_r(L_r R_s + L_s R_r)^2
\end{align*}

and

\begin{align*}
d_1 &= -\omega_s \mu L_{sr}^3 R_r R_s \\
d_2 &= \omega_s L_{sr}^3 R_s(R_r(L_r R_s + L_s R_r) - \omega_s \omega \mu L_r) \\
d_3 &= \omega_s \mu L_{sr}^3 R_s(-2\omega_s R_r(L_r R_s + L_s R_r) - \omega_s \omega^2 \mu L_r - \omega R_r(L_r R_s - L_s R_r)) \\
d_4 &= \omega \omega_s L_{sr}^3 L_r R_s(L_r R_s + L_s R_r) \\
d_5 &= \omega_s \omega_s^2 L_{sr}^3 L_r R_s (\omega L_s R_r - (\omega_s - \omega)(L_r R_s + L_s R_r)) \\
&\quad + \omega_s L_{sr}^2 R_s R_r(L_r R_s + L_s R_r)(\omega_s(L_r R_s + L_s R_r) + \omega L_r R_s) \\
d_6 &= \omega_s L_{sr} R_s R_r(R_s R_r - (\omega_s - \omega)\omega_s \mu)((L_r R_s + L_s R_r)^2 + \omega^2 \mu L_s L_r)
\end{align*}
To obtain an expression for the stability region becomes complicated. As a first result, conditions can be plotted for a numerical case. Using the machine parameters of Section VI, equations (40) and (41) are obtained for different values of the mechanical speed. Figure 6 shows the stability regions for the mechanical speed.

In order to bound the stability region, it is possible to find the asymptote of (40) and (41). Similarly to the case for an unknown rotor resistance presented in the previous section, the asymptotes for the $\Delta_2 = 0$ and $\Delta_3 = 0$ are equal and given by

$$k_I = \frac{L_r R_s + L_s R_r}{\mu} k_P + \frac{\omega L_s R_r - \omega_s (L_r R_s + L_s R_r)}{L_{sr}}.$$  \hspace{1cm} (75)

Note that the slope of the stability region does not depend on the mechanical speed. As the worst case is when $\omega = 0$, the stability for the PI controller proposed can be ensured by setting

$$k_I < \frac{L_r R_s + L_s R_r}{\mu} k_P - \frac{\omega_s (L_r R_s + L_s R_r)}{L_{sr}}.$$  \hspace{1cm} (76)

VI. SIMULATIONS

The proposed controller was tested in numerical experiments using Matlab. In order to obtain an accurate simulation of the real plant, the model was built with the SimPowerSystems toolbox, and effects due to a digital implementation were considered.

The DFIM parameters are $R_s = 0.492\Omega$, $R_r = 0.442\Omega$, $L_s = 7.25\text{mH}$, $L_r = 7.15\text{mH}$ and $L_s = 7.1\text{mH}$. The DIFM is connected to a 380V and 50Hz power grid. This corresponds to a constant stator voltage vector $v_s = [380\sqrt{3}, 0]\text{V}$ and $\omega_s = 100\pi\text{rad/s}$ in the stator-voltage oriented frame.

A first simulation consists in a comparison between the two presented control schemes. For this test, the dq-model (6)-(7) is in a closed loop with the control schemes (18) and (55). The knowledge of all the parameters is assumed for the feedback linearizing control scheme and the control gains are fixed as $k_P = 1$, $k_I = 0.1$. The direct PI controller gains are $k_P = 1$, $k_I = 150$. The mechanical speed is set to $\omega = 325\text{rads}^{-1}$ and the initial conditions are $i_s(0) = [-3, 0]\text{A}$. The reference is changed from the initial value to $i_s^* = [-5, 1]\text{A}$ at $t = 0.02s$. As shown in Figure 7, both algorithms perform in a similar way. This result suggests the use of the direct PI controller because it is easily implementable; rotor currents are not needed, and the knowledge of the machine parameters is not required.
The second set of simulations only test the so-called direct PI controller. In this case, the model is implemented using the SimPowerSystems toolbox of Matlab, which contains a library with realistic implementations of some elements such as sources, electrical machines and measurement elements. For this simulation, the Blondel-Park transformations had to be coded, as well as the obtained control action, \( v_r \), that is converted to the three-phase voltage, \( V_r \), to be applied to the rotor side of the DFIM. Also, the effects of a real implementation are included, \textit{i.e.} sampling, quantification and saturation. The sampling frequency is set at 10kHz, all the variables are quantified as in a 16bit processor, and the rotor voltages which are saturated at \( \pm 120V \), are delayed with one sampling time period. The control parameters and the mechanical speed are \( K_f = 1 \) and \( K_I = 150 \), and \( \omega = 325\text{rads}^{-1} \). The desired dq-stator currents are shown in Figure 8, where several scenarios are considered changing the stator current references, \( i_{sd}^* \) and \( i_{sq}^* \) (also simultaneously, at \( t = 0.5s \)). Note that the reference change at \( t = 0.7s \) implies a switch of the power flow direction (from generating to absorbing).

