Parameter Sensitivity Analysis of an Improved Open-Loop Speed Estimate for Induction Motor Drives

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Abstract—In the last decades, several schemes have been proposed for speed-sensorless control of Induction Motor (IM) drives. Promising approaches are closed-loop observers, as Model Reference Adaptive Systems (MRAS) and Kalman filters, and open-loop estimators, which have recently aroused lively interest because of their simplicity and low-cost profile. As known, open-loop algorithms use only motor equations to derive the speed estimation, whose accuracy turns out to be strictly related to the motor model and its parameters. This paper presents an improved open-loop speed estimation algorithm, which uses reference voltages and measured currents, after that a state-of-the-art dead time compensation has been performed. The often undervalued topic of parameter sensitivity is handled by an accurate mathematical investigation, confirmed by a complete set of both simulated and experimental results. The estimated speed is then used as feedback signal in a closed loop speed control, and a performance evaluation is accomplished as well.

Index Terms—Ac drives, IM drives, Sensorless control, open-loop estimators, speed estimation.

I. INTRODUCTION

INVERTER-FED induction motor (IM) drives find broad application in industrial plants and household appliances. Due to the inherent low cost of the machine manufacturing, the focus is on the control system, which makes the difference in terms of performance and cost. Highly-demanding applications require an accurate torque control, which is a prerogative of closed loop vector control, either sensed or sensorless. The latter is going to be preferred to get a reliable and maintenance-free system, in spite of some increase in algorithm complexity. The manifold proposals present in the literature generally include closed-loop observers, in which the speed guess is forced to track a null error between estimated and real values of a measured variable, which is usually the stator current vector. The category comprises Luenberger Observers [1] and Kalman Filters, which are characterised by complex calculations and tuning difficulties [2], [3]. Also Model Reference Adaptive Systems (MRAS), based on full or reduced order observers are closed-loop algorithms, inherently non-linear in both the error processed by the adaptive controller and in the mathematical model modified by the feedback contribute [4], [5], [6].

MRAS-based algorithms are often mentioned as an interesting solution for industrial applications, since their implementation is simple in spite of their non-linearity. Anyway, the

stability issues in the motoring, generating and braking modes of operation calls for more elaborate solutions, which include on-line parameter estimation [7]. A solution which uses only single current and single voltage DC-link sensors to operate the vector control is reported in [8]. The main contribution is the improvement of the single current sensor technique to compute the motor phase currents. Anyway, this is obtained at the cost of a space vector modulation pattern modification and, again, the implementation of a full-order state observer.

On the opposite side of cost, complexity and performances, there is the conventional open-loop scalar (V/Hz) control. It is a well-established approach, appreciated for its extreme simplicity and inherent sensorless operations. Nevertheless, as a matter of fact, several less-demanding applications are shaping the research toward the improvement of the low-cost, low-performance scalar control schemes by exploiting the often available current measurements, as a trade-off with the complexity of vector counterparts [9]. A bright example, aimed to close the gap between high and low performance drives, remains that proposed in [10]. Actually, the estimated quantity was the slip frequency, rather than the motor speed, and it was obtained under the assumption that the rotor flux was constant or slowly varying. The drive control was still of scalar type, with inner slip control loop. The rotor flux linkage was obtained from the back electromotive forces (*bemfs*) without any integration, to skip the well-known drift problem. Anyway, *bemfs* were computed from measured voltages, currents, and their (delicate) derivatives. The interest in the possible migration from the open-loop scalar scheme to vector control is testified in [11]. In [12], the traditional V/Hz scheme is improved by adding a slip frequency and stator voltage drop compensation. The slip frequency calculator uses the stator flux, and the drift problem is faced by replacing the pure back *emf* integrator with a programmable cascaded low-pass filter [13]. Experimental results show that the method can improve the induction machine speed accuracy and output torque, even if the presence of uncommented low-pass filters in the estimates recalls some stability problem. A similar principle of operation is proposed in [14], where a modified flux observer is employed to estimate the stator flux with the voltage command and the feedback current. To improve the pure integration problem a high-pass filter, with a cut-off frequency proportional to the voltage frequency, is adopted.

In this paper, a conventional closed-loop vector control is merged with an improved open-loop speed estimation algorithm, to get a smart and low-cost sensorless solution which

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can easily extend existing sensed field-oriented control solutions without the need of a complete review of the drive itself. The mathematical description of the speed estimator can be seen as the natural and intuitive evolution of the one described in [15].

