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Direct Torque Control of Induction Motors with Stator Flux Correction Applied to the Low-floor Tramcars

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Original scientific paper

The paper presents a modified direct torque control (DTC) method of an induction motor, which is applied in the main drive of the low-floor tramcars series TMK2200 manufactured by the Croatian consortium CROTRAM – *Končar* d.d., TŽV *Gredelj* d.d. and *Đuro Đaković* d.d. The modification consists in the correction of stator flux vector by means of appropriate stator current component. Simplicity of the basic DTC is maintained because the only required machine parameter is stator resistance, yet the method enables stable operation in the whole speed range (including standstill). In comparison with other DTC schemes, the proposed method is more reliable because the rotor parameters are not used and the rotor quantities are not calculated. The method has been fully verified through exploitation of vehicles, and the simulation results are compared with the experiments on the laboratory model of the drive.

Key words: tramcar, induction motor, direct torque control, estimation

1 INTRODUCTION

The most important task of induction motor vector control in tramcars is to ensure required traction and braking characteristics. Also, it has to insure safe and reliable operation and comfortable drive. Generally, tramcar drives have very wide regulation range regarding speed and load. For example, in tramcars series TMK2200 at maximum speed of 70 km/h, motor speed is 250 % of rated value (the vehicle has to be tested also at 10 % higher speed, i.e. 77 km/h) and in some operation regimes motor load is 230 % of rated one. The required traction and braking characteristics are presented in Figures 1 and 2. Figure 1 shows diagrams of the tractive effort (F) vs. vehicle speed (v) and figure 2 shows diagrams of the normalized torque of induction motor (M/M_n) vs. normalized speed of the motor (n/n_n), where M_n and n_n are rated torque and motor speed, respectively.

With regard to such a wide speed and torque range in which induction motor operates, it is desirable to choose very simple and reliable vector control method which uses as less as possible machine parameters. It is especially desirable to com-

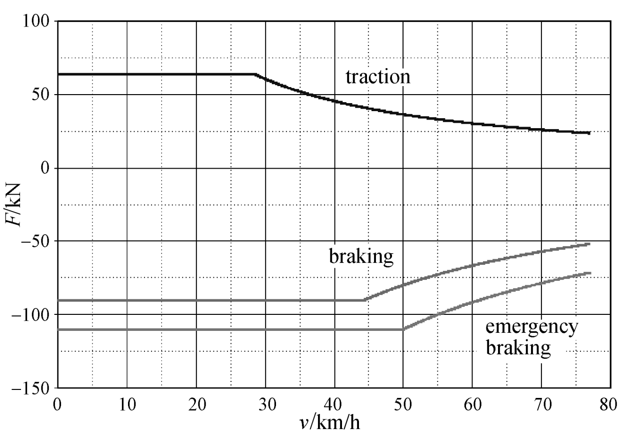


Fig. 1 Tractive effort diagram of tramcar series TMK2200

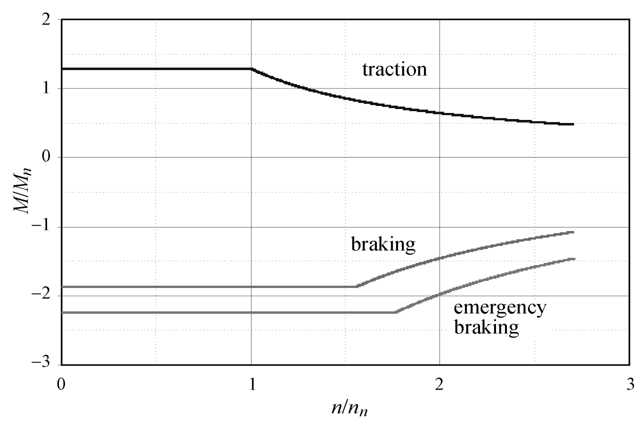


Fig. 2 Traction and braking torques of induction motor in tramcar series TMK2200

pletely avoid usage of rotor parameters because they are very machine temperature and load dependent and they can not be measured during operation. Such a method is the DTC method that has its basics in [1, 2, 3]. It has electromagnetic torque and stator flux regulation loops, hysteresis regulators and it does not require any current regulators and related coordinate transformations. The torque and stator flux are estimated only from measured stator quantities and stator resistance parameter.

A method described in [1, 2] is so called direct self-control (DSC). It is aimed firstly for high power drives with very low switching frequency, for instance electric locomotives [4, 5, 6]. An interesting fact is that in these papers more complex regulation algorithm is used in the low speed range (from standstill up to approximately 20–30 % rated speed as is mentioned in [7]). This algorithm requires knowledge of magnetizing inductance, stator leakage inductance, rotor leakage inductance and rotor resistance, so, all machine parameters are used. The reason for this is minimum thyristor conduction interval (the voltage that is small enough can not be supplied to the motor). In [6] is only mentioned, without any explanation, that the use of more complex model for stator flux estimation is necessary at low speeds. The DTC method for drives which are not limited with switching frequency is presented in [3]. On the basis described in this paper, after approximately 10 years, whole series of inverters are developed and commercialized ([8, 9, 10]). But the problem with stable operation at low speeds still was not solved. In [3] is mentioned, to be quoted: »… integrators cause some drift and operation errors at low-velocity regions. The reason for this is that the operation of integrators can not be done perfectly at zero velocity because of no induced electromotive force in the motor, so that control of Ψ_1 (stator flux) might be unstable when R_1 (stator resistance) deviates from the correct value. These problems are negligible at relatively high velocities such as above 2 Hz.« Therefore, the stator flux is at low speeds estimated by means of stator inductance, main inductance and stator and rotor currents. In [8] is emphasized that stator flux estimation in such a way (with current correction feedback), to be quoted: »… highly improves the stator flux estimate, especially at low speed.«, but it was not explicitly mentioned up to which speed and frequency this feedback should be used and does it or not depend on motor load.

In this paper we present a modification of the direct torque control method based on the stator flux correction. The method is extension of our previous work presented in [11, 12]. The method reveals all characteristics of the basic DTC method but it insures stable and reliable operation in the whole speed region, including start-up from standstill. The modification of the stator flux is made by means of appropriate component of the stator current. In this way, the rotor parameters are not used and the rotor quantities do not have to be calculated.

The paper is organised as follows: the second section presents the related work in the field, the third section presents the basic DTC method and proposed modifications and the forth section gives results of simulation and experimental testing.

2 RELATED WORK ON DIRECT TORQUE CONTROL

The basic DTC method introduced in [3] became the most researched vector control of induction motors in the middle nineties of last century. There are a lot of papers which analyze some DTC aspects and most important of them are briefly analysed below.

