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Design and analysis of a novel bounded nonlinear controller for three-phase ac/dc converters

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***Abstract*— A bounded nonlinear dynamic controller suitable for the regulation of the duty-ratio input of a three-phase AC/DC voltage source converter (VSC) is proposed. The controller design is based on the complete nonlinear average model of the power converter and takes into account that the controlled inputs -duty-ratio signals- should be saturated in a disk area on the input phase portrait while the grid voltages constitute the uncontrolled inputs. The proposed design approach fully satisfies the input constraint while simultaneously achieves precise DC bus voltage regulation and unity power factor operation. In addition, opposite to the conventional controllers, the present development is totally independent from the system parameters. A detailed nonlinear analysis shows that the closed-loop system operating under the uncontrolled external inputs has the input to state stability (ISS) property while the analysis is further extended to prove convergence to the desired nonzero equilibrium when constant external inputs act on the system. Simulation results verify the effectiveness of the controller under sudden DC bus voltage reference and load variations.**

I. INTRODUCTION

In recent decades, the use of three-phase AC/DC voltage source converters (VSC) is continuously increased due to their application in industrial environments and in renewable energy applications. Using pulse-width-modulation (PWM) techniques, VSCs can achieve precise regulation of the DC bus voltage, low harmonic distortion, bidirectional power flow, high power factor operation and simple control implementation. Therefore, since in modern industrial environments the traditional phase-controlled rectifiers have been almost completely substituted by the VSCs, the research interest of their control design has been further increased.

Several control methods have been proposed for the control design of power systems with a three-phase AC/DC VSC. In order to analyze and investigate the system behavior, it has been shown that the average model description [1], [2] provides an adequate dynamic description of an AC/DC converter, providing a nonlinear model which however increases the difficulty of the control design. Traditional proportional-integral (PI) controllers [3], [4] have been proposed which in the most common scenario include a PI controller for achieving unity power factor operation, while a cascaded PI controller is utilized to regulate the DC bus voltage indirectly through controlling the input current. These control schemes are dependent from the system parameters, since they also include appropriate decoupling terms in order to provide a linear model which can be easily handled [4]. As a result, these control methods are sensitive to modeling errors. Various intelligent control techniques [5] such as fuzzy control methods [6] are often used to improve the complete system performance, which however are not based on a solid theoretical analysis and therefore cannot guarantee closed-loop system stability.

Other control techniques that have been also proposed are sliding mode [7] and feedback linearization methods [8], [9], which are based on the average model and provide an enhanced performance in simulation tests but are sensitive to model uncertainties. Since the idea of the average VSC model description has been derived through Hamiltonian theory, several researchers have proposed passivity-based control (PBC) techniques [10], [11], [12] that include suitable Lyapunov functions in order to result in a stable closed-loop system, but their design is still dependent from the system parameters and states.

Another significant drawback of all the aforementioned methods is that none of them can guarantee the required bounded input operation and therefore external limiters are needed for the controlled input implementation. This is an inherent problem arising due to the fact that in all the power electronic controlled devices, the duty-ratio is the only controlled input [13].

In this paper, a new nonlinear controller design approach is developed, based on the average model of the three-phase AC/DC converter that represents the accurate nonlinear dynamic model -adequate for control design- while a resistance load is assumed to be connected to the converter output. The controller tasks are to achieve unity power factor operation and precise DC bus voltage regulation. Therefore, a novel nonlinear dynamic controller is proposed which simultaneously satisfies both tasks. However, since the proposed controller scheme acts on the duty-ratio signal inputs of the converter, it is appropriately designed to achieve specific technical requirements, i.e. saturated control variables in the desired and permitted by the linear modulation range [13]. Furthermore, due to the structure of the proposed nonlinear controller, only the regulated variables are needed to be fed back, leading to a fully independent scheme from the system and load characteristics. An extensive mathematical analysis verifies the controller performance by proving that the closed-loop system is input to state stable (ISS) in an area

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where a desired equilibrium caused by a constant external input exists and further proving that the solution converges to this equilibrium. The complete system is finally simulated and the controller efficiency is tested under several DC bus voltage reference values or under sudden load changes.