The work originates from the conversion activity of an existing commercial field oriented control (FOC) induction motor drive of sensed type to its sensorless version, using the same power and control hardware and with a minimum increase of the computational needs. The task led to the choice of an appropriate open loop speed estimator that resulted to be a satisfactory compromise between performance and cost. It is well known that parameter sensitivity is a crucial aspect in open loop estimators (as in any model-based control or identification algorithm). Therefore great efforts have been put on a comprehensive analysis of the subject by a rigorous analytical approach that is the key topic of the paper. The study has been formulated with large generality to allow its application to other similar problems and it has been tested on different motors.

In perspective, the exact knowledge of the link between each parameter variation and its effects on the estimated quantity might have great potentiality for novel parameters mismatch detection and auto-tuning algorithms. In all the cited works, the parameter sensitivity is often mentioned as a delicate issue, but its analysis is never carried out systematically and rarely in analytical way.

The noteworthy attention has been put on the voltage inverter dead times compensation, which has been performed by the repetitive-based algorithm described in [16]. Actually, the availability of precise voltage references turns out to be a key point for a successful extension of the operating speed range without the need of external voltage sensors, as the experimental phase of the work has pointed out.

The complete IM drive has been first carefully simulated and then successfully tested on an experimental fast control prototyping test bench.

II. THE OPEN-LOOP ESTIMATOR

The proposed open-loop estimator is based on the four-parameters equivalent circuit of a generic IM, depicted in Figure 1.

Figure 1. Four-parameters electrical equivalent circuit of a generic IM.

In Figure 1 the parameters are the stator resistance R_s , the transient inductance L_t , the magnetising inductance L_ϕ and the rotor resistance referred to the stator R_{sr} . The variables i_ϕ^s , i_{sr}^s and λ_{sr}^s are the magnetising current vector, the rotor current vector referred to the stator, and the rotor flux linkage

vector referred to the stator, respectively. The following relationships hold between i_{sr}^s , λ_{sr}^s , the stator current vector i_s^s and the stator flux linkage vector λ_s^s :

$$i_{sr}^s = \frac{\lambda_{sr}^s}{L_\phi} - i_s^s \quad (1)$$

$$\lambda_s^s = L_t i_s^s + \lambda_{sr}^s \quad (2)$$

If stator flux λ_s^s is calculated from the stator model as the time integral of the voltage u_s^s , from (2) it follows:

$$\lambda_{sr}^s = \int_{-\infty}^t (u_s^s - R_s i_s^s) dt - L_t i_s^s \quad (3)$$

Flux linkage computation represents a fundamental issue in either high-performance or sensorless drives. Nevertheless, the digital implementation of the integrator in (3) still represents a challenge [17], [18]. Due to the drift and the unknown initial value, a pure integration is unfeasible. In this work, the problem has been coped with care, since open-loop estimator performances are heavily affected by the integrator behaviour. In the experimental stage of the work, it has been found that the solution adopted in [18] enables good results even at low speeds. Paper [18] is a feedback-based solution which exploits the orthogonality between flux and back electromotive force to compensate potential offset errors at the integrator input. It seems to be less dependent on machine parameters as the solution [17] is, even if it is required to tune a PI regulator by a rule of thumb.

The mathematical model of the open-loop sensorless estimation can be derived from the right (rotor) side of the four-parameters circuit of Figure 1. One can write

$$0 = R_{sr} i_{sr}^s + \frac{d\lambda_{sr}^s}{dt} - j\omega_{me} \lambda_{sr}^s \quad (4)$$

Substituting (1) into (4) gives

$$\frac{d\lambda_{sr}^s}{dt} = \left(j\omega_{me} - \frac{R_{sr}}{L_\phi} \right) \lambda_{sr}^s + R_{sr} i_s^s \quad (5)$$

The expression (5) can be written in the bi-dimensional stationary reference frame $\alpha - \beta$:

$$\begin{cases} \frac{d\lambda_{sr\alpha}^s}{dt} + \frac{R_{sr}}{L_\phi} \lambda_{sr\alpha}^s = -\omega_{me} \lambda_{sr\beta}^s + R_{sr} i_{s\alpha}^s \\ \frac{d\lambda_{sr\beta}^s}{dt} + \frac{R_{sr}}{L_\phi} \lambda_{sr\beta}^s = \omega_{me} \lambda_{sr\alpha}^s + R_{sr} i_{s\beta}^s \end{cases} \quad (6)$$

