Data-driven control design applied to uninterruptible power supplies*

André Corleta[1], Charles Lorenzini[1], Jeferson Vieira Flores[1] and Lucíola Campestrini[2]

***Abstract*— This paper presents the application of the Virtual Reference Feedback Tuning (VRFT) method to the control of UPS systems. The VRFT method is a data-driven method which is able to estimate the parameters of a controller structure chosen by the user using only one batch of input-output data collected on the process, without using a process model. Two scenarios are presented for the UPS control: a resonant controller is designed using the standard VRFT methodology, with proper control structure and reference model selection; a current feedback gain and a resonant controller are designed, enhancing the transient response behavior compared to the first scenario. In order to apply the VRFT to current feedback design, the VRFT method is adapted since the controller is in the feedback loop, an unusual topology to data-driven applications. Simulated results show the applicability of data-driven methods to the control of UPSs and highlight the response enhancement with the current feedback control.**

I. INTRODUCTION

A popular way to protect critical loads against grid disturbances or failures is to feed them through an uninterruptible power supply – UPS. The control of UPSs have been attracting special attention from control community [1] since these systems are subject to tight constraints in both transient and steady-state performance. To regulate the UPS output voltage, control structures such as proportional-integral-derivative – PID [2], resonant [3] and repetitive [4] controllers have been reported in the literature, just to cite a few.

Multiloop strategies have also been applied to UPS [3], [5], [6], showing a significant improvement in tracking performance can be achieved through both voltage and current feedback. In [3], the problem is cast in a state-feedback framework and controller is designed by the solution of an optimization problem under Linear Matrix Inequalities constraints. In [5], a PI controller in the voltage loop is added to a proportional current feedback to control the parallel operation of single phase UPSs. Finally, load current decoupling methods are compared in [6] and a novel current reconstruction method is proposed. All design methods associated to these control strategies are based on the precise knowledge of the plant model, which can be a drawback specially in systems already in field operation.

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Data-driven control methods are constructed directly upon batches of input-output data collected from the process to be controlled and contrast with model-based control design mainly in two fundamental aspects: they are not based on the knowledge of a process model and they do not intend to freely determine the controller transfer function [7]. Based on an optimization criterion, these methods seek for the parameters of a controller structure in order to obtain a closed-loop response which is as close as possible to the desired response, determined by the user.

Different methodologies emerged in the past two decades: some of them are iterative, where usually a large number of experiments are realized to obtain a good response [8]; others are one-shot, that is, they are based on only one batch of input-output data [9]. One-shot methods solve a model reference control problem in order to obtain reference tracking. Several applications are presented in the literature, as process control (level, flow, temperature) [7], electro-hydrostatic actuator [10], control of wastewater treatment plants [11], among others. All of them are based on a desired response for step changes, therefore used to obtain PI or PID controllers. However, to the best of our knowledge, none of the data-driven methodologies have been applied to periodic tracking.

This paper presents the application of the one-shot Virtual Reference Feedback Tuning (VRFT) method to the control of UPS systems in order to obtain reference tracking for sinusoidal signals. Two controllers are designed: a resonant controller and a current feedback gain. In order to apply the VRFT to current feedback design, the VRFT method is adapted since the controller is in the feedback loop, an unusual topology to data-driven applications. Simulated results show the applicability of data-driven methods to the control of UPSs and highlight the transient response enhancement with the current feedback control.

II. CONTROL PROBLEM

In this work, the aim is to control a single-phase half-bridge DC-AC inverter usually employed in the output stage of UPSs [12] as illustrated in Fig. 1. We consider that the UPS input stage provides a constant voltage to the DC link (denoted by V_{cc}) and that the LC output filter parameters L_f and C_f were properly designed based on the UPS nominal power and switching frequency [13]. In particular, we assume that both inductor current i_L and capacitor voltage v_C are measurable and available for feedback. Control signal u_c is implemented through a Pulse Width Modulation – PWM such that the controller output is compared to a triangular wave to yield the signal that drives the switching of S_1 and S_2.

