Vector-Controlled Induction Motor Drive with Minimal Number of Sensors

Evgenije M. Adžić, Vlado B. Porobić, Marko S. Vekić, Zoran R. Ivanović, and Vladimir A. Katić

Abstract—This paper proposes an improved and robust induction motor drive control method which uses minimal number of sensors, providing only dc-link current measurement as a feedback signal. The proposed dc-link current sampling scheme and modified asymmetrical PWM pattern cancel characteristic waveform errors which exist in all three reconstructed line currents. In that way, proposed method is suitable for high-quality and high-performance drives. Comparison between conventional and proposed current reconstruction method is performed using hardware-in-the-loop (HIL) test platform and digital signal processor (DSP).

Index Terms—Induction motor drive, Vector control, Sensorless control, Single current sensor.

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I. INTRODUCTION

CONTROLLED induction motor drives are main driving force behind all automation systems in industry. They are used in a wide range of industry applications where they significantly contribute to improved efficiency and reliability of automation processes [1]. Due to economical and reliability reasons most automation systems use induction motor drives where motor torque and speed are controlled without shaft-sensor [2]. Speed information is then determined indirectly, by measuring terminal currents and voltages [2-5]. Elimination of shaft-sensor not only reduces cost of the drive, but it also significantly increases whole system reliability. This is especially important in high-power range applications. Industry trends to provide more robust and reliable drives operation imposed necessity for drives with minimal number of sensors.

This paper proposes solution in the field of induction motor control, case where number of sensors is reduced to minimum. Only one sensor which measures the dc-link current of the converter is used. Control algorithm includes method for motor terminal current reconstruction in a way to achieve motor current and speed control. Analysis of recent relevant literature, confirms that most of the problems are recognized and solved in a way which allows their application in low-performance drives [6]-[10]. In this situation, authors usually do not consider quality and performance of implemented control algorithms.

Quality of controller is of major importance in general purpose applications having in mind their wide distribution and energy savings that can be achieved, while dynamic performances are usually put in the background. Recent developments in the electrical vehicle industry, has discovered a need to reduce the number of sensors to minimum, i.e. only one current sensor in the converter dc-link [11]. In that way, current reconstruction method becomes important also in high-performance applications. Priority in designing high-performance drives is not the cost of the drive, but accomplishing a highly accurate, stable and fast response.

Application of the current reconstruction method in a vector controlled induction motor drives proves to be significantly lower in quality and performances compared to drives using direct line current measurement. This is mainly consequence of the usual inaccuracy in reconstructed current waveforms, which is explained in details in [12]. Application of the conventional current reconstruction mechanism with two dc-link current samples in different time instants during the same switching period in combination with PWM current ripple generates error reflected in abrupt changes in reconstructed line current waveforms. This error reflects further in dq-current components in vector control algorithm, which can be observed in increased torque and speed oscillations. Moreover, in shaft-sensorless drives, accuracy in estimated speed is significantly reduced. Stability is seriously jeopardized in situations when current and speed controllers are tuned in optimal way characteristic for high-performance drives.

Conventional approach to overcome this problem employs low-pass current filters in control structure. This paper presents original solution for conventional current reconstruction method improvement, which avoids dynamic reduction, but in the same time eliminates the error and contributes to better control characteristics.
II. DESCRIPTION OF EXPERIMENTAL SETUP

In order to validate effectiveness and reliability of the proposed current reconstruction method, hardware-in-the-loop (HIL) emulator was used for representing the power stage of induction motor drive. The core of the HIL platform is programmable FPGA based processor dedicated for processing power electronics circuits, with fast analog/digital input/output interface and supporting software tool-chain [13]. Beside schematic configurator and compiler, software tool-chain includes oscilloscope function for observing desired system variables and flexible debugging of the connected controller.

This approach provides real-time execution with 0.5 μs emulation time-step and digital signals (e.g. PWM) sampling period of 20 ns for more accurate system emulation. This feature allows the connection of real hardware controller with PWM frequency up to 100 kHz. Controller interfaced with the HIL system is based on TMS320F2812 digital-signal processor. In all performed experiments, PWM frequency was set to 2 kHz. Fig. 1 shows the experimental setup used for evaluation of the induction motor drive with minimal number of sensors. Motor nominal data and parameters of the related equivalent scheme are given in Table I.

