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An adaptive Linear Quadratic Regulator for Three-phase UPS system Powering Nonlinear Loads

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Abstract

This paper presents a control strategy applied to three-phase uninterruptible power supplies with a low switching frequency (1500 Hz). In The controller design, the gains are determined by minimizing a cost function, which reduces the tracking error and smoothes the control signal. A recursive least square estimator identifies the parameters model at different load conditions. Then the linear quadratic controller gains are adapted periodically. The output voltage is the only state variable measured. The other state variables are obtained by estimation process. Simulation results show that the proposed control strategy offers good performances for either linear and non-linear loads with low total harmonic distortions (THD) even at low frequencies making it very useful for high power applications.

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Keywords: Three-phase UPS. Total harmonic distortion. Linear quadratic regulator. Non linear loads.

Nomenclature

UPS	uninterruptible power supplies
THD	total harmonic distortions
LQR	linear quadratic regulator
RLS	recursive least square estimator
PWM	pulse width modulation

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1. Introduction

THD of the output voltage and dynamic response of uninterruptible power supplies (UPS) system are among the most important performance features of the UPS .They depend mostly on the control strategy applied to the UPS inverter. In this sense, the inverter's control strategy immediately follows the UPS topology as the second most important factor, which determines the overall performance of the UPS. The simplest method to deal with a high output voltage THD is increase the switching frequency and to use appropriate LC filter will be smaller. However, with increasing switching frequency, losses increase accordingly, which is why for high-power rating applications, the switching frequency are limited usually up to 2 KHz [1].

With a decreasing cost of microcontrollers and digital signal processors (DSP), the use of digital control technique in power converter has increased. However, high power converters are usually operated at low switching frequencies in order to reduce switching losses. Therefore, advanced control strategies are required to overcome these difficulties [2], [3], [4], [5]. To design the closed loop control, the model of the system has an important role in the conception of the controller. Some linear models for single phase PWM inverter system have been reported in literature [2],[3]. The output voltage and its derivative, that is proportional to the capacitor current, can be used as the state variables, as well as the output voltage and the inductor current. However, modeling errors and unmodelled dynamics are quite common. They may be a result of simplifications on the model, which can degrade the performance of the system [4].

Many discrete time controllers have been reported in the literature to control a three phase inverter for use in UPS, such as predictive control [6], [7], repetitive control [8], [9], optimal state feedback [10] and selective harmonic compensation [11], [12]. Even if most of these schemes offered high performance feedback control results, they still relay on high switching frequencies and involve considerable computational over heads.

In this paper a three phases UPS with a low switching frequency is proposed in order to minimize switching losses and improve system efficiency. An adaptive quadratic regulator for single phase UPS application is proposed. The regulator is a useful tool in modern optimal control design. For the proposed controller, a recursive least square estimator identifies the plant parameters which are used to compute the regulator gains periodically. The quadratic cost function parameter is chosen in order to reduce the energy of the control signal. Only the output voltage can be measured and the inductor current is not measurable. As a result, an observer is used to estimate the inductor current. Using a suitable filter the effect of disturbances on the response of the system will be decreased. The simulations were carried out using MATLAB Simulink.

This paper is organized as follows: After the introduction, the global model of the plant is described in section(2), theoretical analysis of the controller, RLS estimator and Kalman filter descriptions are reported in sections (3), (4) and (5) respectively. Simulation results and discussion are presented in section (6) followed by the conclusion in the final section.

2. Plant description and modeling

The Three-phase PWM inverter is shown in Fig.1, the LC filter and the resistive load R are considered to be the plant of the system. The inverter is controlled by the unipolar PWM. The power switches are turned on and off at the carrier frequency.

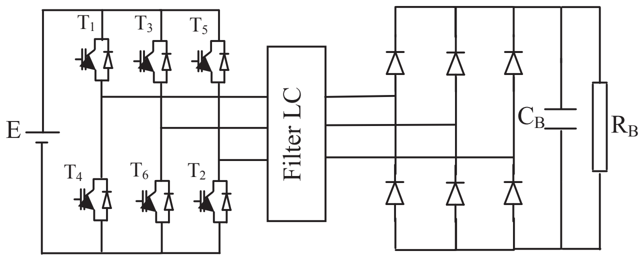

Fig. 1. Inverter, filter and load.

