Dissipativity-Based Adaptive and Robust Control of UPS in Unbalanced Operation

Gerardo Escobar Valderrama, Aleksandar M. Stanković, *Member, IEEE*, and Paolo Mattavelli, *Member, IEEE*

***Abstract*—In this paper, we investigate the output voltage control for three phase uninterruptible power supply (UPS) using controllers based on ideas of dissipativity. To provide balanced sinusoidal output voltages even in the presence of nonlinear and unbalanced loads, we first derive a dissipativity-based controller using a conventional $\alpha\beta$ (fixed frame) representation of system dynamics and a frequency-domain representation of system disturbances. Adaptive refinements have been added to the controller to cope with parametric uncertainties. Second, based on the structure of the first adaptive controller, we propose another controller that leads to a linear time-invariant (LTI) closed loop system which is directly connected to synchronous frame harmonic voltage control. This controller, denoted as robust, avoids the most computationally demanding parameter estimation during adaptation, and offers important advantages for implementation. For the proposed robust controller, a sufficient condition in terms of the design parameters is presented to guarantee stability of the desired equilibrium and robustness against certain parametric uncertainties. Finally, simulation and experimental results on a three-phase prototype show effectiveness and advantages of the proposed class of controllers.**

***Index Terms*—Adaptive control, dissipative systems, nonlinear systems, power supplies, uninterruptible power systems.**

I. Introduction

THE most important performance specifications for uninterruptible power supplies (UPS) systems include voltage regulation, total harmonic distortion, output impedance, transient response and operation with nonlinear/distorted loads. In addition, UPS systems are usually affected by parametric uncertainties and expected to operate under unbalanced conditions. The problem of designing an appropriate UPS control strategy that fulfills all requirements is thus clearly challenging. The growing importance of UPS systems has motivated a flourishing development of different control schemes found in the literature [1]–[9], [13]–[15]. Some controllers rely on single voltage loop using PI, dead-beat [5] or sliding mode controllers as compensators (see [8], [11], [14] for a brief survey on conventional control techniques for UPS). Other solutions proposed in the literature include a nested connection of output voltage and inductor current control loops, usually two PI's or possibly a PI plus a high gain controller like a sliding mode controller [2], [9]. Although these techniques are able to ensure a good transient response, the distortion on the output voltage due to nonlinear loads is typically not compensated completely. Other nonconventional approaches have emerged to overcome these limitations, like repetitive control which has the capability to compensate for periodic disturbances [4], [7], [13], [15]. To complete the design and to ensure acceptable transient response, this technique has to be combined/embedded with another control design approach (e.g., model reference adaptive control or pole placement) thus yielding controllers that are complicated for implementation even in the single phase case.

This paper aims to provide an alternative solution for the reduction of unbalance and distortion in UPS applications, and explores a family of controllers following the dissipativity approach. Starting from frequency domain descriptions of disturbances, our solution is able to perform precise voltage tracking, even with distorting loads. This feature is shared by some frequency domain techniques, such as repetitive control [15] and synchronous frame harmonic control [18]. We model the system dynamics using stationary frame quantities and the load currents (disturbance) with slowly varying phasors, with positive sequence *and* negative sequence quantities (to include unbalanced conditions).

Using dissipativity ideas, we first derive an adaptive controller that guarantees system stability under parametric uncertainties. The controller realizes a partial inversion of the system, and adds the needed damping. The resulting system contains a disturbance term due to the uncertainty in system parameters, and this term is addressed via adaptation. Due to the complexity of this controller, we also propose a simple rotational transformation so that the computation complexity can be significantly reduced, similarly to [16]. Motivated by the form of the first controller, a second controller that preserves the same structure is proposed. In the second case we fix one of the parameter estimates to a certain predefined value. The resulting controller is easier to implement as it is linear and time invariant (LTI), so that stability tests may be performed with traditional tools, like the Routh-Hurwitz criterion. Similarly to other frequency domain techniques, a group of selected harmonics is taken into account for parameter adaptation and voltage regulation. The resulting scheme is directly connected to our previous work [17], where, due to the phasor dynamic modeling of the entire dynamic system, a set of approximations were needed for controller implementation. Our solution proposed here is based on a new, more complete theoretical framework, as only the disturbance terms are represented in the frequency domain and no approximations are needed for final control implementation. Fi-

Manuscript received June 4, 2001; revised February 1, 2003. Recommended by Associate Editor M. A. Rahman.