Stator currents are depicted in Figures 9 and 10. The evolution of the measured three-phase stator currents are displayed in Figure 9. Figure 10 shows how the obtained dq-stator currents stabilize at the desired values under several reference changes. It is worth mentioning that the controller is able to operate for both signs of the \( i_{sd} \) current, showing that this algorithm can
be used for both power generation or consumption (as in driving applications).

Figure 11 contains a detailed zoom for each reference change for the a-phase stator voltage and current. One can see how the phase changes depending on the value of the q-stator current component. The first step at $t = 0.1\text{s}$ only modifies the current amplitude and the phase remains in opposite-phase (power is flowing out of the DFIM, that is, the DFIM is in generation mode). The second change, at $t = 0.3\text{s}$, implies that reactive power (with $\cos \phi = 0.6$) appears. The third reference change ($t = 0.5\text{s}$) returns to zero reactive power and reduces the active power delivered. Finally, the fourth change ($t = 0.7\text{s}$) affects only the direction of the power flow, with zero reactive power (voltage and current are in phase, or in opposite phase), and the current amplitude is kept.

**VII. Conclusions**

In this paper, a particularly simple controller for the DFIM has been presented. It consists of a PI regulator for the stator currents and (possibly) a feedback linearizing term. As the proposed scheme is defined in the stator voltage reference frame, the active and reactive powers are directly related to the d and q stator currents, respectively, and the power regulation does not require extra loops or computations. Moreover, no stator flux estimation is required. Consequently, the algorithm is simpler
than classical vector control.

In contrast with the standard decoupling controllers, the PI action is defined with a skew--symmetric gain matrix, $J_2$, which is required for stability purposes. The stability is analyzed with a Routh-Hurwitz test for polynomials with complex coefficients. This tool provides a simple analysis to determine the stability regions of the control gains.

The paper presents two approaches. The first algorithm consists of a feedback linearization stage plus a PI action. This scheme is particularly attractive because the resulting system is linear and independent of the mechanical speed. The influence of an incorrect estimation of the rotor resistance is also studied. The second algorithm only uses the PI term, so that it is extremely easy to implement. It does not require either the knowledge of the machine parameters or measurements of the rotor currents. As opposed to the previous algorithm, the stability of the second approach is based on the assumption of constant mechanical speed. However, this restriction is widely assumed and can be accepted when the time constant of the mechanical dynamics is much higher than the electrical one.

Finally, the proposed controllers are verified in simulations. First, a dq-model of the DFIM is used to compare both controllers, resulting in a similar behavior. Secondly, the direct PI algorithm is tested in a more accurate scenario. The model used in the simulation contains some parasitic elements and non-ideal effects such as sampling, quantification and some delay, in order to emulate a real experiment. The results, demonstrating good, validate the proposed control scheme.

REFERENCES

APPENDIX

**Proof of Proposition 1.** From Lemma 1, the stability conditions are \( \Delta_1 > 0 \) (automatically fulfilled due to \( L_r, R_s > 0 \)), \( \Delta_2 > 0 \) and \( \Delta_3 > 0 \). The proof contains two parts. First, it is shown that, for \( k_I > 0 \), conditions (35) and \( k_P > 0 \) are derived. Secondly, it is proved that there is no stability for negative values of \( k_I \).

Let us start with the case \( k_I > 0 \). Since (35) comes from \( \Delta_3 > 0 \), the proof only requires to show that \( \Delta_2 > 0 \) is satisfied whenever \( \Delta_3 > 0 \) holds, and that \( k_P \) should be positive.

Note that \( \Delta_2 > 0 \) if and only if
\[
k_P > \frac{k_I \mu (2 \omega_s L_r R_s + k_I L_{sr})}{L_r R_s (\omega_s L_r R_s + k_I L_{sr})}.
\]

Furthermore, consider the function \( f(k_P) \) defined by
\[
f(k_P) = k_P^2 L_{sr} L_r R_s - k_P k_I \mu L_{sr} - k_I \omega_s \mu^2.
\]
and note that, with \( k_I > 0 \), \( \Delta_3 > 0 \) holds true if and only if \( f(k_P) > 0 \). \( f(k_P) \) is a quadratic function such that \( f(k_P) = 0 \) for

\[
k_P = \frac{k_I \mu L_{sr} + \sqrt{(k_I \mu L_{sr})^2 + 4k_I \omega_s \mu^2 L_{sr} L_r R_s}}{2L_{sr} L_r R_s}.
\] (79)

Given that there are two real roots, one positive and the other negative, and the later is inconsistent with (77), it follows that \( f(k_P) > 0 \) if and only if

\[
k_P > \frac{k_I \mu L_{sr} + \sqrt{(k_I \mu L_{sr})^2 + 4k_I \omega_s \mu^2 L_{sr} L_r R_s}}{2L_{sr} L_r R_s},
\] (80)

which also implies \( k_P > 0 \).