Apart from the basic DTC principle, some derivations are summarized in [7]. A DTC method which uses switching table, the DSC method, and the DTC method with constant switching frequency are described in detail. The method with switching table is defined as a basic, and it is pointed out that many modifications of the basic DTC scheme are aimed at improving starting conditions and very low speed operation, torque ripple reduction, noise level attenuation, improving overload condition, increasing the number of available voltage vectors, switching frequency control and rotor flux control.

The DTC method and field-oriented vector control (FOC) are compared in [13]. Simplicity of the basic DTC scheme is emphasized because it does not use PI regulators, coordinate transformations, current regulators and special timers for PWM signals. But the DTC presents some disadvantages: difficulty to control torque and stator flux at very low speed, high current and torque ripples, variable switching behaviour, high noise level and lack of direct current control.

In [14] is described a DTC method which uses combination of so called voltage model and current model for stator flux estimation. Furthermore, an additional carrier signal (500 Hz, square wave) is delivered to torque comparator input. This injected signal improves the waveform of stator flux and current and guarantees a robust start and operation at zero speed region.

A method described in [15, 16] uses rotor flux vector as a regulation quantity instead of stator flux vector. The rotor flux is estimated by means of stator voltages and currents, all machine parameters and difference between estimated and reference rotor flux value. The proposed flux estimator uses complex gain which is function of the working conditions. It is also pointed out that the stator resistance errors and current sensor offsets significantly degrade the drive behaviour in the low speed range.

In order to minimize torque oscillations, torque changes when zero voltage vector is supplied to the motor are analyzed in [17]. Although the basic DTC scheme with hysteresis torque and flux comparators is used, in the whole speed range the switching frequency is maintained constant, and the stator flux vector and the rotor flux vector have to be estimated. So, all machine parameters must be known.

During the start-up and at low speeds the basic DTC method causes demagnetisation phenomena. A new algorithm for avoiding this phenomenon is presented in [18]. Not only the torque, but also the stator flux is kept between hysteresis bands. Only extensive simulations prove effectiveness of the proposed method.

As a result of the uneven voltage vector contribution to the control of torque and stator flux in a direct torque controlled induction machine, the stator flux magnitude drops particularly when machine is running at low speed. A simple strategy which rotates the basic DTC switching sector is presented in [19]. Thereby, stator flux drops are reduced and also current waveform is improved.

In [20] is emphasized that the delay in feedback signals (inherent to every digital system) causes that neither the torque error nor the stator flux error can be restricted within hysteresis bands. Decreasing the width of hysteresis bands can increase the switching frequency only to some level. The method proposed in the paper introduces a dithering technique by superposing triangular waves with high frequency and minute amplitude on the torque and the flux error. As a result, the torque ripple, the stator flux ripple and the acoustic noise are significantly reduced.

The influence of the torque and flux hysteresis bands to the switching frequency is analyzed in [21]. The switching frequency can be predicted from machine parameters, operating conditions and sampling time and it can be limited in order not to excess allowed value. With proposed method, it is also possible to improve current harmonics of induction motor.

The duty-cycle control scheme is proposed in [22]. The method equals the average torque and the lower torque hysteresis limit and reduces the variation in switching frequency. Torque ripples are also reduced, especially in the low speed region.

The novel DTC technique which apply voltage vector not for the whole sampling period but only for the time interval needed by the torque to reach the upper (or lower) limit of the hysteresis band is presented in [23]. By this approach, the control system emulates the operation of a torque hysteresis controller of analog type since the application time of the inverter voltage vector is dictated by the allowed torque excursion and not by the sampling period. The torque ripples are considerably reduced by predicting the motor torque one sampling period ahead.

A new direct torque and flux control based on space-vector modulation is introduced in [24]. The proposed strategy reduces the acoustical noise, the torque and the current ripples. The switching frequency is constant and controllable. But the method is more complex than basic one, it uses four PI regulators with different proportional and integral gains (some of them are dependent on operation conditions) and all machine parameters have to be known. A similar method is proposed in [25]. Instead of hysteresis regulators and switching table, PI regulators are used for the torque and the stator flux control, and the switching frequency is constant. The basic DTC method and its combination with the space-vector modulation are compared in [26]. The combination also uses PI regulators.

Although a method in [27] is described as the direct torque control with constant switching frequency, a lot of modification can be observed. The method does not use hysteresis regulators and the reference value of the stator flux vector is calculated by means of the reference value of the torque and the rotor flux. The method also utilizes coordinate transformation and space-vector modulation technique. It is emphasized that the inverter dead-time compensation highly improves the drive performance, especially at low speeds.

A sort of discrete space-vector modulation is introduced in [28]. Using a standard voltage-source inverter topology, the number of voltage vectors can be increased by subdividing the cycle period in three equal time intervals. In comparison with the basic DTC scheme, the torque and current ripple are significantly reduced but the control algorithm is more complex.

The similar approach is used in [29]. Two novel switching techniques are proposed: in the first, the nonzero and zero voltage vectors are applied to the motor, each of them half a sampling period, and in the second technique, the complex plane is divided in 12 sectors and two nonzero voltage vectors are used during the sampling period, also each half a sampling period.

The method described in [30] is based on a variable duty cycle. The appropriate voltage vector is not applied for the whole sampling period, and its duration is calculated according to the machine electromotive force.

In spite of torque hysteresis regulator, the use of PI regulator, two comparators and two triangular waveform generators is proposed in [31]. The proposed controller replaces the three-level hysteresis comparator and is capable of reducing the torque ripple and maintaining a constant switching frequency.

In the drives with the speed control, it is possible to correct the stator flux with the speed error. The method, which is especially effective in the low speed region, is presented in [32].

The switching characteristics of the DTC are assessed in [33]. The dependencies of the switching frequency on the supply angular frequency, inverter dc link voltage and the sampling period are analyzed. All theoretical results are confirmed by simulations and experiments on three different motors.

3 PROPOSED DIRECT TORQUE CONTROL METHOD FOR TRAMCAR MAIN DRIVE

From all listed papers one can conclude what are the main disadvantages of the basic DTC method (the method with possible change of switching state only in predefined time period and with the use of adequate switching table):

- the difficulties with the torque and the flux control in the low speed range are emphasized almost in all listed papers, without the explicit explanation what is the reason for such a behaviour of the drive and from which speed the DTC method operates stable and reliable;
- the variable switching frequency increases the torque pulsations, the current ripples and the noise level, and the dependence of the switching frequency is in some drives undesirable;
- sometimes is the fact that the inverter currents cannot be directly controlled very big disadvantage;
- for some drives it is undesirable that the average value of the torque does not match the reference value.

Which undesirable characteristics that are aforementioned have an important and significant influence on the tramcar operation? The torque and the flux control in the low speed range are very important because traffic conditions make the start-up from standstill practically the most frequent tramcar operation state. Regardless of the position of the drive controller, and hence regardless of the reference value of the torque, a comfortable start without any oscillations and undesirable jerks is to be achieved.