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Digital Object Identifier 10.1109/TPEL.2003.813768

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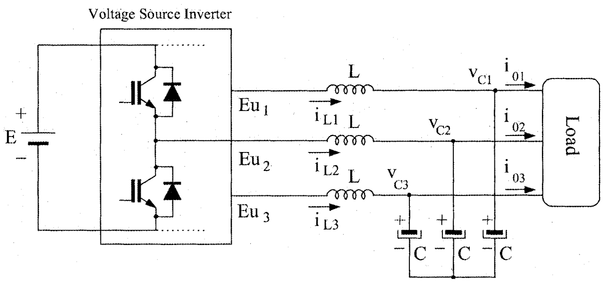


Fig. 1. UPS inverter system.

nally, the proposed control scheme has been implemented using a fixed-point single-chip digital signal processor (ADMC401 by Analog Devices). Experimental results for the proposed robust controller are presented, and compared with those of a conventionally tuned PI controller displaying the advantages of our solution.

II. System Configuration and Problem Formulation

The basic setup for the UPS application discussed in this paper is shown in Fig. 1.

The system dynamics are described by the following expressions

$$\frac{d}{dt}i_L = -\frac{1}{L}v_C + \frac{E}{L}u \tag{1}$$

$$\frac{d}{dt}v_C = \frac{1}{C}(i_L - i_0) \tag{2}$$

- L Inductance;
- C Capacitance;
- E voltage source;
- i_L inductor currents;
- v_C capacitor voltages;
- u control;
- i_0 load current;

where i_L, v_C, u and i_0 are vector quantities of the form $x = [x_\alpha, x_\beta]^\top$ expressed in $\alpha\beta$ coordinates. Parameters L, C, E are all assumed unknown constants, or slowly varying, except for possible step changes following structural changes in the system. Current i_0 is an unbalanced periodic signal which can be expressed as the combination of a fundamental component (at a fixed frequency w) and its harmonics of higher order, that is, we can represent i_0 as

$$i_0 = \sum_{k\in\mathcal{H}} e^{\mathcal{J}wkt} I_{0,k}^p + \sum_{k\in\mathcal{H}} e^{-\mathcal{J}wkt} I_{0,k}^n \tag{3}$$

where vectors $I_{0,k}^p$, $I_{0,k}^n \in \mathbb{R}^2$ are the k^{th} harmonic coefficients for the positive and negative sequence representation, they are also assumed unknown constants (or slowly varying); $\mathcal{H} = \{1,2,3,\ldots\}$ is the set of multiples of the harmonic components considered and $\mathcal{J} = -\mathcal{J}^\top = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$.

The *control objective* is to track a balanced voltage reference $v_C^* = e^{\mathcal{J}wt}[V_d, 0]^\top$ (which is a purely sinusoidal vector signal), in spite of the presence of harmonic disturbances. Here and in what follows $(\cdot)^*$ will be used to denote references and $(\bar{\cdot})$ for values in an equilibrium. The control objective thus implicitly includes two problems, reference tracking in the fundamental harmonic and disturbance attenuation of the output voltage response to higher harmonics mainly introduced by the load current.

The equilibrium point of the overall system (1), (2) by forcing $v_C^* = e^{\mathcal{J}wt}[V_d, 0]^\top$ is given by

$$\bar{i}_L = i_0 + \mathcal{J}wCv_C^*. \tag{4}$$

Note that, in order to perform the voltage regulation, the inductance current must provide the harmonic content of the load current.

III. Adaptive Strategy

Let us write the system (1), (2) in incremental terms as

$$L\frac{d}{dt}\tilde{i}_L = -\tilde{v}_C + Eu - v_C^* - L\frac{d}{dt}i_L^* \tag{5}$$

$$C\frac{d}{dt}\tilde{v}_C = \tilde{i}_L + i_L^* - (i_0 + \mathcal{J}wCv_C^*) \tag{6}$$

where $\tilde{i}_L \triangleq i_L - i_L^*$, $\tilde{v}_C \triangleq v_C - v_C^*$ and we have used the fact that $(d/dt)v_C^* = \mathcal{J}wv_C^*$.

In the case of known parameters, i.e., $\bar{i}_L$ substitutes i_L^* with C, i_0 (and its first time derivative) and L known, the following controller stabilizes the system at the desired equilibrium point

$$Eu = -R_1(i_L - \bar{i}_L) - R_2\tilde{v}_C + v_C^* + L\frac{d}{dt}\bar{i}_L$$

where $R_1 = R_1^\top > 0$ and $R_2 = R_2^\top > 0$ are design parameters and $i_L^* = \bar{i}_L$.

The dissipativity-based control design can now proceed as follows. First, a copy of the system is constructed and evaluated in the desired steady state. Second, we add the required damping by feeding back the errors through gains R_1 and R_2. Finally, from the resulting expression, we solve for the controller u; see [12] for further details in the passivity-based control design.

