VECTOR CONTROL OF AN INDUCTION MACHINE IN A STATOR FLUX REFERENCE FRAME

A. Schwery, B. Kawkabani, M. Tu Xuan, J.J. Simond

Swiss Federal Institute of Technology, Electrical Engineering Dept., CH-1015 Lausanne
tel: 4121/6934690, fax: 4121/6932687, e-mail: alexander.schwery@epfl.ch
SWITZERLAND

Abstract

This paper describes a **stator flux** oriented control strategy using state space controllers. The optimal determination of the different controllers' parameters leads to an independent control of the stator flux and the torque. The effects of undesired offsets on the calculated stator flux are discussed and a solution for eliminating these offsets is presented. Verification by computer simulation and results of the control implementation on a Motorola DSP 56002 are presented.

1 Introduction

In the last few years, the control of induction machines by the field oriented method has become the subject of several publications [1], [2]. Many papers describe controllers of induction motor drives based on the **rotor** flux orientation [3]. The main advantage of this method is that decoupling between torque and flux can easily be achieved. Nevertheless, the exact knowledge of the machine parameters (especially the rotor time constant) is necessary to guarantee the orientation of the rotor flux vector along the real axis, and methods to balance the variations of these parameters have to be provided. The stator flux orientation, less sensitive to the variations of the machine parameters, has the disadvantage of a more complicated decoupling between flux and torque. That's probably the reason why less papers have been published on this strategy.

In the present paper, a new control scheme based on the **stator** flux oriented method is presented (figure 1). Contrary to the rotor flux strategy, the decoupling between the currents in the two axes doesn't imply a decoupling between torque and flux in the stator flux strategy. Therefore it's necessary to add a flux controller to the control scheme. The input of this controller is obtained by integration of the measured stator voltages.

The practical realization has shown that the obtained fluxes are affected by **offsets** which introduce undesired oscillations on the quantities in the rotating reference frame. This problem is examined and a solution is proposed. The command outputs of the controllers are equipped with limiters protecting the controlled system from overload. During operation of a feedback loop in limitation, the integral component of its controller is corrected (anti - reset - windup).

The new aspect of this paper in the **stator** flux orientation strategy consists of using a flux controller as well as the optimal determination of the controllers' parameters in order to obtain a good flux and torque decoupling. A new method to eliminate undesired offsets on the flux acquisition is presented.

The control strategy is implemented on a Motorola DSP 56002. Results showing a good dynamic behavior of the studied system in different operating conditions are presented.

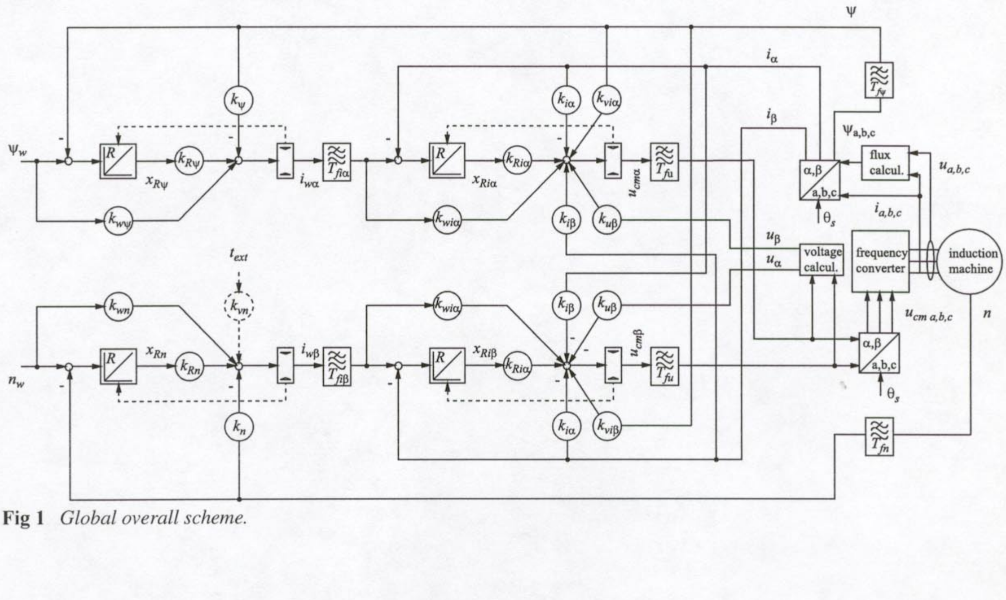

Fig 1 *Global overall scheme.*

2 Studied system

The following paragraph gives a short description of the global system. A more detailed description can be found in [4]. Figure 1 represents the studied system which can be divided into three parts: The **induction machine,** the **frequency converter** which amplifies the signals calculated by the **controller part**.