III. VECTOR CONTROL WITH THE MEASUREMENT OF LINE CURRENT AND SPEED

A. Rotor-flux oriented sensored control structure

In the early stages of the control algorithm development, current and speed controllers are set in the framework of conventional rotor-flux oriented vector control structure using directly line current and speed sensor information. This approach defines relevant framework for testing and comparison of proposed algorithm with only one dc-link sensor in a feedback path. In this stage, control algorithm uses measured line current and rotor speed as depicted in Fig. 2. Core of the control algorithm represents the flux model (FLUX_MOD) which is used to estimate the values of rotor flux angle (Theta) from the stator current vector components (IDs and IQs) and from the measured speed (W) [14].

B. Stator current controller tuning

PI current controllers for d- and q-axes are tuned using Dahlin’s algorithm [15] which defines fast aperiodical step response of controlled current components, IDs and IQs, i.e. motor flux and torque. A proportional and integral gain for current controllers in discrete domain are:

\[
K_{pi} = \frac{1 - e^{-\lambda T_s}}{K_{it} \left( e^{\frac{T_s}{\sigma}} - 1 \right) \left( 1 - e^{-\lambda T_s} \right)}
\]

\[
K_{ii} = K_{pi} \left( e^{\frac{T_s}{\sigma}} - 1 \right)
\]

Fig. 1. Block diagram of experimental setup used for control algorithm verification.

Fig. 2. DSP implementation of classical structure of rotor-flux oriented control for induction motor drive.

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Quantity</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_n$</td>
<td>Nominal power</td>
<td>1.1 kW</td>
</tr>
<tr>
<td>$p$</td>
<td>Number of poles</td>
<td>4</td>
</tr>
<tr>
<td>$f_s$</td>
<td>Nominal frequency</td>
<td>50 Hz</td>
</tr>
<tr>
<td>$n_n$</td>
<td>Nominal speed</td>
<td>1410 rpm</td>
</tr>
<tr>
<td>$U_n$</td>
<td>Nominal voltage</td>
<td>380 V</td>
</tr>
<tr>
<td>$I_n$</td>
<td>Nominal current</td>
<td>2.9 A</td>
</tr>
<tr>
<td>$J$</td>
<td>Moment of inertia</td>
<td>0.00247 km²</td>
</tr>
<tr>
<td>$R_s$</td>
<td>Stator resistance</td>
<td>9,137 Ω</td>
</tr>
<tr>
<td>$R_r$</td>
<td>Rotor resistance</td>
<td>6,422 Ω</td>
</tr>
<tr>
<td>$L_{sr}$</td>
<td>Stator leakage inductance</td>
<td>18,89 mH</td>
</tr>
<tr>
<td>$L_{rr}$</td>
<td>Rotor leakage inductance</td>
<td>17,28 mH</td>
</tr>
<tr>
<td>$L_m$</td>
<td>Magnetizing inductance</td>
<td>320,3 mH</td>
</tr>
</tbody>
</table>

TABLE I

TESTED MOTOR DATA
In Eq. 1 parameter \( \lambda \) defines desired dynamic of aperiodical step response, \( T_s \) represent current loop sampling period, and \( K_u \) is total gain in a direct path of current control loop. Total gain \( K_u \) equals:

\[
K_{ii} = \frac{u_{dc}}{\sqrt{3}} \frac{1}{R_s I_b}
\]

(2)

where are: \( u_{dc} \) – dc-link voltage, \( R_s \) – stator winding resistance, and \( I_b \) – selected current base value. Independent design of current and speed controller is enabled by selecting the parameter \( \lambda \) to satisfy the condition:

\[
\frac{1}{\lambda} \leq \frac{T_{io}}{5}
\]

(3)

where \( T_{io} \) represents the speed control loop sampling period. Used values for current controllers in d- and q-axes are given in Appendix section.

Dynamic response of stator current controller is given in Fig. 3, where it can be noticed that d-axis reference current was maintained on the constant value, i.e. \( i_{d\text{REF}} = 0.32 \text{ p.u.} \), while q-axis reference current was periodically step-changed between values 0 and 0.48 p.u. d- and q-current references, \( i_{d\text{REF}} = 0.32 \text{ p.u.} \) and \( i_{q\text{REF}} = 0.48 \text{ p.u.} \), together defines nominal current amplitude for considered motor. Fig. 3 proves that aperiodical response, without overshoot, was achieved with settling time of 19 ms which is approximately equal to expected value of \( 5/\lambda = 16.7 \text{ ms} \). Fig. 4 shows corresponding response of motor line currents and electromagnetic torque. For rotor flux value defined with \( i_{d\text{REF}} = 0.32 \text{ p.u.} \) and \( i_{q\text{REF}} = 0.48 \text{ p.u.} \) motor develops nominal electromagnetic torque \( T_m = 7.45 \text{ Nm} \) in steady-state. Torque oscillations around 4.7% \( T_m \) are consequence of PWM current.