The main reason to use UPSs is to provide a sinusoidal voltage with the same amplitude and frequency of the electrical grid to critical loads, hence this system is subjected to severe performance requirements defined in standards such as the *IEC* 62040-3 [14]. Both transient and steady-state performance are evaluated based on two output measurements, the Total Harmonic Distortion – THD and the settling time of the RMS output voltage when a load transient occurs. System behavior is evaluated with respect to additive or subtractive steps of linear (purely resistive) load, while the steady state response is analyzed with respect to nonlinear (full wave rectifier and a RC filter) loads. In tests with linear loads to evaluate transient performance, the *IEC* 62040-3 establishes tolerance envelopes limiting the output RMS deviation accordingly to transient duration.

The main control objective associated to UPSs is to ensure the periodic reference tracking of sinusoidal signals. A natural choice is to consider an output feedback dynamic controller driven by the error signal between the output voltage and a reference signal mimicking the power grid voltage. This approach is illustrated in Fig. 2 disregarding the inner control loop, i.e., with Controller 2 equals to zero. From a theoretical point of view, controllers based on the internal model principle such as resonant controllers [3] are adequate to ensure periodic reference tracking. In the case of purely sinusoidal signals, the idea behind resonant control is to insert a controller with infinite gain at the reference frequency, resulting in unitary gain for the closed-loop transfer function between $r(t)$ and $v_C(t)$.

Moreover, if inductor current $i_L(t)$ is available for feedback one may consider a multiloop strategy as illustrated in Fig. 2. Notice that sinusoidal reference tracking is still guaranteed by a resonant controller (Controller 1) in the voltage loop, while the current controller (Controller 2) in the inner loop acts as an additional degree of freedom to improve transient performance.

In the sequel, we propose the use of a data-driven approach to design controllers 1 and 2, where the controller design is based solely on input-output data collected from experiments in the plant, and no process model is used whatsoever.

III. DATA-DRIVEN APPROACH

Data-driven control methods can be seen as controller identification methods, as an analogy to system identification: based on input-output data collected on the process and a chosen class for the controller structure, an algorithm is used to obtain the controller parameters [7]. The formulation of these methods are based on system identification formulation, which considers that a batch of data is collected with periodic sampling and the process can be described in the discrete time domain, therefore obtaining digital controllers. Methodologies applied to obtain reference tracking often consider the use of a one-degree of freedom controller, different from the topology we are going to employ on the UPS system, as presented in Fig. 2. Let us present the usual formulation first, and then we show how we can estimate both controllers under the data-driven approach.

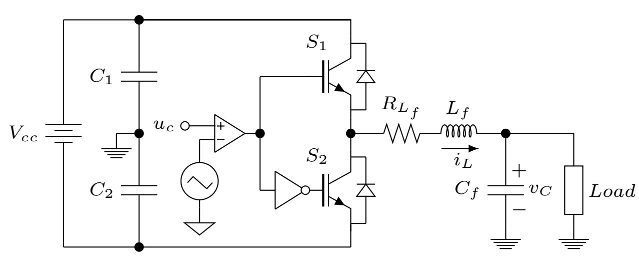

Fig. 1. Schematic representation of the UPS with load.

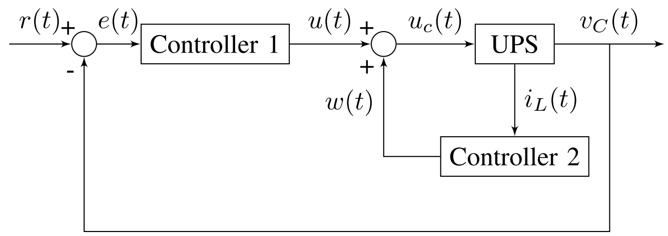

Fig. 2. Multiloop control.