In Section II, the model of a three-phase AC/DC converter feeding a resistance load is presented and analyzed. In Section III, the proposed control scheme is provided and the closed-loop system is investigated. In Section IV, closed-loop system stability is proven using a suitable theoretical analysis. In Section V, simulation tests are conducted for different DC bus voltage reference signals under sudden changes of the load in order to investigate the effectiveness of the proposed controller. Finally, in Section VI, some conclusions are drawn.

II. Model of Three-phase AC/DC Converter With Resistance Load

The complete system consisting of a symmetrical three-phase ac power supply, a boosting inductor circuit, a dc filter capacitor, a three-phase AC/DC voltage source converter and a resistance load, is shown in Figure 1.

The three-phase AC/DC converter consists of six IGBT switching elements which are capable of conducting current and power in both directions and operates using PWM. The resistance and inductance of the boosting inductor are described by the parameters R and L respectively, C is the dc-side filter capacitor while R_L is the load resistance. The converter input voltages and currents are expressed by the notations V_i and I_i, $i = a, b, c$ respectively.

Finally, the three-phase power supply is considered to provide the following three-phase voltages:

$$\begin{aligned} U_a &= U_m \cos\left(\omega_s t\right) \\ U_b &= U_m \cos\left(\omega_s t - \frac{2\pi}{3}\right) \\ U_c &= U_m \cos\left(\omega_s t + \frac{2\pi}{3}\right) \end{aligned} \tag{1}$$

where U_m is the peak value of the supply phase voltage and ω_s is the angular frequency of the power supply.

Since PWM operation introduces switching functions, the three-phase system model contains discontinuous input terms which increase the difficulty of the analysis. In order to overcome this problem, the average model analysis is used [2]. Indeed, for control purposes, both the average models of PWM regulated converters and the circuit representation on the $d-q$ reference frame can be suitably used [1]. Particularly, using the synchronous rotating, voltage-oriented, $d-q$ reference frame, all sinusoidal quantities can be transformed into dc quantities at steady-state operation. As a result, suitable controllers can be effectively applied in order to regulate the quantities of the $d-q$ reference frame at their desired steady-state values. It should be noticed that the higher the PWM frequency the more accurate (practically identical) is the resulting model. Thus, one can perform the

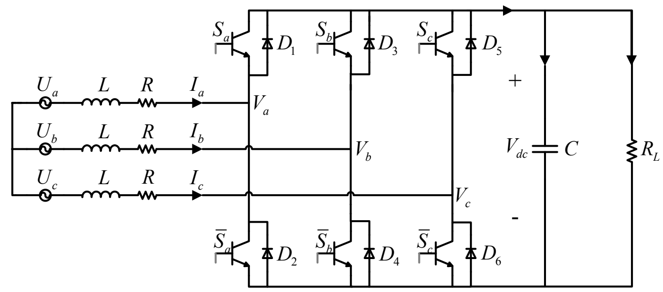

Fig. 1: Schematic diagram of three-phase ac/dc converter with resistance load

Park's Transformation [13]:

$$K_s = \frac{2}{3}\begin{bmatrix} \cos\left(\omega_s t\right) & \cos\left(\omega_s t - \frac{2\pi}{3}\right) & \cos\left(\omega_s t + \frac{2\pi}{3}\right) \\ -\sin\left(\omega_s t\right) & -\sin\left(\omega_s t - \frac{2\pi}{3}\right) & -\sin\left(\omega_s t + \frac{2\pi}{3}\right) \\ 0.5 & 0.5 & 0.5 \end{bmatrix}$$

to produce the average model of the converter in the synchronously $d-q$ rotating frame [10], [11].