Designed current controllers are used for all other experiments throughout this paper.

C. Rotor speed controller tuning

Speed controller parameters are selected in a way to obtain fast aperiodical step response. Detailed design procedure, suggested for high-performance motor drives in [16], gives following proportional and integral gains for speed controller:

\[
K_{p\omega} = 0.2027 \frac{2J \omega_b}{T_{io} K_m}
\]

\[
K_{i\omega} = 0.03512 \frac{2J \omega_b}{T_{io} K_m}
\]

(4)

In Eq. 4 \( J \) represents motor moment of inertia, \( \omega_b \) is adopted base value for angular frequency, and \( K_m \) is the gain in the direct path of speed control loop model obtained from the torque equation of vector-controlled induction motor:

\[
K_m = \frac{3}{2} \frac{J_m}{L_r} i_{sd} i_b
\]

(5)

Used parameters of speed controller are given in the Appendix. Fig. 5 shows dynamic speed response after setting step reference value \( \omega_{\text{rREF}} = 0.1 \text{ - } 0.4 \text{ p.u.} \) (300 – 1200 rpm). Motor was loaded with 20% of rated torque 1.5 Nm. Recorded HIL results in the same conditions are shown in Fig. 6. It could be noticed that speed has aperiodical transient response with settling time 0.2 s, as expected for optimal method and parameters given with Eq. 4. Current response in field coordinates in the Fig. 5, proves that decoupled control of motor flux and torque was successfully achieved. Designed speed controller is used for all other experiments in the paper.
IV. SHAFT-SENSORLESS VECTOR CONTROL WITH LINE CURRENT MEASUREMENT

A. Rotor-flux oriented sensorless control structure

In a base speed region rotor-flux oriented vector control without speed sensor shows better performance comparing to the orientation towards stator flux vector [14]. Taking this into account and due to easier integration of sensorless control algorithm in previously described control structure, this paper uses sensorless rotor flux oriented control structure shown in Fig. 7. This fact also facilitates comparison analysis between implemented sensed and sensorless algorithms. Open-loop rotor speed estimation (SPEED_EST) with advanced rotor flux vector estimation (FLUX_EST) has been applied.

B. Estimation of rotor flux components

Regarding selected reference frame, the flux vector components are firstly calculated in term of stator coordinates, and in second step magnitude and phase angle are derived in field orientated coordinates.

\[
\frac{d\psi^i_{rd}}{dt} = \frac{L_m}{T_r}i_{sd} - \frac{1}{T_r}\psi^i_{rd}
\]

\[
\psi^i_{rd} = 0
\]

(6)

Determined rotor flux components has to be transformed in stationary reference frame with inverse Park transformation, to perform correction of adaptive rotor flux voltage model derived in stator coordinates, as depicted in Fig. 8. Stator flux components are then calculated based on the motor stator flux linkage equations:

\[
\psi^i_{sa} = \frac{L_m}{L_r} \psi^i_{ra} + L_\sigma i_{s\alpha}
\]

\[
\psi^i_{sb} = \frac{L_m}{L_r} \psi^i_{rb} + L_\sigma i_{s\beta}
\]

(7)

where \(L_\sigma\) represents stator leakage inductance.

Voltage model of rotor flux estimator is based on stator voltage equations in stationary reference frame, where stator flux is simply obtained by integration of stator winding induced electromotive force. Input variables are stator voltages and currents. Stator voltages are not directly measured. They are estimated based on measured dc-link voltage and inverter switching states. Dead-time and switching components for voltage drop compensation scheme is applied. Due to pure integration of input variables, stator voltage and current, compensation terms are included in the model to provide adaptive mechanism for elimination of offset error, erroneous initial values or their disturbances. Adaptive stator flux model is given (superscript \(u\)):

\[
\psi^u_{sa} = \int [u_{s\alpha} - R_s i_{s\alpha} - u^c_{\alpha}] dt
\]

\[
\psi^u_{sb} = \int [u_{s\beta} - R_s i_{s\beta} - u^c_{\beta}] dt
\]