The plant can be modelled by the state space variable V_C and i_L .Where the first simple phase is shown in Fig. 2,

$$\begin{bmatrix} \dot{v}_c \\ i_L \end{bmatrix} = \begin{bmatrix} \frac{-1}{RC} & \frac{1}{C} \\ \frac{-1}{L} & 0 \end{bmatrix} \begin{bmatrix} v_c \\ i_L \end{bmatrix} + \begin{bmatrix} 0 \\ \frac{1}{L} \end{bmatrix} u, y = \begin{bmatrix} 1 & 0 \end{bmatrix} \begin{bmatrix} v_c \\ i_L \end{bmatrix} \tag{1}$$

Or

$$\dot{x} = Ax + Bu, y = Cx \tag{2}$$

Then, a discrete time model of the plant obtained by the forward method and sample time T_S is given by:

$$x(k+1) = A_d x(k) + B_d u(k), y(k) = C_d x(k) \tag{3}$$

Where

$$x(k) = \begin{bmatrix} v_c(k) & \hat{i}_L(k) \end{bmatrix}^T, A_d = I + T_S A, B_d = TsB$$

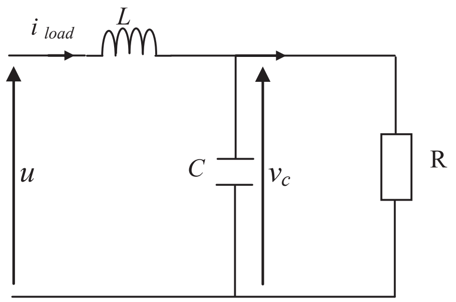

Fig. 2. Inverter, filter and load for the first simple phase.

3. Linear quadratic regulator

The adaptive linear quadratic regulator controller has the objective of tracking the discrete sinusoidal r(k) reference in each sample instant.

The system output y(k) is the capacitor voltage in the discrete form $v_c(k)$. The state variables used in the (LQR) are the measured output voltage $v_c(k)$, the estimated inductor current $\hat{i}_L(k)$, the integrated tracking error v(k); all with a feedback action and the discrete reference r(k) and its derivative $\acute{r}(k)$ with a feed forward action. Each state variable has weighting K_i tuned according to θ(k), which contains the plant parameters identified by the RLS estimator. The control system shown in Fig.3 is therefore proposed.

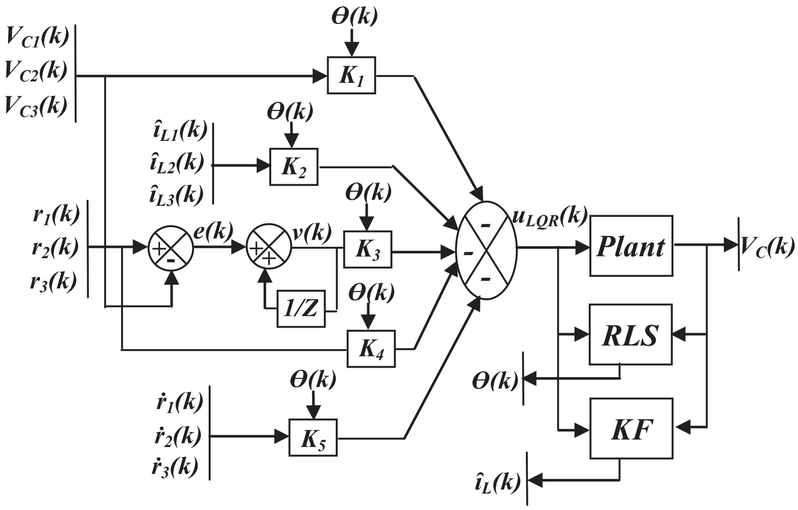

Fig. 3. Block diagram of the control system.

