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A controller for the active filter considering load and line impedances

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Abstract— **The paper presents a controller for an active filter to compensate reactive power and current harmonic distortion in a single phase system, i.e., to guarantee a power factor close to unity. The proposed controller considers the negative effects caused by the interaction between load and line impedances, which may lead to instabilities. In particular, the scheme provides a solution in the critical scenario, when the load is composed by a capacitor connected in parallel to a distorted current source. The rationale behind the solution consists in the introduction of a lead compensator with a gain that is adjusted by adaptation, which replaces the conventional proportional term. This modification improves the stability conditions when the load and source impedances are considerable. Special attention is given to the current control loop because it is precisely in this loop where the instability problems arise. Realistic numerical results are provided to illustrate the benefits of the proposed solution.**

I. INTRODUCTION

Many domestic and industrial loads are of nonlinear nature, that is, they are composed of electronic devices with a nonlinear behavior, which introduce harmonic distortion into the electric grid, and thus they are also referred as distorting loads. A direct consequence is a low power factor, deterioration of devices connected to the net, overheating of transformers and interferences to the nearby costumers, among others. The shunt active filters arise as an effective solution for the compensation of reactive power, harmonic distortion and current unbalance due to distorting loads. Different control solutions for active filters have appeared in [1]-[4]. Usually, the models used for the control design do not consider the dynamics of the load nor the line impedances, then it is expected that direct application of an active filter under this controller may produce oscillations and an instability scenario is prone to happen. It has been observed that this issue is particularly worsen when the active filter intends to compensate higher order harmonics [5]. As a consequence of these unmodeled dynamics a resonance effect arises, which is produced by the interaction between a predominantly capacitive load[1], the line impedance and the connection of the shunt active filter. This phenomenon may induce instability and frequent firing of protections, damaging the bank of capacitors and the line isolation. It is clear that the accelerated growth of harmonic distortion sources worsen these problems. Some works have explained the instability mechanism when the aforementioned impedances are considered. In [5]-[8] the authors explain the instability mechanism that arises when a dynamical load is connected to the electric grid. Then, it is shown that the conventional strategies, using the load and line current detection method, may conduce to an unstable operation. In [7] the authors show that current detection methods may become unstable when a capacitor is connected in parallel to the load. In [5] and [8] the authors present a voltage detection method that somehow alleviates this issue. On the other hand, a solution commonly used in practice consists in introducing an inductor in series to the load, with the idea of making the equivalent load predominantly inductive [9]. However, this solution, although effective in most cases, may result expensive.

This paper proposes a controller to guarantee the compensation of harmonic distortion in spite of the presence of impedances in the line and load. In particular, the case studied considers an inductance plus a resistor as the line impedance and a capacitor as the load impedance, which is a critical situation that has received much attention in the last few years [5]-[8]. The idea behind this solution consists in replacing the proportional term used in the current loop of recently proposed controllers [4], [10]-[12] by a lead compensator. It is analytically shown that, after this modification, the stability conditions are improved. Then, it is proposed to use an adaptation to estimate the associated gain of the lead compensator to enhance the robustness against parameters variation. Finally, to show the benefits of the proposed controller, numerical results using PSCAD 4.0 have been included using a single-phase full-bridge shunt active filter.

II. SYSTEM DESCRIPTION

A. Shunt Active Filter

Figure 1 shows the topology of the shunt active filter, which is designed to compensate reactive power and harmonic distortion in a distribution system. This topology is composed of a voltage source inverter (VSI) connected to the line via an inductor L. A distorting nonlinear load is connected to the voltage source v_S producing a distorting current i_{LT} which is considered as a disturbance. The voltage in the point of common connection is represented by v'_S. Traditionally, for control design purposes, it is assumed that the load current is static, in the sense that, it can be considered as a simple distorted current source without any associated impedance. It is also assumed that the impedances

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[1]It is common in practice to use capacitors or passive filters to correct the power factor in industry.

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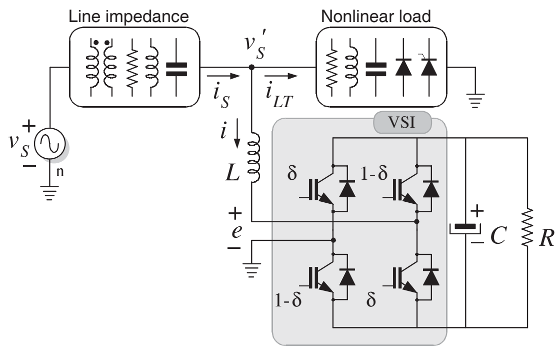

Fig. 1. Single-phase full-bridge shunt active filter connected to a distribution system.

associated to voltage source and transmission line are both negligible. A capacitor C is connected on the dc-side of the VSI.