A. System description

Consider a linear time-invariant discrete-time single-input-single-output process

$$v_C(t) = G(z)u(t) + \nu(t), \tag{1}$$

where z is the forward-shift operator, $G(z)$ is the process transfer function, $u(t)$ is the control input and $\nu(t)$ is a quasi-stationary noise process. This process is controlled by a linear time-invariant controller which belongs to a given - user specified - controller class $\mathcal{C}$ that is linearly parametrized: $\mathcal{C} = \{C(z,\rho) = \rho^T \bar{C}(z), \rho \in \mathbb{R}^n\}$, where $\bar{C}(z)$ is a n-column vector of fixed causal rational functions, whose poles are strictly inside the unit circle except for possible poles at $|z| = 1$. This class is such that $C(z,\rho)G(z)$ has positive relative degree for all $C(z,\rho) \in \mathcal{C}$; equivalently, the closed loop is not delay-free. The control action $u(t)$ can be written as $u(t) = C(z,\rho)(r(t) - y(t))$, where $r(t)$ is a reference signal, which is assumed to be quasi-stationary and uncorrelated with the noise [15]. The system in closed loop becomes

$$\begin{aligned} v_c(t,\rho) &= T(z,\rho)r(t) + S(z,\rho)\nu(t), \\ T(z,\rho) &= \frac{C(z,\rho)G(z)}{1 + C(z,\rho)G(z)} = C(z,\rho)G(z)S(z,\rho). \end{aligned}$$

B. Control design

Model Reference control design consists of specifying a “desired” closed-loop transfer function $T_d(z)$, which is known as the *reference model*, and then solving the following optimization problem for a specified reference signal $r(t)$:

$$\min_{\rho} J^{MR}(\rho) \tag{2}$$

$$J^{MR}(\rho) \triangleq \bar{E}\left[(T(z,\rho) - T_d(z))r(t)\right]^2, \tag{3}$$

where $\bar{E}[f(t)] \triangleq \lim_{N\to\infty} \frac{1}{N}\sum_{t=1}^{N} E[f(t)]$, E meaning expectation [15].

The *optimal controller* is defined as $C(z,\rho^*)$ with

$$\rho^* = \arg\min_{\rho} J^{MR}(\rho).$$

The data-driven approach is based on the assumption that the user can collect a batch of input and output data from the process and the *optimal parameters* of the controller $C(z,\rho^*)$ are estimated from these data, without the use of a mathematical model of the plant.

In this paper, we use the Virtual Reference Feedback Tuning (VRFT) method to estimate the controllers for the UPS system.

C. Standard VRFT method

Among several data-driven methodologies presented in the literature, the Virtual Reference Feedback Tuning (VRFT) is a one-shot method. Besides, it minimizes a different cost function, which allows the minimization to be realized through a least-squares algorithm.

Consider the noise free case, that is $\nu(t) = 0$ in (1). Through either an open-loop or a closed-loop experiment, input data $u(t)$ and output data $v_C(t)$ are collected on the process. Given the measured $v_C(t)$, the virtual reference signal $\bar{r}(t)$ is defined such that $T_d(z)\bar{r}(t) = v_C(t)$, and the virtual error is given by $\bar{e}(t) = \bar{r}(t) - v_C(t)$, as shown in Fig. 3.

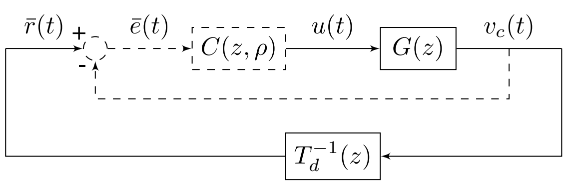

Fig. 3. Closed-loop block diagram and the virtual system's signals for the VRFT method.