Therefore, the complete system consisting of the converter and the resistance load is given as:

$$\begin{aligned} L\dot{i_d} &= -Ri_d + \omega_s L i_q - 2m_d V_{dc} + U_d \\ L\dot{i_q} &= -Ri_q - \omega_s L i_d - 2m_q V_{dc} + U_q \\ C\dot{V}_{dc} &= 3\left(m_d i_d + m_q i_q\right) - \frac{V_{dc}}{R_L} \end{aligned} \tag{2}$$

where the state vector is $x = \begin{bmatrix} i_d & i_q & V_{dc} \end{bmatrix}^T$ and i_d, i_q are the $d-$ and $q-$axis components of the line current, U_d and U_q are the $d-$ and $q-$axis components of the supply voltage. Additionally, $m_d = \frac{V_d}{2V_{dc}}$ and $m_q = \frac{V_q}{2V_{dc}}$ are the $d-$ and $q-$axis components of the switching duty ratios which represent the control input variables.

However, since U_d, U_q are constant (due to the transformation), the only controlled inputs are the duty-ratios m_d, m_q which appear in nonlinear terms.

The two control variables can be used to provide the modulation index m_a and the phase angle $\Delta\phi$ which are required for the PWM operation of the converter as shown below:

$$m_a = \sqrt{m_d^2 + m_q^2} \tag{3}$$

$$\Delta\phi = \arctan\left(\frac{m_q}{m_d}\right). \tag{4}$$

In a common three-phase AC/DC converter operation, the converter is desired to operate using linear modulation, i.e.:

$$m_a \leq 1 \Rightarrow m_d^2 + m_q^2 \leq 1. \tag{5}$$

Clearly, inequality (5) restricts the control inputs m_d and m_q to take values in a permitted range lying in a disk with unity radius. Also, from (5), it can be seen that each of m_d, m_q have as upper and lower bounds the 1 and -1, respectively.

On the other hand, the active and reactive power at the supply input, in the $d-q$ reference frame, [4], [14], are given by the expressions:

$$P = \frac{3}{2}\left(U_d i_d + U_q i_q\right) \tag{6}$$

and

$$Q = \frac{3}{2}\left(U_q i_d - U_d i_q\right). \tag{7}$$

In a typical AC/DC converter control application [14], the $q-$axis is aligned with phase$-a$ of the supply voltage and therefore it holds:

$$U_d = 0 \text{ and } U_q = U_m.$$

Under this assumption and in order to achieve unity power factor, i.e. operation with $Q = 0$, equation (7) constrains i_d to be zero [10], [11], [14].

Assuming, now, unity power factor operation and DC bus voltage regulation at V_{ref} there exists a desired steady-state equilibrium point x^* of (2).

III. The Proposed Nonlinear Controller

The two main tasks of the controller that are to achieve unity power factor and precise DC bus voltage regulation, are achieved by regulating the $d-$axis current to zero and the DC bus voltage V_{dc} to the desired level V_{ref}. Traditional techniques include a PI controller for the power factor control and a cascaded PI controller for regulating the DC bus voltage V_{dc} to the desired level V_{ref} through the axis current [3], [4]. However, in the most commonly used control schemes, additional decoupling terms are added in order to transform the nonlinear initial system (2) into a linear one [4]. In this way, a stable controller can be designed which however suffers by any mismatch on the system parameters (inductance L and angular velocity ω_s), since this leads to a different system from that used for the linear controller design and analyzed for stability. Therefore, undesired phenomena or instability may occur. Additionally, there is not any guarantee that the conventional PI controllers can provide the desired duty-ratio inputs satisfying the constraint of (5). Hence, external limiters are used which, as it is obvious, should be much more conservative than the limits of constraint (5) while require a very difficult gain tuning in order to avoid saturation.