The system dynamics of the scheme shown in Fig. 1, where the impedances of line and load have been omitted, are described by

$$L\frac{di}{dt} = v_S - e \tag{1}$$

$$\frac{C}{2}\frac{d(v_C^2)}{dt} = ei - \frac{v_C^2}{R} \tag{2}$$

$$i_S = i + i_{LT} \tag{3}$$

where i_S and i represent line and injected currents, respectively, v_C is the capacitor voltage on the dc-side; $e \triangleq uv_C$ is the injected voltage representing the actual control input, with $u \in [-1, 1]$ the duty ratio of a PWM generated switching sequence $\delta \in \{0, 1\}$ of a relatively high frequency. This is referred in the power electronics literature as the average model [13]. For security purposes, it is usual, in normal active filters, to connect a large resistor to discharge the capacitor whenever the system is turned off. This resistor, together with switching and other losses are lumped in the model as an unknown constant resistive element R.

To facilitate the control design, it is common in practice to assume that the current dynamics (1) responds much faster than the dynamics involving the capacitor (2), and thus, they can be decoupled from one another invoking time scale separation. This assumption allows to split the control design in two loops, namely, current inner loop and voltage outer loop. Based on this idea, and following the *energy shaping plus damping injection* technique of the passivity-based control approach, a solution was presented in [4], which is briefly revisited below as it will serve as the basis for the proposed controller. The interested reader is referred to [4] for additional details.

Current inner loop. The control objective of this control loop consists in injecting the necessary current so that, the currents i_S is forced to be proportional to the source line voltage v_S. The proportionality constant, denoted as scalar g, represents an equivalent conductance observed at the point of common connection. In other words, the objective consists in driving $\tilde{i}_S = i_S - i_S^*$ to zero, where i_S^* is the current reference given by

$$i_S^* = gv_S \tag{4}$$

where g is determined by the voltage regulation loop, as will be shown later.

A solution for this tracking objective consists in the construction of a control signal u that cancels v_S, adds a damping term to reinforce the stability, and introduces a bank of resonant filters tuned at the harmonics under compensation. The expression of such a current tracking loop is given by

$$e = v_S + k_1\tilde{i}_S + \sum_{k\in\mathcal{H}} \frac{2\gamma_k s}{s^2 + k^2\omega_0^2}\tilde{i}_S \tag{5}$$

where s denotes the Laplace complex variable, k_1 and γ_k, with $k \in \mathcal{H}$, are positive control design parameters, and $\mathcal{H} = \{1, 3, 5, ...\}$ represents the set of harmonics indexes considered for compensation, in this case the odd harmonics. It is well known that single phase distorting loads produce mainly odd harmonics of the fundamental frequency. Notice that the resonant filters are tuned at the k-th harmonic component, i.e., $k\omega_0$. In [4] the authors show that controller (5), in closed loop with system (1)-(3), guarantees that $\tilde{i}_S$ goes to zero asympotically, as far as there are enough resonant filters to compensate each harmonic component of the disturbances, and under the assumption of nonlinear static disturbances.

The infinite gain, provided by the resonant filters in the control scheme (5), represents a potential source of instability. To alleviate this issue, in this paper, and as it is common practice, the resonant filters are replaced by bandpass filters of the form

$$\frac{(k\omega_0 A_k/Q_k)s}{s^2 + (k\omega_0/Q_k)s + k^2\omega_0^2}\ , \quad k \in \mathcal{H} \tag{6}$$

where $A_k > 0$ and $Q_k > 0$ are the desired gain and the quality factor of the k-th bandpass filters, respectively. In this way, the resonance peaks have a limited gain of value A_k.