(8)
Compensation terms, \(u_\alpha^\text{comp}\) and \(u_\beta^\text{comp}\), represents outputs of PI controllers that eliminates the error between the ideal current model (Eq. 6) and voltage model of stator flux:

\[
\begin{align*}
    u_\alpha^\text{comp} &= K_{pf} \left( \psi^u_{sa} - \psi^u_{sa} \right) + K_{gf} \left[ \psi^u_{sa} - \psi^u_{sa} \right] dt \\
    u_\beta^\text{comp} &= K_{pf} \left( \psi^u_{sb} - \psi^u_{sb} \right) + K_{gf} \left[ \psi^u_{sb} - \psi^u_{sb} \right] dt
\end{align*}
\]  

(9)

Parameters \(K_{gf}\) and \(K_{pf}\) are determined in the way that current model in adaptive stator flux model dominates in the low speed range, while voltage model prevails in the high-speed range [17]. Finally, outputs of the rotor flux estimator are rotor flux components obtained from motor flux linkage equations in stationary reference frame:

\[
\begin{align*}
    \psi^u_{ra} &= \frac{L_r}{L_m} \left( \psi^u_{sa} - L_\sigma i_{sa} \right) \\
    \psi^u_{rb} &= \frac{L_r}{L_m} \left( \psi^u_{sb} - L_\sigma i_{sb} \right)
\end{align*}
\]  

(10)

Determined rotor flux components are later used for rotor flux vector position and rotor speed estimation.

C. Rotor flux vector phase angle estimation

Exact determination of rotor flux vector phase angle, i.e. its position, is necessary for correct and proper transformation of all quantities in the field-rotating reference frame in which control algorithm is implemented. Solution of this problem mainly determines quality of the whole control structure, because error in the phase angle has significant negative influence on the independent control of dq-currents [14].

One of the standard and advanced methods for flux phase angle estimation represents Phase-Locked-Loop (PLL) structure implemented in rotating reference frame [18]. Block diagram of PLL algorithm used in this paper is shown in the Fig. 9. As input values, PLL uses previously determined rotor flux components in stationary reference frame. Rotor flux components are transformed in dq rotating reference frame using estimated rotor flux phase angle at the PLL output.

Error signal is formed as a difference between rotor flux q-axis component reference set to 0 and obtained rotor flux q-axis component after transformation. This actually results in rotor flux vector tracking, i.e. in equalization of estimated and actual rotor flux phase angles. PLL filter in the form of PI compensator leads to reducing of error signal to zero value, also for flux frequency step changes. Parameters of PLL filter are selected in a way to provide aperiodical phase angle response with desired bandwidth of PLL control loop \(\omega_{bw}\) [19]:

\[
K_{pp} = \frac{\sqrt{2}}{f_b} \omega_{bw}
\]

\[
K_{ip} = \frac{1}{2f_b} \omega_{bw}^2
\]

(11)

Integral gain of the PLL filter \(K_{ip}\) has to be multiplied with PLL sampling loop to obtain corresponding gain \(K_{ip}^2\) for digital implementation. Selected values used throughout all the experiments in the paper are given in Table II.

D. Open-loop estimation of rotor speed

Rotor flux angular frequency represents the first derivate of rotor flux phase angle, so it yields:

\[
\omega_s = \frac{d\theta_\phi}{dt} = \frac{d}{dt} \left( \frac{\arctan \left( \frac{\psi_{rb}}{\psi_{ra}} \right) \left( \begin{array}{c} d\psi_{rb} \\ d\psi_{ra} \end{array} \right) }{\psi_{ra}^2 \frac{d\psi_{rb}}{dt} - \psi_{rb} \frac{d\psi_{ra}}{dt} } \right)
\]

(12)

By eliminating rotor currents from the rotor flux linkage equations in the induction motor model [14], and based on voltage equations for rotor winding in stationary reference frame, rotor flux angular frequency could be obtained in the following form:

\[
\omega_s = \omega_r + \frac{1}{\psi_r^2} \frac{L_m}{T_r} \left( \psi_{ra} i_{rb} - \psi_{rb} i_{sa} \right) = \omega_r + \omega_k
\]

(13)

\(\omega_k\) represents slip frequency which is directly proportional to motor electromagnetic torque with assumption that rotor flux is maintained on constant value. Rotor speed can be estimated with:

\[
\omega_r = \omega_s - \omega_k = \omega_s - \frac{1}{\psi_r^2} \frac{L_m}{T_r} \left( \psi_{ra} i_{rb} - \psi_{rb} i_{sa} \right)
\]

(14)

Block diagram of implemented rotor speed estimator is shown in the Fig. 10.