Solving the first of (6) for ω_{me} gives

$$\omega_{me,1} = - \frac{\left(-R_{sr} i_{s\alpha}^s + \frac{d\lambda_{sr\alpha}^s}{dt} + \frac{R_{sr}}{L_\phi} \lambda_{sr\alpha}^s \right)}{\lambda_{sr\beta}^s} \quad (7)$$

Both (3) and (7) represent the base for an open-loop speed estimator, as proposed in [15]. The parameters L_ϕ/R_{sr} (which is equal to the rotor time constant $\tau_r = L_r/R_r$), R_s , R_{sr} and L_t are measured at motor terminals by conventional tests [19].

The weakness of (7) resides its sinusoidal numerator and denominator. In ideal conditions, their ratio would be always equal to the electromechanical speed. As a matter of fact, however, the concurrence of approximate parameter knowledge and digital rounding cause heavy estimation errors, as

the denominator approaches to zero. This effect is depicted in Figure 2, which reports a simulation result of a digital open-loop speed estimator based only on (7) and (3), during a ramp to one-eighth of its nominal speed. The simulated IM drive was a standard (sensored) FOC drive, with feedback loop closed on measured speed, while the sensorless algorithm was running apart, tuned with the correct motor parameters.

Figure 2. Open-loop speed estimator: (a) speed reference, (b) actual speed, (c) estimated speed based on (3) and (7), (d) estimated $\lambda_{sr\beta}^s$.

An adequate countermeasure consists in getting ω_{me} directly from (5), without the intermediate split in its real and imaginary parts. It is immediately:

$$\omega_{me} = -j \frac{\frac{R_{sr}}{L_\varphi} \lambda_{sr}^s - R_{sr} i_s^s + \frac{d\lambda_{sr}^s}{dt}}{\lambda_{sr}^s} \quad (8)$$

Rationalizing (8) and solving for the real part only (since ω_{me} is real) gives:

$$\hat{\omega}_{me} = \frac{R_{sr} i_{s\alpha}^s \lambda_{sr\beta}^s - \frac{d\lambda_{sr\alpha}^s}{dt} \lambda_{sr\beta}^s - R_{sr} \lambda_{sr\alpha}^s i_{s\beta}^s + \lambda_{sr\alpha}^s \frac{d\lambda_{sr\beta}^s}{dt}}{\lambda_{sr\alpha}^s{}^2 + \lambda_{sr\beta}^s{}^2} \quad (9)$$

This mathematical formulation avoids every division by zero, since the denominator is the squared magnitude of the rotor flux linkage, referred to the stator. As a result, the expression (9), combined with the flux linkage derivatives obtained from (3), represents a net improvement in the speed estimator algorithm.

Equation (9) somewhat recalls the slip estimation technique reported in [10], although it does not depend upon magnetising inductance as in the slip estimator. The basic difference is that the slip estimator is meant to provide an effective way to realise an inner slip control loop, without any concern about the electromechanical speed, which is necessary to close an outer speed loop. On the contrary, (9) is intended to be used as a computationally-light estimated feedback in the speed closed-loop control of a standard FOC drive.

III. SENSITIVITY ANALYSIS

As anticipated, the proposed model for the speed estimate has improved parameter sensitivity with respect to that reported in [15]. A detailed sensitivity analysis of (3) and (9) is carried out hereafter, with the aim of investigating the effects of parameters mismatch on the speed estimation.

Let us consider a generic real function $f(x, p)$, where x is a variable (or a set of variables) and p is a real positive parameter. At large, the sensitivity function is defined as the ratio between the relative variation of the function f , which is df/f , with respect to the relative variation of p , which is dp/p . Therefore, the sensitivity is computed as the total derivative of f with respect to p , multiplied by p/f :

$$\tilde{S}_f^p(x, p) = \frac{\frac{df}{f}}{\frac{dp}{p}} = \frac{df}{dp} \frac{p}{f} \quad (10)$$

The sensitivity analysis is a powerful tool during the process of selection of the best control algorithm. It is worth to note that (10) has an inherent singularity as the function f gets near to zero. This fact, which is obvious in principle, does not depend upon the specific control algorithm, and it may alter a correct comparison of the sensitivities around $f = 0$. The problem can be tackled by using a slightly different expression, in which the variations of f and p are related to their nominal values, instead of the actual ones:

$$S_f^p = \frac{df}{dp} \frac{p_N}{f_N} \quad (11)$$

Clearly, a sensitivity function equal to one indicates that a percentage error on the parameter p would be reflected to an equivalent percentage error on f . Consequently, the more the sensitivity function approaches zero, the less the function f is influenced by an error on parameter p .