Then, in the proposed system, the state vector z(k) is defined as:

$$z(k) = \begin{bmatrix} v_c(k) & \hat{i}_L(k) & v(k) & r(k) & \dot{r}(k) \end{bmatrix}^T \tag{4}$$

And the LQR control signal is given by

$$u_{LQR}(k) = -Kz(k) \tag{5}$$

To design the optimal gains $K_1, K_2, ..., K_5$, the system must be represented in the form:

$$z(k+1) = Gz(k) + Hu_{LQR}(k) \tag{6}$$

Where each state variable is calculated by a difference equation. The two first variables of vector z(k) are obtained by (3). The signal v(k) is:

$$v(k+1) = e(k+1) + v(k) \tag{7}$$

Where the error is given by:

$$e(k) = r(k) - y(k) \tag{8}$$

From (3), (7) and (8) results the difference equation for

$$v(k+1) = v(k) + r(k) + T_S\,\dot{r}(k) - C_d A_d x(k) - C_d B_d u_{LQR}(k) \tag{9}$$

The continuous time reference variables are:

$$\begin{bmatrix} \dot{r}(t) \\ \ddot{r}(t) \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ -\omega^2 & 0 \end{bmatrix} \begin{bmatrix} r(t) \\ \dot{r}(t) \end{bmatrix}, \dot{r} = Rr \tag{10}$$

This system generates a sinusoidal reference when initiated with initial values:

$$r(0) = 0, \dot{r}(0) = \omega V_p \tag{11}$$

Where V_P is the sine wave amplitude and w is the angular frequency.

In the discrete form, using a sample period T_S , the subsystem (10) is given by:

$$n(k+1) = R_d n(k) \tag{12}$$

Where

$$n(k) = \begin{bmatrix} r(k) & \dot{r}(k) \end{bmatrix} \tag{13}$$

$$R_d = I + T_S R \tag{14}$$

Then, using the state equations (3), (9) and (12), the closed loop system representation becomes:

$$\begin{bmatrix} x(k+1) \\ v(k+1) \\ n(k+1) \end{bmatrix} = \begin{bmatrix} A_d & 0 & 0 \\ -C_d A_d & 1 & C_d R_d \\ 0 & 0 & R_d \end{bmatrix} \begin{bmatrix} x(k) \\ v(k) \\ n(k) \end{bmatrix} + \begin{bmatrix} B_d \\ -C_d B_d \\ 0 \end{bmatrix} u_{LQR}(k)$$

$$y(k) = \begin{bmatrix} C_d & 0 & 0 \end{bmatrix} \begin{bmatrix} x(k) & v(k) & n(k) \end{bmatrix}^T \tag{15}$$

The optimal gains of the control law (5) are those minimizing the following cost function:

$$J = \frac{1}{2} \sum_{k=0}^{\infty} \left\{ z^T(k) Q z(k) + u^T(k) R_u u(k) \right\} \tag{16}$$

Where Q and R_u are chosen as positive definite matrixes that set the weighting of states and the control signal respectively.

The K gains can be obtained by the evaluating the Riccati equations [13] presented as follows:

$$S(k) = G^T S(k+1) G + Q - \left[H^T S(k+1) G\right]^T \left[R_u + H^T S(k+1) H\right]^{-1} \left[H^T S(k+1) G\right] \tag{17}$$

$$K(k) = R_u^{-1} H^T \left(G^T\right)^{-1} \left(S(k) - Q\right) \tag{18}$$

A good flexibility in the design of the controller is provided by the selection of Q and R_u matrixes.

4. Recursive least square estimator

To estimate the plant parameters when the load conditions are variable, a RLS algorithm is used [14]. The discrete plant model with a zero order hold is given by:

$$\frac{y(z)}{u(z)} = \frac{\theta_3}{z^2 + \theta_1 z + \theta_2} \tag{19}$$

The difference equation of the estimated output is:

$$y(k) = -\theta_1 y(k-1) - \theta_2 y(k-2) + \theta_3 u(k-2) \tag{20}$$

Or

$$\hat{y}(k) = \theta^T(k) \Psi(k-1) \tag{21}$$

Where

$$\theta(k) = [\theta_1 \quad \theta_2 \quad \theta_3] \tag{22}$$

And

$$\Psi(k) = [-y(k-1) \quad -y(k-2) \quad u(k-2)] \tag{23}$$

The RLS gains are calculated using:

$$L(k) = \frac{p(k-1)\Psi(k)}{1+\Psi^T k)p(k-1)\Psi(k)},$$

(24)