Instead of the conventional scheme, in this paper, the following bounded and parameter-free control scheme is proposed:

$$m_d = z_1 \tag{8}$$

$$m_q = z_2 \tag{9}$$

with dynamics given as:

$$\dot{z} = A_{contr}\left(x, z\right) z \tag{10}$$

where

$$A_{contr} = \begin{bmatrix} 0 & 0 & -k_1 i_d \\ 0 & 0 & -k_2\left(V_{dc} - V_{ref}\right) \\ k_1 i_d & k_2\left(V_{dc} - V_{ref}\right) & -c\left(z_1^2 + z_2^2 + z_3^2 - r_0^2\right) \end{bmatrix} \tag{11}$$

and states $z = \begin{bmatrix} z_1 & z_2 & z_3 \end{bmatrix}^T$ represent the controller dynamics, k_1, k_2 are two non-zero constant gains and c, r_0 are positive constants. Parameters V_{dc} and V_{ref} are the measured and desired DC bus voltage respectively while i_d is the measured $d-$axis current.

If we assume that V_{dc} and i_d are bounded, then the structure of the proposed controller (10) with the term $-c\left(z_1^2 + z_2^2 + z_3^2 - r_0^2\right)$ in (11) implies that the control law acts as an attractive limit cycle for the controller states z_1, z_2, z_3 on the surface of a sphere C_r with center the origin and radius equal to r_0, i.e.

$$C_r = \left\{z_1, z_2, z_3 : z_1^2 + z_2^2 + z_3^2 = r_0^2\right\}.$$

After being attracted on the surface of sphere C_r, states z_1, z_2, z_3 will stay on this surface for all future time, independently from their initial values. To proceed with our analysis in Section IV and without loss of generality, r_0 is selected as:

$$r_0^2 \leq z_1^2(0) + z_2^2(0) + z_3^2(0) \;\; \forall\, z_1(0), z_2(0), z_3(0) \tag{12}$$

arbitrarily selected.

This means that initial conditions are chosen on the surface or outside sphere C_r. The controller states are attracted on sphere C_r (c is the rate of attractiveness) and according to the convergence definition [15], there exists a time instant $T \geq 0$, after which the controller states have 'practically' converged on sphere C_r where they remain thereafter.

As a conclusion of the aforementioned controller analysis, we point out that after an initial time period, the following hold true:

Control states $z_1(t)$, $z_2(t)$ and $z_3(t)$ are bounded each in the range $[-r_0, r_0]$.

Defining in our case $r_0 = 1$ then it is obvious that:

$$z_1^2(t) + z_2^2(t) = r_0^2 - z_3^2(t) = 1 - z_3^2(t)$$

with $0 \leq z_3^2(t) \leq 1$. This means that $z_1(t)$ and $z_2(t)$ lie on a disk D with radius $r_0 = 1$ (Fig. 2).

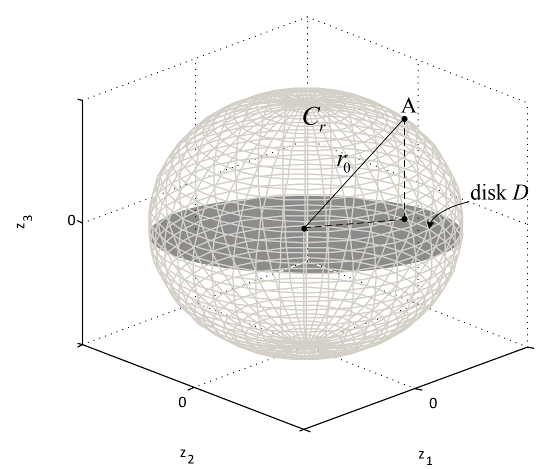

Fig. 2: Sphere C_r in $z_1 - z_2 - z_3$ space and disk D on the $z_1 - z_2$ plane

Thus, the duty-ratio input becomes:

$$m_a = \sqrt{z_1^2(t) + z_2^2(t)} = \sqrt{1 - z_3^2(t)} \tag{13}$$

Hence, it is guaranteed that always: $0 \leq m_a \leq 1$.

The controller acts as a three-state oscillator with angular velocity depended on the difference between the measured and the desired values for V_{dc} and i_d correspondingly. Furthermore, assuming that $i_d \to i_d^{ref} \equiv 0$ and $V_{dc} \to V_{ref}$ constant, there exists an equilibrium point $\tilde{x}^*$ corresponding to some duty-ratio signals $m_d^* = z_1^*$ and $m_q^* = z_2^*$.