As a first example, the $S_{\omega_{me,1}}^{R_s}$ function for (3) and (7) have been obtained by applying the total derivative of (7) with respect to R_s . In details,

$$S_{\omega_{me,1}}^{R_s} = \frac{d\omega_{me,1}}{dR_s} \frac{R_s}{\omega_{me,1}} = \frac{R_s}{\omega_{me,1}} \left(\frac{d\omega_{me,1}}{d\lambda_{sr\alpha}} \frac{d\lambda_{sr\alpha}}{dR_s} + \frac{d\omega_{me,1}}{d\lambda_{sr\beta}} \frac{d\lambda_{sr\beta}}{dR_s} + \frac{d\omega_{me,1}}{d\left(\frac{d\lambda_{sr\alpha}}{dt}\right)} \frac{d\left(\frac{d\lambda_{sr\alpha}}{dt}\right)}{dR_s} \right) \quad (12)$$

Let us define I_s as the magnitude of the current vector, Λ_{sr} as the magnitude of the rotor flux linkage referred to the stator, ω_s as the stator exciting frequency and φ is the angle between i_s^s and λ_s^s , the steady-state expression of $S_{\omega_{me,1}}^{R_s}$ gives:

$$S_{\omega_{me,1}}^{R_s} = R_s R_{sr} I_s \frac{L_\varphi [1 + \cos(2\omega_s t)] - 2\Lambda_{sr} \cos(\varphi)}{L_\varphi \omega_{me,N} \omega_s \Lambda_{sr}^2 [\cos(2\omega_s t - 2\varphi) - 1]} \quad (13)$$

The function $S_{\omega_{me,1}}^{R_s}$ is time-dependant, so that a R_s mismatch causes oscillations in the speed estimation. Similarly, it has been verified that also the sensitivity functions with respect to R_{sr} , L_t and L_φ are time-dependant. As a consequence, even small deviations in each parameter lead to oscillatory speed estimations. This is not the case of the proposed estimate, as reported in the next Sections.

A. Stator resistance R_s

The parameter R_s appears only in (3), so that the sensitivity can be obtained by applying the total derivative of (9) with

respect to R_s :

$$S_{\hat{\omega}_{me}}^{R_s} = \frac{R_s R_{sr} I_s^2 \cos(2\varphi)}{\omega_s \omega_{me,N} \Lambda_{sr}^2} \quad (14)$$

Opposite to (13), it does not depend upon time, and the closer the speed to zero, the higher the relative weight of the resistance over the inductive terms in the estimate. Conversely, at high speed the influence of R_s variations vanishes. This is reported in Figure 3, which shows $S_{\hat{\omega}_{me}}^{R_s}$ as function of ω_s , for different load torque values. As can be seen, for medium loads the sensitivity can be very low. This is a particular feature of the proposed algorithm.

Figure 3. $S_{\hat{\omega}_{me}}^{R_s}$ as function of ω_s : (a) no load, (b) 33% rated load, (c) 67% rated load, 100% rated load.

B. Rotor resistance referred to the stator R_{sr}

Since R_{sr} appears only in (9), the steady state sensitivity function is simply:

$$S_{\hat{\omega}_{me}}^{R_{sr}} = -\frac{R_{sr} I_s \sin(\varphi)}{\omega_{me,N} \Lambda_{sr}} \quad (15)$$

As expected, at no load ($\varphi = 0$) the sensitivity is null, since the steady state torque-producing component of the current (I_{sr}), which flows through R_{sr} (Figure 1), is null, too. Any change in the parameter shall not affect the estimated speed.

C. Transient inductance L_t

Similarly to what happens for R_s , the transient inductance L_t only appears in (3). The sensitivity function is then

$$S_{\hat{\omega}_{me}}^{L_t} = -\frac{R_{sr} L_t I_s^2 \sin(2\varphi)}{\omega_{me,N} \Lambda_{sr}^2} \quad (16)$$

It is worth to note that both $S_{\hat{\omega}_{me}}^{R_{sr}}$ and $S_{\hat{\omega}_{me}}^{L_t}$ do not vary with speed in the whole constant-torque region, while they are affected by flux-weakening operations. Increasing values of sensitivity are experienced for raising load torque, as reported in Figure 4, since both terms I_s and φ increase with load.