The RLS covariance matrix is given by:

$$p(k) = p(k-1) - \frac{p(k-1)\Psi(k)\Psi^T(k)p(k-1)}{1+\Psi^T(k)p(k-1)\Psi(k)} \tag{25}$$

and the plant parameters θ are recursively obtained by:

$$\hat{\theta}(k) = \hat{\theta}(k-1) + L(k)\left[y(k) - \Psi^T\hat{\theta}(k-1)\right] \tag{26}$$

Where:

$$\hat{A}_d = \begin{bmatrix} 0 & -\hat{\theta}_2 \\ 1 & -\hat{\theta}_1 \end{bmatrix}, \hat{B}_d = \begin{bmatrix} \hat{\theta}_3 \\ 0 \end{bmatrix}, C_d = [0 \quad 1] \tag{27}$$

Then, it is possible to identify the plant parameters to a range of different loads through the substitution of matrixes (27) into system (15) and proceed there often with the LQR gains design in real time.

5. Kalman filter

Since only the output voltage is measured, a Kalman filter [13], [15] is used to estimate the inductor current state.

$$\begin{aligned} x(k+1) &= A_d x(k) + B_d u(k) + w(k) \\ y(k) &= C_d x(k) + v(k) \end{aligned} \tag{28}$$

The random variables w(k) and v(k) represent the process and measurement noise respectively. They are assumed to be independent of each other and with normal probability distributions such that:

$$\begin{aligned} E[w(k)^T \quad w(k)] &= R_W \rangle 0 \\ E[v(k)^T \quad v(k)] &= R_V \rangle 0 \\ E[w(k)^T \quad v(k)] &= 0 \end{aligned} \tag{29}$$

In practice, the process noise covariance and measurement noise covariance matrices might change with each time step or measurement. However, here, it is assumed that they are presented below [15].

The Kalman gains are given by:

$$K_G(k) = \left(M(k)C_d^{\ T}\right)\left(C_d M(k) C_d^{\ T} + R_v\right)^{-1} \tag{30}$$

and the estimated variable, the inductor current, is:

$$i_L = \hat{x}_2(k) = \begin{bmatrix}0 & 1\end{bmatrix}\hat{x}(k) \tag{31}$$

The following recursive equations are used:

$$P_K(k) = M(k) - K_G(k) C_d M(k), \tag{32}$$

and

$$M(k) = \left(A_d P_K(k) A_d^{\ T}\right) + \left(B_d R_W B_d^{\ T}\right), \tag{33}$$

After each time and measurement update pair, the process is repeated with the previous posterior estimates used to project or predict the new a priori estimates.

6. Results and discussions

The plant controller parameters, algorithm constants and other system specifications are presented in table 1.

For a linear load, the first input and output voltage waveforms, estimated and measured inductor currents as well as estimated parameters are shown in Fig. 4, 5 and 6 respectively.

The carrier signal and u_{LQR} signal of the first simple phase for a linear load are shown in Fig.7.

A linear load output voltage and the reference with values of R and K (gains) taken from table1 are illustrated in Fig. 8 and the output voltage frequency spectrum is presented in Fig. 9. From this spectrum, the THD is calculated and the obtained value is 1.90% showing a high quality output voltage.

The efficiency of the LQR regulator is well illustrated in Fig. 10. It is shown that the output voltage follows efficiently the reference voltage in case of linear load disturbance.

Fig. 11 shows the output voltage tracking the reference voltage efficiently in the case of linear load disturbance. From this figure, it is clear that the proposed LQR regulator is efficient.

For a nonlinear load, the output voltage, the output current, the output voltage frequency spectrum and the output DC voltage of the diode bridge rectifiers with RC load are shown in Fig. 12 , 13 and 14 respectively. The THD obtained from the voltage spectrum is equal to 2.89% proving a high quality output voltage.

Table 1. System parameters.

