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Robust Model Reference Adaptive PID Controller: Development and Application on Grid-tied Systems

Alexandre S. Lucena¹, Cairo T. G. B. Oliveira¹, Vitória de S. Fabricio¹,
Wagner B. da Silveira², Bernardo Andres¹, Paulo J. D. de O. Evald^{1*}

¹Federal University of Pelotas, Intelligent Systems and Control Group, Pelotas – RS, Brazil.

²Federal University of Santa Maria, Group of Power Electronics and Control, Santa Maria – RS, Brazil.

e-mail: alexandre.lucena@ufpel.edu.br; cairo.tulio@gmail.com; vitoria.sfabricio@gmail.com; wasilveira91@hotmail.com; adoandres@gmail.com; paulo.evald@ufpel.edu.br*.

* Corresponding author.

ABSTRACT This work presents a novel adaptive control structure named robust model reference adaptive proportional integral derivative controller. The controller does not require precise information about system parameters, and requires only the system output feedback to continuously update its gains. Experimental results of this controller regulating the three-phase grid-injected currents of a voltage source inverter with an LCL filter under uncertain grid conditions are presented. The controller demonstrated robustness, feasibility for application on grid-tied power systems, and fast current tracking. The total harmonic distortion of the grid-injected currents is 3.23%, 3.78%, and 3.41% in phase A, B, and C, respectively.

KEYWORDS Inverters, LCL Filters, Renewable Energy, RMRAC-based Controller.

I. INTRODUCTION

Grid-tied voltage source inverters (VSIs) are largely employed as an interface between energy sources and the utility grid. To comply with interconnection standards such as IEEE 519 [1], which limit current total harmonic distortion (THD) to 5%, output filters are required to suppress harmonics generated by switching dynamics. Among filter topologies, the LCL filter is the industry standard. Compared to the L filter, this topology offers better harmonic attenuation (-60 dB/dec), while it can also be built with smaller inductors and compact system designs [2]. However, managing resonance interactions remains a design trade-off, requiring methodologies that balance efficiency and stability [3].

Despite structural advantages, the LCL filter introduces control complexity due to its resonance peak, resulting from the interaction between filter capacitors and grid inductance [4], in variable frequencies, since it depends on the grid condition. Insufficient damping leads to oscillations, compromising closed-loop stability [5]. While passive damping is a simple solution, it induces power losses that reduce efficiency [6]. Consequently, active damping techniques are preferred in high-power applications [3], since there is no power dissipation due to the addition of passive elements into the filter structure. Control challenges increase as grid conditions weaken. In these scenarios, there is a shift in the resonance frequency towards lower frequencies where it interacts with the controller bandwidth [7]. This variation can destabilize fixed-gain controllers designed for stiff grids. Taking this information into consideration, adaptive controllers are an adequate control solution for this system.

The recent literature on adaptive controllers for grid-tied systems include observer-based strategy [8], which estimates grid impedance and grid voltage for adaptive disturbance rejection, and artificial neural network-based controller [9], which does not require datasets for training. Although these strategies effectively regulate the system, they have considerable computational burden due to estimation or learning techniques, which can be an issue for dated microcontrollers. In another approach, an adaptive notch filter-based controller was introduced in [10]. This strategy uses the synchronous-reference-frame phase-locked loop to track the resonance frequency, while current regulation is done using proportional-integral (PI) controllers. Differently, a combined model-free predictive controller and adaptive gain extended state observer was proposed in [11]. This framework uses double second-order generalized integrator-based pre-stage signal filter before the observer estimates capacitor voltage, inverter-side current, parameter perturbations, grid disturbances, and resonance effects. Although satisfactory results were obtained, these strategies are multi-loop control frameworks, requiring precise tuning to ensure alignment of each controller in their tasks. With a simpler main architecture, an optimized proportional-resonant plus PI controllers was presented in [12]. Although the integrated techniques are simple, the authors employed the least mean square (LMS) algorithm for gains update, increasing the controller computational burden and complexity.

Differently from the literature, this work presents a novel robust model reference adaptive proportional-integral-derivative (RMRAPID) controller. This control strategy is

a single-loop framework, which does not need learning algorithms, pre-training stage, or require extra sensors. Its gains are updated by a gradient algorithm, which is lighter than LMS algorithms, using solely the grid-injected currents and voltages at the point of common coupling (PCC) measurements to execute, being compatible with dated microcontrollers. The main contributions of this work are:

- A novel robust adaptive PID controller for partially modeled systems;
- A robust adaptive current controller for grid-tied VSI with an LCL filter that requires only grid-injected current feedback;
- Experimental validation of the RMRAPID controller in a grid-tied power system.

II. CONTROL PROBLEM STATEMENT

The VSI equipped with an LCL filter model is largely established in the literature. Therefore, it will be briefly presented. Figure 1 shows the electrical diagram, where L_c , C_f , and L_{g1} are the filter elements, L_{g2} and r_{g2} is the grid impedance, whose voltage is v_g , while r_{g1} and r_{g2} are parasitic resistances. Besides, V_{dc} is the DC link voltage and i_{ga} , i_{gb} , and i_{gc} are the grid-injected currents. The currents (i_a , i_b , and i_c) are measured in the PCC to feedback the controller, where the line-to-line voltages (v_{ab} and v_{bc}) are also measured to synchronize the inverter with the grid.

