

Fractional-order observer for model-free predictive current control of eighteen-pulse-fed induction motor drives

SALIMA KHAOULA REGUIEG^{1*}, MOHAMED NOUR², ABDELKRIM BENALI², NOUREL HOUDA CHEGGOUFI³

¹LSETER Laboratory, Nour El-Bachir University Centre
El Bayadh, Algeria

²Department of Electrical Engineering Nour El-Bachir University Centre
El Bayadh, Algeria

³LIMA Laboratory, Nour El-Bachir University Centre
El Bayadh, Algeria

e-mail: {s.reguieg/m.nour/a.benali/n.cheggoufi}@cu-elbayadh.dz

Abstract: High-performance induction motor drives are penalised at both terminals: on the supply side by low-order current harmonics injected by the diode front end, and on the machine side by the parameter dependence and the weighting-factor tuning burden of finite-control-set model predictive control. This paper studies a terminal-to-terminal architecture that addresses both. An eighteen-pulse phase-shifting autotransformer rectifier cancels every characteristic harmonic below the seventeenth, and its residual dc-link ripple is measured and fed forward into the machine-side observer. The machine-side controller replaces the motor model by a first-order ultra-local model whose lumped disturbance is reconstructed by a fractional-order extended state observer, while the ultra-local input gain is identified on line by a gradient law; the resulting predictive current law is single-objective and therefore needs no weighting factor. Convergence is established with a Lyapunov argument and Matignon's criterion. Simulation of a 4-kW drive shows the eighteen-pulse front end reducing the idealised supply-current distortion from 30.8% to 9.8%, and a stator-current distortion of 1.34% for a controller that uses no machine parameter at all, against 1.53% for a well-tuned model-based benchmark. Three deliberately negative findings are reported: the input-gain sensitivity is strongly asymmetric, so that the gain should be biased low rather than centred; the fractional order brings no measurable benefit over the integer observer under the bandwidth parameterisation used here; and the ripple feed-forward is largely redundant once the eighteen-pulse rectifier is present.

Key words: eighteen-pulse rectifier, fractional-order extended state observer, induction motor drive, model-free predictive current control, total harmonic distortion, ultra-local model

1. Introduction

The induction motor remains the workhorse of industrial variable-speed drives, yet the power-electronic interface that energises it limits performance at both of its terminals. At the point of common coupling, the front-end diode rectifier of a conventional voltage-source drive draws a quasi-square current rich in harmonics of order $6k \pm 1$, routinely exceeding the five-percent current-distortion ceiling of the harmonic-control standard [12]. On the machine side, the finite-control-set model predictive control paradigm, although prized for its conceptual transparency and its fast-transient response, inherits two structural weaknesses. The literature has addressed these three deficiencies in three separate communities, and the present paper is an attempt to treat them as one problem.

Model-based predictive current and torque control predict the future machine state through a discrete model parameterised by stator and rotor resistances, leakage and magnetising inductances, and rotor flux. Under thermal drift, magnetic saturation and skin effect these parameters migrate, the prediction degrades, and both steady-state error and ripple grow [1, 8], [23]. The remedies proposed fall into two families. The first estimates the offending parameters on line, by recursive least squares, Kalman filtering or model-reference adaptive schemes; the estimator then inherits an observability requirement of its own, and its transient is superimposed on the transient of the drive. The second family discards the model.

To decouple the controller from the machine, the ultra-local model of model-free control replaces the physical model by a first-order input-output relation whose lumped, time-varying disturbance absorbs all parametric and unmodelled dynamics [9]. Reconstructing that disturbance with an extended state observer yields model-free predictive current control, now established for permanent-magnet machines [3], doubly-fed generators [4] and, more recently, induction motors [5, 6, 22]. Two residual dependencies persist. The ultra-local input gain is itself a function of the stator inductance and is almost universally fixed to a nominal constant, so that the claim of a parameter-free controller is, strictly speaking, incomplete. Moreover, the conventional integer-order observer is reputed to exhibit poor steady-state disturbance observation and pronounced sensitivity to measurement noise [10], which has motivated a growing interest in fractional-order extensions of the extended state observer [11, 24].

The second weakness is the weighting factor. Predictive torque control minimises a cost function that aggregates torque error and flux error, quantities of incommensurable units. Their relative importance is fixed by a weighting factor whose optimal value is operating-point dependent and remains, by broad consensus, an open problem [2, 7]. A substantial literature is devoted to selecting it — by exhaustive search, by algebraic argument, by fuzzy inference or by neural approximation — and a smaller literature is devoted to eliminating it, by sequential evaluation of the cost terms, by ranking, or by reformulating the objective in a single physical unit. The reformulation route is the cleanest, because it removes the degree of freedom rather than automating its tuning.

A third gap concerns the supply. The two strands above treat the machine-side controller in isolation, implicitly assuming a stiff, ripple-free dc link. In practice a distorted supply

propagates second- and sixth-harmonic ripple onto the bus, which the inverter reproduces as torque pulsation. Multi-pulse rectifiers suppress supply harmonics passively and without the resonance hazards of tuned filters, and the eighteen-pulse phase-shifting autotransformer rectifier offers an attractive compromise between the twelve-pulse and the twenty-four-pulse topologies [13, 14]. An alternative, active route regulates the dc-link voltage itself through a controlled rectifier, which tunes the voltage vector available to the machine-side switching table and thereby attenuates torque and flux ripple at its origin [15]. Their interaction with a model-free inverter controller, however, has not been examined as an integrated whole.

This paper assembles the two ends into one architecture and then subjects each of its claimed benefits to a numerical test. The contributions are the following.

A terminal-to-terminal topology is proposed in which the measured dc-link ripple is routed into the machine-side observer as a known disturbance, so that the front end and the current controller are designed against a common disturbance budget.

A fractional-order extended state observer is formulated for the induction motor ultra-local model, with a bandwidth parameterisation that preserves dimensional consistency and reduces to the classical linear extended state observer as the order tends to unity. Its convergence is established by Matignon's criterion.

An online input-gain adaptation law is derived from a Lyapunov argument, with the intent of removing the last machine parameter from the current loop.