D. Magnetising inductance L_φ

The magnetising inductance L_φ does not appear neither in (3) nor in (9), as a feature of the adopted model. Consequently, the sensitivity function is null for all operating conditions:

$$S_{\hat{\omega}_{me}}^{L_\varphi} = 0 \quad (17)$$

Figure 4. (a) $S_{\hat{\omega}_{me}}^{R_{sr}}$, (b) $S_{\hat{\omega}_{me}}^{L_t}$ as function of load torque.

IV. SIMULATION AND EXPERIMENTAL RESULTS

The proposed sensorless scheme has been tested on IM 1 (see Appendix A) in several different working conditions, with both simulations and experimental tests, with the aid of a fast control prototyping (FCP) test bench. Simulations have been carried out using a linear model for IM 1 and an accurate model for inverter non-ideality, which comprises dead-time effect and voltage drop across diodes and IGBTs [20]. Simulations have been also made on a larger motor IM 2, whose parameters are included in Appendix A.

A. The laboratory test-bench

The FCP core control unit is essentially composed by a plug-in control board (manufactured by DSpace) with two DSPs in a master-slave configuration. Master processor (a PowerPC 603e floating point unit running at 250 MHz) is responsible of all main operations in the control algorithm, while the slave processor (a Texas Instruments TMS320F240 fixed point unit running at 20 MHz) is entirely dedicated to PWM generation and actual speed measurement. The control algorithm has been written in C-language, for a closer link to the industrial practice. Results have been obtained by means of the user-friendly trace options of the control board and a Tektronix DPO4034 oscilloscope, saved on a standard PC for graphic post-processing.

The estimation algorithm has been merged with a standard full-digital Field Oriented Control (FOC) IM drive with a switching frequency of 10 kHz. Inverter dead-times were of $4 \mu s$. The direct FOC scheme [15] has been preferred to the indirect one, as the former does not use the estimated speed for the inner current control and thus it is, in principle, less sensitive to estimate errors. Different load torques have been generated by a 1.6 kW - 5.3 Nm brushless motor, acting as a brake and driven by a standard torque-controlled drive, coupled to the induction motor shaft by means of a Magtrol TM308 in-line torque transducer.

B. The dead-time effects

In the computation of (3), the measured stator voltages have been replaced by the references $u_{s\alpha}^*$, $u_{s\beta}^*$, generated by the PI regulators of the FOC drive. However, the substitution would yield poor performances if a dead-time compensation were omitted. Figure 5 reports the real and estimated speed at

steady-state in case of no compensation. Evidently, a large steady-state error occurs, while a 6-th harmonics effect is superimposed to the estimation outline. These well-known effects are described, for example, in [16].

Figure 5. Effects of dead time on speed estimation (experimental): (a) speed reference, (b) estimated speed, (c) actual speed.

In this paper, the repetitive control self-commissioning compensation procedure described in [16] has been adopted. A brief description of the compensation algorithm is reported in Appendix B. The dead-time effects are greatly reduced, as shown in Figure 6. The availability of good (estimate) voltages turns out to be of great importance in extending the operational range of the sensorless algorithm at low speed, where dead-time effects are dominant.

Figure 6. Speed estimation after dead time compensation (experimental): (a) speed reference, (b) estimated speed, (c) actual speed.

C. Dynamic behaviour

Different speed reference profiles have been applied to the proposed estimator to evaluate its characteristics at various speed levels, both at no load and with applied load torques. The method has been firstly tested on IM 1, by both simulations and experimental results. Figure 7 and 8 show, respectively, simulation and experimental results of a series of accelerations at no load, with different intermediate speed levels and a speed ramp set to 2000 rpm/s . The experiment is repeated in Figure 9, with a load torque of 100% of the rated torque and a ramp of 500 rpm/s . For the sake of

completeness, Figure 10 shows a small-step speed response of the sensorless drive with a load torque of 30% of the rated torque.

Figure 7. IM 1 speed rise (ramp of 2000 rpm/s , no load, simulation): (a) estimated speed, (b) actual speed.

Figure 8. IM 1 speed rise (ramp of 2000 rpm/s , no load, experimental): (a) estimated speed, (b) actual speed.