In the case parameters L and C and signal i_0 are unknown, we propose the following adaptive controller to which adaptation has been added to compensate for parameter uncertainties

$$Eu = -R_1(i_L - \hat{i}_L) - R_2\tilde{v}_C + v_C^* + \hat{L}\frac{d}{dt}\hat{i}_L \tag{7}$$

where $(\hat{\cdot})$ stands for estimated quantities, and we have redefined $\tilde{i}_L = i_L - \hat{i}_L$. Notice that $\hat{i}_L$ is being used as the estimate for the unknown signal i_L^*.

Update of $\hat{L}$ is done by a single gradient law while $\hat{i}_L$ is reconstructed indirectly by the estimation of its Fourier coefficients as shown below. Let us assume that $\bar{i}_L$, considered unknown, has

$$\bar{i}_L = \sum_{k\in\mathcal{H}} e^{\mathcal{J}wkt}\bar{I}_{L,k}^p + \sum_{k\in\mathcal{H}} e^{-\mathcal{J}wkt}\bar{I}_{L,k}^n \tag{8}$$

inherited from the previously defined i_0 in (3).

An estimation for this signal represented by $\hat{i}_L$ is

$$\hat{i}_L = \sum_{k\in\mathcal{H}} e^{\mathcal{J}wkt}\hat{I}_{L,k}^p + \sum_{k\in\mathcal{H}} e^{-\mathcal{J}wkt}\hat{I}_{L,k}^n \tag{9}$$

where $\hat{I}^{p,n}_{L,k}$ are the estimates for the coefficients $\overline{I}^{p,n}_{L,k}$.

Thus the estimation error signal $\varepsilon_L \triangleq \hat{i}_L - \overline{i}_L$ becomes

$$\varepsilon_L = \sum_{k\in\mathcal{H}} e^{\mathcal{J}wkt}\mathcal{E}^p_{L,k} + \sum_{k\in\mathcal{H}} e^{-\mathcal{J}wkt}\mathcal{E}^n_{L,k} \tag{10}$$

where $\mathcal{E}^{p,n}_{L,k} \triangleq \hat{I}^{p,n}_{L,k} - \overline{I}^{p,n}_{L,k}$.

The closed loop system becomes

$$L\frac{d}{dt}\tilde{i}_L = -R_1\tilde{i}_L - (1+R_2)\tilde{v}_C + \tilde{L}\frac{d}{dt}\hat{i}_L \tag{11}$$

$$C\frac{d}{dt}\tilde{v}_C = \tilde{i}_L + \varepsilon_L \tag{12}$$

where $\tilde{L} \triangleq \hat{L} - L$.

Following the Lyapunov approach we get the following adaptive laws:

$$\dot{\hat{I}}^p_{L,k} = -\gamma_k e^{-\mathcal{J}wkt}\tilde{v}_C, \quad k\in\mathcal{H} \tag{13}$$

$$\dot{\hat{I}}^n_{L,k} = -\gamma_k e^{\mathcal{J}wkt}\tilde{v}_C, \quad k\in\mathcal{H} \tag{14}$$

$$\dot{\hat{L}} = -\gamma_0\tilde{i}_L^\top\frac{d}{dt}\hat{i}_L \tag{15}$$

where $\gamma_0 > 0$ and $\gamma_k > 0$, $(k\in\mathcal{H})$ are scalar design parameters. These adaptations make negative semi-definite the time derivative of the energy storage function

$$W = \frac{L}{2(1+R_2)}|\tilde{i}_L|^2 + \frac{C}{2}|\tilde{v}_C|^2 + \frac{1}{2\gamma_0(1+R_2)}\tilde{L}^2 + \frac{1}{2}\sum_{k\in\mathcal{H}}\frac{1}{\gamma_k}\left[\left(\mathcal{E}^p_{L,k}\right)^2 + \left(\mathcal{E}^n_{L,k}\right)^2\right]$$

where $|\cdot|$ represents the module of a vector, thus $|X|^2 = X^\top X$. As a first conclusion we have that $\tilde{i}_L \to 0$. Then invoking standard LaSalle's theorem arguments [10] we obtain an invariant set described by

$$\frac{C\tilde{L}}{1+R_2}\frac{d^2}{dt^2}\tilde{v}_C = \tilde{v}_C$$

and since $\tilde{v}_C$ is bounded, then the only possible solution is $\tilde{v}_C = 0$ which in its turn implies $\tilde{L} = 0$ and $\varepsilon_L = 0$.