Thus, the proposed controller provides a bounded duty-ratio control input, exactly in the range permitted by the linear modulation of the PWM applied on the VSC device. The attractiveness of the controller operation guarantees robustness and controller states' response on the sphere C_r. In order to hold all these controller properties true, it is essential to prove that i_d and V_{dc} are bounded while as $i_d \to 0$ and $V_{dc} \to V_{ref}$, the resulting z_1^* and z_2^* coincide with a stable solution of the entire system at the desired equilibrium. This is a cumbersome task and the proof is presented in the following section.

Applying the proposed controller in (2) and considering as $\tilde{x} = \begin{bmatrix} x^T & z^T \end{bmatrix}^T$ the entire state vector of the closed-loop system, the following dynamic model is obtained:

$$\begin{aligned}
L\dot{i}_d &= -Ri_d + \omega_s L i_q - 2z_1 V_{dc} \\
L\dot{i}_q &= -Ri_q - \omega_s L i_d - 2z_2 V_{dc} + U_m \\
C\dot{V}_{dc} &= 3(z_1 i_d + z_2 i_q) - \frac{V_{dc}}{R_L} \\
\dot{z}_1 &= -k_1 i_d z_3 \\
\dot{z}_2 &= -k_2 (V_{dc} - V_{ref}) z_3 \\
\dot{z}_3 &= k_1 i_d z_1 + k_2 (V_{dc} - V_{ref}) z_2 - c\left(z_1^2 + z_2^2 + z_3^2 - r_0^2\right) z_3
\end{aligned} \tag{14}$$

As one can see from (14), the closed-loop system is an autonomous nonlinear system with the external voltage U_m as an uncontrolled input.

IV. Closed-Loop System Stability Analysis

It is essential to prove that (14) remains stable under the proposed control operation. Now, to proceed with our analysis, the following Proposition is first proven which establishes the ISS property for the closed-loop system.

Proposition 1: Closed-loop system given by (14) is input to state stable (ISS) with respect to the external input U_m.

Proof:

To proceed with the proof, we formulate system (14) as follows:

$$\dot{\tilde{x}} = A(\tilde{x})\tilde{x} + B(\tilde{x})u \tag{15}$$

where $A(\tilde{x})$ has the following block diagonal structure

$$A(\tilde{x}) = \begin{bmatrix} A_{plant} & | & 0_{3\times3} \\ - & & - \\ 0_{3\times3} & | & A_{contr} \end{bmatrix} \text{ and } B = \begin{bmatrix} B_1^T & | & 0_{1\times3} \end{bmatrix}^T \tag{16}$$

with A_{contr} given by (11),

$$A_{plant} = \begin{bmatrix} -\frac{R}{L} & \omega_s & \frac{2z_1}{L} \\ -\omega_s & -\frac{R}{L} & \frac{2z_2}{L} \\ -\frac{3z_1}{C} & -\frac{3z_2}{C} & -\frac{1}{R_L C} \end{bmatrix},\ B_1 = \begin{bmatrix} 0 & \frac{1}{L} & 0 \end{bmatrix}^T$$

and external uncontrolled input $u = U_m$.

Let the unforced closed-loop system (17), i.e. the system without the external input ($u = 0$):

$$\dot{\tilde{x}} = A(\tilde{x})\tilde{x}. \tag{17}$$

Consider the following Lyapunov function for (17):

$$\tilde{V} = \frac{1}{2}Li_d^2 + \frac{1}{2}Li_q^2 + \frac{1}{3}CV_{dc}^2 + \frac{1}{2}z_1^2 + \frac{1}{2}z_2^2 + \frac{1}{2}z_3^2. \tag{18}$$