Voltage outer loop. To accomplish the regulation objective, the capacitor voltage v_C should be maintained at a constant voltage level V_d. This regulation objective is solved by suitable designing the scalar g, which, as seen in equation (4), is used to construct the reference i_S^*. This control loop is formed by a proportional term of limited bandwidth plus an integral term of the form

$$G = \frac{k_i}{s}\tilde{z} + \frac{k_p}{\tau_1 s + 1}\tilde{z} \tag{7}$$

$$g = \frac{G}{v_{S,RMS}^2} \tag{8}$$

where $\tilde{z} \triangleq \frac{V_d^2}{2} - \frac{v_C^2}{2}$; k_p and k_i are the proportional and integral gains, respectively, and $v_{S,RMS}$ is the root mean square value of the source voltage. The scale factor $1/v_{S,RMS}^2$ in (8) is introduced to avoid numerical errors in the computation of g. In fact, G represents an approximate of the total power delivered by the power supply to the system composed by active filter and load.

B. Stability analysis considering line and load dynamics

In contrast to the previous case, the output impedance of the power supply and the associated impedance of the line, as well as the effect of the associated passive elements on the load side, are considered in this case. Figure 2 shows the equivalent circuit of the shunt active filter considering impedances in both the line and the load. In this system, an impedance $Z_S(s)$ is connected in series with the internal voltage v_S. The load is represented by a Norton equivalent circuit, where the static current generator i_L represents the purely distorting load and the impedance $Z_L(s)$ models its associated passive components. The active filter is composed of a voltage source e connected in parallel to the overall load by means of an inductor L.

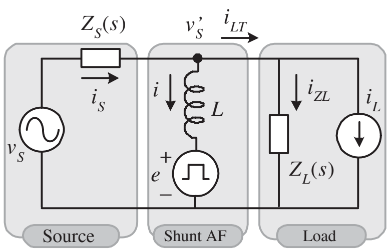

Fig. 2. Equivalent circuit of a shunt active filter connected to a more realistic distribution system considering load and line impedances.

From the equivalent circuit of Fig. 2, the source current can be computed as

$$i_S = \frac{Z_L(s)}{Z_S(s)+Z_L(s)}\left(\frac{v_S}{Z_L(s)}+i+i_L\right) \tag{9}$$

Notice that, in the ideal case where the line is an ideal conductor, the source impedance is negligible, and the load current is static, then $Z_S(s)=0$ and $Z_L(s)=\infty$, out of which $i_S=i_L+i$.

The new definition of the source current (9) in terms of impedances $Z_S(s)$ and $Z_L(s)$ defines the new current dynamics. Figure 3 shows the closed loop diagram of the new definition of the source current dynamics (9) with the controller (5), and after rearranging the terms. Notice that, in contrast with the ideal case, the gain loop includes an extra transfer function denoted by

$$G_Z(s) = \frac{Z_L(s)\,(1+gZ_S(s))}{Z_S(s)+Z_L(s)} \tag{10}$$

The characteristic polynomial must include now the poles associated to impedances $Z_S(s)$ and $Z_L(s)$. Therefore, some poles risk to be unstable if the original controller scheme is preserved, as it will become clear later.

A critical case that has been studied lately consists in taking the line impedance as an inductor in series with a resistor, and the load impedance as a capacitor [7], that is, they have the frequency domain representation $Z_S(s)=L_Ss+R_S$ and $Z_L(s)=1/(C_Ls)$. In this case, instability problems have been experimentally observed even for arbitrarily small C_L.

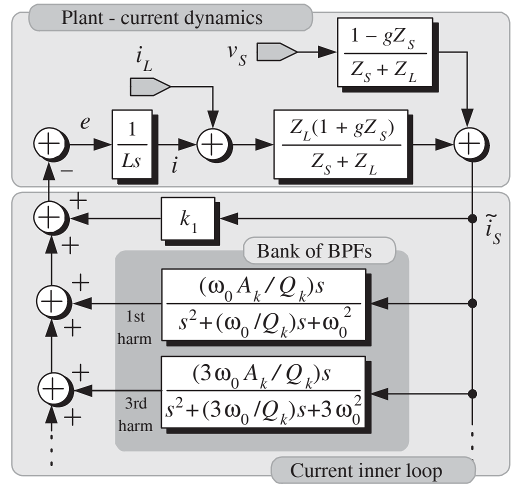

Fig. 3. Block diagram of the current loop considering load and line impedances.