The variable switching frequency is not big disadvantage. Namely, total losses that are dissipated on the semiconductor module consist of switching and conducting losses. Switching losses depend on switching frequency and on currents and conducting losses depend only on currents, i.e. on the load. So, it is possible to control the switching frequency with the change of the hysteresis bands, in order to decrease it in the low speed range (the maximum torque references are bigger) and vice versa at higher speeds. From the point of noise, the variable switching frequency is actually desirable, because the drive in that case reveals the characteristics of the white noise. If a DTC method with constant switching frequency has to be applied, with regard to the IGBT components that are used and with regard to the available space in the vehicle, the switching frequency should not exceed approximately 2,5 kHz. Therefore, in all methods that use some kind of combination between the DTC and the space-vector modulation, the switching period has to be approximately 400 μs or more. This period is too long period for good and reliable control (in the most drives that were analyzed in the listed papers the sampling periods are up to 200 μs, so, the switching frequencies are 5 kHz or more).

A direct current control is desirable in order to avoid overcurrents during operation. However, with limitation of the maximum traction and braking torques, avoidance of fast changes of stator flux reference value and its appropriate adjustment to the speed and dc link voltage, it is possible to control the currents indirectly. Therefore, the stable operation in the whole speed region and at all conditions can be achieved without any overcurrents. The mismatch between the average and the reference torque value practically does not have any influence, because the torque needed for acceleration or deceleration of the vehicle is set by the driver.

Therefore, it can be concluded that the basic DTC is appropriate vector control method for the tramcar main drive and generally for the drives with very wide regulation range regarding speed and load. However, the stable and reliable drive in the low speed region has to be achieved. It is also desirable to avoid control schemes which use rotor quantities and rotor parameters (this would additionally increase the unreliability and the uncertainty of the drive), and to avoid the use of different control algorithms in the low and in the high speed range (this approach is more complex and makes the analysis of the drive during operation more difficult).

3.1 Basic direct torque control method

In the DTC method an electromagnetic torque and a stator flux vector are regulated quantities. The stator flux vector is defined with phase values, [34, 35]:

$$\vec{\Psi}_s = \Psi_{s1} + \Psi_{s2} e^{j\frac{2\pi}{3}} + \Psi_{s3} e^{j\frac{4\pi}{3}} = = \mathrm{Re}(\vec{\Psi}_s) + j\,\mathrm{Im}(\vec{\Psi}_s). \quad (1)$$

Ψ_{si} (i = 1, 2, 3) is the phase stator flux of the motor and can be calculated according to:

$$\Psi_{si} = \int (u_{si} - R_s i_{si}) \mathrm{d}t. \quad (2)$$

u_{si} is motor phase voltage and is defined with dc link voltage, switching state applied in observed time interval and with voltage drops on semiconductors. R_s is motor phase stator resistance and i_{si} is motor phase stator current.

The developed electromagnetic torque (the torque in the air-gap) can be calculated from estimated fluxes and measured currents:

$$m_{elm} = \sqrt{3} p (\Psi_{s1} i_{s2} - \Psi_{s2} i_{s1}) \quad (3)$$

where p is the number of pole pairs of the motor.

The torque and the stator flux can be regulated by means of an appropriate voltage vector, i.e. with adequate switching state. It is possible to apply total eight combination which result with six non--zero voltage vectors and with two zero voltage vectors (in this case the voltages of the motor are determined with the voltage drops on a semiconductors). The stator voltage vector is defined analogous to (1):

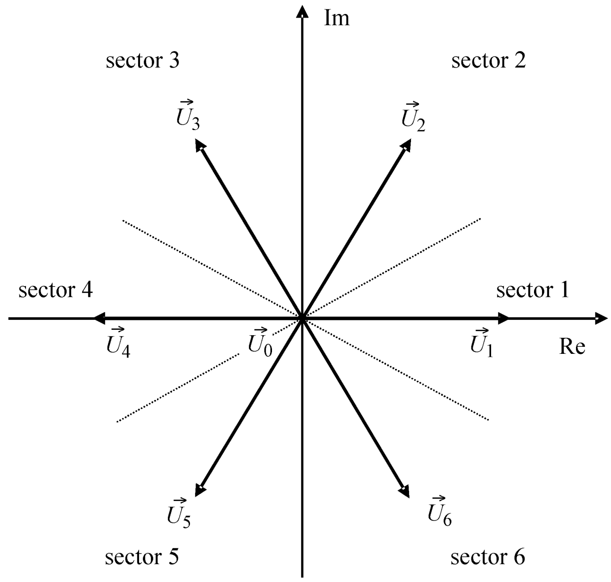


Fig. 3 Stator voltage vectors in complex plane

$$\vec{u}_s = u_{s1} + u_{s2} e^{j\frac{2\pi}{3}} + u_{s3} e^{j\frac{4\pi}{3}} = \mathrm{Re}(\vec{u}_s) + j\,\mathrm{Im}(\vec{u}_s) \quad (4)$$

and is presented in the complex plane for aforementioned combination in Figure 3. In figure is also marked a division of the complex plane into six sectors (sextants), 30° around every voltage vector.

The stator voltage vector is a discrete quantity. It can have only six discrete values with a module determined mostly with the dc link voltage and one value at which the module is approximately zero. On the other hand, the stator flux vector is a continuous quantity, whose module and position can be influenced by appropriate control of inverter switching state. If the module of the stator flux vector is at some speed constant, the currents and all fluxes in the machine will be nearly sinusoidal, and the torque change can be very fast due to a possible fast change of the angle between the stator and rotor flux vector ([3]). In such a way, the machine torque is controlled directly and the value and frequency of the fundamental component of inverter output voltage are controlled indirectly.

A position of the stator flux vector has to be calculated in the DTC method, i.e. it has to be determined in which sector the stator flux vector lies. Furthermore, the hysteresis torque and flux regulators are used. If one designates τ as an output of the torque regulator and ϕ as an output of the flux regulator, and if *1* means that the regulated quantities are bellow lower hysteresis boundary (the reference value minus hysteresis width) and *0* means that the regulated quantities are above higher hysteresis boundary (the reference value plus hysteresis width), the control algorithm of inverter switching state will be defined. This algorithm is presented in Table 1, [3, 36]. The table is valid only for one direction regarding speed (counter--clockwise to the figure 3), and for the opposite speed direction the switching state table should be defined completely analogously.