All parameters and variables in this paper are expressed in per unit *[p.u.]* except the time that is expressed in seconds *[s]*.

2.1 Induction machine

The equations describing the induction machine are expressed in a rotating reference frame (**stator** flux reference frame). The real axis of this reference frame is aligned with the stator flux. Thus, the imaginary part of the stator flux is zero. The reference frame rotates with the synchronous angular speed. The real and imaginary axes are denoted with the indexes α and β respectively.

2.2 Frequency converter

For the determination of the controllers' parameters a first order differential equation is used to take into account the time delay caused by the frequency converter. The simulation program offers the possibility to make simulations either with this simplified model, or with a more realistic model using a vector modulation method to calculate the stator voltages.

Figure 1 shows that stator voltages are used in the current control loop. Instead of measuring the real voltages on the machine they are estimated using the set-point values $u_{cm\alpha}$, $u_{cm\beta}$ and the differential equation describing the frequency converter.

2.3 Control system

Flux linkages in the stator of the machine are determined using the stator resistance value and measurements of stator currents and voltages. Then currents and flux are transformed using Park's transformation at the input of the control system. Four state controllers are used to control torque and flux of the machine. Speed, flux and current controllers are mounted in cascade. Thus, inner quantities may easily be limited. All controllers are equipped with limiters protecting the system from overload. During limitation, the integral components are corrected (anti-reset-windup). Finally, at the output of the control system the two set-point values of the stator voltages are retransformed using the inverse Park transformation. All controllers are designed as being continuous and then transformed into digital controllers by a Padé approximation [6].

2.4 Current controllers

As shown in figure 1 state space controllers are used to control the currents in both axes. The parameters of these two regulators are determined in order to ensure a good dynamic response and the decoupling between the two components of the stator current.

Starting with a closed loop description of the current control loop, a complex transfer function, relying the stator current to its set-point value, can be found. The imaginary part of this transfer function is put equal to zero to decouple the two components of the current (a variation of $i_{w\alpha}$ has no influence on i_β, respectively $i_{w\beta}$ on i_α). Proceeding this way, expressions for the following controller parameters can be found (see figure 1): $k_{u\beta}$, $k_{i\beta}$, $k_{wi\beta}$, $k_{Ri\beta}$, $k_{vi\beta}$, $k_{u\alpha}$, $k_{vi\alpha}$. The remaining parameters $k_{i\alpha}$, $k_{Ri\alpha}$, $k_{wi\alpha}$ are defined by pole placement in order to ensure a good dynamic behavior of the current control loop. These parameters depend on the operating point of the machine. That's the reason why they have to be recalculated at each cycle.

2.5 Flux controller

Contrary to the rotor flux orientation method the decoupling of the currents in the two axes does not imply a decoupling between torque and flux in the stator flux orientation method. More precisely, the stator flux not only depends on the current in the real axis but also on the current in the imaginary axis. A change of the set-point value in the imaginary axis does not influence the value of the current in the real axis (since they are decoupled), but has an effect on the flux. This flux variation has to be compensated by the controller, in order to have a constant flux level. As we will see in the results, the flux controller allows also to reduce the flux level while operating in field weakening range.

How the stator flux is obtained, and the problems related with the integration will be discussed in the second part of the paper. To determine the parameters of the flux and the speed controllers, the current control loop is replaced by a first order transfer function. A transfer function relying the current in the real axes and the flux can be found. The parameters $k_{R\psi}$, k_ψ and $k_{w\psi}$ are then defined by pole placement.

2.6 Speed controller

Using the same transfer function for the current control loop and the mechanical equation describing the machine, a closed loop transfer function linking the speed to its set-point value is expressed. The parameters k_{Rn}, k_n and k_{wn} are also defined by pole placement.

3 Stator Flux

The three stator phase flux linkages are obtained by integration using equation (1). As the machine is fed by PWM voltages, the calculation of equation (1) would require an extremely high sampling frequency. Therefore the calculation is done by hardware.

$$\psi_{a,b,c} = \int_0^t (u_{a,b,c} - r_s i_{a,b,c})dt \qquad (1)$$

The stator resistance value is transmitted to the three acquisition cards. This value is multiplied with the measured currents to obtain the voltage drops over the stator resistances. These voltage drops are substracted from the phase voltages and the results are integrated to obtain the stator flux linkages of each phase.