The time derivative of $\tilde{V}$ is calculated as:

$$\begin{aligned}
\dot{\tilde{V}} &= Li_d\dot{i}_d + Li_q\dot{i}_q + \frac{2}{3}CV_{dc}\dot{V}_{dc} + z_1\dot{z}_1 + z_2\dot{z}_2 + z_3\dot{z}_3 \\
&= -Ri_d^2 + \omega_s L i_q i_d - 2z_1 V_{dc} i_d - Ri_q^2 - \omega_s L i_q i_d - \\
&\quad 2z_2 V_{dc} i_q + 2z_1 V_{dc} i_d + 2z_2 V_{dc} i_q - \frac{2}{3}\frac{V_{dc}^2}{R_L} - k_1 i_d z_1 z_3 + \\
&\quad k_1 i_d z_1 z_3 - k_2 (V_{dc} - V_{ref}) z_2 z_3 + k_2 (V_{dc} - V_{ref}) z_2 z_3 \\
&= -Ri_d^2 - Ri_q^2 - \frac{2}{3}\frac{V_{dc}^2}{R_L} - c\left(z_1^2 + z_2^2 + z_3^2 - r_0^2\right) z_3^2 \leq 0
\end{aligned} \tag{19}$$

Taking into account (12), inequality (19) holds true and since $\dot{\tilde{V}} \leq 0$, it is directly concluded that the unforced closed-loop system (17) is Lyapunov stable, i.e. all its states are bounded.

Furthermore, since (17) represents an autonomous system and the storage function $\tilde{V}(\tilde{x})$ is radially unbounded with non-positive derivative over the whole state space, then in accordance to Global Invariant Theorem 3.5 described in [16], all solutions are uniformly globally asymptotically stable (UGAS) and converge to the largest invariant set M in E. Set E is defined as the set where $\dot{\tilde{V}} = 0$ while as it is evident, invariant set M simply consists of the union of $x = 0$ and z constrained on the limit cycle described by sphere C_r.

Now, taking into account the block diagonal form of matrix $A(\tilde{x})$, one can handle the two subsystems (with state vectors x and z respectively), as separate systems. Since the controller subsystem

$$\dot{z} = A_{contr} z \tag{20}$$

again describes a limit cycle with uniformly bounded states $z(t)$, the other subsystem $\dot{x} = A_{plant}x$, with $A_{plant} = A_{plant}(z(t))$ can be considered as a linear time-varying (LTV) system:

$$\dot{x} = A_{plant}(t)x \tag{21}$$

and therefore UGAS for $x = 0$ is equivalent to globally exponential stability of (21).

At this point, we recall Lemma 4.6 of [15] to prove directly that the system with external input

$$\dot{x} = A_{plant}x + B_1 u \tag{22}$$

is input-to-state stable (ISS).

■

To further clarify the way the proposed controller acts, Fig. 3 is presented. Let $\tilde{x}^*$ be an equilibrium corresponding to some duty-ratio values $m_d^* = z_1^*$ and $m_q^* = z_2^*$. The desired

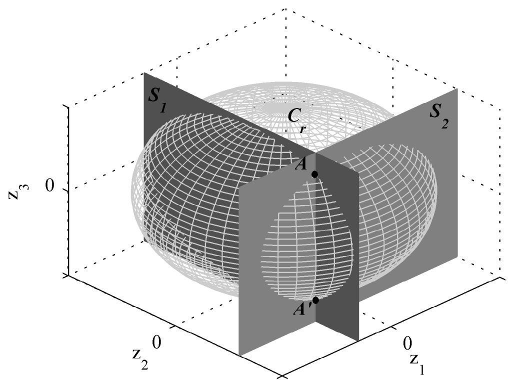

Fig. 3: Controller equilibrium in 3-D state-space

control inputs $m_d^* = z_1^*$ and $m_q^* = z_2^*$ represent two planes S_1 and S_2 respectively in $z_1 - z_2 - z_3$ space. Then, the desired equilibrium $z^* = \left[\begin{array}{ccc} z_1^* & z_2^* & z_3^* \end{array} \right]^T$ of the control states, corresponding to the plant unique equilibrium x^*, are determined from the intersection of sphere C_r with planes S_1 and S_2, which are described by two points A and A' as shown in Fig. 3. Since the controller states z_1, z_2, z_3 can only move on the surface of the sphere and sphere C_r always intersects with planes S_1 and S_2, then the solution of the closed-loop system (14) is bounded in a region where the desired equilibrium $\tilde{x}^*$ exists.