Table 1 **Direct torque control switching state table**

Sector		1	2	3	4	5	6
$\tau = 0$	$\phi = 0$	$\vec{U}_0$	$\vec{U}_7$	$\vec{U}_0$	$\vec{U}_7$	$\vec{U}_0$	$\vec{U}_7$
	$\phi = 1$	$\vec{U}_0,\vec{U}_1$	$\vec{U}_7,\vec{U}_2$	$\vec{U}_0,\vec{U}_3$	$\vec{U}_7,\vec{U}_4$	$\vec{U}_0,\vec{U}_5$	$\vec{U}_7,\vec{U}_6$
$\tau = 1$	$\phi = 0$	$\vec{U}_3$	$\vec{U}_4$	$\vec{U}_5$	$\vec{U}_6$	$\vec{U}_1$	$\vec{U}_2$
	$\phi = 1$	$\vec{U}_2$	$\vec{U}_3$	$\vec{U}_4$	$\vec{U}_5$	$\vec{U}_6$	$\vec{U}_1$

There are two non-zero voltage vectors in Table 1, denoted with $\vec{U}_0$ and $\vec{U}_1$. The only difference between them is that for the first voltage vector all all three upper IGBT switches in the inverter and for the second all three lower switches are used. Two voltage vectors are also combined in the case when the torque decrease and the flux increase are required ($\tau = 0$ and $\phi = 1$). The non-zero voltage vector is used in the low speed region, [34], because use of the zero voltage vector in a large time period can cause a demagnetisation of the machine, especially in breaking mode.

3.2 Modification of the basic direct torque control method

The basic idea of stator flux correction is to use stator currents in an appropriate manner. Namely, both the stator currents and the stator fluxes are consequences of the supplied voltages. They distinguish in the fact that the currents are measured and the fluxes are estimated quantities. The differences between motor currents and their corresponding values in DSP are small due to current transducers errors, a low-pass filtering, which is necessary, and due to errors in A/D conversion. On the other hand, the differences between the real fluxes in the machine and the corresponding estimated values in the DSP could be large. The main reason for this is a calculation of the phase fluxes according to (2) with some numerical methods and with an integration time typically couple of tens microseconds or more, and the use of integer arithmetics. Furthermore, the machine voltages, which are determined with dc link voltage and with voltage drops on semiconductors, are used for the stator flux estimation. The dc link voltage is a measured quantity and, like in the stator currents, there is a small difference between real value and corresponding value in the DSP. The voltage drops on semiconductors have great influence on the drive operation, especially in the low speed range (a zero voltage vector is applied for a relatively long time period and the motor voltages are defined directly with the voltage drops on semiconductors). Therefore, they have to be taken into account for stable operation. For the stator flux estimation, the stator currents and the stator flux resistance are also used. There are aforementioned errors in currents, and the value of resistance can be corrected through winding temperature using thermal transducers.

A consequence of the stator flux estimation in the described way is a difference between real fluxes in the machine and the fluxes calculated in DSP. The electromagnetic torque, the module and the position of the stator flux vector are calculated from estimated values of the stator flux. So, in the case of mismatches of the fluxes, wrong inverter switching state, i.e. wrong voltage vector could be applied. The torque and stator flux reference values would not be maintained and the drive could become unstable. But, if there are torque oscillations, and if the machine fluxes are not sinusoidal and have dc components (although there are not oscillations in estimated torque and stator fluxes calculated in DSP are sinusoidal), the machine currents will not also be sinusoidal or will have dc components. The question is how to use information contained in the currents?

For this purpose, in Figure 4 the stator current vector $\vec{i}_s$ and stator flux vector $\vec{\Psi}_s$ are presented. An angle between them is designated with $\Delta\varphi$ and has positive values in a traction mode and negative values in the braking mode. Namely, the developed electromagnetic torque in (3) can be defined in terms of:

$$m_{elm} = \frac{2}{3} p \left|\vec{\Psi}_s\right| \left|\vec{i}_s\right| \sin(\Delta\varphi). \tag{5}$$

It is possible to divide the stator current vector into a component $\vec{i}_{s\Psi}$ that is collinear with the stator flux vector and a component $\vec{i}_{sM}$ that is perpendicular to it. $\vec{i}_{sM}$ is also the component which creates the torque:

$$m_{elm} = \frac{2}{3} p \left|\vec{\Psi}_s\right| \left|\vec{i}_{sM}\right|. \tag{6}$$

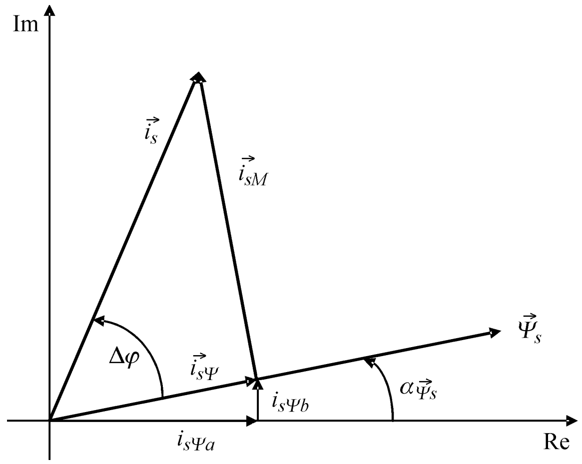


Fig. 4 Stator current vector and stator flux vector in complex plane

Although the $\vec{i}_{s\Psi}$ is stator current component, it can be analyzed as a vector with its real and imaginary part:

$$\vec{i}_{s\Psi} = i_{s\Psi a} + j i_{s\Psi b}. \tag{7}$$

Because the vectors $\vec{\Psi}_s$ and $\vec{i}_{s\Psi}$ are collinear, a real part of the stator flux vector $\vec{\Psi}_s$ and a current component $i_{s\Psi a}$, and an imaginary part of $\vec{\Psi}_s$ and a current component $i_{s\Psi a}$, in the case of correct control and stable drive operation, must be sinusoidal and must be in phase. This fact is used for the stator flux correction. It is enough to compare the instantaneous values of the real and imaginary parts of the vectors $\vec{\Psi}_s$ and $\vec{i}_{s\Psi}$ (with adequate scaling) and the difference (also with scaling) to add to the fluxes calculated according to (2):

$$\Delta\vec{\Psi}_s = k_i \vec{i}_{s\Psi} - \vec{\Psi}_s \tag{8}$$

$$\vec{\Psi}_{sc} = \vec{\Psi}_s + k_\Psi \Delta\vec{\Psi}_s. \tag{9}$$

k_i is scalar with a dimension of an inductance, $\vec{\Psi}_{sc}$ is the corrected stator flux vector, and k_Ψ is scalar. Then, by means of the vector $\vec{\Psi}_{sc}$ components, the torque is calculated according to (3) and the switching states are determined according to the switching table 1.

