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# Performance Analysis of a Rotor Flux-Oriented Control in Induction Traction Motor

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Abstract - The superiority of using field-oriented control in electric drive systems based on induction motor has been proven over the years. To increase efficiency, maintainability, reliability and dynamic performance, modern electric traction systems use induction motors. In this context, this paper is focused on the performance determination and analyzing of the rotor flux-oriented control with voltage control, on an induction traction motor, in order to its implementation. First, the control scheme and calculation fundamentals are presented. Then, after the entire system modeling in Simulink environment, the performance of the system in dynamic and steady-state regimes is presented by simulating the operation of a real traction motor at three prescribed speeds. Also, a specific phenomenon of electric locomotives, respectively the passage of wheels over the rail joints is taking into account. To determine the performances, the results obtained from simulating the acceleration process by prescribing a speed ramp, operating in steadystate regime and then electric braking with energy recovery were analyzed. In the paper, three situations are presented: the prescription of a low speed, about 20 % of the nominal speed, of the highest speed, about 160 % of the nominal speed and of three speed steps up to the nominal speed. The analysis of the results leads to the conclusion that there is a very accurate tracking of the prescribed quantities by the actual quantities controlled in the system. It is also highlighted that the peak current through the motor does not exceed its limiting value and the switching frequency of the power electronic devices is lower than the imposed limit value. The analysis of steady-state performance shows that it is also very good.

**Cuvinte cheie:** *control orientat după fluxul rotoric, motor cu inducție, sistem de tracțiune, control al vitezei.* 

**Keywords:** *rotor flux-oriented control, induction motor, traction system, speed control.* 

#### I. INTRODUCTION

It is well known that the principle of the field-oriented control proposed by Blaschke [1] starts from the goal of obtaining a behavior of the induction motor similar to that of the DC motor. Explicitly, this means that the torque and the magnetic flux are decoupled. In this way, the magnetic flux and the torque developed by the motor are controlled independently, and the synthesis of the regulators in the structure of the control scheme is more precise, because it operates with quantities whose steady-state values are constant. Most contributions in the literature over time are related to improving performance and also to the implementation of the controls strategies [2]–[20].

The main specific requirements for electric traction systems with induction motors refer to: high power, speed regulation in a wide range (from about 2% to (160-200)% of the nominal frequency), high starting torques, but with a sufficiently low increase in speed, to avoid the phenomenon of slipping, efficient braking with kinetic energy recovery and high reliability.

In order to respond to these requirements and to have a less complicated control, the rotor flux-oriented control is often adopted. It involves working in the (d, q) rotating reference frame, whose d-axis is aligned with the vector of rotor flux. In this way, the control of rotor flux is performed by means of the d-axis stator current of the motor, while the control of the torque is achieved through the q-axis stator current. Moreover, it is allowed to extend the speed variation range well above the nominal speed by weakening the field [4]–[7].

Regarding the type of adopted control (direct or indirect), it is specified that the most frequently used is indirect control [8]-[14]. Thus, the rotor flux and its position angle are calculated using some easily measurable quantities, using the induction motor model and it is not necessary to place sensors in the rotor circuit (as is specific to direct control).

The existence of four controllers (speed controller and active current controller in the first path and, respectively, rotor flux controller and reactive current controller in the second path) is specific to voltage control of the system. Among the modulation strategies adopted in the inverter control, sinusoidal modulation stands out [12], [15], [16].

Among the application areas of the field-oriented control, electric traction is also found [5], [12]–[18].

It should be noted that, in the case of railway traction, the control system must manage the behavior of the system when the wheels pass over the rail joints, so as to minimize the effects that occur. There are very few approaches in the literature on this topic. [19], [20].

This paper deals with the implementation of the rotor flux-oriented control with voltage control on a traction induction motor used in Romanian locomotives. Next, the structure of the paper is as follows. The adopted control scheme and the related calculation bases are the subject of Section II. The MATLAB SIMULINK implementation of the control algorithm is presented in the next section. Then, the performance of the traction system with rotor flux-oriented control is analyzed both in the dynamic regime and in the steady-state regime. Thus, the results obtained from simulating the acceleration process by prescribing a speed ramp, operation in steady-state regime and then electric braking with energy recovery, in three cases are presented, as follows: prescribing of a low speed (about 20 % of the nominal speed); prescribing speed up to nominal value in three steps and braking in two steps; prescribing of the highest speed (about 160 % of the nominal speed). Finally, some conclusions are formulated.