Complexity of the expressions in the controller above can be reduced if rotation matrices of the form $e^{\pm\mathcal{J}kwt}$ are avoided. For this, consider the following coordinate transformation:

$$\hat{i}^p_{L,k} = e^{\mathcal{J}kwt}\hat{I}^p_{L,k} \tag{16}$$

$$\hat{i}^n_{L,k} = e^{-\mathcal{J}kwt}\hat{I}^n_{L,k} \tag{17}$$

therefore

$$\hat{i}_L = \sum_{k\in\mathcal{H}}\left(\hat{i}^p_{L,k} + \hat{i}^n_{L,k}\right) = \sum_{k\in\mathcal{H}}\hat{i}_{L,k}.$$

Using (13), (14) their time derivatives are given by

$$\frac{d}{dt}\hat{i}^p_{L,k} = -\gamma_k\tilde{v}_C + \mathcal{J}kw\hat{i}^p_{L,k} \tag{18}$$

$$\frac{d}{dt}\hat{i}^n_{L,k} = -\gamma_k\tilde{v}_C - \mathcal{J}kw\hat{i}^n_{L,k}. \tag{19}$$

The expression for the adaptive controller (7) in terms of the new variables is

$$Eu = -R_1 i_L - \left(2\hat{L}\sum_{k\in\mathcal{H}}\gamma_k + R_2\right)\tilde{v}_C + v_C^* + \sum_{k\in\mathcal{H}}\left[(R_1 + \mathcal{J}kw\hat{L})\hat{i}^p_{L,k} + (R_1 - \mathcal{J}kw\hat{L})\hat{i}^n_{L,k}\right] \tag{20}$$

where $\hat{L}$, $\hat{i}^p_{L,k}$ and $\hat{i}^n_{L,k}$ are computed according to (15),(18), and (19), respectively.

IV. Robust Adaptive Controller

The controller (20) above can be significantly simplified if estimation of L in (15) is avoided. We propose then to substitute the estimate $\hat{L}$ in (20) for some predefined value L_0 with the hope that the error caused by this uncertainty can be absorbed by the robust controller action, that is

$$Eu = -R_1 i_L - \left(2L_0\sum_{k\in\mathcal{H}}\gamma_k + R_2\right)\tilde{v}_C + v_C^* + \sum_{k\in\mathcal{H}}\left[(R_1 + \mathcal{J}kwL_0)\hat{i}^p_{L,k} + (R_1 - \mathcal{J}kwL_0)\hat{i}^n_{L,k}\right] \tag{21}$$

and the adaptations are now reduced to only (18) and (19).

We remark that L_0 is considered to be a design parameter, and not necessarily an estimate of L. This design parameter should fulfill the condition $L_0 \geq L_M$, where L_M is a known lower bound for L, as will become clear later.

The closed loop system with the controller above yields the following LTI dynamics:

$$L\frac{d}{dt}i_L = -R_1 i_L + \sum_{k\in\mathcal{H}}\left[(R_1 + \mathcal{J}kwL_0)\hat{i}^p_{L,k} + (R_1 - \mathcal{J}kwL_0)\hat{i}^n_{L,k}\right] - \left(1 + R_2 + 2L_0\sum_{k\in\mathcal{H}}\gamma_k\right)(v_C - v_C^*) \tag{22}$$

$$C\frac{d}{dt}v_C = i_L - i_0 \tag{23}$$

$$\frac{d}{dt}\hat{i}^p_{L,k} = -\gamma_k(v_C - v_C^*) + \mathcal{J}kw\hat{i}^p_{L,k}, \quad k\in\mathcal{H} \tag{24}$$

$$\frac{d}{dt}\hat{i}^n_{L,k} = -\gamma_k(v_C - v_C^*) - \mathcal{J}kw\hat{i}^n_{L,k}, \quad k\in\mathcal{H} \tag{25}$$

whose equilibria is a periodic orbit given by

$$\overline{v}_C = v_C^*, \quad \overline{i}_L = i_0 + \mathcal{J}wCv_C^*$$
$$\overline{\hat{i}}^p_{L,k} = (R_1 + \mathcal{J}kwL_0)^{-1}(R_1 + \mathcal{J}kwL)\overline{i}^p_{L,k}, \quad k\in\mathcal{H}$$
$$\overline{\hat{i}}^n_{L,k} = (R_1 - \mathcal{J}kwL_0)^{-1}(R_1 - \mathcal{J}kwL)\overline{i}^n_{L,k}, \quad k\in\mathcal{H}$$

where $\overline{(\cdot)}$ is used to represent trajectories in the equilibrium, $\overline{i}^p_{L,k}$ and $\overline{i}^n_{L,k}$ are the k^{th} positive and negative sequence components of $\overline{i}_L$ referred to the fixed frame, i.e., $\overline{i}_L = \sum_{k\in\mathcal{H}}(\overline{i}^p_{L,k} + \overline{i}^n_{L,k})$.