Although an estimated angle $\alpha_{\vec{\Psi}s}$, which vector closes with the real axis (figure 4), is used for calculation of the real and imaginary component of the vector $\vec{i}_{s\Psi}$, and which could be different from the real machine angle, the use of the stator current vector for calculation of an angle $\Delta\varphi$ and a module of the vector $\vec{i}_{s\Psi}$, in some way gives information about the drive contained in the stator currents. Any stator currents distortion and deviation from a sinusoidal waveform will result with non-collinearity of real vectors $\vec{\Psi}_s$ and $\vec{i}_{s\Psi}$, according to (8) an adequate flux correction will be done, and the real and imaginary components of vectors $\vec{\Psi}_s$ and $\vec{i}_{s\Psi}$ will force to be in phase. In the case of a sinusoidal supply, the vector $\Delta\vec{\Psi}_s$ calculated according to (8) would be a vector collinear with the stator flux vector $\vec{\Psi}_s$, and the corrected stator flux vector $\vec{\Psi}_{sc}$ calculated according to (9), would differentiate from the vector $\vec{\Psi}_s$ only regarding modules. But, with correct choosing of the scalars k_i and k_Ψ, it is possible to obtain that the difference between modules of the vectors $\vec{\Psi}_s$ and $\vec{\Psi}_{sc}$ is very small, and that the drive operation is completely stable in the whole speed range and at all conditions.

4 SIMULATION AND EXPERIMENTAL RESULTS

4.1 Laboratory model of tramcar drive

The drive assembled in the laboratory consists of components that are in tramcars: IGBT (1200 A, 1700 V) inverter in the three phase bridge connection and two induction motor which are connected in parallel. The data and parameters of traction motor are given in Table 2:

Table 2 **Data and parameters of tramcar induction motor**

Data		Parameters	
Type	V6AOJ 205-04	Stator resistance	0,044 Ω
Connection	star	Stator leakage inductance	0,263 mH
Power	65 kW		
Voltage	320 V	Magnetizing inductance	8,90 mH
Current	150 A		
Frequency	58 Hz	Rotor resistance	0,025 Ω
Pole pairs	2	Rotor leakage inductance	0,350 mH
Speed	1705 min^{-1}		

The dc link voltage in experiments was equal to the rated value of tramcar supply network in the city of Zagreb (600 V). A load-machine is dc motor that enables setting of operation steady-states with regard to motor speed and load and is also used as a high inertia for dynamic tests. The control algorithm is executed with the digital signal processor (DSP) TMS320F240 and the sampling time is 80 μs. The hysteresis bands are settled in such a way that the switching frequency is in range from 1,5 kHz to 2,5 kHz range.

In simulation model both voltage drops on semiconductors and inverter dead-time is taken into account. The motor is described in rotor flux coordinate system, and is adapted to a voltage source inverter, [35, 36]. For numerical calculation of differential equations an integration time constant is 100 ns. On the other hand, for calculation of the stator fluxes according to (2), the torque according to (3) and for determination of the switching state, the 80 μs time period is used. Integer arithmetics, low-pass filtering of the stator currents and dc link voltage and their A/D conversion are also used.

4.2 Simulation and experimental results

For a presentation of measured results, the data which are in the DSP software and are available through D/A outputs are used. The measured quantities are: the estimated torque, the inverter current, the estimated phase stator flux and the current component of the vector $\vec{i}_{s\Psi}$ which is in the phase with the measured flux (according to Figure 4). The operation and stability of the drive are also additionally controlled with direct measurement of the torque on the shaft (the torque transducer was used) and with the speed measurement (combination of a speed sensor and a toothed wheel was used).

The above described correction method is firstly verified in steady-state for rated operating point (motor speed 1705 min^{-1} and an average value of the torque for each motor 364 Nm). Figures 5 and 6 present simulation and experimental results. The scalar k_i in (8) was 2 mH (module of product $k_i\vec{i}_{s\Psi}$ for rated values was about 30 % of vector $\vec{\Psi}_s$ module), and k_Ψ in (9) was $7 \cdot 10^{-4}$. Therefore, the difference in the vector $\vec{\Psi}_s$ module and its corrected quantity $\vec{\Psi}_{sc}$ is practically negligible. These are also values of the scalars k_i and k_Ψ that are used in the whole tramcar operation range (Figure 2).

Since the basic DTC method has most difficulties in the low speed region, the next experiment was done at 5 % of rated speed, also with rated load. Figures 7 and 8 show simulation and experimental results with the same scalars k_i and k_Ψ like in the experiment at rated steady-state point. It is evident from the pictures that the currents are sinusoidal and that the drive operates stable.

An influence of the correction method is analyzed at the same operating point (5 % of rated speed and rated load), but with amounts of scalars k_i and k_Ψ which are 10 times smaller then in the previous experiment. Figures 9 and 10 present results. Even though there are no oscillations in the torque, and the estimated stator flux is sinusoidal, it can be noted that the inverter current has dc component (a duration of plus and minus half-period are uneven) and, like the current component of vector $\vec{i}_{s\Psi}$ in phase with measured flux, deviates from sinusoidal waveform. Consequently, there were oscillations in motor torque (the torque on the shaft). It is interesting to point out that at this speed the drive operation without correction practically was not possible, i.e. the basic DTC method could not be applied. Namely, the torque oscillations were very large, the motor currents were strongly distorted and the steady-state operation was impossible.

For verification of the applied method, a start-up from standstill was done. A dc load machine was used only as a high inertia, and therefore, it is achieved that time duration of the starting was similar to starting duration on the vehicle. After initial dc magnetisation, approximately maximum torque is set. The reference value of the torque is changed with predefined change of rate in order to limit the drive jerk.

Figure 11 presents experimental results with amounts of scalars k_i i k_Ψ that are adjusted on the vehicle, and Figure 12 presents results with the values which are 10 times smaller. In both cases there are no oscillations in the estimated torques, and the estimated fluxes are sinusoidal. But, in the experiment depicted in Figure 11, there are no oscillations in the torque on the shaft and the currents are completely sinusoidal. However, in Figure 12 there is again a deviation of the currents from sinusoidal waveform and their strong distortion, accompanied with significant oscillations in the torque on the shaft and in the speed. The start-up of the vehicle in such a way would be very uncomfortable with high jerks; actually, it would be completely unacceptable.