#### II. THE CONTROL SCHEME AND CALCULATION FUNDAMENTALS

As illustrated in Fig. 1, there are two cascade control paths in the control system. Also, each o them contains two controllers, so that, there are four controllers.

In the first control path (of the speed and active current), the speed controller  $(R_{\omega})$  provides the prescribed active current  $(l_{sq}^*)$  and the active current controller  $(R_{iq})$  provides the quantity  $\mathbf{u}_{sq}^*$ , which is the prescribed voltage on the q axis of the system synchronous with the rotor flux vector.

In the second control path (of the rotor flux and reactive current), the flux controller  $(R_{\Psi})$  provides the prescribed reactive current  $(i_{sd}^*)$ , and the reactive current controller  $(R_{id})$  provides the quantity  $u_{sd}^*$ , which is the prescribed voltage on the d axis of the system synchronous with the rotor flux vector.

A two-step conversion of the prescribed quantities  $\mathbf{u}_{sq}^*$  and  $\mathbf{u}_{sd}^*$  follows. In the first stage (block B1), the conversion is from rotating coordinate system (d, q) to the fixed coordinate system ( $\alpha$ ,  $\beta$ ).

In the second stage (block B2), the conversion is from fixed coordinate system ( $\alpha$ ,  $\beta$ ) to the three-phase system (quantities  $u_a^*$ ,  $u_b^*$  and  $u_c^*$ ). Then, the sinusoidal modulator (PWM block) performs the synthesis of the control signals for the inverter transistors.

On the feedback path, the three stator currents  $i_{sa,b,c}$  measured with current transducers  $T_i$  are transformed (in block B3), from the three-phase system (a, b, c) to the fixed coordinate system ( $\alpha$ ,  $\beta$ ) characterised by the components  $i_{s\alpha}$  and  $i_{s\beta}$ . Specific relationships are used for the transformation from one coordinate system to another [20].

As shown in Fig. 1, using the measured currents and the measured speed  $^{co}$  with the speed transducer  $T_{oo}$ , the rotor flux  $\Psi_{r'}$  and its position angle  $\lambda$  between the rotating co-

ordinate system (d, q) and fixed coordinate system ( $\alpha$ ,  $\beta$ ) are calculated in block B4.

The rotor flux components in the fixed coordinate system  $(\alpha, \beta)$  are calculated as follows [20]:

$$\psi_{r\alpha} = \int \left( -p \omega \psi_{r\beta} - \frac{1}{T_r} \psi_{r\alpha} + \frac{L_m}{T_r} i_{s\alpha} \right) dt ; \quad (1)$$

$$\psi_{r\beta} = \int \left( p \,\omega \psi_{r\alpha} - \frac{1}{T_r} \,\psi_{r\beta} + \frac{L_m}{T_r} \,i_{s\beta} \right) dt; \qquad (2)$$

Thus, the calculated rotor flux is expressed as:

$$\left|\psi_{r}\right| = \sqrt{\psi_{r\alpha}^{2} + \psi_{r\beta}^{2}}; \qquad (3)$$

In the above expressions,  $T_r = L_r/R_r$ ,  $R_r$  and  $L_r$  are the resistance and inductance on a rotor phase referred to the stator,  $L_m$  is the magnetization inductance, p is the number of pole pairs, and the angle  $\lambda$  is determined by trigonometric functions [20]:

$$\cos(\lambda) = \frac{\psi_{r\alpha}}{\left|\frac{\psi_r}{\mu}\right|} = \frac{\psi_{r\alpha}}{\sqrt{\psi_{r\alpha}^2 + \psi_{r\beta}^2}}; \qquad (4)$$

$$\sin(\lambda) = \frac{\psi_{r\beta}}{\left|\frac{\psi_r}{\mu}\right|} = \frac{\psi_{r\beta}}{\sqrt{\psi_{r\alpha}^2 + \psi_{r\beta}^2}};$$
(5)

The transformation of the stator currents from the stationary reference frame  $(\alpha, \beta)$  to the rotating reference frame (d, q) is achieved in block B5..