A single-objective predictive current law is used in which the weighting factor is structurally absent rather than merely tuned, because both terms of the cost function are currents.

Equally important, and in the interest of a faithful record, three of the mechanisms above are found *not* to deliver what the literature would lead one to expect, and this is reported as such in Section 5. All numerical values quoted in this paper are produced by the simulation described in Section 4; none is transcribed from another source, and no experimental claim is made.

2. System topology and model

The drive is shown in Fig. 1, a phase-shifting autotransformer feeds three six-pulse diode bridges whose outputs are paralleled onto a common dc bus; a two-level voltage-source inverter supplies the induction machine. The measured bus voltage is made available to the machine-side observer.

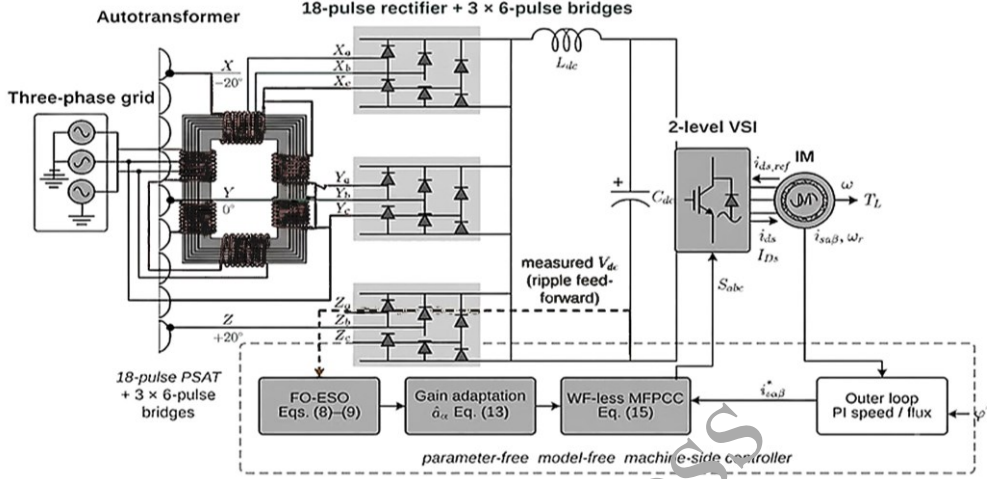


Fig. 1. Proposed hybrid drive. An eighteen-pulse phase-shifting front end supplies a two-level inverter; the measured dc-link ripple is fed forward into the fractional-order observer that drives the weighting-factor-free predictive current controller

Figure 1 shows the complete component-level topology of the drive: the three-phase grid feeds the phase-shifting autotransformer, whose three secondary sets displaced by -20° , 0° and $+20^\circ$ electrical degrees each supply a six-pulse diode bridge; the three bridge outputs are paralleled through interphase reactors onto the common dc link, which supplies the two-level voltage-source inverter and the induction machine. The lower layer of the figure is the weighting-factor-free, model-free machine-side controller, comprising the fractional-order extended state observer, the online input-gain adaptation, and the predictive current law, together with the measured dc-link ripple that is fed forward into the observer.

2.1. Eighteen-pulse phase-shifting front end

A phase-shifting autotransformer generates three balanced secondary sets displaced by $\delta_k \in \{-20^\circ, 0^\circ, +20^\circ\}$, each feeding a six-pulse diode bridge; the outputs are paralleled through interphase reactors onto the common bus. The reactors absorb the instantaneous voltage differences between the three bridges and force an approximately equal current sharing, so that the primary current is the phasor sum of three equal, mutually displaced bridge currents. Referred to the primary, the h -th harmonic of the three bridge currents recombines with the factor

$$c_h = 1/3 |1 + 2\cos[(h \mp 1)\delta]|, \quad \delta = 20^\circ, \quad (1)$$

in which the upper sign applies to harmonics of order $6m + 1$ and the lower sign to those of order $6m - 1$. The factor vanishes for every characteristic harmonic except

$$h = 18k \pm 1, \quad k = 1, 2, \dots, \quad (2)$$

so that the lowest surviving orders are the seventeenth and the nineteenth. The fifth, seventh, eleventh and thirteenth harmonics, which dominate the six-pulse spectrum and which the harmonic standard penalises most heavily, are cancelled by superposition rather than by dissipation.

The cancellation can be read directly from the physical action of the transformer. The primary winding carries the vector sum of three secondary currents that are equal in magnitude but mutually displaced by twenty electrical degrees. For a harmonic of order h the three contributions are phase-shifted by $(h-1)\delta$ and $(h+1)\delta$; when h is a characteristic six-pulse harmonic other than $18k \pm 1$ the three phasors are spaced by 120 electrical degrees and sum to zero, whereas for $h = 18k \pm 1$ they are co-phased and add constructively. The autotransformer therefore realises a passive, loss-free vector cancellation of the fifth, seventh, eleventh and thirteenth harmonics, which is the mechanism quantified by the factor of Eq. (1). Because the unit is an autotransformer rather than an isolation transformer, only the differential volt-amperes between primary and secondaries are processed magnetically; the through power flows directly. For an eighteen-pulse phase-shifting autotransformer the processed rating is typically in the range of one fifth to one third of the load rating, so the added magnetic cost is of the order of 20 to 35 percent of a plain six-pulse front end, and a correspondingly smaller fraction of the total drive cost once the inverter, machine and control electronics are included. This is the price paid for removing the low-order harmonics without the resonance hazards of tuned passive filters or the switching losses and control burden of an active front end.

The bus voltage is not constant. Writing it as a mean value plus a ripple,

$$V_{dc}(t) = \bar{V}_{dc} + \tilde{v}_{dc}(t), \quad \tilde{v}_{dc} = \sum_m \tilde{V}_m \cos(m\omega_g t + \varphi_m), \quad (3)$$

where $\tilde{v}_{dc}$ collects the bus harmonics of order $18k$ and, under a distorted or unbalanced supply, residual terms at $2\omega_g$ and ω_g . Equation (3) is measured and exploited in Section 3. Regulating $\bar{V}_{dc}$ actively, as in [15], is a complementary measure that this paper does not pursue: the front end here is uncontrolled, and only the ripple is exploited. The eighteen-pulse rectifier attenuates the ripple at its source; the feed-forward path attenuates what remains. Whether both are needed is examined in Section 5.5.