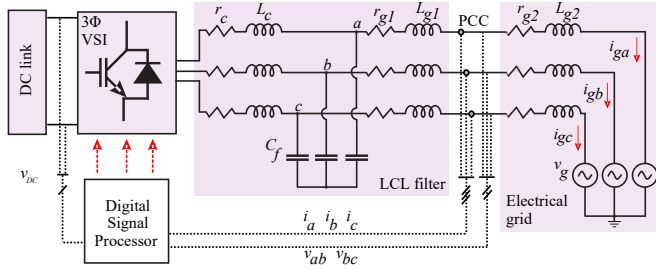


FIGURE 1. Grid-tied VSI with passive filter.

The control design focuses on the LCL filter model, whose system is modeled in $\alpha\beta 0$ coordinates via Clarke transformation to simplify the LCL filter modeling and current control design. Under the assumption of a balanced grid, the zero-sequence component is neglected, since there is no current flowing through it, allowing the system to be represented by two decoupled single-phase equivalent circuits. By defining the total grid impedance as $r_g = r_{g1} + r_{g2}$ and $L_g = L_{g1} + L_{g2}$, the transfer function $G(s)$ relating the grid-injected current i_g to the modulated control signal u is [7]

$$G(s) = \frac{i_g(s)}{u(s)} = \frac{b_0}{a_0 s^3 + a_1 s^2 + a_2 s + a_3}, \quad (1)$$

where $a_0 = L_g L_c C$, $a_1 = (r_g L_c + r_c L_g)C$, $a_2 = L_c + L_g + r_g r_c C$, $a_3 = r_g + r_c$, and $b_0 = 1$. Figure 2 shows the experimental setup.

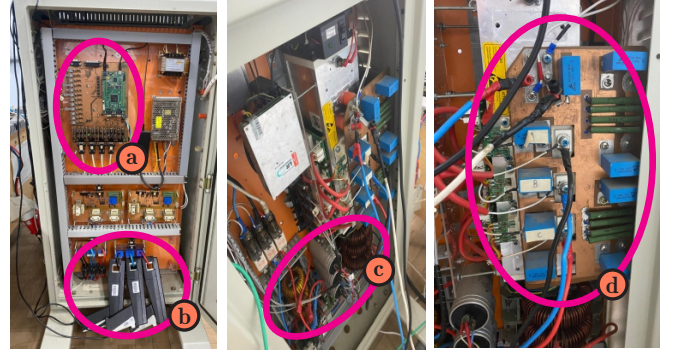


FIGURE 2. Experimental setup: (a) DSP, (b) Sensors, (c) LCL filter, and (d) VSI.

III. RMRAPID CONTROLLER

Consider the classical PID formulation in Laplace domain,

$$U(s) = K_p E(s) + \frac{K_i}{s} E(s) + K_d s E(s), \quad (2)$$

where K_p , K_i , and K_d are the proportional, integral, and derivative gains, respectively. Besides, U and E represent the control action and tracking error, respectively. By applying the backward Euler approximation $s \approx (1 - z^{-1})/T_s$ in (2), multiplying both sides of the (2) by $(1 - z^{-1})$, it follows:

$$u(k) - u(k-1) = \left(K_p + T_s K_i + \frac{K_d}{T_s} \right) e(k) + \left(-K_p - \frac{2K_d}{T_s} \right) e(k-1) + \frac{K_d}{T_s} e(k-2), \quad (3)$$

where T_s is the sampling time.

Replacing $e(k) = y_m(k) - y(k)$, where $y_m(k)$ is the output of the reference model $W_m(z)$, into (3) and dividing it by c_1 , the equation turns out to be

$$\frac{1}{c_1} u(k) - \frac{1}{c_1} u(k-1) = y_m(k) - y(k) + \frac{c_2}{c_1} e(k-1) + \frac{c_3}{c_1} e(k-2), \quad (4)$$

where $c_1 = K_p + T_s K_i + K_d/T_s$, $c_2 = -K_p - 2K_d/T_s$, and $c_3 = K_d/T_s$. Considering $y_m(k) = \delta y_m(k) + r(k)$, where $r(k)$ is the reference signal, which is necessary in RMRAC formulation, (4) can be rewritten as

$$\frac{1}{c_1} u(k) - \frac{1}{c_1} u(k-1) = \delta y_m(k) + r(k) - y(k) + \frac{c_2}{c_1} e(k-1) + \frac{c_3}{c_1} e(k-2). \quad (5)$$