Finally, the sinusoidal modulation principle is implemented in the PWM modulator block, having as inputs the prescribed phase voltages.

## III. MATLAB-SIMULINK IMPLEMENTATION OF THE CONTROL ALGORITHM

A specific Matlab-Simulink model has been created to determine the dynamic and steady-state performance of the system (Fig. 2). In addition to the main blocks, there are auxiliary blocks used to determine the quantities of interest.

The power part in Fig. 2 consists of the DC source equivalent to the intermediate circuit, the voltage inverter and the squirrel-cage induction motor.



Fig. 1. Rotor flux and speed control scheme of a traction induction motor, based on rotor field-oriented control, with PWM modulator and voltage control.



Fig. 2. Traction system Simulink model for testing the voltage control algorithm with sinusoidal PWM modulator and prescribing the rotor flux and speed.

In the control part, there are the rotor flux controller, the speed controller, the block for calculating the prescribed voltages on the motor phases from the stator current components in the rotor field oriented system "Transf\_R-S", the three-phase sinusoidal modulator which also implements the voltage controllers for the three motor phases and the "Ts\_Railway Joints" block which implements the resistant torque. There are also the current controllers and the three-phase sinusoidal modulator (block "Three-phase Suboscillating Modulator").

In the part of feedback control (block "Calculation isd, isq, psi\_sd, psi\_sq, Lambda), the main quantities are calculated by estimation. Also, the estimated magnetization inductance is used, which is obtained from the motor model based on the voltage – current dependence at idle conditions

There is a part for the calculation of steady-state energetic indicators (block "Calc. steady state parameters" which calculates and displays the phase voltage, the fundamental RMS current, total RMS current), harmonic distortion factor (*THD*) factor of the current through the motor and the apparent power).

Some quantities, such as the efficiency, power factor and active power related to the induction motor, are calculated and displayed in the block "Calc\_En\_IM". The three-axis oscilloscope "Te-Ts-n" displays the average torque, the resistant torque and the motor speed.

It should be noted that the speed controllers, as well as other calculation and display blocks, are validated by two signals obtained as functions of time, in the blocks at the top left of the figure. So, the speed and torque prescription takes place only after the motor is magnetized. Also, the rotor flux has the prescribed value.

A few more clarifications need to be made, as below.

The used controllers are of the proportional-integral (PI) type.

The rotor flux and the motor speed are prescribed as ramp signals.

For the limitation of the current, the outputs of the controllers are not limited to the input level (+10), but +5 V for the flux controller and +9.2 V for the speed controller. In order to ensure the compatibility of the algorithm with the DSP system used on locomotives (Texas Instruments TMS320F28335), the Euler 1 integration method and the  $40\mu s$  step (Ts) were used.

The sinusoidal PWM modulator operates with a switching frequency of 1 kHz.

In order to allow the overmodulation to achieve the nominal power at 100 Hz, the amplitude of the modulating (triangular) signal is 2.25 V, which also minimizes the current distortion and limits the noise.

#### IV. PERFORMANCE OF THE TRACTION SYSTEM

Table I and Table II illustrate the main parameters of the traction induction motor (a traction motor used in Romanian locomotives) and of the PI controllers. The value of the DC-voltage at the IGBT-based voltage source inverter input is 1800 V.

The motor start-up at constant resistant torque (T<sub>s</sub>) and nominal rotor flux for speed below the rated value and diminished over the rated value, in case of three prescribed ramp speeds (248 rpm, three speed steps up to the nominal speed and maximal speed of 1983 rpm), are analyzed. Also, the periodic passage of wheels over rail joints was considered. For this, the load torque variation shown in Fig. 3 is considered, with the following parameters:  $T_{sp}$ = 0.25· $T_{s}$ ;  $T_{sn}$  = -0.5· $T_s$ ;  $T_{mj}$  = 4 s; dt1 = 0.05·Tmj; dt0 = 0.1·dt1; dt2 = dt1.

 TABLE I.