Even though the plant $G(z)$ is unknown, when it is fed by $u(t)$ (the measured input signal), it generates $v_C(t)$ as output. So, a "good" controller is one that generates $u(t)$ when fed by $\bar{e}(t)$. Since both signals $u(t)$ and $\bar{e}(t)$ are known, the controller design can be seen as the identification of the dynamical relation between $\bar{e}(t)$ and $u(t)$. As a result of this reasoning, VRFT minimizes the following criterion

$$J^{VR}(\rho) = \bar{E}\left\{L(z)[u(t) - C(z,\rho)\bar{e}(t)]\right\}^2, \qquad (4)$$

where $L(z)$ is a filter used to approximate the minima of $J^{VR}(\rho)$ and $J^{MR}(\rho)$ [9]. This filter is given by

$$|L(e^{\jmath\omega})|^2 = |T_d(e^{\jmath\omega})|^2 |S(e^{\jmath\omega},\rho)|^2 \frac{\Phi_r(e^{\jmath\omega})}{\Phi_u(e^{\jmath\omega})},\ \ \forall\omega\in[-\pi,\pi]$$

where $\Phi_r(e^{\jmath\omega})$ is the spectrum of the reference signal $r(t)$ we want to apply to the closed-loop system and $\Phi_u(e^{\jmath\omega})$ is the spectrum of the applied input signal $u(t)$. If both functions are alike, so are their minimum. However, since $S(z,\rho)$ is unknown, the filter is approximated by

$$|L(e^{\jmath\omega})|^2 = |T_d(e^{\jmath\omega})|^2 |1 - T_d(e^{\jmath\omega})|^2 \frac{\Phi_r(e^{\jmath\omega})}{\Phi_u(e^{\jmath\omega})},\ \forall\omega\in[-\pi,\pi] \qquad (5)$$

where the approximation $|S(e^{\jmath\omega},\rho)|^2 \approx |S_d(e^{\jmath\omega})|^2$ was made.

If the controller is linearly parametrized, $J^{VR}(\rho)$ is quadratic and can be easily minimized, which is one of the main advantages over other data-driven methods. This methodology can be used to estimate the resonant controller for the UPS system. However, it can not be directly used to estimate the current feedback gain. We show how we can adapt the VRFT methodology to cope with this problem.

IV. MAIN RESULTS

A. Adapted VRFT for Current Feedback

Consider the subsystem presented in Fig. 4. Input data

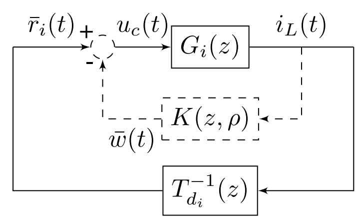

Fig. 4. Closed-loop block diagram and the virtual system's signals for the adapted VRFT method.

$u_c(t)$ and output data $i_L(t)$ are collected on the process. Given the measured $i_L(t)$, the virtual reference signal $\bar{r}_i(t)$ is defined such that $T_{d_i}(z)\bar{r}_i(t) = i_L(t)$. Then, the controller output signal can be obtained as $\bar{w}(t) = \bar{r}_i(t) - u_c(t)$.

Following the same reasoning of the standard VRFT formulation, a "good controller" is the one that generates $\bar{w}(t)$ when fed by $i_L(t)$. Since both signals are known, the controller design is again seen as the identification of the dynamical relation between $\bar{w}(t)$ and $i_L(t)$, which can be written as

$$J_i^{VR}(\rho_i) = \bar{E}\{L_i(z)[\bar{w}(t) - K(z,\rho_i)i_L(t)]\}^2, \qquad (6)$$

where $L_i(z)$ is a filter used to approximate the minima of $J_i^{VR}(\rho_i)$ and the model reference cost criterion to the control topology presented in Fig. 4. Notice that if $K(z,\rho_i)$ is linear in the parameters, then the controller is estimated again through the least squares method. In this case, the model reference cost criterion can be written as

$$J_i^{MR}(\rho_i) \triangleq \bar{E}\left[\left(\frac{G_i(z)}{1 + K(z,\rho_i)G_i(z)} - T_{d_i}(z)\right) r_i(t)\right]^2$$

where the *ideal controller* $K_d(z)$ is the one that allows $T(z,\rho_i)$ to match exactly $T_{d_i}(z)$.