2.2. Inverter and machine model

The inverter applies one of eight voltage vectors,

$$\mathbf{v}_i = 2/3 V_{dc} (S_a + S_b e^{j2\pi/3} + S_c e^{j4\pi/3}), \quad S_{a,b,c} \in \{0,1\}, \quad (4)$$

of which two are null. For benchmarking only, the stator-current dynamics in the stationary frame are

$$\frac{d\mathbf{i}_s}{dt} = -\frac{R_s L_r^2 + R_r L_m^2}{\sigma L_s L_r^2} \mathbf{i}_s + \frac{L_m}{\sigma L_s L_r} \left(\frac{1}{T_r} - j\omega_r \right) \boldsymbol{\psi}_r + \frac{\mathbf{v}_s}{\sigma L_s}, \quad (5)$$

with $\sigma = 1 - L_m^2/(L_s L_r)$ and $T_r = L_r/R_r$, where R_s and R_r are the stator and rotor resistances, L_s , L_r and L_m the stator, rotor and magnetising inductances, ψ_r the rotor flux linkage and ω_r the electrical rotor speed. Equation (5) builds the model-based comparator only; the proposed controller never uses it.

2.3. Machine-side ultra-local model

The proposed controller treats the inverter-machine subsystem as a single-input single-output plant per axis,

$$\frac{di_{s\xi}}{dt} = F_\xi(t) + \alpha u_\xi, \quad \xi \in \{\alpha, \beta\}, \quad (6)$$

where u_ξ is the applied voltage component, α is the unknown, inductance-dependent input gain, and the lumped disturbance

$$F_\xi(t) = f_\xi(i_s, \psi_r, \omega_r) + (\alpha - \hat{\alpha})u_\xi + d_{dc,\xi}(\hat{v}_{dc}) + d_{nl,\xi}, \quad (7)$$

absorbs every machine parameter, the gain-identification error, the measured ripple contribution and the inverter nonlinearities such as dead time and device voltage drops. The objective is to reconstruct F_ξ and identify α on line, so that (6) becomes fully known without any machine parameter. The price of the ultra-local formulation is that F_ξ is fast: it contains the back-electromotive force, and therefore varies at the fundamental frequency and above. The observer must consequently be fast in turn, which is what the extended state formulation of the next section provides.

3. Proposed control law

3.1. Fractional-order extended state observer

Treat the disturbance as an extended state. With $x_1 = i_{s\xi}$, $x_2 = F_\xi$ and $\dot{x}_2 = h_\xi(t)$ bounded, a fractional integrator of order $\mu \in (0,1]$ is assigned to the disturbance channel:

$$\dot{\hat{x}}_1 = \hat{x}_2 + \hat{\alpha} u_\xi + \beta_1 e_\xi, \quad (8)$$

$$D^\mu \hat{x}_2 = \beta_2 e_\xi, \quad e_\xi = i_{s\xi} - \hat{x}_1, \quad (9)$$

where D^μ is the Caputo derivative [19]. Bandwidth parameterisation [17] places both gains with a single knob $\omega_o > 0$:

$$\beta_1 = 2\omega_o, \quad \beta_2 = \omega_o^{1+\mu}. \quad (10)$$

The exponent $1 + \mu$ preserves the dimensional consistency of the fractional channel and recovers the classical $\beta_2 = \omega_o^2$ as $\mu \rightarrow 1$. With $\tilde{x}_i = x_i - \hat{x}_i$ the commensurate error system reads

$$\begin{bmatrix} \dot{\tilde{x}}_1 \\ D^\mu \tilde{x}_2 \end{bmatrix} = \begin{bmatrix} -\beta_1 & 1 \\ -\beta_2 & 0 \end{bmatrix} \begin{bmatrix} \tilde{x}_1 \\ \tilde{x}_2 \end{bmatrix} + \begin{bmatrix} 0 \\ h_\xi \end{bmatrix}. \quad (11)$$

By Matignon's theorem [18] the unforced system is asymptotically stable if and only if every eigenvalue λ of the state matrix satisfies

$$|\arg(\lambda)| > \frac{\mu\pi}{2}. \quad (12)$$

The characteristic polynomial $\lambda^2 + \beta_1\lambda + \beta_2$ has roots in the open left half-plane for every $\omega_o > 0$; since $\mu \leq 1$, the stability sector of (12) is no smaller than in the integer case, and the estimation error is ultimately bounded by a ball whose radius shrinks with ω_o . The bound is not asymptotic to zero, because h_ξ is bounded but not vanishing; this is the standard limitation of a second-order extended state observer confronted with a time-varying disturbance, and it is visible as the amplitude attenuation reported in Section 5.3.

3.2. Online input-gain adaptation

Define the gain error $\tilde{\alpha} = \alpha - \hat{\alpha}$ and the tracking error $\varepsilon_\xi = i_{s\xi}^* - i_{s\xi}$. A gradient update driven by the tracking error and the applied input,

$$\dot{\hat{\alpha}} = \gamma \sum_{\xi} \varepsilon_{\xi} u_{\xi}, \quad \gamma > 0, \quad (13)$$

together with the Lyapunov candidate $V = 1/2 \sum_{\xi} \varepsilon_{\xi}^2 + 1/2\gamma \tilde{\alpha}^2$, yields $\dot{V} = -k \sum_{\xi} \varepsilon_{\xi}^2 \leq 0$ for a positive loop gain k , so the augmented error system is stable and $\varepsilon_{\xi} \rightarrow 0$ by Barbalat's lemma.

Remark. The Lyapunov result guarantees boundedness of the current error, not exact identification of α . Convergence of the parameter estimate would additionally require persistency of excitation in the regressor u_{ξ} . Because the observer lumps any residual gain error $(\alpha - \hat{\alpha})u_{\xi}$ into $\hat{F}_{\xi}$, the gain is only weakly excited in the prediction error: the observer and the adaptation law compete for the same signal, and the faster of the two wins. Section 5.3 confirms that $\hat{\alpha}$ does not converge to α , and quantifies the consequence.