The transfer function $G_Z(s)$, for this particular case, is given by

$$G_Z(s) = \frac{gL_Ss+gR_S+1}{L_SC_Ls^2+C_LR_Ss+1} \tag{11}$$

which adds, in the gain loop, a couple of complex-conjugate poles and a zero located at $\lambda_Z=-(gR_S+1)/(gL_S)$. Notice that, for $1 \gg gR_S$, as normal in practice, the zero location is approximately in $-1/(gL_S)$.

To show in a simpler form the influence of the impedances consideration, consider first the unperturbed system with a proportional damping term only as the controller. This yields the following characteristic polynomial

$$1+\frac{gL_Ss+gR_S+1}{Ls(L_SC_Ls^2+C_LR_Ss+1)}k_1=0 \tag{12}$$

In this simplified case the Routh-Hurwitz criterion provides the following necessary and sufficient stability condition $0<k_1<R_SL/L_S$, which depends directly on the unknown line parameters R_S and L_S. Notice that, for practical values of L, R_S and L_S, the admissible value of k_1 can be considerably limited.

III. Proposed controller

To overcome the instability problem induced by the presence of line and load impedances, it is necessary to modify the inner loop (5) of the controller. A first proposal consists in multiply the damping term by a lead compensator as shown in (13)-(14). This modified current loop controller includes also a term that cancels v'_S and a bank of bandpass filters tuned at the harmonics under compensation as before.

$$\begin{aligned} e &= v'_S + k_1F(s)\tilde{i}_S \\ &+ \sum_{k\in\mathcal{H}} \frac{(k\omega_0A_k/Q_k)s}{s^2+(k\omega_0/Q_k)s+k^2\omega_0^2}\tilde{i}_S \end{aligned} \tag{13}$$

$$F(s) = \frac{\tau_zs+1}{\tau_ps+1} \tag{14}$$

where the lead compensator $F(s)$ introduces a zero in $-1/\tau_z$ and a pole in $-1/\tau_p$. Usually $\tau_p \ll \tau_z$ to place the pole far to the left of the zero, hence the pole would have a small effect on the resonant dominant part of the root locus.

It has been observed that this simple modification allows, not only a wider range of values for k_1, but also the compensation of higher harmonics. However, it can be shown that the admissible range of k_1 strongly depends on system and bandpass filters parameters. In other words, if these parameters are known, then a suitable gain, say k_1^*, can be fixed to preserve stability of the equilibrium point.

As parameters are usually unknown and possibly time varying, the selection of k_1^* could be incorrect, which may entail a poor performance and instability. To alleviate these issues, the following adaptive estimator for the parameter k_1^* is proposed

$$\phi = F(s)\tilde{i}_S \quad (15)$$

$$\dot{\hat{k}}_1 = \mu\phi\tilde{i}_S \quad (16)$$

where $\hat{k}_1$ is an estimate of k_1^*, μ is a positive design parameter, ϕ is the so-called regressor, and $F(s)$ is the transfer function of the lead compensator.

The structure of the previous adaptive law, is justified as follows. Let us define the estimation error $\tilde{k}_1 = \hat{k}_1 - k_1^*$. In Fig. 3, add the filter $F(s)$ and replace the constant gain by its estimate $\hat{k}_1$. Then, decompose the estimate into the ideal (unknown) gain k_1^* and the parameter error $\tilde{k}_1$ and "pull-out" the later from the diagram. Figure 4 shows the block diagram considering all these changes, where, for simplicity, the resonant filters have been removed. Computing the transfer function from the signal $\tilde{k}_1\phi$ to $\tilde{i}_S$ (neglecting the additional inputs due to v_S and i_L) yields

$$\tilde{i}_S = -H(s)[\tilde{k}_1\phi] \quad , \qquad H(s) = \frac{G_Z(s)}{Ls + k_1^* F(s) G_Z(s)}.$$

Replacing the expression above in (16), and noting that $\dot{\tilde{k}}_1 = \dot{\hat{k}}_1$, yields

$$\dot{\tilde{k}}_1 = -\mu\phi[H(s)(\tilde{k}_1\phi)]. \quad (17)$$

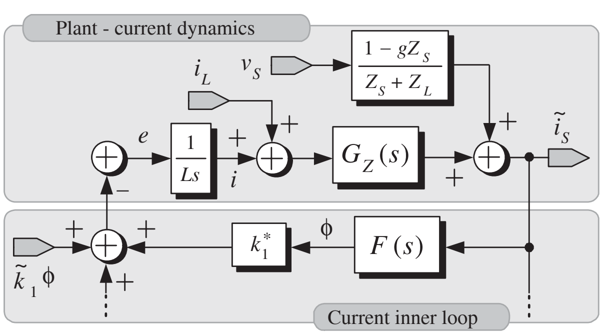

Fig. 4. Block diagram of the modified current loop.