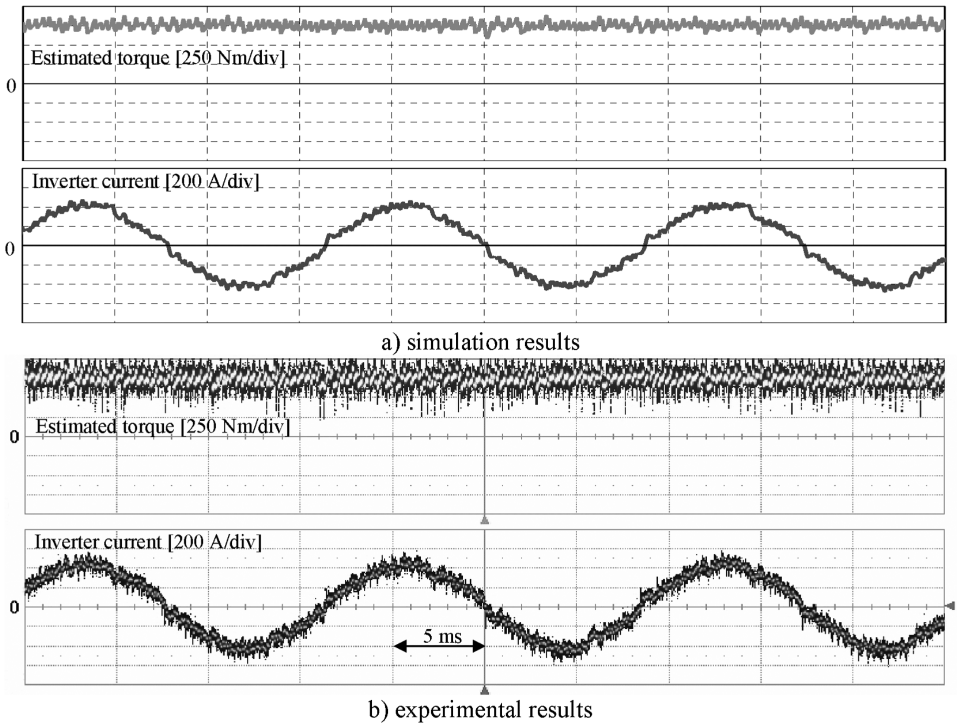


Fig. 5 Estimated torque and inverter current at rated speed and at rated load; use of correction method with k_i and k_Ψ like on the vehicle

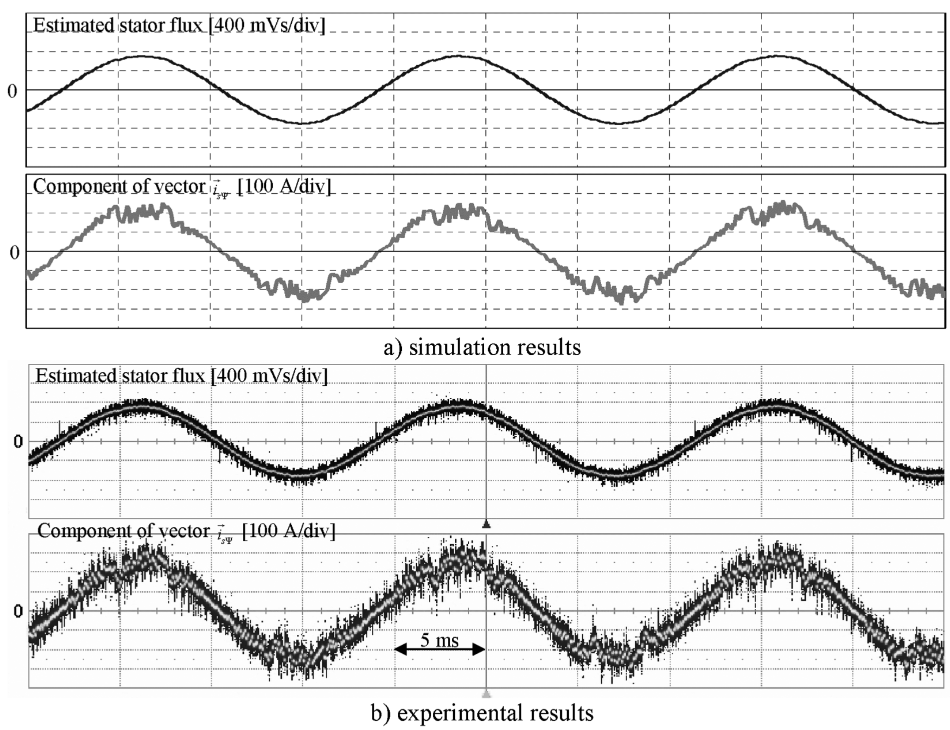
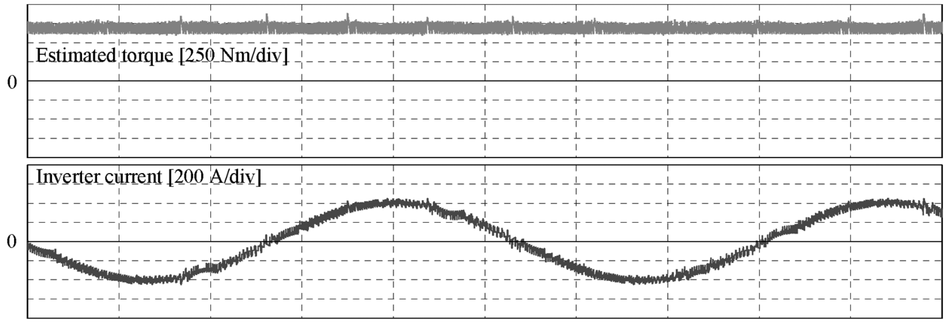
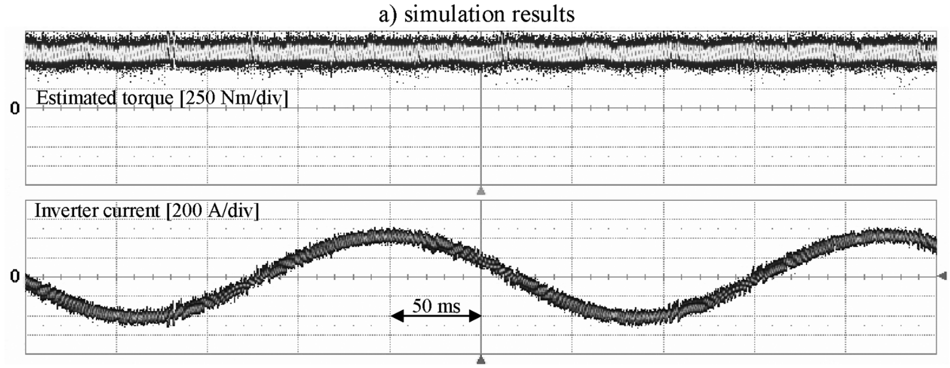


Fig. 6 Estimated stator flux and component of $\vec{i}_{s\Psi}$ at rated speed and rated load; use of correction method with k_i and k_Ψ like on the vehicle