System parameters	Values
DC input voltage	E=750V
Reference voltage	Vref1= 311 V (peak),60Hz
Sample time	Ts= 1/I8000s
States weightings	Q=diag [50 100 150 1 1]
Control weighting	Ru=100
For linear load:	
Filter inductance	L= 5.3 mH
Filter capacitance	C= 100 μF
Linear load	R= 32Ω
LQR gains	K= [9.0014 33.8617 -1.0248 -10.8893 -0.0036]
For non linear load:	
Non linear load	Diode bridge rectifiers with RC
Filter inductance	L= 5.3 mH
Filter capacitance	C=1 00 μF
LQR gains	K= [9.1300 34.0653 -1.0237 10.8806 -0.0036]
Switching frequency	f =1500 Hz

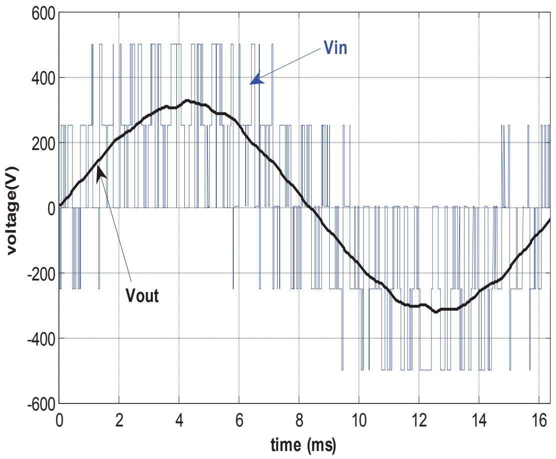

Fig. 4. Input and output voltage of the first phase for a linear load.

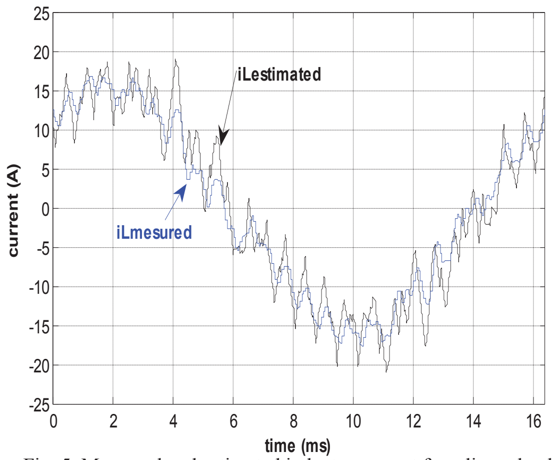

Fig. 5. Measured and estimated inductor current for a linear load.

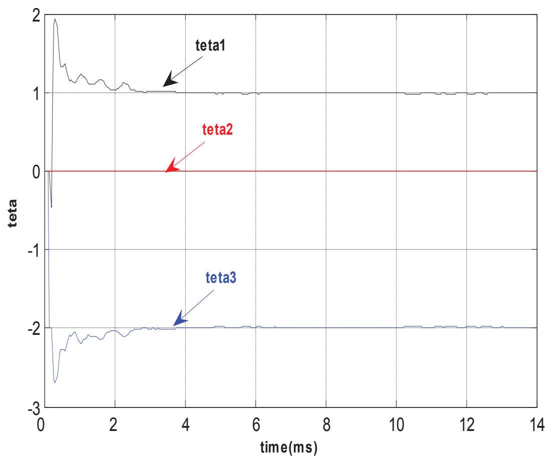

Fig. 6. The estimated parameters for a linear load.

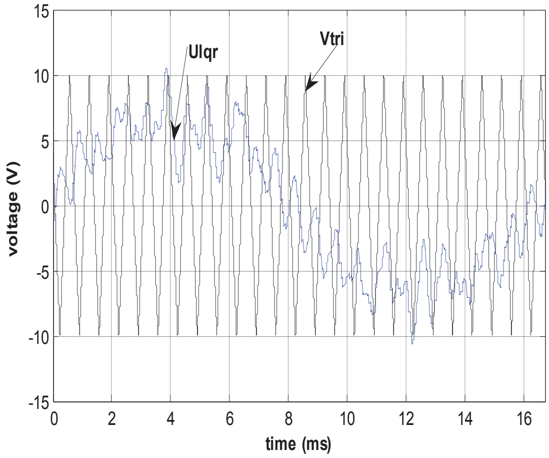

Fig. 7. Carrier and Ulqr signal of the first simple phase for a linear load.