Since, according to the analysis already discussed, system (22) is ISS, we can further apply Theorem 3 as given by the authors in [17] because all the other Assumptions mentioned in [17] are satisfied. Under these conditions, it becomes clear from [17] that there always exists a suitable storage function for system (22) with a constant external input $u = U_m$ satisfying Theorem 3. Hence, finally, the following Proposition is directly proven.

Proposition 2: System (2) with controlled input given by (8)-(12) and external input U_m constant is stable and its solution converges to the desired equilibrium.

It is clear that Proposition 2 integrates Proposition 1 and guarantees a stable and constant steady state system performance.

V. Simulation Results

In order to verify the effectiveness of the proposed nonlinear control scheme, the closed-loop system is simulated. The system parameters are given in Table I.

The controller gains are:

$$k_1 = 10A^{-1},\ k_2 = 0.01V^{-1},\ c = 1000,\ r_0 = 1$$

while the initial values of the controller states are chosen in a manner that (12) holds true as equality i.e.,

$$z_1(0) = 0.2,\ z_2(0) = 0.6,\ z_3(0) = 0.7746.$$

Initially, the system is assumed to be at any initial state with power factor different from unity. At time-instant $t = 0s$ the reference DC bus voltage is set to $V_{ref} = 450V$ while at time instant $t = 5s$ it changes to $V_{ref} = 500V$. Finally, at time instant $t = 10s$, a 20% step increase is performed at the resistance load R_L (from 300Ω to 360Ω), in order to investigate the effectiveness of the proposed controller under sudden load changes.

Figures 4a and 4b illustrate the time responses of the $d-$ and $q-$axis currents respectively. One can observe that the $d-$axis current i_d is always regulated to zero implying unity power factor operation. This can be verified from Figure 5 which describes the time response of the reactive power Q injected from the grid which is indeed regulated to zero independently from the several DC bus voltage reference changes or the load variations.

The DC bus voltage is also suitably regulated to its reference value (Fig. 4c) verifying the effectiveness of the proposed controller even under sudden changes of the load. This also verifies the stability analysis already described in the previous section, proving that the proposed controller acts fully independently from the system parameters and produces a stable solution. Additionally, Figures 4d, 4e and 4f show the time response of the two control inputs m_d and m_q defined by the proposed control law and the third control state variable, respectively.

Finally, Figure 6 shows the responses of the two control inputs on $z_1 - z_2$ plane, verifying the fact that m_d and m_q will always lie on disk D thus achieving linear modulation.

TABLE I: Three-phase AC/DC converter with resistance load

Symbol	Quantity	Value
U_m	maximum supply phase voltage	$200V$
R	resistance of the boosting inductor	0.1Ω
L	inductance of the boosting inductor	$3mH$
C	DC-filter capacitor	$470\mu F$
ω_s	power source angular frequency	$2\pi 50 rad/s$
R_L	load resistance	300Ω

VI. Conclusions

A novel nonlinear controller design for three-phase AC/DC voltage source converters has been developed. The proposed scheme is fully independent from the system parameters and states and suitably regulates the DC bus voltage to the desired level ensuring unity power factor operation. A detailed mathematical analysis shows that the proposed controller design guarantees a stable closed-loop system operation with convergence to the desired equilibrium, while its effectiveness is verified through extended simulation results.