Applying Parseval's Theorem in the reference model cost we get

$$\begin{aligned} J_i^{MR}(\rho_i) &= \frac{1}{2\pi}\int_{-\pi}^{\pi}\left|\frac{K_d G_i^2 - K(\rho_i)G_i^2}{(1+K(\rho_i)G_i)(1+K_d G_i)}\right|^2 \Phi_{r_i} d\omega, \\ &= \frac{1}{2\pi}\int_{-\pi}^{\pi} |K_d - K(\rho_i)|^2 |T_{d_i}|^2 |T(\rho_i)|^2 \Phi_{r_i} d\omega, \end{aligned} \tag{7}$$

where we have omitted the dependence on ω for the sake of space.

For the adapted virtual reference cost (6), consider the signal $\bar{w}(t)$. It can be written as

$$\begin{aligned} \bar{w}(t) &= T_{d_i}^{-1}(z)i_L(t) - G_i^{-1}(z)i_L(t) \\ &= K_d(z)i_L(t). \end{aligned} \tag{8}$$

Substituting (8) into (6) and applying Parseval's theorem yields

$$J_i^{VR}(\rho_i) = \frac{1}{2\pi}\int_{-\pi}^{\pi} |L_i|^2 |K_d - K(\rho)|^2 \Phi_{i_L} d\omega. \tag{9}$$

It is clear that if

$$|L_i(e^{\jmath\omega})|^2 = |T_{d_i}(e^{\jmath\omega})|^2 |T(e^{\jmath\omega},\rho_i)|^2 \frac{\Phi_{r_i}(e^{\jmath\omega})}{\Phi_{i_L}(e^{\jmath\omega})} \;\; \forall\omega\in[-\pi,\pi],$$

then (7) and (9) are alike and thus present the same minimum. Notice that the filter depends on the unknown quantity $T(e^{\jmath\omega},\rho_i)$. If we use the same idea as the one used in the standard VRFT and approximate $T(e^{\jmath\omega},\rho_i) \approx T_d(e^{\jmath\omega})$, then the filter is given by

$$|L_i(e^{\jmath\omega})|^2 = |T_{d_i}^2(e^{\jmath\omega})|^2 \frac{\Phi_{r_i}(e^{\jmath\omega})}{\Phi_{i_L}(e^{\jmath\omega})} \;\; \forall\omega\in[-\pi,\pi]. \tag{10}$$

Again, if the controller is linear in the parameters, that is $K(z,\rho_i) = \rho_i^T \bar{K}(z)$, where $\bar{K}(z)$ is a vector of fixed causal rational functions, then the solution for $J_i^{VR}(\rho_i)$ is given by the least squares method.

B. Controller Class and Reference Model Design

In order to apply data driven approaches presented here, one should define both the controller class and the closed-loop reference model that are going to be used. The main goal of the feedback control is to provide, in steady state, a sinusoidal voltage with the same amplitude and frequency of the electrical grid. This means that performance requirements should be translated into a desired output voltage $v_{C_d}(t)$.

1) Resonant Design: In order to follow a sinusoidal reference of a fixed frequency, either the controller or the process should contain the poles related to the reference signal, in this case, complex poles with $|z| = 1$. Therefore, the chosen controller class for Controller 1 in Fig. 3 is given by

$$C(\rho,z) = \frac{\rho_1 + \rho_2 z^{-1} + \rho_3 z^{-2} + \rho_4 z^{-3} + \rho_5 z^{-4}}{1 + 2\cos(\Omega)z^{-1} + z^{-2}}, \tag{11}$$

which is linear in the parameters, and $\Omega = \omega_0 T_s$ with T_s being the sampling period, and ω_0 the reference signal frequency.