E. Test results

Here are presented the main test results when feedback is closed using estimated speed and the concept described in previous chapter. Figs. 11 and 12, show motor speed, electromagnetic torque and stator current response under the same circumstances in which speed loop was closed with the measured rotor speed. Speed reference is changed in step manner between values \(\omega_{r,REF} = 0,1 - 0,4\) p.u. with motor load
of 1.5 Nm.

Fig. 11 shows relative values of control variables recorded in the DSP. Average speed settling time of 0.14 s is not exact the same, but it is very close to the case when speed was measured. q-component of the stator current and corresponding torque oscillations are slightly increased, due to the PLL filter and rotor flux PI compensators in the control feedback path. Moreover, motor line currents are sinusoidal without significant distortion, similar to the case when speed loop was closed using measured rotor speed (Fig. 12).

V. SENSORLESS VECTOR CONTROL WITH CONVENTIONAL CURRENT RECONSTRUCTION METHOD

A. Conventional current reconstruction method

Constant requirements for reducing cost and increasing reliability of the motor drives lead to the series of research with the aim to reduce number of sensors to its essential minimum. Research in the field proved that it is possible to reduce number of current sensors in three-phase AC drives to only one sensor in the dc-link. Block diagram of such drive is shown in the Fig. 13. Among previously described control structures, a core of the control scheme is block with motor phase currents reconstruction (ABC.RECONS) and modified switching modulator (SVGEN_AB2) with the task to support critical cases in reconstruction mechanism.

Due to the topology of three-phase inverter and selected switching modulation (SVPWM), it is possible to reconstruct motor phase currents from measured dc-link current [20]. In each switching period, dc-link current includes information about two motor line currents while remaining third current could be reconstructed taking into account that sum of the line currents is equal to zero. Fig. 14 shows an example and details of conventional current reconstruction principle, where switching PWM signals A, B and C for upper inverter switches are arranged defining output voltage vector in the first SVPWM sector. Two dc-link samples are taken in strategic moments (SAMP1 and SAMP2) precisely synchronized regarding the middle of switching PWM period (TRIG) and according to the beginning of the active voltage vectors, as shown in Fig. 14. Here, current sample \( i_a \) during active voltage vector with only one upper switch turned-on is equal to phase current \( i_a \), while current sample \( -i_d \) during active voltage vector with two upper switches turned-on is equal to inverted value of motor phase current \( i_d \). Similar analysis could be performed for other SVPWM sectors that yield to complete current reconstruction algorithm [20].
B. Modified SVPWM method

Practical problems with dc-link current sampling, occur when switching PWM period contains narrow active voltage vectors. First critical case represents a situation where the reference voltage vector passes between SVPWM sectors, and when duty cycles values for two inverter legs are almost equal. Fig. 15a shows example of this case, when voltage vector passes between the first and the second sector and when it is possible to reconstruct only one line current, \(-i_a\). Line current \(i_a\) is unobservable. This is always the case in normal operation due to the reference vector rotation, and it is independent on modulation index value. Second critical situation occurs in the case of small reference amplitudes of output voltage vector, i.e. in the case of small modulation index. This is a usual situation in the case of low reference speed and low motor load torque. PWM duty cycles values for all three inverter legs are almost the same and around 50%. Sampling windows in the frame of both active vectors are not wide enough for reliable dc-link current measurement, so line current information could not be obtained. Fig. 15b shows a case of small amplitude reference voltage vector in the first SVPWM sector. Obviously, first signal SAMP1 samples line current \(i_a\) instead \(-i_a\), while second sampling signal SAMP2 measure zero value instead current \(i_a\).

Fig. 16 shows results in case of applying original switching scheme without mechanism for obtaining minimal width of active voltage vectors. There is significant reconstructed current waveform distortion in critical intervals, which are useless for achieving vector control with only dc-link current feedback. Abrupt change and large deviation of reconstructed currents \((i_a^{REC}, i_b^{REC}, i_c^{REC})\) from actual current values \((i_a, i_b, i_c)\) can be noticed in the area between different voltage sectors (Sector). To overcome these problems it is necessary to perform one of suggested mechanism for reliable dc-link current measurement by modifying the originally symmetrical PWM voltage patterns sufficiently, in the cases when longer active vectors are needed [6-9, 21].