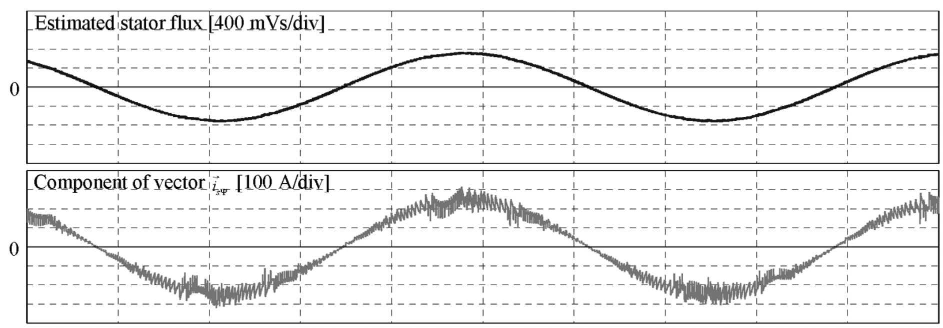


a) simulation results

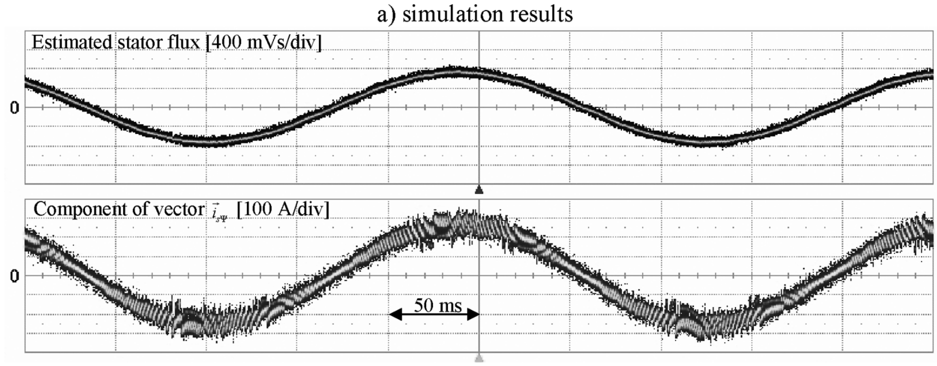


b) experimental results

Fig. 7 Estimated torque and inverter current at 5 % of rated speed and rated load; use of correction method with k_i and k_Ψ like on the vehicle



a) simulation results



b) experimental results

Fig. 8 Estimated stator flux and component of $\vec{i}_{s\Psi}$ at 5 % of rated speed and rated load; use of correction method with k_i and k_Ψ like on the vehicle

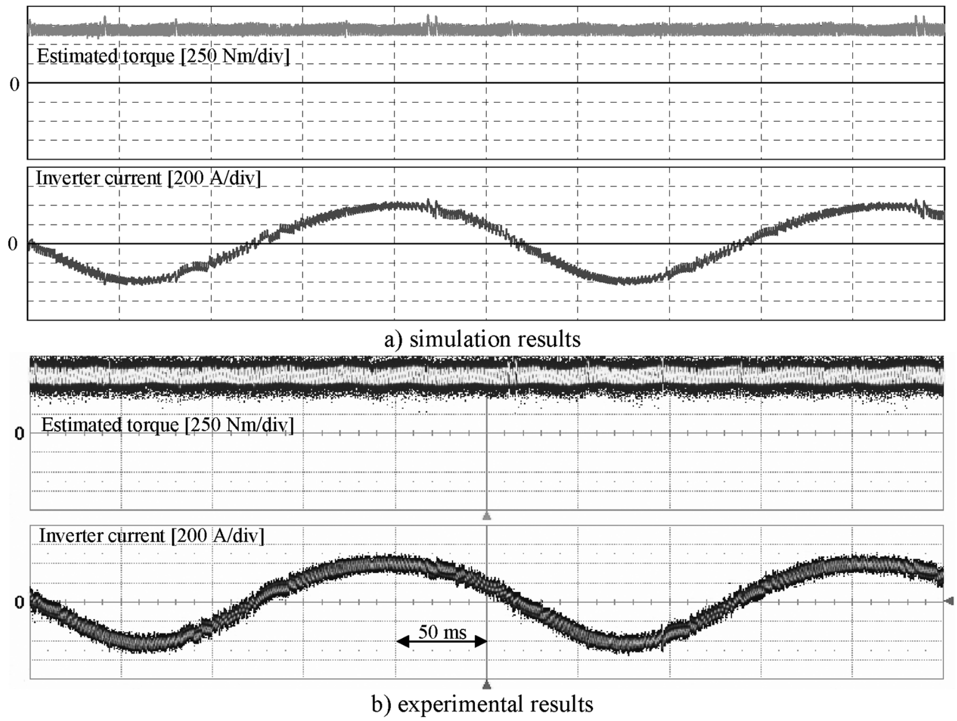


Fig. 9 Estimated torque and inverter current at 5 % of rated speed and rated load; use of correction method with k_i and k_Ψ 10 times smaller than on the vehicle

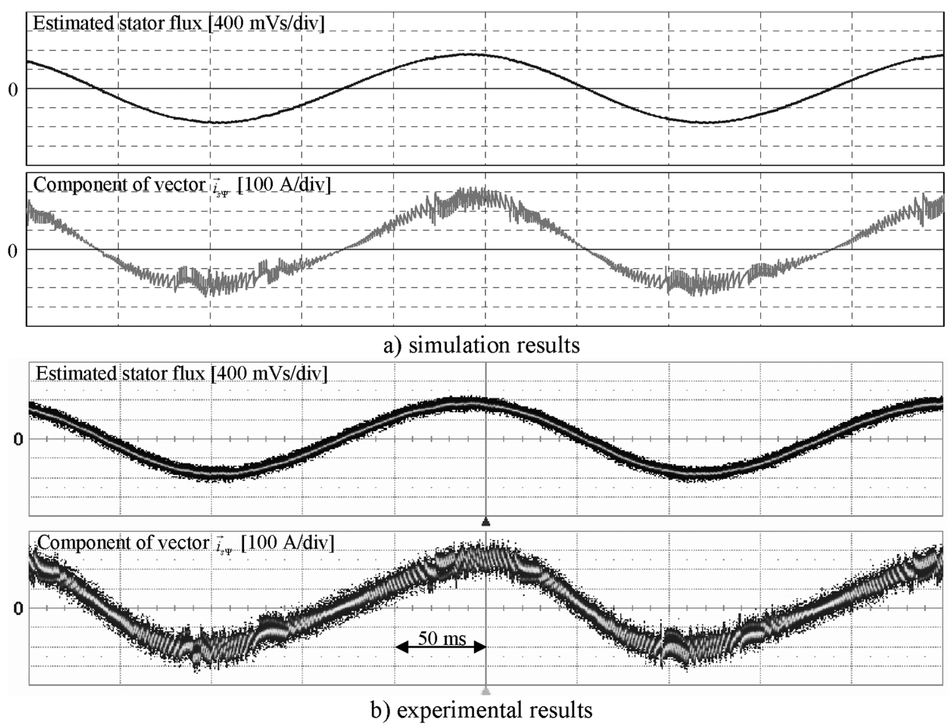


Fig. 10 Estimated stator flux and component of $\vec{i}_{s\Psi}$ at 5 % of rated speed and rated load; use of correction method with k_i and k_Ψ 10 times smaller than on the vehicle