3.3. Weighting-factor-free predictive current law

With F_{ξ} reconstructed and α estimated, (6) is fully known. A forward Euler discretisation at the control period T_c , with the standard two-step delay compensation, gives the prediction associated with a candidate vector $\mathbf{v}_i$:

$$i_{s\xi}^p(k+1) = i_{s\xi}(k) + T_c [\hat{F}_{\xi}(k) + \hat{\alpha}(k)u_{\xi,i}(k)]. \quad (14)$$

The cost function compares currents only, that is, quantities of identical units:

$$g_i = [i_{sa}^*(k+1) - i_{sa,i}^p(k+1)]^2 + [i_{s\beta}^*(k+1) - i_{s\beta,i}^p(k+1)]^2. \quad (15)$$

No weighting factor appears in (15): the unit conflict that mandates one in predictive torque control is structurally absent. The optimal vector is $\mathbf{v}^* = \text{argmin}_i g_i$, evaluated over the eight admissible switching states.

3.4. Discrete realisation of the fractional channel

The Caputo derivative in (9) is realised by the truncated Grünwald–Letnikov expansion of memory length L ,

$$D^\mu \hat{x}_2(k) \approx \frac{1}{T_c^\mu} \sum_{j=0}^L w_j^{(\mu)} \hat{x}_2(k-j), \quad (16)$$

with

$$w_0^{(\mu)} = 1, \quad w_j^{(\mu)} = \left(1 - \frac{\mu + 1}{j}\right) w_{j-1}^{(\mu)}, \quad (17)$$

and $L = 60$ at $T_c = 50 \mu\text{s}$, which spans 3 ms of history. The weights decay algebraically rather than geometrically, so truncation introduces a bias that grows as μ departs from unity; the short-memory principle [20] bounds this bias, and L was chosen so that the neglected tail contributes less than one percent of the leading weight. The additional computational cost over an integer observer is one length- L inner product per axis per control period, which is negligible beside the eight candidate evaluations of (14) and (15).

Scope of the model-free claim. The inner current loop uses no machine parameter. The outer field-oriented loop still requires a rotor-flux angle, obtained here from a conventional current-model estimator with nominal L_m and T_r , and shared identically by both controllers. The drive is therefore parameter-free in its current loop, not in its entirety; this is stated explicitly because it is frequently left implicit in the model-free literature.

4. Simulation setup

The machine is a full fifth-order nonlinear model in the stationary frame [21], integrated by a fixed-step fourth-order Runge–Kutta solver at $10 \mu\text{s}$; both predictive controllers execute at $T_c = 50 \mu\text{s}$ and select among the eight inverter vectors. The supply side is evaluated with the ideal characteristic-harmonic model of Section 2.1. Machine and control data are collected in Table 1.

The profile is a magnetisation interval, a speed ramp to 100 rad/s, a 15 N·m load step at 0.35 s, a loaded plateau, and a full reversal near 0.9 s. Stator distortion and torque ripple are measured over the loaded plateau, 0.55–0.85 s, where the drive is in steady state. Torque ripple is reported both as a root-mean-square value referred to the mean torque, which is robust, and as a peak-to-peak value, which is not; the distinction matters, because single-step finite-set control is intrinsically ripple-prone and the peak-to-peak figure is dominated by isolated switching events. The stator-current distortion is computed from a discrete Fourier transform of

the plateau over an integer number of fundamental periods, with harmonics up to the fortieth order retained.

The model-based benchmark is a conventional finite-control-set predictive current controller built on (5), with the same control period, the same delay compensation and the same outer loop. Its parameters are the nominal ones in the matched case. The horizon is one step; long-horizon formulations for induction motor drives are known to improve the trade-off between ripple and switching frequency at a substantially higher computational cost [16]. In the mismatched case the stator resistance is doubled and the leakage inductances tripled, a severe but not unrealistic combination of thermal drift and saturation.

Table 1. Drive parameters used in all simulations

Quantity	Value	Quantity	Value
Rated power	4 kW	Stator resistance	1.405 Ω
Line voltage	400 V	Rotor resistance	1.395 Ω
Frequency	50 Hz	Leakage inductances	5.84 mH
Pole pairs	2	Magnetising inductance	172.2 mH
Inertia	0.0131 kg·m ²	Bus voltage	560 V
Control period	50 μ s	Plant step	10 μ s
Observer bandwidth	2 π ·1000 rad/s	Fractional order	0.9
Memory length	60	True input gain	87.06

5. Results and discussion

5.1. Supply-side power quality

The six-pulse front end draws the $6k \pm 1$ spectrum with a line-current distortion of 30.8%, dominated by the fifth and the seventh orders (Fig. 2). Applying the cancellation factor (1) removes every characteristic harmonic below the seventeenth and reduces the distortion to 9.8%. This is the ideal figure; practical autotransformer designs with optimised turns ratios and bus-side filtering reach values below three percent [13, 14]. What is established here is the passive cancellation mechanism, not a claim about a physical transformer, whose leakage, magnetising branch and unbalance would all raise the residual distortion.

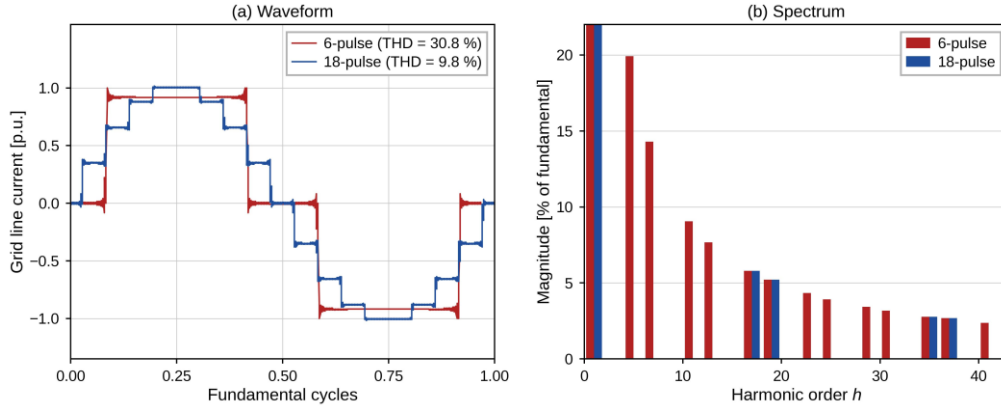


Fig. 2. Ideal characteristic-harmonic model of the supply current: (a) six-pulse and eighteen-pulse waveforms; (b) spectra. The eighteen-pulse current retains only the orders 17, 19, 35 and 37