This differential equation is ubiquitous in adaptive control and has been studied extensively [16], [17]. To get an idea of the nature of the equation consider an ideal case where $H(s)$ is a simple positive gain, say h, (this is the case when ϕ is "slow" with respect to the bandpass filtering of $H(s)$ and μ is small, so $\hat{k}_1$ changes also slowly). Then, the equation reduces to the non–autonomous linear equation

$$\dot{\tilde{k}}_1 = -\mu h \phi^2(t)\tilde{k}_1,$$

that can be explicitly solved as

$$\tilde{k}_1(t) = \exp^{-\mu h \int_0^t \phi^2(\tau)d\tau}\, \tilde{k}_1(0).$$

It is clear that the parameter estimation error satisfies $|\tilde{k}_1(t)| \leq |\tilde{k}_1(0)|$ for all $t \geq 0$, showing that the search tends to reduce the estimation error as desired. Under some conditions on $\phi(t)$—namely, that it is not square integrable—it can be also shown that $\tilde{k}_1(t) \to 0$.

The approximative explanation given above can be formalized using averaging analysis and the notion of average positive realness [18] of the transfer function leading to the following proposition, whose proof is a direct application of Theorem 3.8 of [18].

Proposition III.1 Consider the dynamical system described by (17). Assume

(i) $H(s)$ is a stable transfer function.
(ii) $\phi(t)$ is a nonzero T–periodic integrable function of t.
(iii) The average positive real condition

$$\sum_{i=-\infty}^{i=\infty} \mathcal{R}_e\{H(j\omega_i)\}|c_i|^2 > 0,$$

is satisfied, where c_i are the coefficients of the exponential Fourier series of ϕ and $\omega_i = \frac{2\pi i}{T}$.

Then there exists $\mu^* > 0$ such that, for all $\mu \in (0, \mu^*)$, $\tilde{k}_1(t)$ converges exponentially to zero. □

In words, the proposition states that, if

- μ is sufficiently small, i.e., slow adaptation, and
- the energy of the spectrum of ϕ is concentrated on the frequency range when $\mathcal{R}_e H(j\omega) \geq 0$, that is, in the range when the phase shift is smaller that $\frac{\pi}{2}$,

then the overall system will have some suitable stability properties. Notice that we have neglected in this analysis the presence of the signals v_S and i_L and concentrated our attention on the adaptation loop.

A block diagram of the overall proposed controller is shown in Fig. 5.

IV. Numerical results

The proposed controller shown in Fig. 5 that uses the estimator of k_1 (16) has been tested in the simulation program PSCAD 4.0 with the following parameters: $L = 5$ mH, $C = 2200\,\mu$F, $V_d = 380$ V_{DC}, and the switching frequency $f_{sw} = 20$ kHz. A voltage source of $127\,V_{RMS}$ at $f = 60$ Hz is considered with source impedance parameters $L_S = 2.65$ mH and $R_S = 2.59\,\Omega$ (same as the ones above used). The load is composed of a single-phase diode rectifier with a capacitor of $10\,\mu$F with an associated load

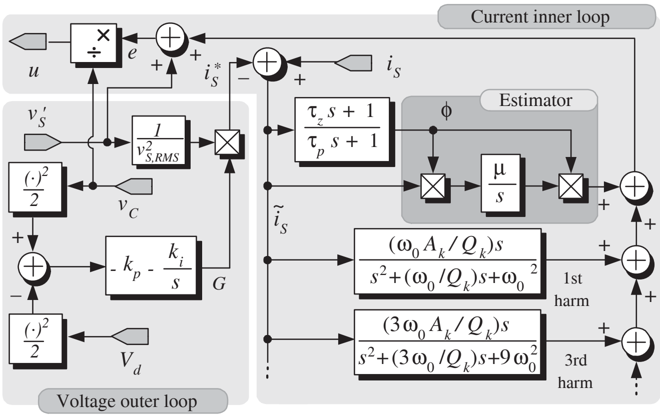

Fig. 5. Block diagram of the overall proposed controller.

resistance on the dc-side. The critical case, connecting a capacitor of $82\,\mu$F in parallel to a distorted current load, has been considered. The design parameters of the proposed controller are selected as follows: $A_k = 50$ and $Q_k = 40$ $\forall\, k = \{1, 3, 5, ..., 17\}$, $k_p = 0.4$, $k_i = 0.06$, $\tau_z = 0.004$, $\tau_p = 0.0001$ and $\mu = 0.001$.