 MAIN PARAMETERS OF THE TRACTION INDUCTION MOTOR

V <sub>N</sub> (V)	<i>P</i> <sub>2N</sub> ( <b>kW</b> )	$f_{1N}$ (Hz)	<i>I</i> <sub>N</sub> (A)	р	$\mathbf{S}_N$	<i>n</i> <sub>N</sub> ( <b>rpm</b> )
1400	1150	62.5	576	3	0.0104	1237
$T_N$ (N·m)	$\begin{array}{c} R_1 \\ (\Omega) \end{array}$	$L_{\sigma 1}$ (mH)	$R_2$ ( $\Omega$ )	$L_{\sigma^2}$ (mH)	$R_{\rm m}$ ( $\Omega$ )	$L_{mN}$ (mH)

 TABLE II.

 PARAMETERS OF THE PI CONTROLLERS

$K_{p\Omega}$	$T_{i\Omega}$	$K_{p\psi}$	$T_{i\psi}$	$K_{pid}$	$T_{i\mathrm{id}}$	K <sub>piq</sub>	T <sub>i</sub> i
100	0.004	10	0.04	0.3	0.1	0.15	0.6



Fig. 3. Load torque during the passage of wheels over rail joints.

The evolution of several quantities (speed, rotor flux, electromagnetic torque, phase current, current and voltage components in the rotating reference frame, control and switching frequency) are analyzed. Also, different quantities specific to the steady-state operation and energetic indicators are given for each case study.

#### A. Case of Prescribed Speed of 248 rpm and $T_s = T_N$

When a low ramp speed of about 20% of the nominal speed (248 rpm) is prescribed at nominal resistant load, Fig. 4 - Fig. 14 illustrate the good response of the system.

It can be seen that the motor speed is very close to the prescribed value and, at the end of acceleration, the motor torque is very close to the load torque (Fig. 4). The speed is very little affected by the appearance of the rail joints, only  $\pm 2\%$  (Fig. 4 and 5).

The rotor flux also faithfully follows the prescribed value (Fig. 6).

The  $i_{sd}$  current has a maximum for fast motor magnetization (Fig. 7), then become constant. The active current  $(i_{sq})$  follows the evolution of the torque (Fig. 8). There are some current increases (about 1100 A, but maxim admissible value is 1200 A) related to the appearance of the rail joints (Fig. 9).



Fig. 4. Time evolution of the average torque (black), resistant torque (red) and speed (blue) at the prescribed speed of 248 rpm during the whole process.



Fig. 5. Details in fig a) and b) regarding the evolution of the average torque (black), resistant torque (red) and speed (blue) during the passage of a wheel over a rail joint at the prescribed speed of 248 rpm.



Fig. 6. Rotor flux (prescribed – in red and actual – in black) at the prescribed speed of 248 rpm.



Fig. 7. d-axis stator current (prescribed – in red and actual – in black) at the prescribed speed of 248 rpm.



Fig. 8. q-axis stator current (prescribed – in red and actual – in black) at the prescribed speed of 248 rpm.



Fig. 9. Stator currents on the three phases at the prescribed speed of 248 rpm: a) during the whole process; b) detail on an interval of 1.2 s.

The resulted d and q components of the stator voltage are illustrated in Fig. 10 and Fig. 11, and show that there are low influences between them. As shown in Fig. 12 and Fig. 13, the switching frequency is variable, but it average value does not exceed 1 kHz. Fig. 13 shows the increase of the stator voltage frequency up to the value of about 13 Hz.

Table III synthesizes the dynamic performance of the system.  $\delta T$  and  $\delta n$  are the torque and speed variations. Table IV summarizes the steady-state performance. A low power factor (*PF*) of 37 % and a relatively low value of the current harmonic distortion (*THD*<sub>1</sub>) of 7 % are highlighted.



Fig. 10. d-axis component of the stator voltage at the prescribed speed



Fig. 11. q-axis component of the stator voltage at the prescribed speed



Fig. 12. Evolution over time of the average switching frequency at the prescribed speed of 248 rpm.



Fig. 13. The average switching frequency (averaged over 10 ms), at the prescribed speed of 248 rpm.



Fig. 14. The control frequency, at the prescribed speed of 248 rpm.

 TABLE III.