Only after the integration, the flux is sampled and read by the DSP. To calculate the orientation of the flux space vector (the orientation of the rotating reference frame), the three phase flux linkages are transformed into an orthogonal stator fixed two axes system.

3.1 Offsets

Due to measurement imperfections and the use of a differential amplifier for the integration, the obtained flux linkages are affected by undesired offsets. The influence of these offsets on the control strategy have been examined with the simulation program. Figure 2 shows the starting-up of an induction machine. An offset of *2%* has been added to the calculated flux linkages.

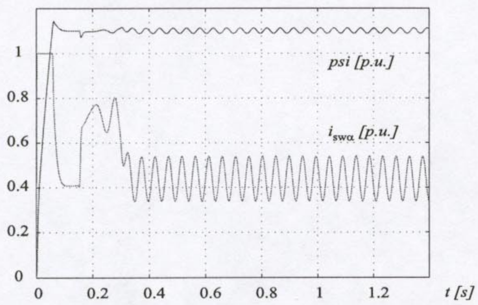

Fig 2 *Oscillations introduced by an offset on the sampled stator flux.*

It can be seen that the offset introduces oscillations on the current in the real axis as well as on the flux. The oscillations frequency corresponds to the angular speed of the reference frame which is about *0.4 p.u.* in this example. Oscillations can also be observed on the speed as well as on the current in the other axis. In the following chapters a method to eliminate the offsets is presented.

3.2 Filter

The offsets can be eliminated by high-pass filtering after integration. At high speed, a small filter time constant can be chosen in order to suppress any offset after a short laps of time. At low speed though, filtering with a small time constant would introduce an error by eliminating not only the offset but also a part of the 'useful signal'. On the other hand, at low speed, the introduced oscillations are at a low frequency, so that the error on the speed can be corrected by the speed regulator. Therefore a compromise would be to filter with a small time constant at high speed and a big time constant at low speed. By choosing a time constant depending on speed as given by equation (2), the amplitude and time delay errors introduced by the filter are constant.

$$T_h(n) = \frac{a}{n} \tag{2}$$

Where a is a parameter, n the mechanical speed and T_h the filter time constant. The following results have been obtained with $a = 0.1$. Normally, the correction of the amplitude and the time delay introduced by the filter should be considered as being a function of the stator frequency and not of the mechanical speed. But due to the difficulty in obtaining a good stator frequency estimation, we prefer the small error introduced by making a correction depending on the speed (rather than on the stator frequency).

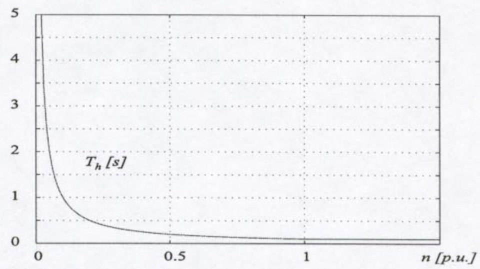

Fig 3 *High-pass filter time constant versus speed.*

3.3 Modifying the filter time constant

Considering the discrete version of a first order low-pass filter (3), it can be seen that any change of the filter time constant introduces a step on the filtered signal.

$$\psi_{ft}[k+1] = \psi_{ft}[k]\left(1 - e^{-\frac{\Delta t}{T_h}}\right) + \psi[k]e^{-\frac{\Delta t}{T_h}} \tag{3}$$

Where ψ_{ft} is the flux after low-pass filtering and Δt the sampling period. ψ_{ft} is the offset that has to be substracted from ψ at each sampling interval. The effect of a sudden change of the filter time constant is shown in figure 4.

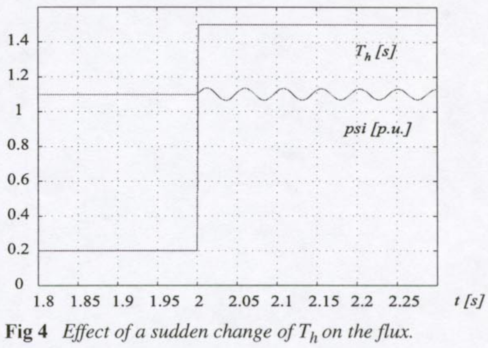

Fig 4 *Effect of a sudden change of T_h on the flux.*

The step on the offset due to a change of T_h is given by equation (4).

$$step = e^{-\frac{\Delta t}{T_h}} - e^{-\frac{\Delta t}{(T_h + \Delta T_h)}} \tag{4}$$

The idea consists in finding relations for ΔT_h giving a step that is smaller than a certain limit, which called *lim*. Table 1 summarizes the maximum and minimum ΔT_h values that have to be respected when changing ΔT_h.