Figure 6 shows the responses of the system under the basic controller (5), after the connection of a capacitor C_L on the load side. This figure shows (from top to bottom) the line voltage v_S', the compensated line current i_S, the load current i_L and the current i injected by the active filter. Notice that, before the connection of capacitor C_L, the responses have reached the desired equilibrium with a given k_1, in particular, the compensated current i_S is almost sinusoidal and in phase with the line voltage v_S', thus guaranteeing a power factor close to unity. After the connection of capacitor C_L, the selected k_1 falls outside the admissible region, i. e., resonant dominant poles get a positive real part. As a consequence, a resonance effect is produced causing oscillations that considerably distort all signals. In contrast, Fig. 7, displaying the same signals as before, shows that, after a relatively small transient due to the connection of capacitor C_L, the compensated source current i_S recovers its almost sinusoidal shape with the same phase as the line voltage v_S'.

Figure 8 shows the transient responses of (top) the equivalent conductance g and (bottom) the capacitor voltage v_C when capacitor C_L is connected on the load side and under the proposed controller. It is observed that, after a relatively small transient, the capacitor voltage is maintained close to the reference $V_d = 280$ V_{DC}, while the equivalent conductance g converges towards a certain positive constant value.

Figure 9 shows the transient responses of (top) the estimate $\hat{k}_1$ and (bottom) the regressor ϕ after the connection of the capacitor C_L on the load side. This figure shows that, after a relatively short transient, the estimate $\hat{k}_1$ asymptotically converges towards a constant, while the regressor ϕ reaches practically zero.

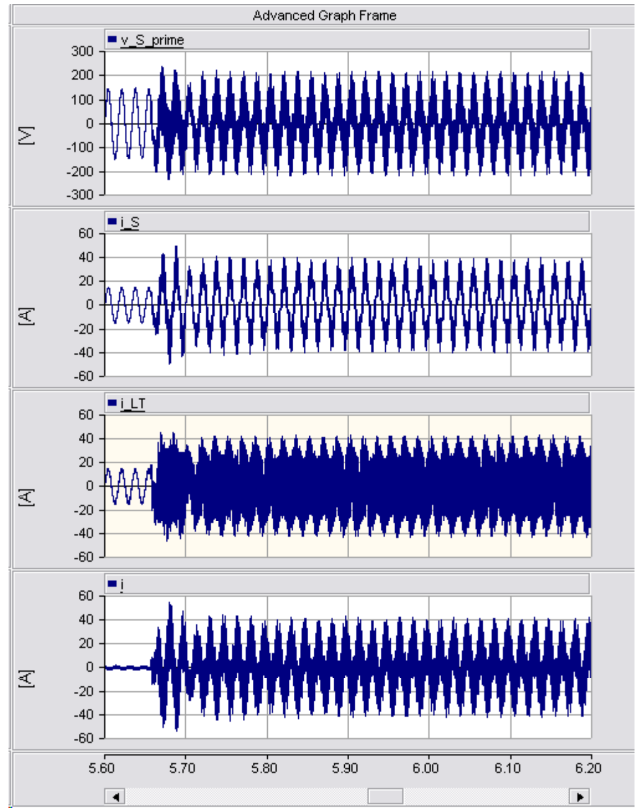

Fig. 6. Transient responses, under the original compensator (5) and (7), during the connection of capacitor C_L on the load side of: **(from top to bottom)** line voltage v_S', compensated line current i_S, load current i_L and active filter current i.

V. Concluding Remarks

The paper presented a modification to a typical controller for an active filter to guarantee compensation of reactive power and current harmonic distortion despite of the presence of a critical dynamical and distorting load. The critical case, that has been studied here, consisted in the connection of a capacitor in parallel to the original load. This represents a compensation mechanism commonly used in practice for power factor correction, which explains the interest given here and in previous works. A model of the overall system has been presented considering the load and source impedances. It was shown that the interaction between the line and load impedances, in this critical case, produces a resonance effect, which induces instability. The proposed solution included an adaptive implementation to enhance the robustness against system parameters uncertainties. Numerical results have been provided to illustrate the benefits of this solution.