Direct adaptive controllers are generally formalized in a vector notation $\theta(k)^T \omega(k) + r(k) = 0$, where $\theta(k)$ is a vector of adaptive gains and $\omega(k)$ is the vector of dynamics that ponder these gains. Therefore, rearranging (5), and defining the dynamics of interest as $\omega(k) = [u(k) \ u(k-1) \ y(k) \ e(k-1) \ y_m(k) \ e(k-2)]^T$. Then, there are six adaptive gains $\theta(k) = [\theta_1(k) \ \theta_2(k) \ \theta_3(k) \ \theta_4(k) \ \theta_5(k) \ \theta_6(k)]^T$, which replace the constant coefficients of (5). Thus, it can be

written that $\theta_1 u(k) + \theta_2 u(k-1) + \theta_3 y(k) + \theta_4 e(k-1) + \theta_5 y_m(k) + \theta_6 e(k-2) + r(k) = 0$. Consequently, it follows:

$$u(k) = -\frac{1}{\theta_1(k)}(\theta_2(k)u(k-1) + \theta_3(k)y(k) + \theta_4(k)e(k-1) + \theta_5(k)y_m(k) + \theta_6(k)e(k-2) + r(k)). \quad (6)$$

A. Gains adaptation algorithm

The gains θ are continuously updated in real time. This adaptation is provided by the gradient algorithm, given by

$$\theta(k+1) = \theta(k) - T_s \sigma(k) \Gamma \theta(k) - T_s \kappa \frac{\Gamma \zeta(k) \epsilon(k)}{\bar{m}^2(k)}, \quad (7)$$

where Γ is a positive definite matrix, while κ is a positive scalar, both used for defining the convergence rate of the gains. Furthermore, $\bar{m}(k)$ is a majorant signal and $\sigma(k)$ is the σ -modification, both discussed later.

The gradient algorithm also computes the product of the regressor vector filtered by the reference model $W_m(z)$, which results in $\zeta(k)$, and the augmented error $\epsilon(k)$. These internal signals are calculated as

$$\zeta(k) = W_m(z)\omega(k), \quad (8)$$

$$\epsilon(k) = -e(k) + \theta^T(k)\zeta(k) - \nu(k), \quad (9)$$

$$\nu(k) = W_m(z)[\theta^T(k)\omega(k)]. \quad (10)$$

B. Robustness mechanisms

A majorant signal is used to ensure boundedness of closed-loop signals. The majorant signal is

$$\bar{m}^2(k) = m^2(k) + \zeta^T(k)\Gamma\zeta(k), \quad (11)$$

where $m(k+1) = (1 - T_s\delta_0)m(k) + T_s\delta_1(1 + |u(k)| + |y(k)|)$, being δ_0 and δ_1 positive constants, which must be designed to ensure that $m(0) \geq \delta_1/(1 - \delta_0)$. Besides, the $\sigma(k)$ -modification is implemented to avoid drift in the gains while they are converging. It is described by

$$\sigma(k) = \begin{cases} 0 & \text{if } \|\theta(k)\| \leq M_0 \\ \sigma_0 \left(\frac{\|\theta(k)\|}{M_0} - 1 \right) & \text{if } M_0 < \|\theta(k)\| < 2M_0, \\ \sigma_0 & \text{if } \|\theta(k)\| \geq 2M_0 \end{cases} \quad (12)$$

where $M_0 > \|\theta^*\|$, where θ^* is the vector of ideal gains, which is unknown a priori. Therefore, M_0 can be oversized. Additionally, σ_0 is the upper bound of σ -modification.

C. Disturbance rejection terms

When the system is subjected to periodic exogenous disturbances, two additional gains are designed to provide the rejection of these dynamics. They are related to the phase $V_s(k)$ and quadrature $V_c(k)$ components of the periodic disturbance. In this sense, $\omega(k)$ is expanded to $\omega(k) =$

$[u(k) \ u(k-1) \ y(k) \ e(k-1) \ y_m(k) \ e(k-2) \ V_s(k) \ V_c(k)]$, and, consequently, the vector of gains is modified to $\theta(k) = [\theta_1(k) \ \theta_2(k) \ \theta_3(k) \ \theta_4(k) \ \theta_5(k) \ \theta_6(k) \ \theta_s(k) \ \theta_c(k)]$, where $\theta_s(k)$ and $\theta_c(k)$ are weighted by $V_s(k)$ and $V_c(k)$, respectively. The new control action is

$$u(k) = -\frac{1}{\theta_1(k)}(\theta_2(k)u(k-1) + \theta_3(k)y(k) + \theta_4(k)e(k-1) + \theta_5(k)y_m(k) + \theta_6(k)e(k-2) + r(k)) - \frac{1}{\theta_1(k)}(\theta_s(k)V_s(k) + \theta_c(k)V_c(k)). \quad (13)$$

Figure 3 depicts the closed-loop system with the controller.

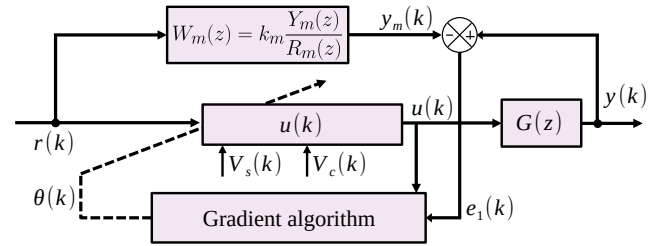


FIGURE 3. Block diagram of the RMRAPID controller.

D. Stability analysis

The controller developed in this work can be considered as an extension of the controller introduced in [13], which is a robust model reference adaptive PI controller. In this sense, the stability proof of the RMRAPID controller follows exactly the same steps demonstrated in Section 3 of [13], considering the same assumptions, starting with the same Lyapunov candidate function and achieving the same stability condition. Therefore, this proof will be omitted here to avoid textual overlap.