$\Delta T_h < 0$		$\Delta T_h > -\dfrac{\Delta t}{\ln\left(e^{-\frac{\Delta t}{T_h}} - lim\right)} - T_h$
$\Delta T_h > 0$	$T_h < -\dfrac{\Delta t}{\ln(1 - lim)}$	$\Delta T_h < -\dfrac{\Delta t}{\ln\left(e^{-\frac{\Delta t}{T_h}} + lim\right)} - T_h$
$\Delta T_h > 0$	$T_h > -\dfrac{\Delta t}{\ln(1 - lim)}$	$\Delta T_h = \infty$

Tab 1*Limits that have to be respected when changing T_h.*

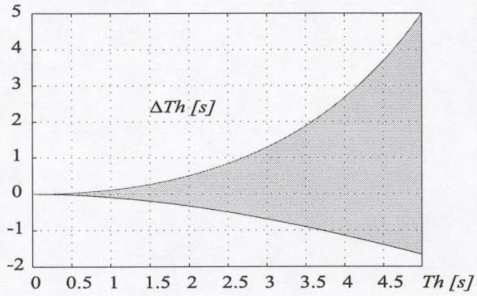

Fig 5 *Area of acceptable ΔT_h.*

Figure 5 shows the possible values for ΔT_h for lim = *0.0001*. For this limit the third case of table 1 is only reached for $T_h > 10$ and can thus not be seen on figure 5. When the above method is applied to the example shown in figure 4, the same change of T_h has a much smaller influence on the flux, as it can be seen in figure 6.

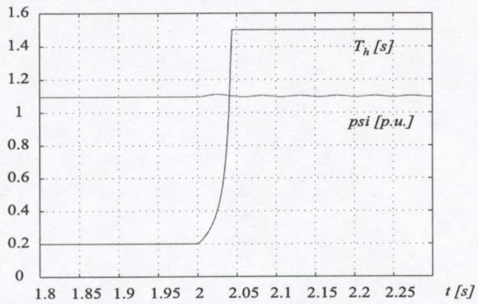

Fig 6 *Effect of a change of T_h on the flux using the described method.*

To illustrate the effect of the method, the starting-up simulation shown in figure 2 is repeated. Respecting the limits given in table 1, an additional algorithm modifies the filter time constant according to figure 3. Figure 7 shows that the initial amplitude of the oscillations on $i_{w\alpha}$ is smaller compared to figure 2. Furthermore, the oscillation is now damped in a short laps of time. The remaining oscillations on the flux can be neglected.

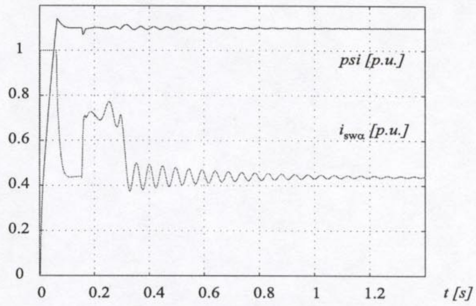

Fig 7 *Starting-up with an offset on the sampled stator flux using the described method.*

4 Results

4.1 Experimental setup

The described control algorithm has been realized and tested. Figure 8 shows the block diagram of the test bench. An induction machine (see appendix) is fed by a frequency converter whose IGBTs are controlled by the control unit. The control algorithm has been implemented on a Motorola 56002 DSP and can be downloaded and modified with a PC. The induction machine is connected to a direct current machine which is fed by a bidirectional current converter. Thus the induction machine can work as generator and motor in both directions of speed.

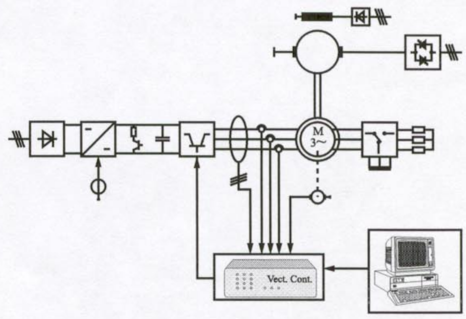

Fig 8 *Block diagram of the experimental setup.*

4.2 Field weakening example

The first example describes an application of field weakening. Figures 9 and 11 show measurements of the set-point values of speed and flux with the corresponding speed and flux. Figures 10 and 12 give the current set-point values in the corresponding axes. At t = *2.5 s*, the speed set-point value changes from *0.4 p.u.* to *-0.7 p.u.*. To reach the *-0.7 p.u.* without changing the current limitations and the dcvoltage level of the frequency converter, the flux level has to be decreased. For this reason the flux level is changed to *0.8 p.u.* before changing the speed. During the entire test, the machine is loaded with the same load torque which is about 66% of the rated torque. This means that the induction machine works first as a motor and than as a generator.