To deal with the stability study we need to compute the characteristic polynomial of the linear system above. The system order has to be reduced to make this symbolic calculation tractable, so we interpret the matrix $\mathcal{J}$ as analog of the complex number $j = \sqrt{-1}$ and we consider all design matrices as scalars. We observed however, that the resulting polynomial has real coefficients, thus using the standard Routh-Hurwitz criterion, we can establish the following condition:

$$L_0 > L - (1+R_2)\frac{\phi_1}{\phi_2}$$

where $\phi_1 > 0$ and $\phi_2 > 0$ are two rather involved and positive functions of the parameters. Hence, it is enough to select an L_0 such that

$$L_0 > L_M \tag{26}$$

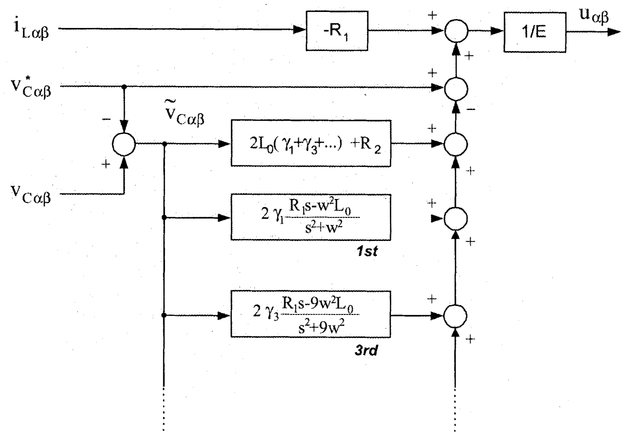


Fig. 2. Block diagram of the proposed controller.

to guarantee asymptotic stability of the equilibrium. Notice that gain R_2 helps to relax this condition.

The controller expression (21) with adaptive laws (18), (19) can be also expressed in a more familiar form by considering the following transformations:

$$\eta_k^p = -(R_1 + kwL_0\mathcal{J})\hat{i}_{L,k}^p$$
$$\eta_k^n = -(R_1 - kwL_0\mathcal{J})\hat{i}_{L,k}^n.$$

This yields the following expression for the controller

$$\begin{aligned} Eu =& -R_1 i_L - \left(2L_0 \sum_{k\in\mathcal{H}} \gamma_k + R_2\right)\tilde{v}_C + v_C^* \\ & - \sum_{k\in\mathcal{H}} (\eta_k^p + \eta_k^n) \\ \dot{\eta}_k^p =& (R_1 + \mathcal{J}kwL_0)\gamma_k \tilde{v}_C + \mathcal{J}kw\eta_k^p \\ \dot{\eta}_k^n =& (R_1 - \mathcal{J}kwL_0)\gamma_k \tilde{v}_C - \mathcal{J}kw\eta_k^n. \end{aligned}$$

By expressing the dynamical part of the controller in the form of a transfer function be can also write the controller above as

$$\begin{aligned} Eu =& -R_1 i_L - \left(2L_0 \sum_{k\in\mathcal{H}} \gamma_k + R_2\right)\tilde{v}_C + v_C^* - \eta_{\alpha\beta} \\ \eta_{\alpha\beta} =& \sum_{k\in\mathcal{H}} \left(\frac{R_1 s - k^2 w^2 L_0}{s^2 + k^2 w^2}\right) 2\gamma_k \tilde{v}_C \end{aligned} \tag{27}$$

where s is the complex variable. Fig. 2 presents the block diagram of the proposed controller (27). Very interesting is the fact that the compensation of harmonics in each second order filter requires the introduction of a zero on the right hand side of the complex plane.

Concerning the DSP implementation, we recall that our solution requires the sensing of output voltages and inductor currents, so that the requirement in term of hardware peripheral devices is exactly the same as in a conventional multi-loop scheme. Our computational requirements, as shown in Fig. 2, are notably higher, since two second-order filters are needed for each compensated harmonic (one for the α component and one for the β component). Such signal processing requirements are not, however, a serious limitation for modern control DSP's, even for the compensation of a large number of harmonics. Our nonoptimized implementation of such filters, for example, requires around 2.5 μs for each harmonic component.