E. Parametrization procedure

This work implements a recent real-time auto-tuning technique, which was first introduced in [14]. This technique consists in feedbacking the system with a virtual LCL model with the same system parameters before the inverter synchronization, allowing to tune the gains with real world information, but maintaining the inverter inhibited during this stage. Once the convergence of gains is achieved, the inverter is uninhibited.

IV. EXPERIMENTAL RESULTS

The experiments were conducted on a grid-tied VSI with an LCL filter, evaluating the controller's performance under three distinct operational phases: (I) real-time auto-tuning using a virtual plant, (II) grid synchronization and steady-state operation, and (III) robustness assessment under load steps and parametric variations. The last event is forced by the opening of a contactor, which makes the current flow for another way that has an additional inductor, representing

the increase of grid impedance.. The control algorithm was implemented on a digital signal processor (DSP), the TMS320F28335 DSP microcontroller from Texas Instruments. Data acquisition was performed via an internal buffer and an oscilloscope to visualize and validate the results.

The nominal parameters of the LCL filter are $L_c = 1$ mH, $C_f = 62$ μ F, and $L_{g1} = 0.3$ mH, designed following the steps presented in [15], and with estimated $R_c = R_{g1} = R_{g2} = 50$ m Ω . The grid voltage is 150 V_{RMS} at 60 Hz, adjusted using a variable transformer. Additionally, the nominal value of L_{g2} is 0.3 mH, but has been increased to 1.3 mH to evaluate the controller robustness. The sampling frequency and switching frequency are both equal to 5040 Hz. The VSI synchronization is done using a Kalman filter-based phase-locked loop [16] and the control action is synthesized by means of space vector modulation [17]. The fixed controller parameters are $\gamma = 10$, $\kappa = 100$, $\delta_0 = 0.7$, $\delta_1 = 1.0$, $M_0 = 5$, $\sigma_0 = 0.1$, adjusted similar to other direct adaptive controllers from the literature. The initial adaptive gains are tuned directly in the DSP by the auto-tuning method [14]. The reference model is $W_m(z) = 0.7/(z - 0.3)$.

A. Auto-tuning, grid synchronization, and overall performance

During the auto-tuning stage, presented in Figures 4 and 5, the inverter is not tied to the grid yet. However, the disturbance is estimated using a Kalman filter and the controller receives this information to discover the adequate gains to reject properly during the nominal operation. Although the auto-tuning stage occurs real-time, this phase is virtual, since there is no current being injected into the grid. It is important to highlight that it allows us to use any reference signal rich in frequency, and not necessarily the reference signal that will be tracked during the controller actuation (a sinusoidal one). During this phase ($t < t_{sync} = 0.2$ s), the adaptive gains, shown in Figure 4(b) for the α coordinate and Figure

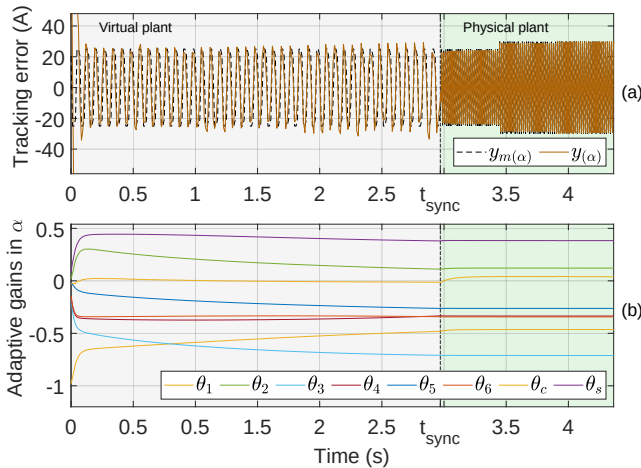


FIGURE 4. Experiment (synchronization and physical plant) in α : (a) $W_m(z)$ tracking and (b) gains evolution.

5(b) for the β coordinate, have a rapid adaptation process. After synchronization, the controller tracks a sinusoidal current reference of 25 A/60 Hz.

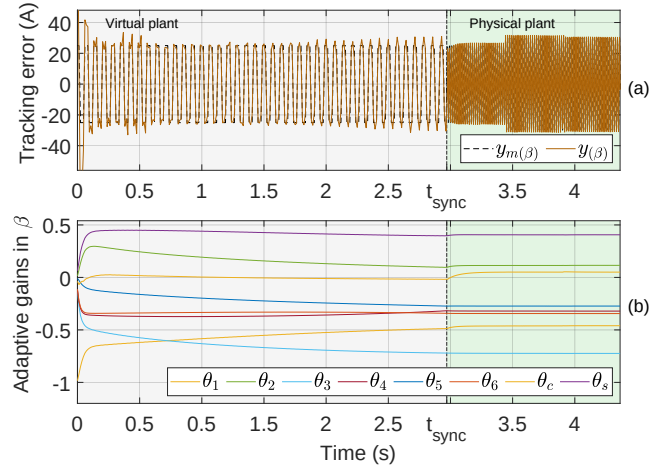


FIGURE 5. Experiment (synchronization and physical plant) in β : (a) $W_m(z)$ tracking and (b) gains evolution.