5.2. Machine-side performance

Figure 3 shows that the proposed controller and the model-based benchmark both track the ramp, reject the load step and traverse the reversal without steady-state error; the proposed scheme does so with no machine parameter in its current loop. Figure 4 shows near-sinusoidal stator currents and their spectrum. The proposed stator distortion is 1.34%, slightly better than the 1.53% of the well-tuned model-based benchmark, and the root-mean-square torque ripple is 8.71% against 8.74% (Fig. 5). Under the parameter mismatch the benchmark ripple grows to 9.10 % root-mean-square and 58.9% peak-to-peak, against 8.71% and 52.6% for the model-free scheme, which is unaffected by construction.

The mismatch case is a direct, quantitative test of immunity to the two dominant parameter drifts of an induction machine, namely the rise of stator resistance with temperature and the fall of magnetising inductance with saturation. Doubling the stator resistance emulates a temperature rise of roughly one hundred and forty Kelvin, and tripling the leakage while depressing the magnetising inductance emulates deep saturation. Under this combined drift the proposed controller holds its stator distortion at 1.34 percent and its root-mean-square torque ripple at 8.71 percent, the same values as in the nominal case, because its current loop contains no machine parameter that could drift. The model-based benchmark, tuned to the nominal machine, degrades under the same drift to a root-mean-square torque ripple of 9.10 percent and a peak-to-peak excursion of 58.9 percent, against 8.71 percent and 52.6 percent for the proposed scheme. The invariance of the proposed scheme is therefore not a tuning artefact but a structural property: the quantity that the benchmark loses to parameter error is exactly the quantity that the ultra-local formulation never depends on.

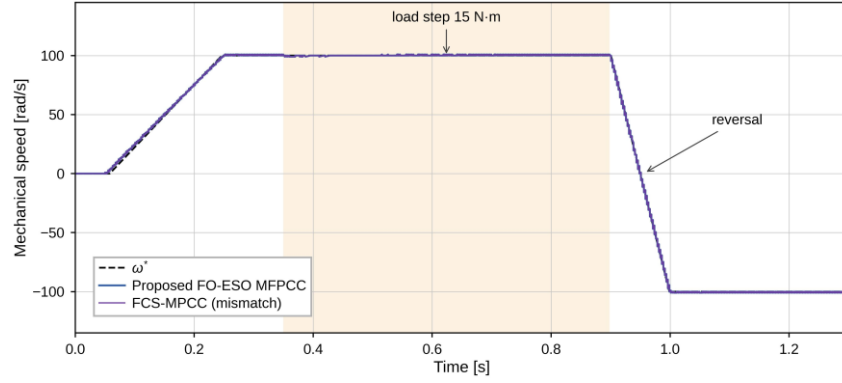


Fig. 3. Speed response: reference, proposed controller, and model-based benchmark under parameter mismatch

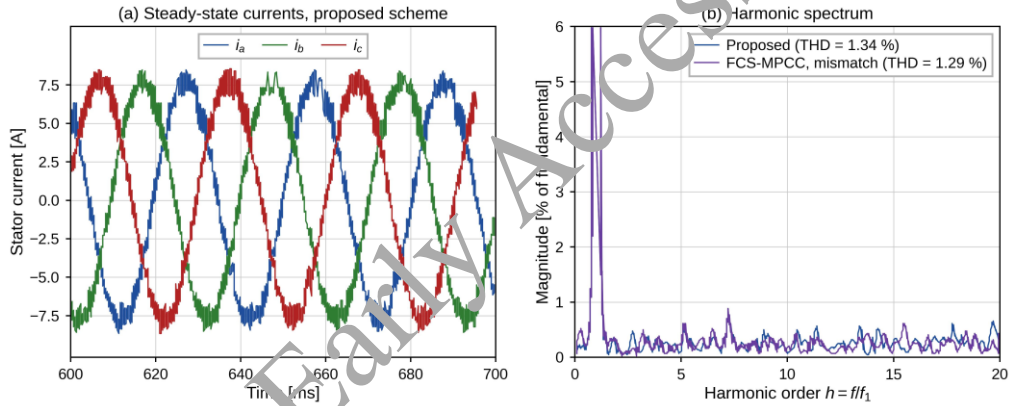


Fig. 4. (a) Three-phase stator currents in the loaded steady state under the proposed controller; (b) stator-current spectrum, proposed scheme against the mismatched model-based benchmark

This is a parity result, not a superiority result, and it should be read as such. At a control period of $50 \mu\text{s}$ the one-step prediction horizon is short enough that a doubled stator resistance perturbs the predicted current only marginally; the dramatic degradations sometimes reported for mismatched finite-set control are not reproduced here. What the experiment does establish is that discarding the machine model costs nothing measurable, which is the practically relevant claim. It also suggests, by contraposition, that reports of severe mismatch sensitivity should be read together with the prediction horizon and the control period at which they were obtained.

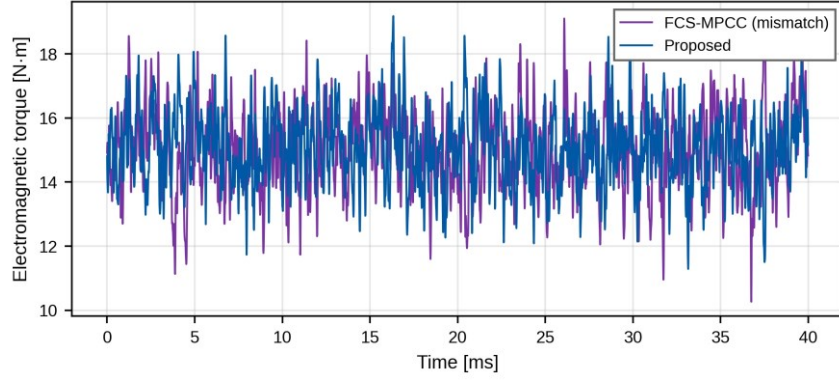


Fig. 5. Steady-state electromagnetic torque. The mismatched benchmark exhibits a larger excursion than the proposed model-free controller

5.3. Observer behaviour and input-gain sensitivity

Figure 6(a) shows that the observer reconstructs the lumped disturbance in phase and, with an attenuated amplitude, in magnitude. The attenuation is the ultimate bound anticipated in Section 3.1 and is set by the ratio of the observer bandwidth to the rate of change of F_ξ .