This resonant controller is the one designed aiming to obtain the desired output voltage signal $v_{C_d}(t) = T_d(z)r(t)$, where $T_d(z)$ should be chosen accordingly. A reference model that represents null steady state error for a sinusoidal signal of a fixed frequency is one that presents unitary gain and null phase in that frequency. Besides, the poles of the reference model can be defined in order to achieve a desired settling time. The lower the value of the poles, the faster is the system response. Thus, the reference model to be used is given by

$$T_d(z) = k\frac{(1 - z_0 z^{-1})^2 z^{-1}}{(1 - pz^{-1})^4}, \tag{12}$$

where $0 < p < 1$ is defined by the user, while zeros z_0 and the gain k are calculated such that $|T_d(e^{\jmath\Omega})| = 1$ and $\angle T_d(e^{\jmath\Omega}) = 0$ in the frequency of interest.

2) Current feedback design: In order to improve the transient response of the UPS, the inductor current is fed back through a proportional gain. Thus, the controller class for the current feedback is given by $K(z,\rho_i) = K_i$. Notice, however, that the formulation of the adapted VRFT method presented in Section IV-A allows other controller structures, provided that linear in the parameters.

Since performance requirements are over the output voltage $v_C(t)$ and not on $i_L(t)$ itself, it is not direct how we should choose the reference model for the current signal. However, considering the UPS schematic on Fig. 1, one may find out the inductor current as a function of the output voltage. Therefore, a reference model for the desired inductor current can be determined as a function of the reference model for the output voltage $T_d(z)$. Using Euler's method to represent the derivative relation between the voltage and the current on the capacitor we obtain

$$i_L(t) = C_f \frac{1 - z^{-1}}{Ts} v_C(t) + Y_0 v_C(t). \tag{13}$$

Substituting $v_C(t)$ by the desired signal $v_{C_d}(t)$, we obtain a formulation for $i_{L_d}(t)$, the desired inductor current. Besides, if we substitute $v_{C_d}(t) = T_d(z)r(t)$, then we have a relation between $i_{L_d}(t)$ and the reference signal $r(t)$, which gives

$$i_{L_d}(t) = T_{d_{i_L}}(z)r(t) = [C_f \frac{1 - z^{-1}}{T_s} + Y_0]T_d(z)r(t).$$

If C_f and Y_0 are unknown, then we can approximate $\bar{T}_{d_{i_L}}(z) = (1 - z^{-1})T_d(z)$,where at least the derivative relation between voltage and current signals is preserved.

V. SIMULATION RESULTS

To illustrate and validate our method, we will consider a simulation environment combining softwares Matlab and PSIM, with PSIM responsible to simulate the power system (with all nonlinearities associated to inverter switching) while control implementation and design is performed in Matlab/Simulink. The numerical parameters considered in our simulations (see Table I) are based on a commercial 3.5kVA UPS system from Schneider Electric. Linear and nonlinear load parameters have been determined based on the IEC 62040-3, Annex E.

TABLE I

UPS AND LOAD PARAMETERS

Parameter	Symbol	Value
Filter inductance	L_f	1.0mH
Inductor resistance	R_{L_f}	15.0mΩ
Filter capacitance	C_f	300.0μF
Nominal Load admittance	Y_0	0.1519S
Reference frequency	f_0	60Hz
DC Link Voltage	V_{cc}	520V
PWM frequency		21.6kHz
Linear load resistor 20%	R_{l_1}	32.92Ω
Linear load resistor 80%	R_{l_2}	8.23Ω
Nonlinear load series resistor	R_{snl}	0.18Ω
Nonlinear load parallel resistor	R_{pnl}	10.39Ω
Nonlinear load capacitor	C_{nl}	12028.04μF

A. Performance requirements and user choices

The closed-loop system should present as output a voltage signal with null steady-state error to a 60 Hz sinusoidal reference. Besides, considering load addition and load removal, the settling time of the voltage signal should be such that the transient response remains inside an envelope defined by the standards. These performance criteria are translated into a reference model as presented in (12), where the position of the four poles (all in the same location) are chosen in order to satisfy the desired settling time, and gain k and zeros z_0 are calculated in order to obtain the steady-state behavior.