Figures 6(c) and 7(c) show the acquired data from the buffer, specifically the grid-injected current and adaptive gains in α and β coordinates. Figures 6(b) and 7(b) show the instant when the current reference changes to 30 A/60

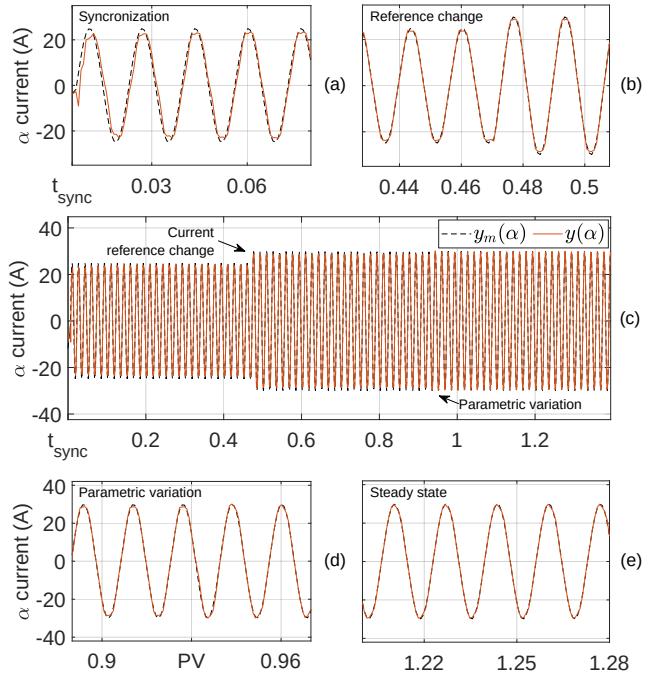


FIGURE 6. Current tracking and adaptive gains in α coordinate: (a) virtual plant (synchronization phase), (b) current reference change, (c) overall experiment, (d) parametric variation and (e) steady state.

Hz, in α and β coordinates, respectively. As can be seen, the tuned controller has a fast response to this event. Besides, the parametric variation is shown in Figure 6(d) and 7(d) as

well as the steady state in 6(e) and 7(e). In these figures, it is also observed that tuned controllers are robust to grid inductance variation, which was quickly corrected with only small overshoots. These graphics demonstrate the stability of the closed-loop system. The tracking accuracy is discussed later in detail. Figure 8 depicts the current tracking errors.

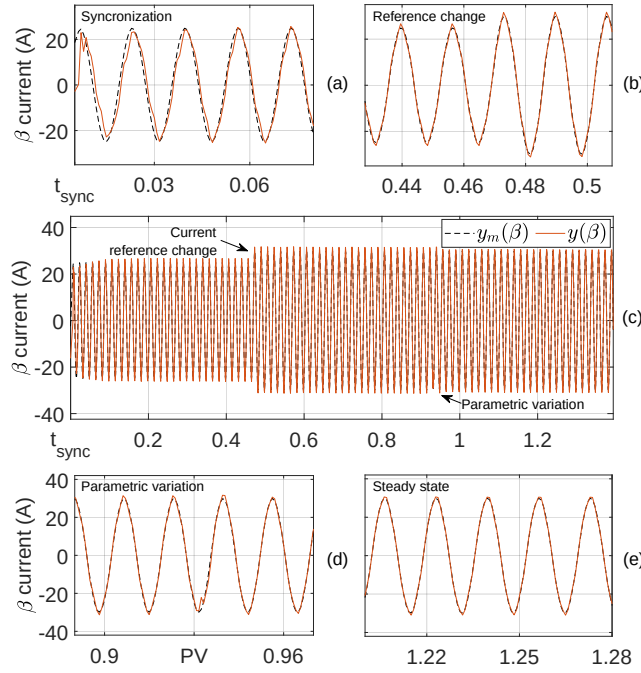


FIGURE 7. Current tracking and in β coordinate: (a) virtual plant (synchronization phase), (b) current reference change, (c) overall experiment, (d) parametric variation and (e) steady state.

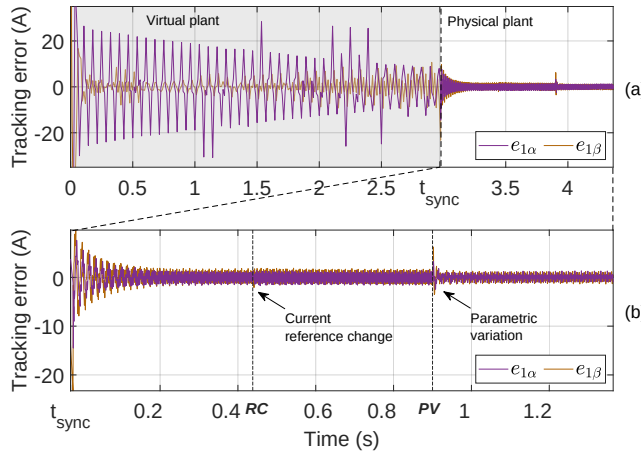


FIGURE 8. Current tracking errors in (a) α and β coordinate, and (b) zoom after synchronization.