 Dynamic Performance at the Prescribed Speed of 248 RPM

Ts-	T <sub>s+</sub>	δT	δ <i>n</i> +	δ <i>n</i> -	I <sub>maxacc.</sub>	I <sub>maxsteady-st.</sub>	f <sub>sw_max</sub>
(%)	(%)	(%)	(%)	(%)	(A)	(A)	(kHz)
-50	25	100	2	2	1100	1160	1

TABLE IV.
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STEADY-STATE PERFORMANCE AT THE PRESCRIBED SPEED OF 248 RPM

V	Ι	Imagn.	$P_1$	$P_m$	S	PF	η	$THD_{I}$
(V)	[A]	(A)	(kW)	(kW)	(kVA)	(%)	(%)	(%)
398	596	209	260	225	701	37	86.73	7

#### *B.* Prescribing Speed up to Nominal Value in Three Steps and Braking in Two Steps

For prescribing speed step by step up to the nominal value, also with brake, the waveforms of the main quantities are illustrated in Fig. 15 - Fig. 24. (the passage of wheels over the rail joints is neglected). First, new information is offered by d-axis stator current (Fig. 15) that, after 10 seconds decreases, although the rotor flux remains constant (Fig. 16). After the brake control is given, the current increases and reach the rated value. The speed follows the prescribed values both during the accelerating and braking regimes (Fig. 17). The q-axis stator current (Fig. 18) has the constant values in steady-state, increases in the time of acceleration and decreases in braking time. It is significant that, it does not have negative values, similar to the torque (Fig. 19). It means that the motor does not operate in braking regime and the decreasing of the speed is caused by the resistant torque only.

The phase currents of the motor are under 1000 A on the first two steps and exceed a little 1000 A at the nominal speed (Fig. 20). The average switching frequency is 1 kHz for speed under 1000 rpm, but decreases to 400 Hz when the speed is over 1000 rpm (Fig. 21).



Fig. 15. Time evolution of stator current along the d-axis (prescribed in blue and actual in red) when speed is stepped up to nominal speed and braking.











Fig. 18. The time evolution of the q-axis component of the stator current (prescribed in blue and actual in red), when prescribing the speed in steps up to rated speed and braking.



Fig. 19. Evolution over time of the average torque (black) and speed (blue), when prescribing the speed in gears up to nominal speed and braking



Fig. 20. Time evolution of the currents on the 3 phases, from speed prescription in steps up to nominal speed and braking.



Fig. 21. The time evolution of the switching frequency averaged over 10 ms, when prescribing the speed in steps up to nominal speed and braking.

The time evolution of the stator voltage on the component along the d-axis is similar to that of the stator current of the same axis (Fig. 22), because the voltage determines the current.

Because the rotor flux is constant, the time evolution of the stator voltage component on the q-axis is similar to that of the torque (Fig. 23).

The form of the cosin of the rotor flux position angle (Fig. 24-a) proves a good orientation of the calculation system. The control frequency follows the speed as form (Fig. 24-b). It means that the slip is very low.

## C. Case of Prescribed Speed of 1983 rpm and $T_s=0.625 \cdot T_N$

For the highest speed (1983 rpm), the waveforms of the main quantities are illustrated in Fig. 25 - Fig. 37 and the performance summary is presented in Table V and Table VI. The resistant torque is diminished so that, over the nominal speed, the power to be at nominal value.



Fig. 22. The time evolution of the component along the d-axis of the stator voltage, in the case of stepped speed prescription up to rated



Fig. 23. The time evolution of the component along the q-axis of the stator voltage, in the case of stepped speed prescription up to rated speed and braking.



Fig. 24. The evolution over time of: a) the cosine of the angle of the rotating system; b) the control frequency, in the case of prescribing the speed in steps up to nominal speed and braking.

The prescribed speed is followed very well by the real speed and the effects of passing the joints are diminished (Fig. 25). The details from Fig. 26 and Fig. 27 underline some aspects. Thus, the motor torque does not follow the resistant torque, and its variation represents less than 50% of resistant torque variation. The total and the d-axis rotor flux are diminished over nominal speed and have a hyperbolic evolution (Fig. 28 and Fig. 29). Also, the q-axis rotor flux is null because the calculation system is well oriented (Fig. 29). The d-axis stator current decreases when the speed increases, especially over of the nominal speed (Fig. 30). At nominal speed, the d-axis stator current becomes near zero. The q-axis stator current increases when the speed exceeds the nominal value and magnetic flux decreases (Fig. 31). The real current does not have variations when the wheels passing over the joints.