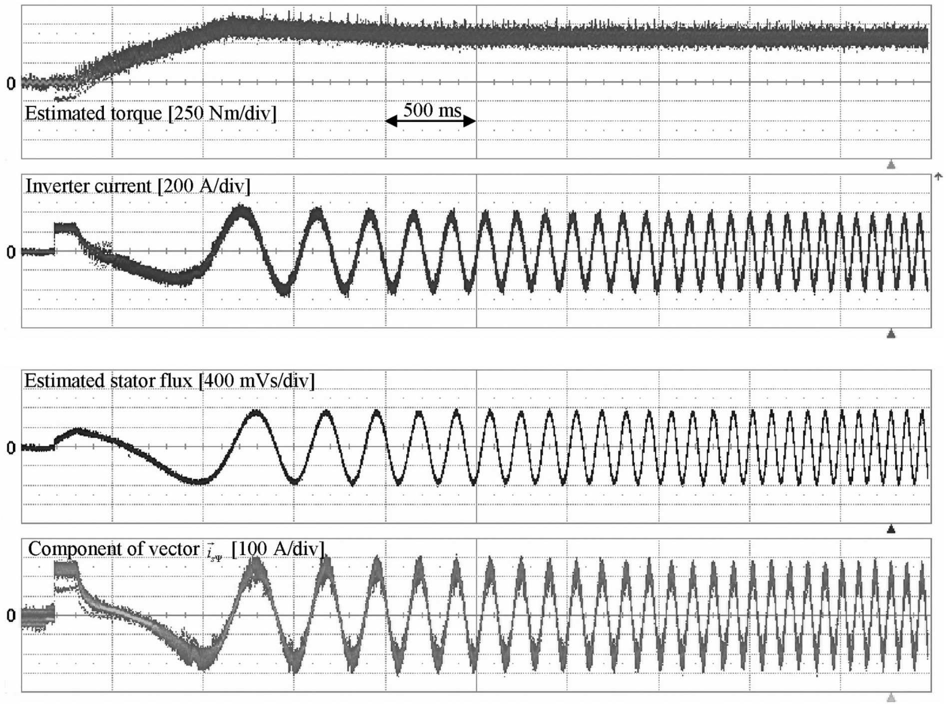


Fig. 11 Estimated torque, inverter current, estimated stator flux and component of $\vec{i}_{s\Psi}$; start-up of drive; use of correction method with k_i and k_Ψ like on the vehicle; experimental results

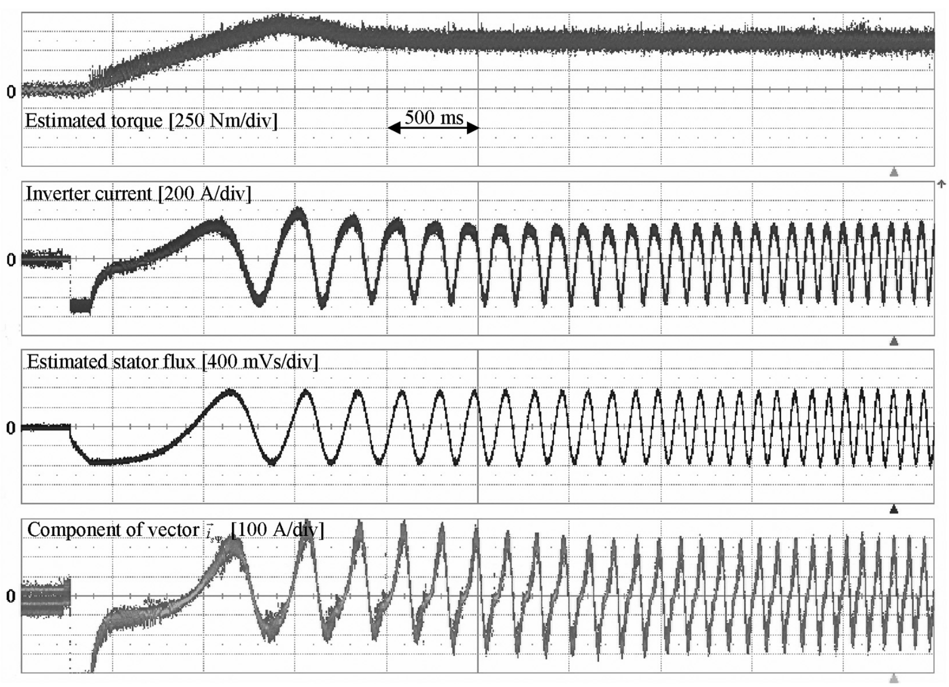


Fig. 12 Estimated torque, inverter current, estimated stator flux and component of $\vec{i}_{s\Psi}$; start--up of drive; use of correction method with k_i and k_Ψ 10 times smaller than on the vehicle; experimental results

5 CONCLUSION

One of the main tramcar drive characteristic is very wide regulation range regarding speed and torque, as well as often starting from standstill. Vector control of tramcar induction motor has to fulfil all regulation requests and insure stable and reliable operation. In the low-floor tramcars series TMK2200 a DTC method was applied. The method is very suitable for tramcar drives because it uses only stator parameters and stator quantities. But, the basic DTC method can not be used in the low speed region (a stable drive operation without some modification can not be achieved without some modification). Therefore, the basic method is modified with the correction of the stator flux vector by means of an appropriate stator current component. With the correction, the oscillations in the torque and current ripples at low speeds are eliminated, and the comfortable start-up is achieved without deterioration of any basic DTC characteristics. In such a way, an use of rotor parameters is avoid, there is no need for calculation of rotor quantities, and the only one control algorithm is used in the whole speed range. The drive operation is, consequently, simpler and more reliable.

The method is verified with the experiments on the laboratory model of the drive and during vehicles exploitation (some vehicles are running more than two years).

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Izravno upravljanje momentom asinkronih motora s korekcijom statorskog toka primijenjeno na niskopodnim tramvajima. U članku je prezentirana modificirana metoda izravnog upravljanja momentom asinkronih motora, primijenjena u glavnom pogonu niskopodnih tramvaja serije TMK2200 koje proizvodi hrvatski konzorcij CROTRAM – Končar d.d., TŽV Gredelj d.d. i Đuro Đaković d.d. Modifikacija se sastoji u korekciji vektora statorskog toka pomoću odgovarajuće komponente statorske struje. Jednostavnost izvorne metode je zadržana budući da je jedini parametar stroja kojeg je potrebno poznavati otpor statorskog namota, s time da je osiguran stabilan rad u čitavom području brzina vrtnje uključujući i mirovanje. U usporedbi s ostalim varijantama izravnog upravljanja momentom, predložena metoda je pouzdanija jer se ne koriste rotorski parametri niti se računaju rotorske veličine. Metoda je u potpunosti provjerena tijekom eksploatacije vozila, a simulacijski rezultati su uspoređeni s eksperimentalnima na modelu pogona u laboratoriju.

Ključne riječi: tramvaj, asinkroni motor, izravno upravljanje momentom, estimacija

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