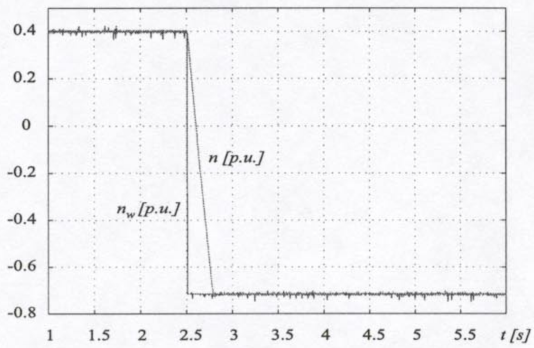

Fig 9 *Measured speed and set-point value of speed.*

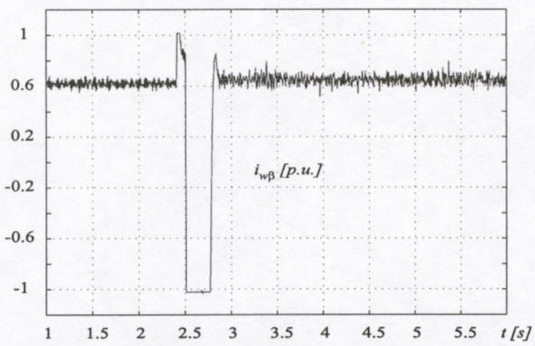

Fig 10 *Current set-point value in the imaginary axis.*

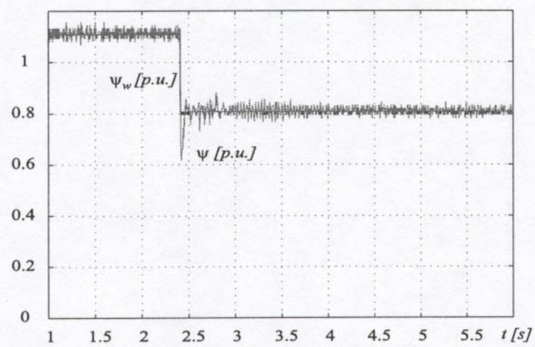

Fig 11 *Measured flux and set-point value of flux.*

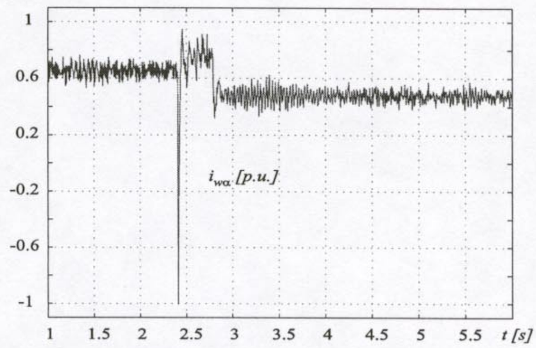

Fig 12 *Current set-point value in the real axis.*

4.3 Low speed example

The second example shows a reverse of speed at constant flux level. The same load torque is applied to the induction machine and kept constant during the test. Figures 13 and 15 show measurements of the set-point values of speed and flux with the corresponding speed and flux. Figures 14 and 16 illustrate measurements of the current set-point values in the corresponding axes. At $t = 2.5\ s$ the speed set-point value changes from *0.4 p.u.* to *-0.05 p.u.*. Due to the high filter time constant at low speed, (see figure 3) the oscillations on $i_{w\alpha}$ are more important compared to the first example. Nevertheless, the speed follows its set-point value because the speed controller has enough time to compensate the oscillations.