The desired result is proved if it can be shown that (80) implies (77), i.e.

\[
\frac{k_I \mu L_{sr} + \sqrt{(k_I \mu L_{sr})^2 + 4k_I \omega_s \mu^2 L_{sr} L_r R_s}}{2L_{sr} L_r R_s} > \frac{k_I \mu (2\omega_s L_r R_s + k_I L_{sr})}{L_r R_s (\omega_s L_r R_s + k_I L_{sr})},
\] (81)

or, equivalently,

\[
(\omega_s L_r R_s + k_I L_{sr})(k_I \mu L_{sr} + \sqrt{(k_I \mu L_{sr})^2 + 4k_I \omega_s \mu^2 L_{sr} L_r R_s}) > 2k_I \mu L_{sr}(2\omega_s L_r R_s + k_I L_{sr}).
\] (82)

This inequality is satisfied if

\[
(\omega_s L_r R_s + k_I L_{sr})\sqrt{(k_I \mu L_{sr})^2 + 4k_I \omega_s \mu^2 L_{sr} L_r R_s} > 2k_I \mu L_{sr}(2\omega_s L_r R_s + k_I L_{sr}) - (\omega_s L_r R_s + k_I L_{sr})k_I \mu L_{sr}
\] (83)

or

\[
(\omega_s L_r R_s + k_I L_{sr})^2((k_I \mu L_{sr})^2 + 4k_I \omega_s \mu^2 L_{sr} L_r R_s) + 4(k_I \mu L_{sr})^2(2\omega_s L_r R_s + k_I L_{sr}) - (\omega_s L_r R_s + k_I L_{sr})^2(k_I \mu L_{sr})^2
\] (84)

or

\[
4k_I \omega_s \mu^2 L_{sr} L_r R_s (\omega_s L_r R_s + k_I L_{sr})^2 > 4(k_I \mu L_{sr})^2(2\omega_s L_r R_s + k_I L_{sr})\omega_s L_r R_s
\] (85)

or

\[
(\omega_s L_r R_s + k_I L_{sr})^2 > (k_I L_{sr})^2 + 2k_I \omega_s \mu L_{sr} L_r R_s.
\] (86)

Since the last inequality is indeed always valid, we may conclude that, for \( k_I > 0 \), \( \Delta_3 > 0 \) implies that \( \Delta_2 > 0 \) and the stability is determined by (35) and \( k_P > 0 \).

It is possible to show that for \( k_I < 0 \) there does not exist a set of \( k_P \) values which satisfies simultaneously \( \Delta_2 > 0 \) and \( \Delta_3 > 0 \). Solutions of \( \Delta_3 > 0 \) with \( k_I < 0 \) imply \( f(k_P) < 0 \). If

\[
k_I < -\frac{4\omega_s L_r R_s}{L_{sr}},
\] (87)
and equation (78) has two negative real solutions

\[ k_{P1} = \frac{k_I \mu L_{sr} + \sqrt{(k_I \mu L_{sr})^2 + 4k_I \omega_s \mu^2 L_{sr} L_r R_s}}{2 L_{sr} L_r R_s} \]  
\[ (88) \]

\[ k_{P2} = \frac{k_I \mu L_{sr} - \sqrt{(k_I \mu L_{sr})^2 + 4k_I \omega_s \mu^2 L_{sr} L_r R_s}}{2 L_{sr} L_r R_s} \]  
\[ (89) \]

Solutions of \( f(k_P) < 0 \) are bounded by \( k_{P2} < k_P < k_{P1} \). On the other hand, from \( \Delta_2 > 0 \), inequality (87) implies

\[ k_P < \frac{k_I \mu (2 \omega_s L_r R_s + k_I L_{sr})}{L_r R_s (\omega_s L_r R_s + k_I L_{sr})}. \]  
\[ (90) \]

It can be shown that the constraint given by (90) does not contain the range \( k_{P2} < k_P < k_{P1} \), i.e.

\[ \frac{k_I \mu (2 \omega_s L_r R_s + k_I L_{sr})}{L_r R_s (\omega_s L_r R_s + k_I L_{sr})} < \frac{k_I \mu L_{sr} - \sqrt{(k_I \mu L_{sr})^2 + 4k_I \omega_s \mu^2 L_{sr} L_r R_s}}{2 L_{sr} L_r R_s}. \]  
\[ (91) \]