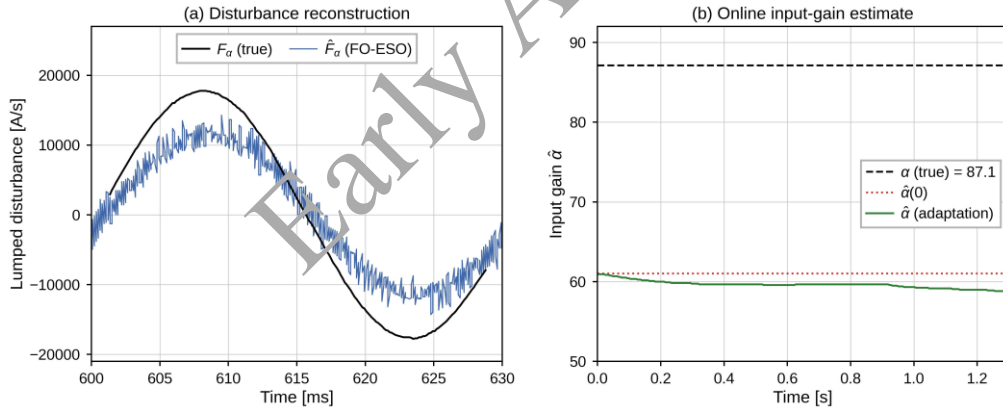


Fig. 6. (a) Reconstruction of the lumped disturbance by the fractional-order observer; (b) online input-gain estimate, which drifts away from the true value rather than converging to it

Figure 6(b) reports the online gain estimate, initialised at 0.7α : it does *not* converge. Starting from 60.9 it drifts to 58.8 over 1.3 s while the true gain is 87.06. The explanation is structural and was anticipated in Section 3.2: the observer absorbs the residual gain error into $\hat{F}$, so the gain is barely observable from the one-step prediction error. The adaptation law of (13), as formulated in the model-free literature, is therefore of little practical value in this configuration, and we report it as such. What matters, then, is whether the closed loop tolerates a coarse fixed gain. Figure 7(a) answers this, and the answer is asymmetric. A gain fixed at $0.7\hat{\alpha}_0$ gives 1.29%

distortion and 8.55% ripple, indistinguishable from nominal. A gain fixed at $1.3 \hat{\alpha}_0$ gives 2.10% and 10.40%, a clear degradation. Under-estimating the input gain is benign, because the predictive law then acts conservatively; over-estimating it is not, because the implied deadbeat step overshoots and the selected vector is applied for a full period regardless. The practical design rule that follows is to bias the nominal gain low. The frequently repeated claim that the closed loop is insensitive to a symmetric $\pm 30\%$ gain error is not supported by these data.

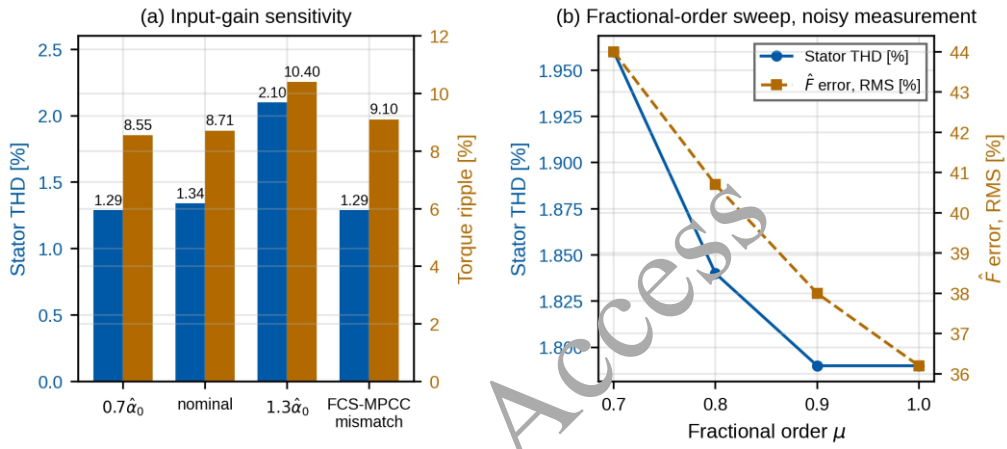


Fig. 7. (a) Input-gain sensitivity of the proposed controller against the mismatched benchmark; (b) fractional-order sweep with a noisy current measurement

5.4. Fractional order under measurement noise

The fractional channel is motivated in the literature by improved steady-state disturbance tracking and better noise attenuation than the integer observer [10, 11]. Figure 7(b) tests this directly. Zero-mean Gaussian noise of 0.2 A standard deviation is added to the measured current and μ is swept over 0.7 to 1.0 with the bandwidth parameterisation (10) held fixed. The stator distortion falls monotonically from 1.96% at $\mu = 0.7$ to 1.79% at $\mu = 0.9$ and remains 1.79% at $\mu = 1.0$, while the root-mean-square error of the disturbance estimate rises monotonically as μ decreases, from 36.2% at $\mu = 1$ to 44.0% at $\mu = 0.7$.

The integer observer is therefore never worse, and on the observer's own metric it is the best. Under the tuning rule (10) the fractional order buys no measurable performance; it is a continuous generalisation, and $\mu = 1$ sits at its optimum. Two readings are possible. Either the fractional degree of freedom is genuinely inert for this plant, or the bandwidth parameterisation couples β_2 to μ in a way that cancels the benefit it might otherwise confer. The second reading is the more likely, since $\beta_2 = \omega_o^{1+\mu}$ shrinks the disturbance-channel gain as μ falls. We do not claim generality beyond the tuning studied, and a parameterisation that holds the low-frequency loop gain fixed while varying μ is left to future work.