During the virtual phase, it exhibited significant magnitudes, reaching peak values over 25 A in α and β coordinates. However, as previously mentioned, this phase is virtual, emulated in the DSP memory with the purpose of

allowing the controller to find a set of gains that minimize the tracking error during synchronization. Therefore, these errors are not a concern, since they are not real tracking errors. Since the virtual phase ends, the synchronization occurs. As the gains converged to a stable set (Figure 4(b) for the α coordinate and Figure 5(b) for the β coordinate), the error decreases drastically, ensuring a seamless transition to the physical plant. The tracking errors remain small during the overall experiment. Naturally, these errors are not null due to the switching dynamics of the VSI.

Figure 9(a) shows the DC bus voltage (V_{dc}) behavior during this transition. At $t_{sync} \approx 2.97$ s, the synchronization occurs, and a sag is observed in the V_{dc} . This sag is corrected by an internal controller of this voltage source, but this loop is outside the scope of this work. Regarding the control actions, they are safely implementable for this system. The available voltage in the DC bus is 500 V, but the control actions demanded less than half of this value for its synthetization, as can be seen in Figures 9(b) and 9(c). The maximum transient peaks of control actions were 152.72 V and 198.66 V for α and β coordinates, respectively. The absence of control saturation aids the AC currents transition smoothly to the physical grid injection. Figure 10 shows the three-phase grid-injected currents acquired from the oscilloscope, corroborating the controller performance.

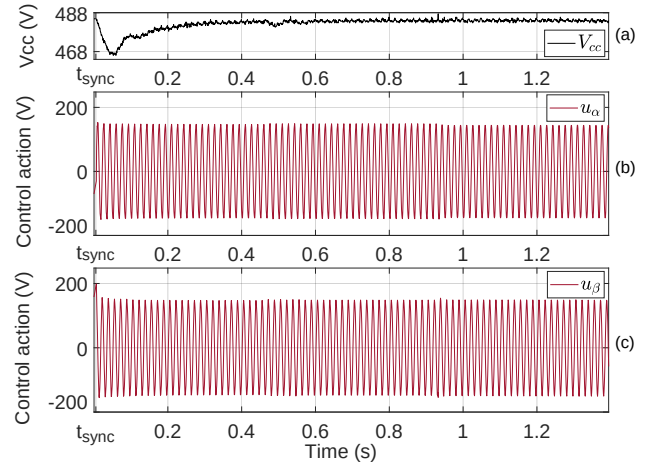


FIGURE 9. Voltage: (a) DC bus voltage, (b) Control action in α coordinate, and (c) Control action in β coordinate.

B. Dynamic performance and robustness

Two events were triggered during the experiment to assess the dynamic response and robustness of the RMRAPID controller. First, a step change in the current reference amplitude is applied at $t \approx 3.44$ s ($t_{sync} + 0.46$) s, increasing from 25 A to 30 A. Figures 6(b) and 7(b) show the current tracking for both coordinates, where it can be inferred that RMRAPID controllers actuated quickly in $\alpha\beta$ coordinates. Naturally, it is reflected in the three-phase grid-injected currents, shown in Figure 10. The controllers were able to allow current