For the results presented here, the poles' position was chosen as $p = 0.83$, which gives

$$T_d(z) = 0.106\frac{(1-0.912z^{-1})^2z^{-1}}{(1-0.830z^{-1})^4}, \tag{14}$$

with a settling time of 1.2 ms, which satisfies the IEC 62040-3 standard. The resonant controller structure to be tuned is given by (11).

When the current feedback controller is used, a static gain K_i is used as controller and the reference model to the inductor current is a function of the voltage desired response, given by the approximate $\bar{T}_{d_i}(z)$.

Besides the performance criteria and the controller structures, one must define the signal to be applied in the experiment. This choice is not arbitrary, and depends on the operation point of the system and the operation modes one wants to excite during the experiment [15]. An open-loop test was realized aiming to collect data from the process. A sum of sinusoidal signals with frequencies from $10Hz$ up to $300Hz$ was utilized. By choosing this range of frequencies the operation point ($60Hz$) was contemplated, as well as the third and fifth harmonic components and the resonance frequency of the LC filter ($290Hz$). The chosen signal is given by

$$\begin{aligned} u_c(t) = 30\,[&\sin(2\pi 10t) + \sin(2\pi 60t) + \sin(2\pi 100t)\\ &= \sin(2\pi 150t) + \sin(2\pi 200t) + \sin(2\pi 300t)]\,. \end{aligned} \tag{15}$$

An experiment with nominal linear load was performed during 1 s, and voltage signal $v_C(t)$ and inductor current signal $i_L(t)$ where measured.

B. Controllers design

Aiming to verify the functionality of the methodology previously described, two different scenarios were tested: one that considers the use of the resonant controller only; other that estimates also the current feedback gain, applying the adapted version of the VRFT method presented in Section IV-A.

1) First scenario: Consider that the inductor current is not available for feedback. So, only the resonant controller is designed aiming to obtain the desired voltage response through the application of the standard VRFT methodology. The procedure is described as follows:

Procedure 1:

- In an open-loop experiment, apply the signal (15) as the input signal of the UPS and collect the voltage signal $v_C(t)$.
- Estimate the resonant controller (11) parameters.

Applying Procedure 1 results in $C_1(z, \hat{\rho})$ with

$$\hat{\rho} = [11.0483 \quad -25.3660 \quad 13.5067 \quad 5.12969 \quad -4.29969]\,.$$

2) Second scenario: Both capacitor voltage and inductor current are available and are going to be used in the UPS control. The idea is to first design the current feedback gain and, from a new experiment, obtain data to design the resonant controller. The procedure is described as follows:

Procedure 2:

- In an open-loop experiment, apply the signal (15) as the input signal of the UPS and collect the current signal $i_L(t)$.
- Estimate the current gain K_i through the adapted VRFT;
- Close the current loop with the estimated K_i and perform a second open-loop experiment for the voltage signal, applying the signal (15) as $u(t)$ (see Fig. 2) and collect the voltage signal $v_C(t)$.
- Estimate the resonant controller (11) parameters.

Applying Procedure 2 results in $K_i = 1.86278$ and $C_2(z, \hat{\rho})$ with

$$\hat{\rho} = [11.9421 \quad -28.6027 \quad 19.7789 - 1.14516 \quad -1.95577].$$

C. Load transient simulations

We considered additive and subtractive load steps from 20% to nominal linear load and from nominal linear load back to 20% as determined by the IEC 620040-3 standard [14]. Fig. 5 presents the output voltage just after the additive linear load step at $t = 0.203s$, where it is clear that reference tracking is achieved for both controllers in less than a quarter of reference cycle.