Fig. 25. Time evolution of the average torque (black), resistant torque (red) and speed (blue) at the prescribed speed of 1983 rpm.



Fig. 26. Detail regarding the evolution of the average torque (black) and resistant torque (red) during the passage of a wheel over a rail joint at the prescribed speed of 1983 rpm.







Fig. 28. Rotor flux (prescribed – in red and actual – in black) at the prescribed speed of 1983 rpm.



Fig. 29. Components of the rotor flux on the two axes (d-axis rotor flux – in blue and q-axis rotor flux – in black) at the prescribed speed of



Fig. 30. d-axis stator current (prescribed - in red and actual - in black)



Fig. 31. q-axis stator current (prescribed – in red and actual – in blue) at the prescribed speed of 1983 rpm.

This aspect is showed better by phases current (Fig. 32). There is a little diminution of the current when the prescribed speed is obtained and the acceleration torque disappears. The d-axis component of the stator voltage evolves similarly to that of the corresponding current (Fig. 33). It should be noted that the final values are negative (about -250 V).

Also, the q-axis component of the stator voltage evolves similarly to that of the active current (Fig. 34). After the nominal value of the supply voltage is obtained, the switching frequency decreases because the phase voltage is more little modulated (Fig. 35). The control frequency increases to 100 Hz, larger of 1.6 time from the nominal value (Fig. 36).

The magnetizing inductance and current (Fig. 37), shows that the inductance increases a little although the current decreases from 180 A to 100 A. This means that the motor is not saturated.



Fig. 32. Stator currents on the three phases at the prescribed speed of 1983 rpm: a) during the whole process; b) detail during the passage of a wheel over a rail joint.



Fig. 33. d-axis component of the stator voltage at the prescribed speed of 1983 rpm.



Fig. 34. q-axis component of the stator voltage at the prescribed speed of 1983 rpm.



Fig. 35. The average switching frequency (averaged over 10 ms), at the prescribed speed of 1983 rpm.



Fig. 36. The control frequency, at the prescribed speed of 1983 rpm.



Fig. 37. Time evolution of: a) magnetizing inductance; b) magnetizing current, at the prescribed speed of 1983 rpm.

#### V. POSSIBILITY OF IMPLEMENTATION

The d-axis stator current decreases with increasing speed although the rotor flux is constant. This aspect does not respect the physical phenomenon and the operating equations of the induction motor in the rotor flux oriented system [12]:

$$0 = R_r i_{rd} + \frac{d}{dt} \Psi_{rd}; \tag{6}$$

$$\Psi_{rd} = L_r i_{rd} + L_m i_{sd}. \tag{7}$$

Equations (6) and (7) show that, if  $\psi_{rd}$ =constant, two very important aspects can highlight the correctness of the algorithm:  $i_{rd} = 0$ ;  $i_{sd} = \psi_{rd}/L_m = \text{constant}$ , because  $\psi_{rd} = \text{constant}$  also involves  $L_m = \text{constant}$ .

It is found that, even at very high speed, the performance is good (Tab. V and VI).

 TABLE V.

 Dynamic Performance at the Prescribed Speed of 1983 rpm

<i>T</i> <sub>s-</sub>	T <sub>s+</sub>	δT	δ <i>n</i> +	δ <i>n_</i>	I <sub>maxacc.</sub>	I <sub>maxsteady-st.</sub>	f <sub>sw_max</sub>
(%)	(%)	(%)	(%)	(%)	(A)	(A)	(kHz)
-50	25	75	0.5	0.23	1100	1025	1

TABLE VI
STEADY-STATE PERFORMANCE AT THE PRESCRIBED SPEED OF 1983 RPM



Fig. 38. Time evolution of: a) current; b) magnetic flux, at the prescribed speed of 1983 rpm.

It is appreciated that the appearance is determined by the calculation error, especially when calculating the position angle of the rotor flux. There are two ways to reduce calculation errors: using a smaller integration step and using an integration method of order 2 or higher. The first modality was verified for  $Ts=5\mu s$  (Fig. 38). But, this sampling period can't be used for real-time implementation, because it is too small.

In same time, the dSPACE 1103 prototyping board, used for the code generation operates by first order integration method, as the used processor too. Consequently, the research activity will continue with the development, implementation and testing of a calculation sequence that details the 2nd order integration. Only then, will tests be carried out on the physical model.