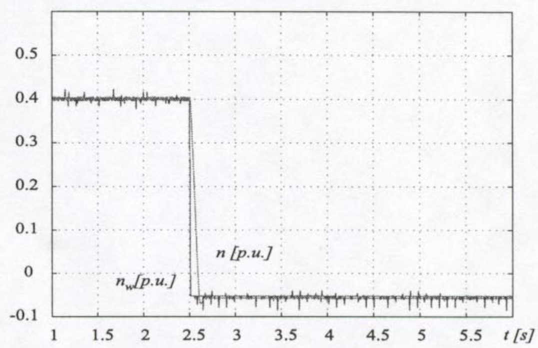

Fig 13 *Measured speed and set-point value of speed.*

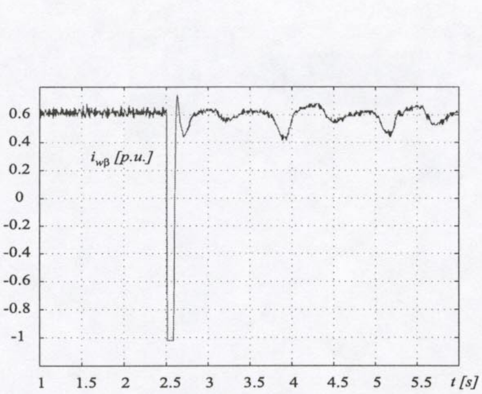

Fig 14 *Current set-point value in the imaginary axis.*

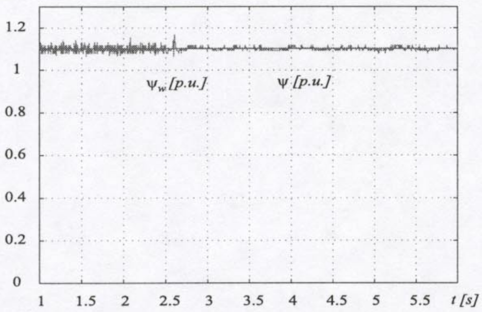

Fig 15 *Measured flux and set-point value of flux.*

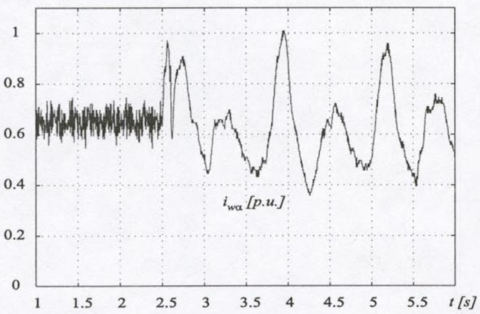

Fig 16 *Current set-point value in the real axis.*

5 Conclusions

In this paper the control of an induction machine using a stator flux strategy has been presented. The need of a flux controller has been explained. The problem of undesired offsets on the sampled stator flux has been discussed. A solution to eliminate these offsets has been given.

Simulation results help to explain the effects of such offsets and confirm that they can be eliminated with the described method.

The control algorithm has been implemented on a DSP 56002. Measured results show that good dynamic performances of the drive as well as an independent control of torque and flux can be obtained by using the proposed strategy.

Appendix

Induction machine with wound rotor, star-connected:

rated voltage	*220/380*	*[V]*
rated current	*8.7/5.0*	*[A]*
rated active power	*2200*	*[W]*
rated speed	*1400*	*[rpm]*
rated power factor	*0.81*	*[1]*
stator resistance	*0.0755*	*[p.u.]*
stator leakage inductance	*0.1195*	*[p.u.]*
magnetizing inductance	*2.5*	*[p.u.]*
rotor resistance	*0.0677*	*[p.u.]*
rotor leakage inductance	*0.1707*	*[p.u.]*
number of poles	*4*	*[1]*
mechanical time constant (with direct current machine)	*0.297*	*[s]*

References

[1] Lajoie-Mazenc E., *Vector control technics of induction machines for high performances adjustable speed drives,* Edition EDF, January 1992

[2] Profumo F., Pastorelli M., Ferraris P., De Doncker R.W., *Comparison of universal field oriented (UFO) controllers in different reference frames,* EPE 1991, Firenze, Italy

[3] J. Moerschell, *Signal processor based field oriented vector control for an induction motor drive*, EPE 1991, Firenze, Italy

[4] A. Schwery, B. Kawkabani, J.J. Simond, *A stator flux oriented vector control for a PWM inverter-fed induction motor drive*, ICEM 1996, Vigo, Spain

[5] M.T. Nguyen, M. Tu Xuan, J.J. Simond, *Digital transient torque measurements for rotating or linear AC machines (real time measurement)*, EPE 1995, Sevilla (Spain), pp. 1.319 - 1.321

[6] Bühler Hansruedi, *Conception de systèmes automatiques*, Presses polytechniques Romandes, 1988, Lausanne, p. 363-376