The closed-loop transient performance is better analyzed in the deviation profiles depicted in Figs. 6 and 7 for additive and subtractive load steps, respectively. With the insertion of the current feedback, the settling time is reduced to around 2ms and the response is less oscillatory. Also, that the maximum deviation is around 10% for the first scenario and is reduced to below 5% for the second.

Simulation results with nonlinear loads (rectifier + parallel RC filter) showed THD readings of 20.3% and 9.36% in the

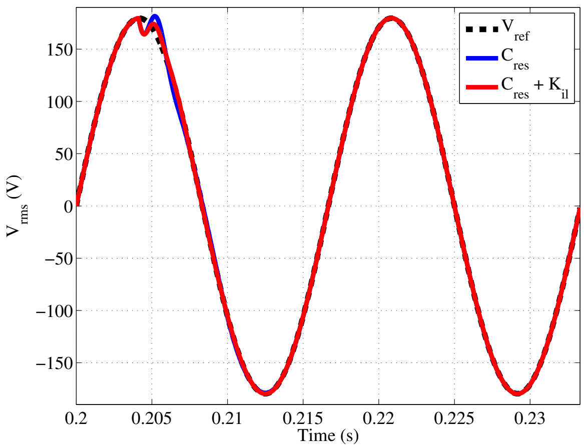

Fig. 5. System response for load transient simulations.

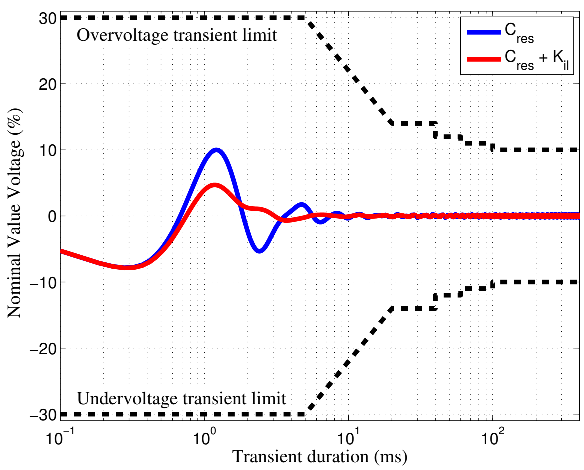

Fig. 6. Transient response for the additive (20% to 100%) linear load step.

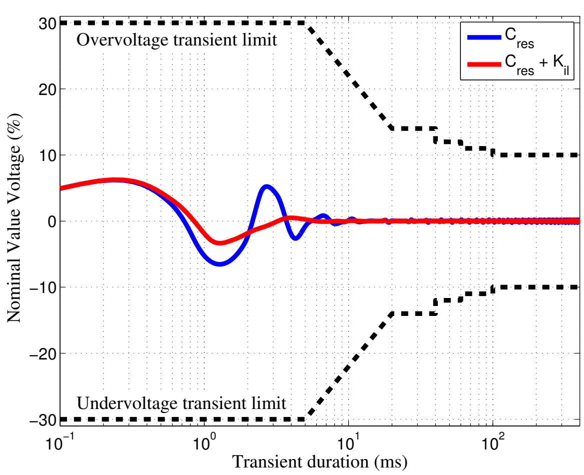

Fig. 7. Transient response for the subtractive (100% to 20%) linear load step.

first and second test scenarios, respectively. A THD of 9.36% is at the same level of performance as obtained with LMI *model-based* design in [3] with the same control structure (resonant + proportional current feedback).

VI. CONCLUSIONS

The application of the VRFT method to UPSs control was presented. Two different scenarios are possible: with or without the usage of a current proportional controller. From the results, an enhancement on the transient performance was obtained using the feedback current gain together with the resonant controller. Although the design focuses on the transient behavior to linear loads, the closed loop considering the resonant controller together with the current gain resulted in lower THD values also for non-linear loads, which shows the importance of this topology. Better results can be obtained by considering multiple-resonant controllers in order to reject the higher harmonics from the non-linear load, which is a subject under investigation.

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