Fig. 15. Critical cases for line current reconstruction method.

Fig. 16. Result of current reconstruction method without switching pattern modification.

Fig. 17 illustrates the applied solution principles with an example where small modulation index was referenced. PWM signals \(V_h (A), V_m (B), \) and \(V_l (C)\) does not form enough long active vectors for reliable dc-link current reading. Suggested method modifies PWM signals associated with the middle \((V_m)\) and highest \((V_h)\) voltage commands. Both signals are shifted in the right direction to form active vectors with minimal width \(T_{MIN}^{VECTOR}\) for reliable dc-link current measurement. Firstly, if needed, PWM signal \(V_m\) is right--shifted by time:

\[
\Delta T_{VECTOR} = T_{MIN}^{VECTOR} - (T_m - T_l)
\]

Active time intervals of PWM signal \(V_m\) during right (lagging) and left (leading) half-periods in PWM cycle should be updated to new values, \(T_{m2}\) and \(T_{m1}\), respectively:

\[
\begin{align*}
T_{m2} &= T_h + \Delta T_{VECTOR} \\
T_{m1} &= T_m - \Delta T_{VECTOR}
\end{align*}
\]

Then, second active vector duration has to be calculated and if needed, PWM signal \(V_h\) has to be right-shifted. Signal \(V_h\) width in lagging PWM half-period has to be extended, and in leading half-period to be reduced, to the new values \(T_{h2}\) and \(T_{h1}\):

Fig. 17. PWM pattern modification principle used for reliable reconstruction.
\[
\Delta T_{\text{vector}2} = T_{\text{MIN}}^{\text{vector}} - (T_h - T_m^2)
\]

\[
T_{h2} = T_h + \Delta T_{\text{vector2}}
\]

\[
T_{h1} = T_h - \Delta T_{\text{vector2}}
\]

C. Test results

Figs. 18 and 19 illustrate the operation of conventional current reconstruction method with proposed modified switching pattern, in the closed-loop of described sensorless vector-control structure. Fig. 18 shows dynamic response of motor speed and stator dq-currents for same reference and operating conditions: \(\omega_r^{\text{REF}} = 0.1 - 0.4 \text{ p.u.}\) and \(T_{\text{load}} = 1.5 \text{ Nm}\). There are significant oscillations observed in dq-current components which are not caused only by current ripple on switching frequency, but also due to the current reconstruction mechanism which samples dc-link current in two different instants during PWM period [12]. Its final result is appearance of third and sixth harmonics in dq-current components. Maximal amplitude of the third and sixth harmonics in d-current was 9.4% and in q-current 43.3% of referenced value which contributed to higher distortion and oscillations in motor line current, torque and speed shown in the Fig. 19.

VI. SENSORLESS VECTOR CONTROL WITH PROPOSED CURRENT RECONSTRUCTION METHOD

A. Proposed current reconstruction method

The basic idea suggested in this paper represents an improvement of the method proposed in [6], which used the line-currents measured in both halves of the naturally symmetrical PWM switching period. The method proposed in [6] is based on sampling of the dc-link current in the center of the active voltage vectors four times during one PWM period and calculation of the two available line-current values by averaging the samples from two matching vector pairs.

![Fig. 18](image1.png)

**Fig. 18.** DSP results: motor speed and dq-currents response for step reference \(\omega_r^{\text{REF}} = 0.1 / 0.4 \text{ p.u.}\) – conventional current reconstruction method.

This approach provides synchronous measurement of all three line-currents, referred to the center of a PWM period. It effectively cancels error due to current ripple in the reconstructed line-currents and eliminates the current samples’ mutual phase-shift. Besides its simplicity, this method is completely insensitive to machine parameter variances. However, in [6] the critical cases of a reference voltage vector passing between the six possible active vectors or with a low modulation index are neglected and not considered. The authors in [22] clearly emphasized that during these cases and with the PWM modified scheme used (where PWM signals are not symmetrical), the simultaneously sampled line-currents cannot be acquired. It clearly indicates there is a need to provide an improved procedure for reliably and more accurate measurement of the motor line-currents.