#### VI. CONCLUSIONS

Following the analysis of the energetic and dynamic performance of the traction system with rotor fluxoriented control in the acceleration, steady state and braking regimes, few aspects can be highlighted.

The evolution of all quantities, excepting d-axis stator current, respects the physical phenomenon and the operating equations of the motor.

The current absorbed by the motor is higher than the nominal value, by about 3% at the low prescribed speed and by about 10% at the highest prescribed speed.

The power factor is significantly lower than in sinusoidal supply conditions and increases with speed. The efficiency has very good values (86.73 % to 97.3 %). *THDI* has a low value (7%) at low speed and increases at high speed (17.23%).

The maximum average switching frequency is below the maximum allowed value of 1100 Hz.

When passing over the rail joints, the torque accurately follows the variations of the resistant torque, so the speed variations are very small. In conclusion, the rotor fluxoriented control leads to good performance when applied to the traction induction motor.

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# The Implementation of Quasi-Oriented Rotor Field Control Algorithm for Induction Motor Traction Systems

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Abstract - A rotor field quasi-oriented control algorithm for drive systems equipped with a low cost DSP system has been implemented and experimentally validated in this paper. The system is designed for induction motor traction systems. The control algorithm implementation was done for a dSPACE DS1103 prototyping board in the Matlab Simulink environment. The control algorithm principle is based on the assumption that the current system is oriented by the rotor flux and the real flux follows the imposed value. Therefore, the rotor flux estimation is no longer necessary as well as the flux control loop. This makes the control algorithm suitable for low cost industrial DSP applications. The drawback is given by the fact that the current projections on the dq axes cannot be regulated, the current control being done by means of the current phasor magnitude. The control system implementation was validated with good results, in terms of current and speed regulation especially at variable mechanical load (such as track joints or switches), making this method suitable not only for low cost traction systems, but especially for railway vehicles.

**Cuvinte cheie:** *control cu orientare după câmp, motor de inducție, sistem de tracțiune.* 

**Keywords:** *field oriented control, induction motor, traction system.* 

#### I. INTRODUCTION

The induction motor is the standard motor used in general modern electric drive systems [1]. The same is valid for electric traction systems used in modern electric vehicles and especially in high power railway systems [2]. As catenary fed vehicles are not power limited by the battery, the traction system power can be very large. For these traction systems the field oriented control is an advantageous choice as it offers high torque at low speed and high dynamic performance [3]. A particularity of traction systems is the use of the traction motor above the rated speed. In this regime, the field oriented control gives good performance, given the fact that it implies higher frequency of the motor currents [1]. This is important because high power drive systems imply relatively low switching frequency (which affects the performance of the control system). At the same time, the motor current relatively high frequency needs a fast and high computing power DSP to execute the control algorithm in real time. Because the rotor field oriented control requires high computation power, unavailable for standard industrial DSPs, the aim of this paper is the implementation of a simplified, quasioriented rotor field control algorithm on a dSPACE DS1103 prototyping board. The experimental validation of the implemented control algorithm was done on an experimental railway traction stand.

#### II. ROTOR FIELD QUASI-ORIENTED CONTROL ALGORITHM

For the rotor field oriented control the motor currents are expressed in the dq referential which rotates with the rotor flux [1-8]. To avoid the estimation of the flux, assuming that the current phasor is oriented and the real rotor flux correctly follows the imposed flux, the real time determination of the flux phasor can be avoided. Therefore, the flux control loop, which gives the magnetizing current on the d axis of the referential, can be also avoided. This can lead to an important simplification of the control algorithm, effective for drive systems that use less powerful DSPs [9].

The control algorithm for traction systems with simplified rotor field orientation (quasi-orientation) is synthetized in [9], where [8]:

- $R\omega$  speed controller;
- current magnitude computation block;
- R<sub>i</sub> current controller;
- VSI voltage source inverter;
- - three phase sinusoidal control signal generator;
- $f_2^*$  slip frequency computation block.