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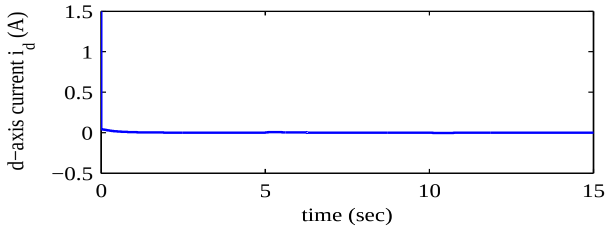
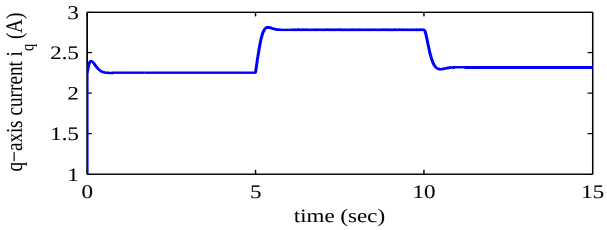
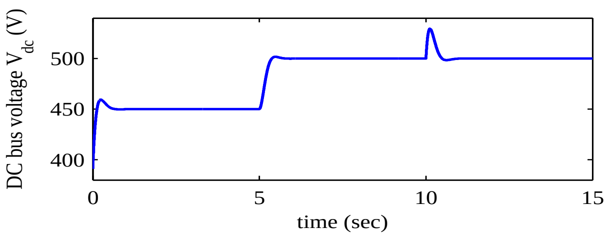


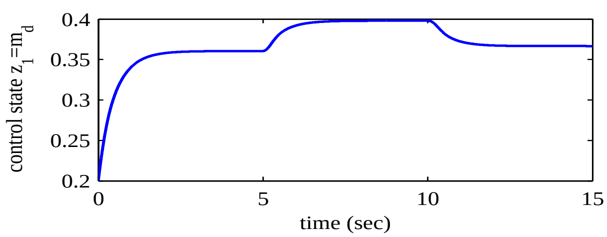
(a) $d-$axis current i_d



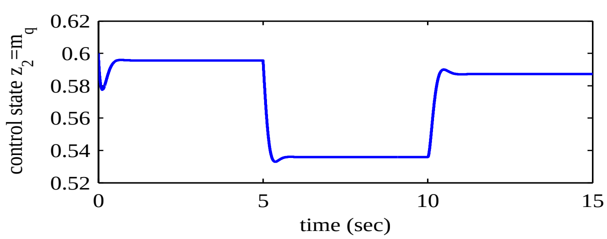
(b) $q-$axis current i_q



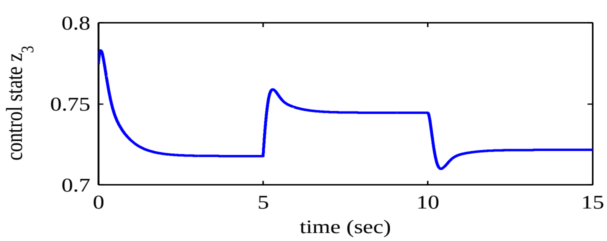
(c) DC bus voltage V_{dc}



(d) $d-$axis switching duty-ratio m_d



(e) $q-$axis switching duty-ratio m_q



(f) controller state z_3

Fig. 4: Time response of the closed-loop system states using the proposed nonlinear controller

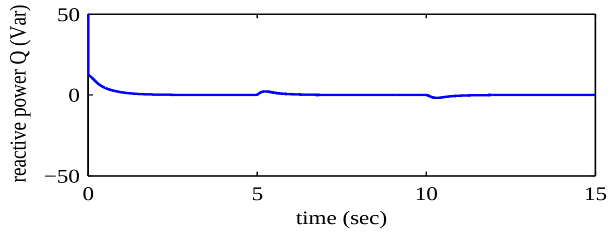


Fig. 5: Reactive power injected from the grid

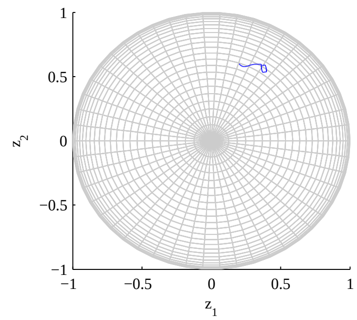


Fig. 6: Convergence of switching duty-ratios on disk D

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