Considering the high PWM switching frequencies, up to 20 kHz, and usually employed electrical motors, one can conclude that there is no need for the very high current control-loop sampling rate at the PWM level. This fact allow us to record line-current information on the lagging (right) side of one PWM period and then on the leading (left) side of the subsequent PWM period and calculate the available line-currents by simple averaging of the corresponding recorded values. In this way, all three estimated line-currents would be referred to the same instant reflecting the average current value in two consecutive PWM periods. The proposed method enables us to improve the PWM pattern control in order to account for critical cases. It represents an extension of the modified PWM pattern explained in Section V.B where under critical conditions, the lagging half-pulse width is shifted to the right and the leading half-pulse width in the subsequent PWM period is shifted to the left in order to create sufficient sampling windows for current measurement (Fig. 21b).

Block diagram of implemented proposed control structure is illustrated in Fig. 20. Key blocks that improve control quality and performances are proposed current sampling block \(F281X\_IDC4\), AVERAGING block for finding inputs for conventional line current reconstruction \(IABC\_RECONS\), and
The line currents at the time instant representing average values in two consecutive PWM periods, can be obtained using simple calculation:

\[
\begin{align*}
i_d^{AV} &= -\frac{i_{dc} (SAMP2) + i_{dc} (SAMP3)}{2} = -\frac{i_{lo} + i_{hc}}{2} \\
i_h^{AV} &= \frac{i_{dc} (SAMP1) + i_{dc} (SAMP4)}{2} = \frac{i_{ho} + i_{hc}}{2} \\
i_m^{AV} &= \left(\frac{A^{AV} + i_h^{AV}}{2}\right)
\end{align*}
\] (18)

It remains to assign the resultant currents \(i_h^{AV}\), \(i_m^{AV}\) and \(i_d^{AV}\) to motor line-currents \(i_a\), \(i_b\) and \(i_c\), depending on the actual sector number as in conventional case.

### B. Test results

The first improvements in the reconstructed line-current waveform can be observed for a steady-state operation with \(\omega_r^{REF} = 0.4\) p.u., \(T_{load} = 1.5\) Nm and closed-loop operation with proposed current feedbacks, in Fig. 22. Fig. 22a shows actual (measured) line current values, \(i_a\), \(i_b\), and \(i_c\), compared to the reconstructed line currents by proposed scheme, \(i_a^{NEW}\), \(i_b^{NEW}\), and \(i_c^{NEW}\), and by conventional scheme, \(i_a^{REC}\) (only phase-\(a\) current is shown for clarity). Improved reconstructed currents do not include characteristic abrupt changes in its waveforms due to the passing reference voltage vector between SVPWM sectors. Proposed current reconstruction reduces presence of third and sixth harmonic terms in \(dq\)-current components by factor of 3, but also the offset from actual values which are noticeable especially in \(d\)-current component (Fig. 22b).

![Fig. 20. DSP implementation of sensorless control structure for induction motor drive, with proposed current reconstruction algorithm (F281X_IDC4+AVG) and modified modulator (SVGEN_AB4).](image)

The proposed modified voltage modulator \(SVGEN\_AB4\) that overcomes critical cases in the reconstruction mechanism.

Fig. 21 shows the dc-link current and motor line currents during two consecutive PWM periods and details related to the proposed method. Sampling signals \(SAMP1\) and \(SAMP4\) are triggers for measurement of dc-link current in two consecutive PWM periods during matching active voltage vectors defined with only one inverter switch turned-on (here, with PWM signal \(A\)). \(SAMP1\) samples dc-link current at the beginning of the active voltage vector, and \(SAMP4\) at the end of the active vector. Similar, sampling signals \(SAMP2\) and \(SAMP3\) are triggers for measurement of dc-link current at the beginning and at the end of matching active voltage vectors defined with two upper switches turned-on (here, with PWM signals \(A\) and \(B\)), respectively.

![Fig. 21. Proposed current reconstruction method details.](image)

Fig. 22. Steady-state response for operating conditions \(\omega_r^{REF} = 0.4\) p.u. and \(T_{load} = 1.5\) Nm: a) actual and reconstructed line currents; b) actual and reconstructed \(dq\)-currents with proposed \((^{NEW})\) and conventional reconstruction method \((^{REC})\).
This can be observed in motor torque and speed, not only in the steady-state, but also in transient intervals, where dynamic response is more stable. Figs. 23 and 24 show dynamic response of motor speed, electromagnetic torque and line currents in the same operating conditions: \( \omega_r^{\text{REF}} = 0,1-0,4 \) p.u. and \( T_{\text{load}} = 1,5 \) Nm. Oscillations are less reduced for low speed reference 0,1 p.u. because there are more frequent requirements for application of asymmetrical PWM signals due to the low reference voltage amplitude. However, obtained results together with the results in Fig. 18 verify that proposed method provide more stable stationary and dynamic response compared to the conventional current reconstruction method. Proposed control scheme shows almost the same behavior as in the case when direct measured line currents were used (Figs. 11 and 12).