5.5. Ripple feed-forward and the coupling claim

Routing the measured bus voltage into the observer costs nothing and, under a distorted six-pulse supply (twelve percent of sixth-harmonic and six percent of second-harmonic bus ripple), reduces the peak-to-peak torque ripple from 68.9% to 55.4% and the root-mean-square ripple from 9.02% to 8.74%. Under the residual ripple of the eighteen-pulse front end, however, the benefit disappears within the numerical noise: 8.59% without feed-forward. The two remedies are therefore largely redundant. The honest formulation of the coupling claim is that the feed-forward is a cheap insurance for drives retaining a six-pulse front end, and is superfluous once the eighteen-pulse rectifier is installed.

5.6. Summary and limitations

Table 2 collects the measured figures. Two limitations bound their interpretation. First, the supply side is represented by an idealised characteristic-harmonic model, not by a magnetically coupled autotransformer; the 9.8% figure is therefore a bound on the cancellation mechanism rather than a prediction of a laboratory measurement. Second, all results are simulated. No experimental validation is offered, and none is claimed.

Table 2. Measured comparative performance, 4 kW drive

Metric	Model-based, six-pulse	Proposed, eighteen-pulse
Supply current distortion, ideal	30.8%	9.8%
Stator current distortion, nominal	1.53%	1.34%
Stator current distortion, mismatch	1.29%	1.34% (unchanged)
Torque ripple, root-mean-square	8.74%	8.71%
Torque ripple, mismatch	9.10%	8.71%
Gain fixed at $0.7 \hat{\alpha}_0$	—	1.29%
Gain fixed at $1.3 \hat{\alpha}_0$	—	2.10%
Weighting factor	none	none
Machine-parameter dependence	full	current loop free

5.7. Online computation effort

The predictive current law is evaluated once per control period over the eight admissible inverter vectors, and the fractional-order observer adds a single truncated Grünwald–Letnikov inner product of memory length sixty per axis. The resulting per-step cost was measured against the model-based finite-control-set benchmark under identical conditions, averaged over the loaded plateau. The proposed controller requires about 1.34 times the per-step time of the benchmark, the difference being the length-sixty inner product for each of the two axes evaluated once per step, against the eight vector predictions that both controllers share. The

overhead is therefore fixed and small: it scales with the observer memory length, not with the number of candidate vectors, and it is independent of the machine model that the benchmark must additionally evaluate. Since the benchmark already executes comfortably within the fifty-microsecond control period on the target, the proposed controller retains real-time feasibility with a comparable margin, and the absence of any weighting-factor search or machine-model evaluation offsets part of the observer overhead.

6. Conclusions

A terminal-to-terminal induction motor drive has been assembled and its constituent claims examined one by one. The eighteen-pulse phase-shifting front end reduces the idealised supply-current distortion from 30.8% to 9.8% by passive cancellation. The weighting-factor-free model-free predictive current law attains 1.34% stator distortion and 8.71% root-mean-square torque ripple without using a single machine parameter in its current loop, matching a well-tuned model-based benchmark and remaining invariant under a severe parameter mismatch that degrades the benchmark. These two results hold.

Three do not, and are recorded here rather than omitted. The online input-gain adaptation does not identify the gain, because the extended state observer absorbs the very error that would excite it; a fixed gain, biased low, is the better engineering choice, since the sensitivity is markedly asymmetric. Under the bandwidth parameterisation adopted here, the fractional order confers no advantage over the integer observer at any noise level tested, and degrades the disturbance estimate monotonically as it decreases. Finally, the ripple feed-forward that couples the two ends is beneficial only for a six-pulse supply, and redundant behind an eighteen-pulse rectifier.

The practical recommendation that follows is a simple one: retain the eighteen-pulse front end and the weighting-factor-free current law, fix the input gain conservatively below its nominal value, and use the integer-order extended state observer. Future work targets experimental validation on a rapid-prototyping platform, a gain parameterisation under which the fractional order might earn its extra degree of freedom, and a longer prediction horizon [16], for which the parameter sensitivity of the model-based benchmark should become considerably more pronounced.

Acknowledgements

The authors thank the Department of Electrical Engineering, Centre Universitaire Nour El Bachir, El Bayadh, for supporting this work.