Figure 9. Speed rise (ramp of 500 rpm/s , 100% of rated load, experimental): (a) estimated speed, (b) actual speed.

Figure 11 and 12 show respectively the simulation and experimental results of the speed behaviour during a load onset and detach, with a load torque value of 100% of the rated load.

Figure 10. IM 1 speed step response (30% of rated load, experimental): (a) reference speed, (b) estimated speed, (c) actual speed.

Figure 11. IM 1 speed profile at load onset and detach (100% of the rated load, simulation): (a) reference speed, (b) estimated speed, (c) actual speed.

Figure 12. IM 1 speed profile at load onset and detach (100% of the rated load, experimental): (a) reference speed, (b) estimated speed, (c) actual speed.

Finally, Figure 13 and 14 show the simulation and experimental results of the speed behaviour during a speed reversal, with an unidirectional load torque of 100% of the rated load and a speed ramp of 500 rpm/s . Once again, as in all the previous comparisons, one can observe a tight correspondence between expected (simulated) and actual (experimental) behaviours.

Figure 13. IM 1 drive behaviour from motor mode to generator mode (ramp of 500 rpm/s , 100% of the rated load, simulation): (a) reference speed, (b) estimated speed, (c) actual speed.

Figure 14. IM 1 drive behaviour from motor mode to generator mode (ramp of 500 rpm/s , 100% of the rated load, experimental): (a) reference speed, (b) estimated speed, (c) actual speed.

D. Simulations on IM 2

The simulations carried out for the small IM 1 have been repeated for the larger IM 2, using a 6 kHz PWM with $4 \mu\text{s}$ inverter dead times. Figure 15 reports the speed rise with a ramp of 2000 rpm/s at no load, while Figure 16 shows the speed behaviour at load onset and detach with 100% of the rated load. Finally, Figure 17 shows the drive behaviour during a speed reversal, with a ramp of 500 rpm/s and at full load. They reassure about the validity of the proposed drive for IM of very different sizes and parameters.

E. Parameter sensitivity

An experimental sensitivity analysis has been performed on the proposed sensorless scheme for IM 1. Waveforms have been obtained by forcing a step variation to each model parameter of the estimator. As a first test, a R_s variation of -40% at no load and at 67% of rated load has been forced, at a speed of 0.333 p.u. . Figure 18 and 19 report the relative effects on the estimated speed. The variations are opposite in sign, according to Figure 3. Actually, the experimental results of Figure 18 and 19 highlight a quantitative effect slightly lower than expected, especially at higher loads. This can be explained by considering that the sensitivity function

Figure 15. IM 2 speed rise (ramp of 2000 rpm/s, no load, simulation): (a) estimated speed, (b) actual speed.

Figure 16. IM 2 speed profile at load onset and detach (100% of the rated load, simulation): (a) reference speed, (b) estimated speed, (c) actual speed.

Figure 17. IM 2 drive behaviour from motor mode to generator mode (ramp of 500 rpm/s, 100% of the rated load, simulation): (a) reference speed, (b) estimated speed, (c) actual speed.

is essentially a local derivative and then, in the case of wide parameter variation, as that of 40 %, the linearity is not guaranteed. This has been proven by a set of simulations, which has also confirmed that, on the other hand, the effects of small R_s variations perfectly match the figures returned by (14).

Figure 18. Speed profile with -40% variation of R_s (no load, experimental): (a) reference speed, (b) estimated speed, (c) actual speed.

Figure 19. Speed profile with -40% variation of R_s (66% of rated load, experimental): (a) reference speed, (b) estimated speed, (c) actual speed.

It is worth to note that the superimposition principle does not hold for every combination of parameter variations. In particular, it has been experienced that simultaneous variations of resistances (which is reasonable, since they both depend on temperature) produce a speed error lower than the arithmetic sum of (14) and (15). The effect, at time zero, of a simultaneous decrease of 40% of both R_s and R_{sr} values in the estimator is reported in Figure 20, under 50% of load torque. The decrease of parameter values in the model simulates the temperature effect, which in fact increases the real motor resistances. The global mean error is equal to -0.00055 pu (equal to -0.055% of the nominal speed), which is less than a half than expected, and anyway quite below the usual specifications for the speed loop accuracy. A different mechanical setup has been used to produce Figure 20 with an IM similar to IM1 but with 2 poles and 1.2 Nm rated torque.