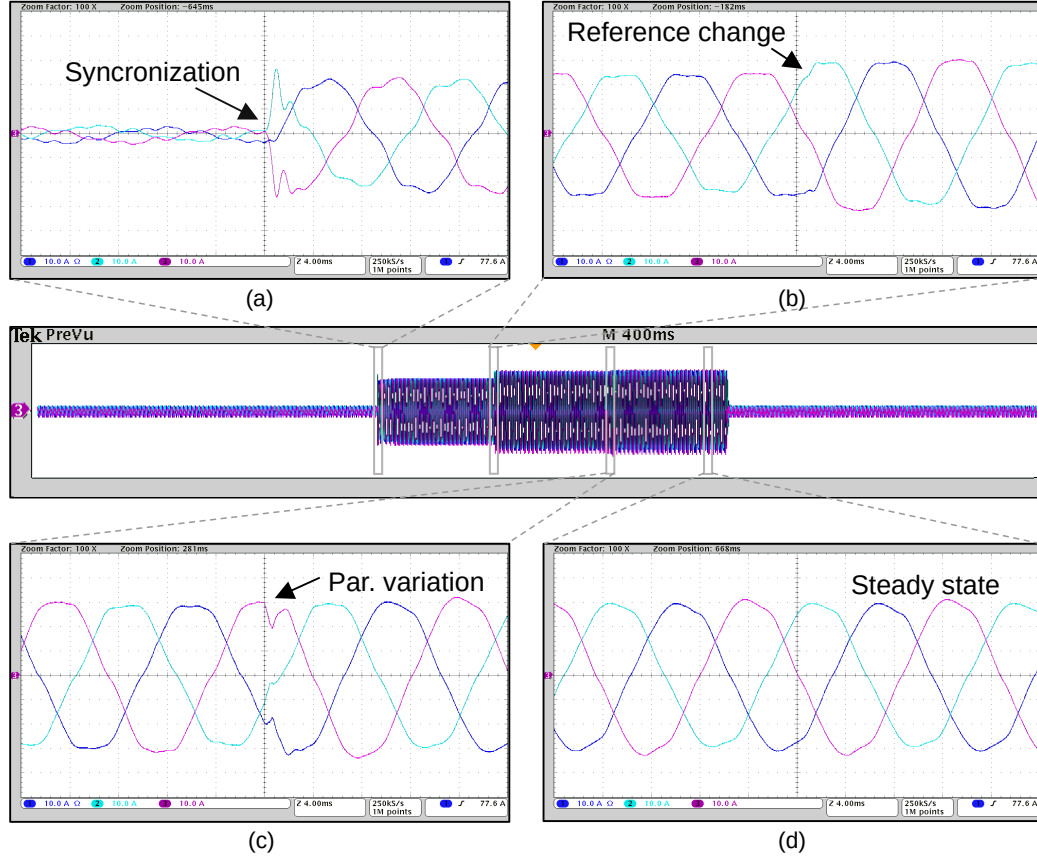


FIGURE 10. Grid-injected currents: overall experiment at the center of figure with zoom at (a) virtual plant (synchronization), (b) current reference change, (c) grid inductance variation, (d) steady state. Data from oscilloscope.

tracking with negligible overshoot and achieving steady state after the current reference update in approximately two grid cycles. Subsequently, a parametric variation is forced to test the controller's robustness against grid impedance changes at $t \approx 3.90$ s ($t_{sync} + 0.93$) s. Figure 10(c) reveals that the system maintains stable operation despite the exogenous disturbance. Although a minor transient occurs, the adaptive mechanism compensates for plant mismatch, ensuring that grid-injected currents remain stable and, with small error, quickly compensated for, as can be seen in Figure 8.

Table 1 summarizes important information for security and quality analysis. First, the maximum values of the three-phase grid-injected currents are 30.04 A, 29.96 A, and 30.91 A for phases A, B, and C, respectively. For this system, the maximum allowable current is 35 A. A superior value activates the protection system, disconnecting the VSI from the electrical grid to ensure its integrity. Thus, the controller safely regulates the grid-injected currents. Regarding the amplitude of tracking errors, the maximum peaks are 7.61 A and 9.70 A in α and β coordinates, respectively, both related with the synchronization instant. The errors during the operation, even with the current reference change and abrupt grid inductance change, are relevantly smaller.

TABLE 1. Quantitative performance summary.

Category	Metric	Value	Unit
Phase currents	Max. current (phase A)	30.04	A
	Max. current (phase B)	29.96	A
	Max. current (phase C)	30.91	A
Max. tracking errors	Max. $e(\alpha)$ [sync.]	7.61	A
	Max. $e(\beta)$ [sync.]	9.70	A

C. Power quality

Figure 10(d) shows the three-phase grid-injected currents during steady state. The currents are balanced and exhibit low distortion. The harmonic spectrum analysis is presented in Figure 11. The mean THD of the grid-injected currents is 3.48%, while the individual THD for currents in phase A, B, and C is 3.23%, 3.78% and 3.41%, respectively. This value complies IEEE 519 standard, which stipulates a maximum THD of 5%. Additional results are presented in the Appendix, comparing RMRAPID with other adaptive controller.

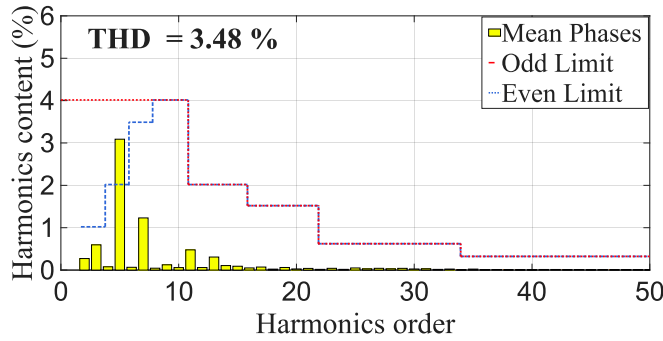


FIGURE 11. Mean total harmonic distortion of the three-phase grid-injected currents.

V. CONCLUSION

This work presented a novel adaptive control algorithm that unifies the intuitiveness of the PID controller with the resilience of RMRAC approaches. The resulting controller, named RMRAPID controller, has a full adaptive gains structure, with a completely mathematically fused structure of RMRAC and PID strategies. This novel controller is an extension of RMRAPI, which ensured system stability in a weak grid, while RMRAPI turned unstable (see Appendix). Experimental results demonstrated feasibility, robustness, and fast current tracking. The individual THD for grid-injected currents in phase A, B, and C was 3.23%, 3.78%, and 3.41%, respectively. The controller limitation is that it does not consider system constraints in the design as predictive strategies or meta-heuristic-tuned controllers. Therefore, future research will be done in this direction.

APPENDIX - PERFORMANCE COMPARISON

This appendix presents the performance comparison of the developed RMRAPID controller against RMRAPI controller [13]. The test was performed in the PSIM software, considering the same Kalman filter-based PLL, space vector modulation, and system parameters. To demonstrate the controller robustness operating in a weaker grid scenario compared to the previous experimental results, this test considers a grid inductance variation from 0.3 mH to 8 mH when time achieves 0.25 s. Figure 12 shows the currents tracking using both controllers. As can be seen, only the proposed RMRAPID controller was able to regulate the grid-injected currents respecting the current limitation of the system, which is 35 A. A zoom is provided inside this figure to highlight it. The RMRAPID achieved 34.47 A, while the RMRAPI presented 38.05 A, which will force the VSI to be disconnected in practice.