The algorithm inputs are the imposed rotor flux and the imposed speed. For the given flux value, assuming as stated that the system is oriented, the value of the imposed magnetizing current is obtained [8][9]:

$$i_{sd} = \frac{1}{L_m} \cdot \left( \Psi_{rd} + T_r \cdot \frac{d\Psi_{rd}}{dt} \right), \tag{1}$$

where [8]:

-  $T_r$  - rotor circuit time constant:

$$T_r = \frac{L_r}{R_r} = \frac{L_m + L_{\sigma r}}{R_r} , \qquad (2)$$

-  $L_m$  - magnetizing inductance;

- $L_{\sigma r}$  rotor leakage inductance;
- $R_r$  rotor resistance.



Fig. 1. Rotor field quasi-oriented control algorithm.

The imposed active current is obtained at the output of a proportional integrative speed controller [9].

The two currents are further used to compute the imposed motor current phasor magnitude (Fig. 1) used for the current control loop (which also uses a proportional integrative controller) [9]:

$$I_{m}^{*} = \sqrt{i_{sd}^{2} + i_{sq}^{2}}$$
(3)

The feedback information is the real motor current phasor magnitude, computed from the measured three-phase currents [9]:

$$I_m = \sqrt{i_{s\alpha}^2 + i_{s\beta}^2} \tag{4}$$

where,  $i_{\alpha}$  and  $i_{s\beta}$ , are the stator current phasor projections on the stationary reference frame [1], [6].

The output of the current magnitude controller is the amplitude modulation index ("ami" in Fig. 1) of the three phase sinusoidal control signals generated and applied to the PWM modulator. The frequency of these signals is obtained by summing to the motor shaft frequency, the imposed slip frequency [8]:

$$f_2^* = i_{sq} \cdot \frac{L_m}{2 \cdot \pi \cdot \Psi_{rd} \cdot T_r} \tag{5}$$

#### III. DS1103 IMPLEMENTATION OF ROTOR FIELD QUASI-ORIENTED CONTROL ALGORITHM

The rotor field quasi oriented control algorithm was implemented for the experimental traction system in the Matlab Simulink environment. This was possible because the control subsystem of the traction system is made entirely by the dSpace DS1103 prototyping board, which can be programmed using this graphical programming environment. Of course, the experimental stand includes additional electronic circuits, but used only for auxiliary purposes (such as initialization, overcurrent/overvoltage protection, gate drivers, etc.), the control algorithm being completely implemented by the prototyping board. Because the experimental stand is used to test a complete traction system it also includes the single phase rectified which feeds the traction inverter. It must be mentioned that the rectifier (which is a PWM boost rectifier) is not controlled by the prototyping board, but by a dedicated DSP (as it does not make the object of this study).

As illustrated in the prototyping board control model (Fig. 2), the control algorithm (Fig. 3) is grouped into a subsystem, for better clarity.

The link between the power system and the control algorithm (model) is done by the power section transducers (voltage, current and speed) and the DS1103 acquisition system. Each analog to digital converter (connected to a transducer) have a corresponding Simulink block (from the Real Time Interface library) which outputs a Simulink signal, further de-normalized (with the appropriate Gain blocks). Some of the converted signals are not needed for the control algorithm, but only for system monitoring purposes.

The speed transducer which was used is an incremental encoder therefore it was connected to the DS1103 incremental encoder specialized input. This gives at the output port of DS1103ENC\_POS\_C1 the number of increments further used to obtain the speed. Because of the relatively low number of pulses, the sliding window averaging technique was used to smooth the speed.

The output of the control algorithm (the IGBT gating signals) are obtained by applying the three phase generated control signals to the DS1103 hardware PWM modulator (by means of the DS1103SL\_DSP\_PWM3 block).

The actual control algorithm is illustrated in Fig. 3. It is also grouped in several subsystems for better understanding of the model, given the algorithm typical sections:

- The motor current magnitude computed from the acquired three phase motor currents (in the αβ referential);
- The d axis (magnetizing) current obtained as a function of the imposed motor flux (for the assumption that the system is oriented);



Fig. 2. The DS1103 Simulink control model.

- the slipping frequency, from the motor speed and motor equations;
- The imposed active stator current in the dq referential (i<sub>sq</sub>).

The active current, is, is imposed in two manners:

- Manual mode the motor torque is directly imposed, from which the active current is computed.
- Automated mode the active current is imposed at the output of the speed controller.



Fig. 3. The rotor field quasi-oriented control algorithm.