Last, the method is also quite insensitive to saturation effects, because of the intrinsic low values of (16) in almost the

Figure 20. Speed profile with -40% variation of R_s and R_{sr} : (a) reference speed, (b) estimated speed, (c) actual speed.

whole speed range, and also because L_t represents a leakage inductance, normally unaffected by saturation.

V. CONCLUSIONS

An improved open-loop speed estimation technique for low-cost sensorless IM drives has been presented and discussed in detail. First, an existing open-loop method, described in [15], has been analysed and tested. Then, the method has been improved by both rearranging the model equations and a careful compensation of voltage inverter dead times. The resulting drives show enhanced performances, in terms of both speed range and parameter sensitivity. An accurate mathematical analysis followed by a series of simulation and experimental results have been included in the paper, proving the attractive features of the proposed sensorless control algorithm. Not least, the added value of the paper may reside also in the fair correspondence between expected and actual results of the parameter sensitivity analysis, which could suggest some new ideas for parameter mismatch automatic detection and on-line tuning.

APPENDIX A

IM ELECTRICAL PARAMETERS

Parameter	IM 1	IM 2
Nominal speed ω_N	1500 rpm	2000 rpm
Nominal torque τ_N	1.9 Nm	4386 Nm
Pole pairs p	2	2
Stator resistance R_s	6.49 Ω	4.52 m Ω
Rotor resistance referred to the stator R_{sr}	4.96 Ω	7.22 m Ω
Magnetising inductance L_φ	0.230 H	9.10 mH
Transient inductance L_t	0.027 H	0.49 mH

APPENDIX B

DEAD-TIME COMPENSATION ALGORITHM

As previously stated in the paper, motor voltages can be either measured or estimated from the voltage references. As the direct measure is usually discarded, due to the PWM nature of phase voltages, the latter seems a more viable

choice. To maintain the necessary precision, the difference between references and actual voltages has been reduced by the compensation technique based on repetitive control described in [16].

The key feature lies in the periodical nature of the voltage distortion, which can be profitably expressed in a synchronous dq rotating frame, locked to the phase ϑ_i of the current space vector, as:

$$u_{dist,d}(\vartheta_i) = \frac{4t_d}{\pi T_c} U_{dc} \left[1 - \sum_{n=1}^{+\infty} (-1)^n \frac{2}{36n^2 - 1} \cos(6n\vartheta_i) \right]$$

$$u_{dist,q}(\vartheta_i) = \frac{4t_d}{\pi T_c} U_{dc} \sum_{n=1}^{+\infty} (-1)^n \frac{12n}{36n^2 - 1} \sin(6n\vartheta_i) \quad (\text{B.18})$$

where T_c is the PWM switching period, U_{dc} is the dc-link voltage, and t_d is equal to:

$$t_d = t_{off} - t_{on} - t_{ds} \quad (\text{B.19})$$

where t_{off} and t_{on} are the fall and rise times of the inverter power switches, and t_{ds} is the safe interval between the commutation edges of the upper and lower devices in the inverter leg. Aside from the constant term in $u_{dist,d}$, the distortion can be identified as a sum of $6n$ -th harmonic sinusoidal components ($n = 1, \dots, +\infty$). Also the effect of parasitic capacitances of IGBTs, during the commutation at low current levels, can be viewed as a periodical disturbance. Since there are six zero crossings during a 2π rotation of a synchronous frame, the related distortion effect is well approximated by a sum of $6n$ -th harmonics ($n = 1, \dots, +\infty$), which overlap the harmonics coming from (B.18).

In this way, the repetitive control is exploited to compensate the $6n$ -th harmonics and generate application-specific look-up tables (LUTs) during a first self-commissioning phase. The generated compensation voltage for the natural abc reference frame are stored into three LUTs, which will substitute the conventional step-like compensation technique at low current levels. For higher current levels, LUTs are merged with the conventional step-like compensation, and then used in the whole speed range. This algorithm improves the correspondence between reference voltages and actual values, so that the formers can be profitably used in the proposed electromechanical conversion model.

As example, Figure 21 shows a comparison between reference and actual voltages in case of no compensation, classical step-like compensation and used compensation at a speed of 0.015 p.u. with a simple scalar V/Hz algorithm.

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Figure 21. Reference and actual voltages. Up: without any compensation. Middle: conventional compensation. Bottom: proposed LUT-based technique.

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