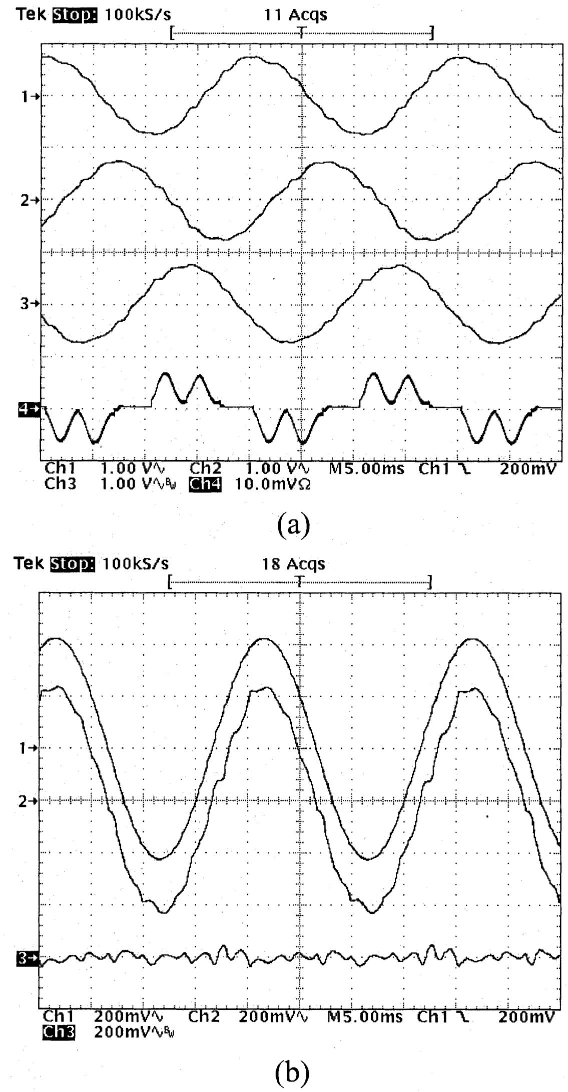
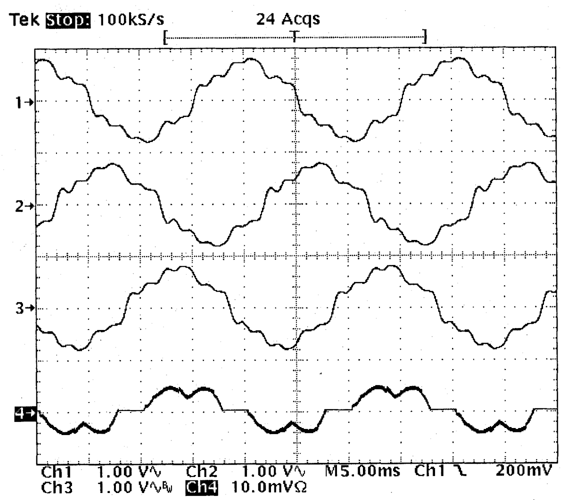


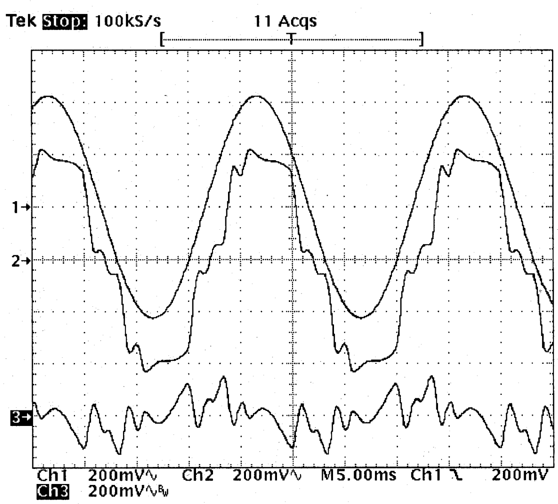
Fig. 3. Three-phase rectifier load with the proposed solution: (a) (from top to bottom) output voltage phase $a-b-c$ (100 V/div) and phase c output current (10 A/div); (b) (from top to bottom) output voltage reference, output voltage in β coordinate (offset to clearly show the difference) and the corresponding error (40 V/div).

Regarding the selection of controller parameters, a set of reasonable approximations can be used for an initial setting of their values.

1) The matrix coefficient R_1 can be set as $R_1 = k_{ip} \cdot I_2$, where k_{ip} is a conventional proportional gain of a PI current controller. Accordingly, we can set k_{ip} to be equal to $2\pi f_{ic} \cdot L$, where f_{ic} is the desired current loop bandwidth, usually 1/10–1/15 of the switching frequency.
2) Parameter L_0 can be set equal to (the possibly rough knowledge of) inductor value L.
3) The matrix coefficient R_1 can be set as $R_2 = k_{vp} \cdot I_2$, where k_{vp} is now a conventional proportional gain of a PI voltage controller in a multiloop solution.
4) Finally, gains γ_k can be set so as to compensate the remaining transfer function that can be roughly approximated as first order pole at designed voltage loop bandwidth with a dc gain equal to $1/k_{vp}$.