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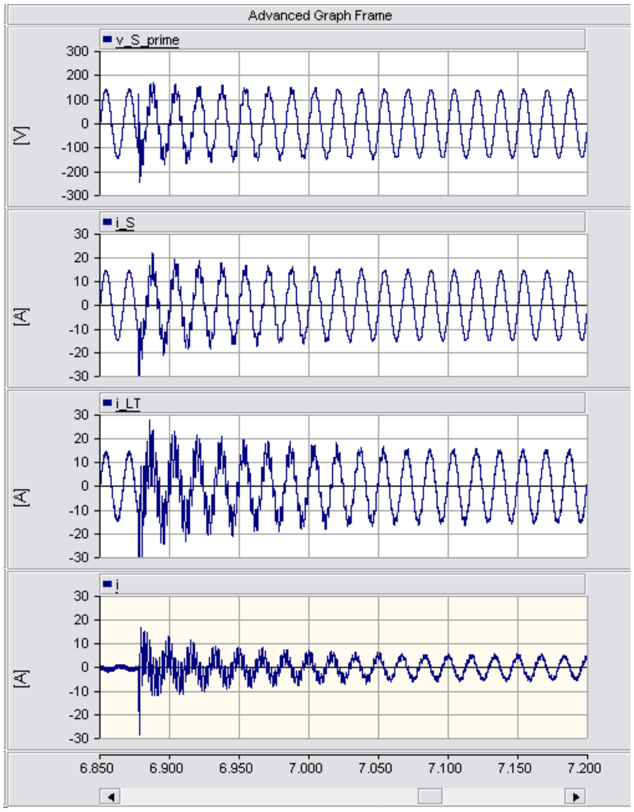

Fig. 7. Transient responses, under the proposed compensator of Fig. 5, during the connection of capacitor C_L on the load side of: **(from top to bottom)** line voltage v_S', compensated line current i_S, load current i_L and active filter current i.

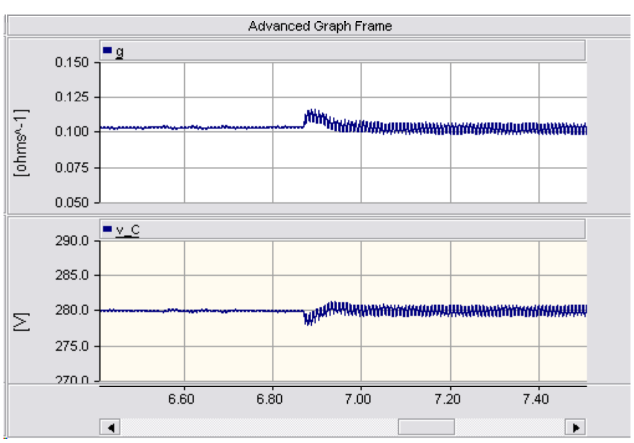

Fig. 8. Transient responses under the proposed controller when capacitor C_L is connected on the load side: **(top)** the equivalent conductance g and **(bottom)** the capacitor voltage v_C.

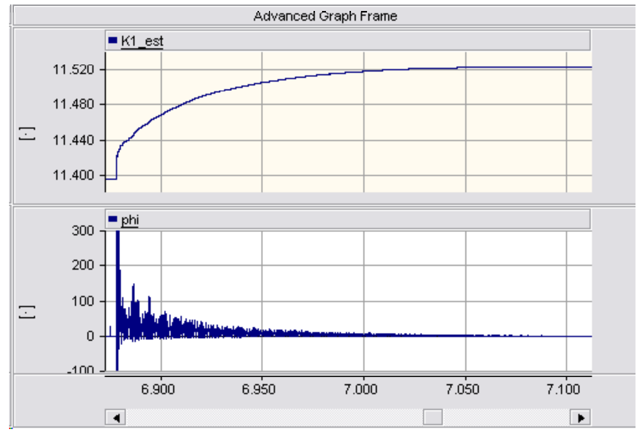

Fig. 9. Transient responses under the proposed controller when capacitor C_L is connected on the load side: **(top)** the estimated damping term $\hat{k}_1$ and **(bottom)** the regressor ϕ.

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