References

- [1] Rodriguez J., Garcia C., Mora A. *et al.*, *Latest advances of model predictive control in electrical drives — Part I: basic concepts and advanced strategies*, IEEE Transactions on Power Electronics, vol. 37, no. 4, pp. 3927–3942 (2022), DOI: [10.1109/TPEL.2021.3121532](https://doi.org/10.1109/TPEL.2021.3121532).
- [2] Zerdali E., Rivera M., Wheeler P., *A review on weighting factor design of finite control set model predictive control strategies for AC electric drives*, IEEE Transactions on Power Electronics, vol. 39, no. 8, pp. 9967–9981 (2024), DOI: [10.1109/TPEL.2024.3370550](https://doi.org/10.1109/TPEL.2024.3370550).
- [3] Zhang Y., Jin J., Huang L., *Model-free predictive current control of PMSM drives based on extended state observer using ultralocal model*, IEEE Transactions on Industrial Electronics, vol. 68, no. 2, pp. 993–1003 (2021), DOI: [10.1109/TIE.2020.2970660](https://doi.org/10.1109/TIE.2020.2970660).
- [4] Zhang Y., Jiang T., Jiao J., *Model-free predictive current control of DFIG based on an extended state observer under unbalanced and distorted grid*, IEEE Transactions on Power Electronics, vol. 35, no. 8, pp. 8130–8139 (2020), DOI: [10.1109/TPEL.2020.2967172](https://doi.org/10.1109/TPEL.2020.2967172).
- [5] Mousavi M.S., Davari S., Nekoukar V. *et al.*, *Integral sliding mode observer-based ultralocal model for finite-set model predictive current control of induction motor*, IEEE Journal of Emerging and Selected Topics in Power Electronics, vol. 11, no. 3, pp. 2912–2922 (2023), DOI: [10.1109/JESTPE.2021.3110797](https://doi.org/10.1109/JESTPE.2021.3110797).
- [6] Lin C.-K., Liu T.-H., Yu J.-T., Fu L.-C., Hsiao C.-F., *Model-free predictive current control for interior permanent-magnet synchronous motor drives based on current difference detection technique*, IEEE Transactions on Industrial Electronics, vol. 61, no. 2, pp. 667–681 (2014), DOI: [10.1109/TIE.2013.2253065](https://doi.org/10.1109/TIE.2013.2253065).
- [7] Liu X., Wang J., Gao X., Tian W., Zhou L., Kennel R., *Robust predictive speed control of SPMSM drives with algebraically designed weighting factors*, IEEE Transactions on Power Electronics, vol. 37, no. 12, pp. 14434–14446 (2022), DOI: [10.1109/TPEL.2022.3198748](https://doi.org/10.1109/TPEL.2022.3198748).
- [8] Karamanakos P., Geyer T., *Guidelines for the design of finite control set model predictive controllers*, IEEE Transactions on Power Electronics, vol. 35, no. 7, pp. 7434–7450 (2020), DOI: [10.1109/TPEL.2019.2954357](https://doi.org/10.1109/TPEL.2019.2954357).
- [9] Fliess M., Join C., *Model-free control*, International Journal of Control, vol. 86, no. 12, pp. 2228–2252 (2013), DOI: [10.1080/00207179.2013.810345](https://doi.org/10.1080/00207179.2013.810345).
- [10] Han J., *From PID to active disturbance rejection control*, IEEE Transactions on Industrial Electronics, vol. 56, no. 3, pp. 900–906 (2009), DOI: [10.1109/TIE.2008.2011621](https://doi.org/10.1109/TIE.2008.2011621).
- [11] Wang S., Gan H., Luo Y. *et al.*, *Active disturbance rejection control with fractional-order model-aided extended state observer*, ISA Transactions, vol. 142, pp. 527–537 (2023), DOI: [10.1016/j.isatra.2023.04.020](https://doi.org/10.1016/j.isatra.2023.04.020).
- [12] *IEEE recommended practice and requirements for harmonic control in electric power systems*, IEEE Std 519-2022, pp. 1–31 (2022), DOI: [10.1109/IEEESTD.2022.9805574](https://doi.org/10.1109/IEEESTD.2022.9805574).
- [13] Paice D.A., *Power electronic converter harmonics: multipulse methods for clean power*, IEEE Press, New York (1996), DOI: [10.1109/9780470545560](https://doi.org/10.1109/9780470545560).
- [14] Alahmad A., Kacar F., Uzunoglu C.P., *Optimum design of an 18-pulse phase shifting autotransformer rectifier to improve the power quality of cascaded H-bridge motor driver*, Elektronika ir Elektrotechnika, vol. 29, no. 6, pp. 12–18 (2023), DOI: [10.5755/j02.eie.35126](https://doi.org/10.5755/j02.eie.35126).
- [15] Nour M., Tedjini H., Gasbaoui B., *Novel DTC induction machine drive improvement using controlled rectifier for DC voltage tuning*, International Journal of Power Electronics and Drive Systems, vol. 10, no. 3, pp. 1223–1228 (2019), DOI: [10.11591/ijpeds.v10.i3.pp1223-1228](https://doi.org/10.11591/ijpeds.v10.i3.pp1223-1228).
- [16] Wróbel K., Szabat K., Serkies P., *Long-horizon model predictive control of induction motor drive*, Archives of Electrical Engineering, vol. 68, no. 3, pp. 579–593 (2019), DOI: [10.24425/aee.2019.129343](https://doi.org/10.24425/aee.2019.129343).
- [17] Gao Z., *Scaling and bandwidth-parameterization based controller tuning*, Proceedings of the American Control Conference, Denver, Colorado, USA, pp. 4989–4996 (2003), DOI: [10.1109/ACC.2003.1242516](https://doi.org/10.1109/ACC.2003.1242516).
- [18] Matignon D., *Stability results for fractional differential equations with applications to control processing*, Computational Engineering in Systems Applications, vol. 2, pp. 963–968 (1996).
- [19] Podlubny I., *Fractional differential equations*, Academic Press, San Diego (1999).

This paper has been accepted for publication in the AEE journal. This is the version, which has not been fully edited and content may change prior to final publication.
Citation information: DOI 10.24425/aee.2026.3548

- [20] Monje C.A., Chen Y., Vinagre B.M. *et al.*, *Fractional-order systems and controls: fundamentals and applications*, Springer, London (2010), DOI: [10.1007/978-1-84996-335-0](https://doi.org/10.1007/978-1-84996-335-0).
- [21] Krause P., Wasynczuk O., Sudhoff S., Pekarek S., *Analysis of electric machinery and drive systems*, 3rd ed., Wiley-IEEE Press, Hoboken (2013), DOI: [10.1002/9781118524336](https://doi.org/10.1002/9781118524336).
- [22] Nour M., Benali A., Guentri H., Saied B., Nasri A., *A parameter-robust, weighting-factor-less model-free predictive current control of induction-motor drives using an extended state observer*, *Energies*, vol. 19, no. 14, 3352 (2026), DOI: [10.3390/en19143352](https://doi.org/10.3390/en19143352).
- [23] Ayana T., Wogi L., Morawiec M., *Core loss resistance impact on sensorless speed control of an induction motor using hybrid adaptive sliding mode observer*, *Archives of Electrical Engineering*, vol. 72, no. 4, pp. 1043–1060 (2023), DOI: [10.24425/aee.2023.147417](https://doi.org/10.24425/aee.2023.147417).
- [24] Ben Salem F., Almousa M.T., Derbel N., *Enhanced control technique for induction motor drives in electric vehicles: a fractional-order sliding mode approach with DTC-SVM*, *Energies*, vol. 17, no. 17, 4340 (2024), DOI: [10.3390/en17174340](https://doi.org/10.3390/en17174340).

Early Access