(a)



(b)

Fig. 4. Three-phase rectifier load with PI control: (a) (from top to bottom) output voltage phase $a-b-c$ (100 V/div) and phase c output current (10 A/div); (b) (from top to bottom) output voltage reference, output voltage in β coordinate (offset to clearly show the difference) and the corresponding error (40 V/div).

Disregarding for simplicity the influence of such pole, we can set the gain γ_k as $\gamma_k = 2.2 * k_{vp}/T_{kr}$, where T_{kr} is the desired response time for each harmonic component (evaluated between the 10% and 90% of a step response of the amplitude of the corresponding sinusoidal perturbation).

V. Experimental Results

The proposed controller has been experimentally tested using a reduced-scale prototypes with the following parameters: $L = 4\ mH$, $C = 15\ \mu F$, =$E = 200\ V$, switching frequency $f_{sw} = 10\ kHz$, output voltage frequency $f = 50\ Hz$, and selected frequencies: first, third, fifth, seventh, and ninth. Following the proposed design guidelines, control parameters have been chosen as follows: $R_1 = 25.1$, $L_0 = 4\ mH$, $\gamma_k \in \{14.1, 1.47, 1.59, 1.9\}$, where $\gamma_1 = 14.1$ for the fundamental component, $\gamma_3 = 1.47$ for the third harmonic component and so on, and $R_2 = 0.256$. The proposed control strategy has been implemented by means of the 16-b fixed point DSP-based controller ADMC401 by Analog Devices. This DSP

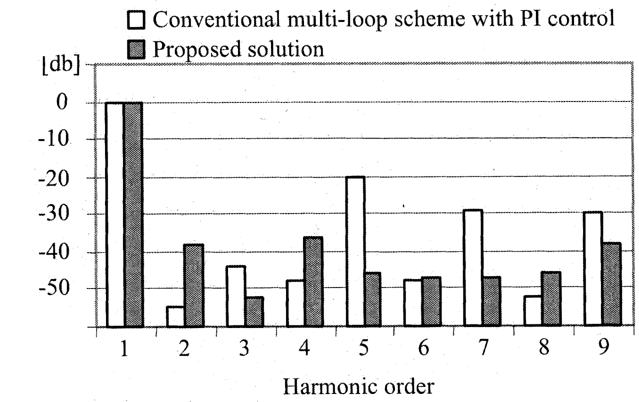
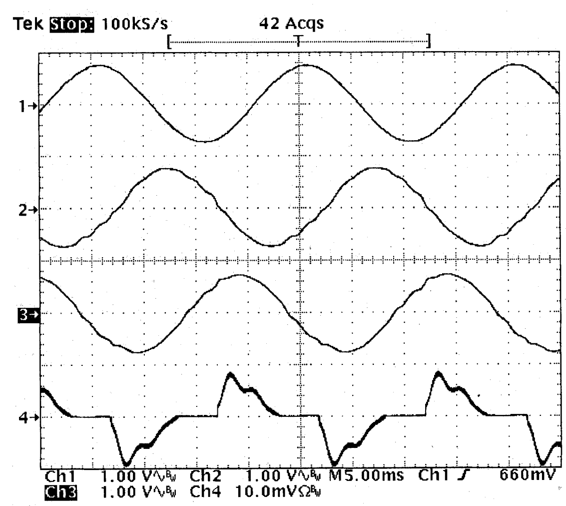
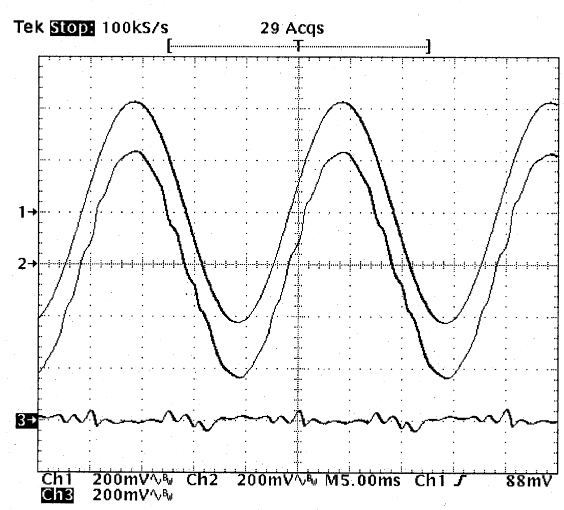


Fig. 5. Normalized output voltage spectrum.