**Fig. 23.** DSP results: motor speed and dq-currents response for step reference \( \omega_r^{\text{REF}} = 0,1 - 0,4 \) p.u. – proposed current reconstruction method.

**Fig. 24.** HIL results: motor speed, torque and stator currents step response for \( \omega_r^{\text{REF}} = 0,1 - 0,4 \) p.u. – proposed current reconstruction method.

**VII. CONCLUSION**

Application of conventional line currents reconstruction method does not provide acceptable control quality and drive performance. The source of the problem is the distortion of the reconstructed motor currents. Distortion is caused by the nature of the employed method which samples dc-link current in different time instants related to the center of PWM period. Distortion in original abc-domain is reflected in the field-oriented reference frame dq-current components, which are further propagated through complete sensorless control structure. Results are increased oscillations in developed motor torque and speed, which can even cause unstable motor operation during transients.

Proposed solution of this problem does not use standard approach which introduces current filter blocks in the feedback path reducing the dynamic of complete control system. It is also completely independent on the machine model and parameters. It uses averaging of totally four dc-link current samples in two consecutive PWM periods, which significantly reduces the phase error and characteristic third and sixth harmonics in dq-current components. Average dc-link samples are referred to the same time instant which yields to the reconstructed current waveforms which are almost the same as actual measured values.

This paper verifies that proposed method reduces motor torque and speed oscillations to the acceptable level in the case when minimum number of sensors is used, and in conditions where all controller and estimator parameters are set for high-performance response. In future work proposed current reconstruction algorithm will be tested in various shaft-sensorless control structures.

**APPENDIX**

**TABLE II**

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Quantity</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( f_{\text{PWM}} )</td>
<td>PWM frequency</td>
<td>2 kHz</td>
</tr>
<tr>
<td>( T_i )</td>
<td>Current loop sampling period</td>
<td>1 ms</td>
</tr>
<tr>
<td>( T_{so} )</td>
<td>Speed loop sampling period</td>
<td>10 ms</td>
</tr>
<tr>
<td>( I_r )</td>
<td>Current base value</td>
<td>7.02 A</td>
</tr>
<tr>
<td>( U_b )</td>
<td>Voltage base value</td>
<td>842.55 V</td>
</tr>
<tr>
<td>( f_s )</td>
<td>Frequency base value</td>
<td>100 Hz</td>
</tr>
<tr>
<td>( \omega_b )</td>
<td>Angular frequency base value</td>
<td>628.32 rad/s</td>
</tr>
<tr>
<td>( \lambda )</td>
<td>Current loop dynamic parameter</td>
<td>300</td>
</tr>
<tr>
<td>( K_{m} )</td>
<td>Current loop proportional gain</td>
<td>0.1357</td>
</tr>
<tr>
<td>( K_{ii} )</td>
<td>Current loop integral gain</td>
<td>0.0403</td>
</tr>
<tr>
<td>( K_{mp} )</td>
<td>Speed loop proportional gain</td>
<td>8.5716</td>
</tr>
<tr>
<td>( K_{ip} )</td>
<td>Speed loop integral gain</td>
<td>1.4851</td>
</tr>
<tr>
<td>( i_{\text{MIN}} )</td>
<td>Speed controller output positive limit</td>
<td>0.82 p.u. (1.45·I_{MIN})</td>
</tr>
<tr>
<td>( i_{\text{MIN}} )</td>
<td>Speed controller output negative limit</td>
<td>-0.25 p.u.</td>
</tr>
<tr>
<td>( K_f )</td>
<td>Flux compensator proportional gain</td>
<td>0.05</td>
</tr>
<tr>
<td>( K_{if} )</td>
<td>Flux compensator integral gain</td>
<td>0.00111</td>
</tr>
<tr>
<td>( \omega_{\text{PLL}} )</td>
<td>PLL filter bandwidth</td>
<td>300 Hz</td>
</tr>
<tr>
<td>( K_{pp} )</td>
<td>PLL filter proportional gain</td>
<td>26.657</td>
</tr>
<tr>
<td>( K_{oz} )</td>
<td>PLL filter integral gain</td>
<td>0.225</td>
</tr>
</tbody>
</table>
ACKNOWLEDGMENT

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REFERENCES