Figure 13 depicts the control action of both controllers. Notably, there are no concerns about the feasibility of implementation of the novel controller, since the DC link voltage is 500 V and the higher voltage value required by the RMRAPID controller was -223.61 V. As the RMRAPI

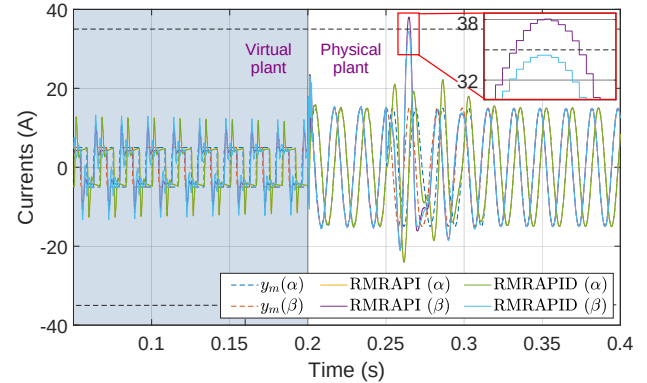


FIGURE 12. Comparison of RMRAPI and RMRAPID controllers: currents tracking.

controller would force the disconnection in practice, it is unfeasible in this case.

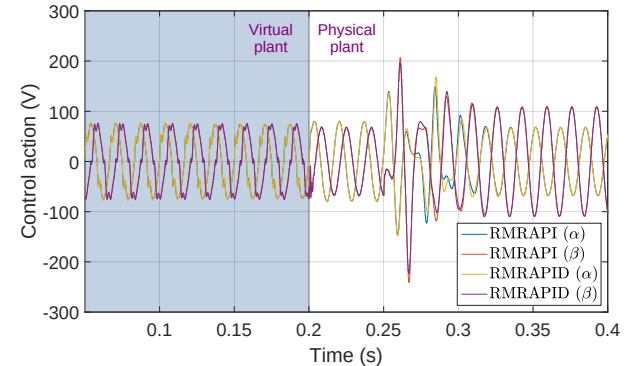


FIGURE 13. Comparison of RMRAPI and RMRAPID controllers: control actions.

Table 2 summarizes the performance comparison between both controllers. Three error metrics were evaluated: integral of absolute error (IAE), integral of squared error (ISE), and integral of absolute error pondered by the time (ITAE). In all metrics, the developed controller presented a slightly superior error index in α coordinate, but largely reduced the tracking error in β coordinate, resulting in a superior performance in the overall evaluation. The difference (diff.) between indexes can be checked in the last column of this table, where + indicates an increase and "-" depicts the reduction. But, in practice, only the developed RMRAPID is able to maintain the system operating connected to the grid. Thus, it is not useful to extend the performance comparison.

AUTHOR'S CONTRIBUTIONS

A.S.LUCENA: Conceptualization, Investigation, Software, Visualization, Writing – Original Draft.
C.T.G.B.OLIVEIRA: Validation, Writing – Original Draft.
V.S.FABRICIO: Writing – Original Draft.
W.B.D.SILVEIRA: Formal Analysis, Methodology,

TABLE 2. Performance comparison between RMRAPID and RMRAPI controllers.

Metric	RMRAC PI		RMRAC PID		Diff. (%)	
	α	β	α	β	α	β
IAE	0.0299	0.0101	0.0305	0.0088	+2.04%	-13.69%
ISE	0.0038	0.0014	0.0039	0.0012	+4.57%	-11.68%
ITAE	0.0065	0.0023	0.0066	0.0019	+2.01%	-18.65%

Resources, Software, Writing – Review & Editing.
B.ANDRES: Writing – Review & Editing. **P.J.D.O.EVALD:**
 Project Administration, Supervision, Validation, Writing –
 Review & Editing.

PLAGIARISM POLICY

This article was submitted to the similarity system provided by Crossref and powered by iThenticate – Similarity Check.

DATA AVAILABILITY

The data used in this research is available in the body of the document.

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BIOGRAPHIES

Alexandre Silva Lucena received the B.Sc. in Control and Automation Engineering at Federal University of Pelotas (UFPEL) in 2026. He is a member of the Intelligent System and Control Group (ISCG/UFPEL).

Cairo Túlio Gomes Barros Oliveira is currently a student in Control and Automation Engineering at UFPEL and a member of the ISGC/UFPEL.

Vitória de Souza Fabricio received the B.Sc. in Control and Automation Engineering at UFPEL in 2026. She is a member of ISCG/UFPEL.

Wagner Barreto da Silveira is a Professor at Federal University of Rio Grande do Sul and a member of ISCG/UFPEL and GEPOC/UFMS.

Bernardo Andres is a Professor at UFPEL and a member of ISCG/UFPEL.

Paulo Jefferson Dias de Oliveira Evald is a Professor at the UFPEL and leader of the ISCG/UFPEL, GEPOC/UFMS, INEP/UFSC and INCT-EPTech.