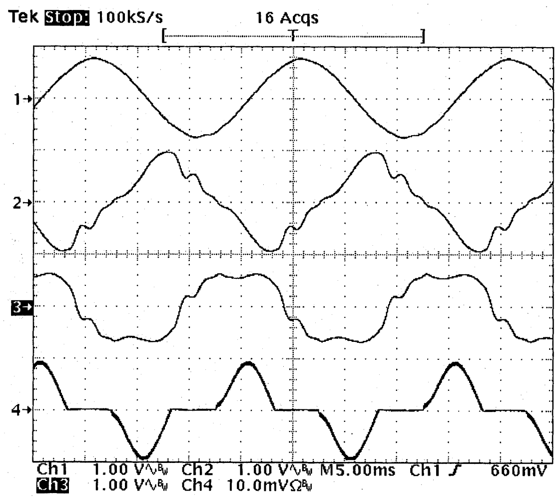


(a)

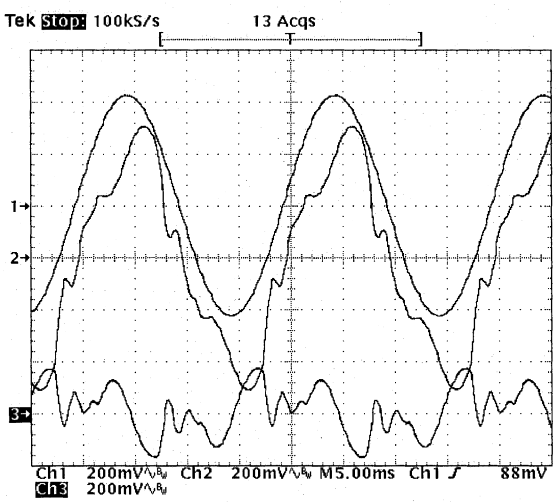


(b)

Fig. 6. Unbalance test (single-phase rectifier) with the proposed solution: (a) (from top to bottom) output voltage phase $a-b-c$ (100 V/div) and phase c output current (10 A/div) and (b) (from top to bottom) output voltage reference, output voltage in β coordinate (offset to clearly show the difference) and the corresponding error (40 V/div).



(a)



(b)

Fig. 7. Unbalance test (single-phase rectifier) with PI control: (a) (from top to bottom) output voltage phase $a-b-c$ (100 V/div) and phase c output current (10 A/div); (b) (from top to bottom) output voltage reference, output voltage in β coordinate and the corresponding error (40 V/div).

unit contains a capable arithmetic unit (26MIPS) and several embedded peripherals, such as a high-resolution PWM modulator, flash 12 b A/D converters, which allow conversions up to eight channels in less than 2 ms. We point out again that the time required to implement the control of each frequency (both for the α and β components) is around 2.5 μs using a nonoptimized assembly code, allowing control of a large number of harmonics.

The results of the proposed control with three-phase diode rectifier loads are reported in Fig. 3 (a) and (b), while the results obtained with conventional PI control are reported in Fig. 4(a) and (b). Note that the quality of the output voltage has been strongly improved respect to the PI control, since the dominant harmonics (i.e., the fifth and the seventh components) have been well compensated by the proposed strategy. Moreover, comparing Fig. 3(b) and Fig. 4(b) it is worth noting that also the fundamental component on the output voltage error has been strongly reduced. The improvement in terms of THD reduction are also evident from Fig. 5, which reports output voltage spectrum of Figs. 3(a) and 4(a). Again, note that the distortion at the selected frequencies has been reduced by control action. The compensation is not entirely complete, since a small residual distortion at the selected frequencies is still present. This phenomenon is mainly due the quantization and rounding errors in the fixed-point DSP implementation.

As a final and very challenging test for unbalanced conditions, we have applied a *single-phase* rectifier load to a phase-to-phase voltage. The results, reported in Fig. 6(a) and (b), are much better than those obtained with conventional multi-loop scheme [see Fig. 7(a) and (b)], highlighting the advantages of the proposed solution.

VI. Conclusion

In this paper, we present a family of dissipativity-based controllers for the output voltage regulation of a three phase uninterruptible power supplies. Each of the proposed controllers is expressed in terms of the conventional $\alpha\beta$ (fixed frame) representation, and provides balanced sinusoidal output voltages even in the presence of nonlinear and unbalanced loads. Adaptation was first added to the basic controller to cope with parametric uncertainties. A simplified (LTI robust) version of the adaptive controller was derived next under assumptions that are easy to satisfy in practice, and its stability demonstrated. The proposed robust controller was implemented and experimentally tested in balanced and extremely unbalanced operation. Comparisons with conventionally tuned PI controllers showed a considerable improvement.

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