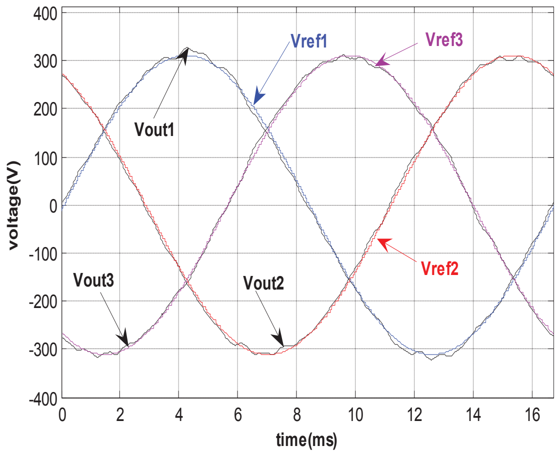

Fig. 8. References voltages, output voltages for a linear load.

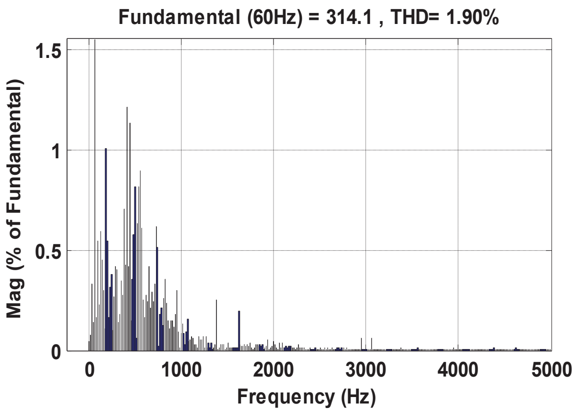

Fig. 9. Spectral analysis of the output voltage for a linear load.

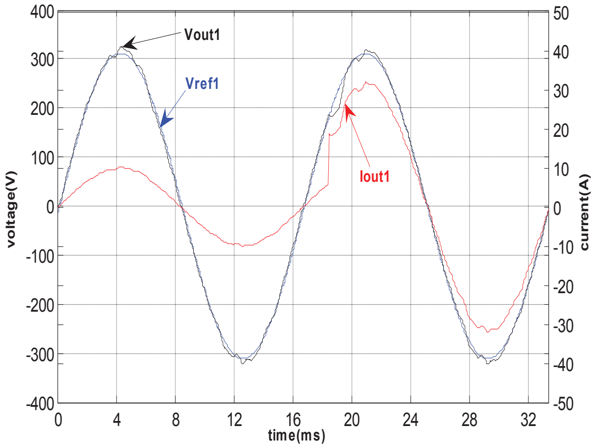

Fig. 10. Reference voltage, output voltage and current with linear load disturbance (From R= 32Ω to R=10Ω).

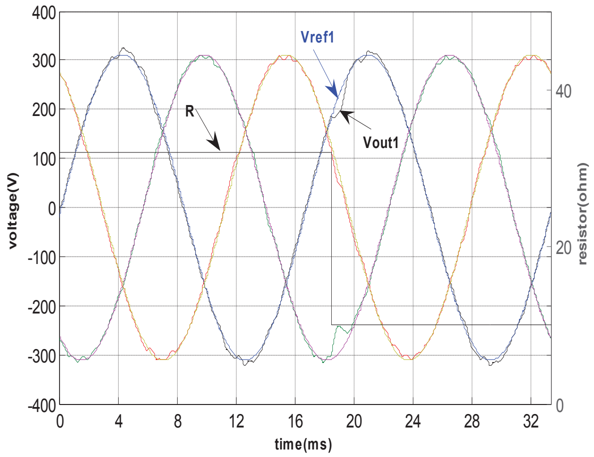

Fig. 11. References voltages, output voltages with linear load disturbance (From R= 32Ω to R=10Ω).

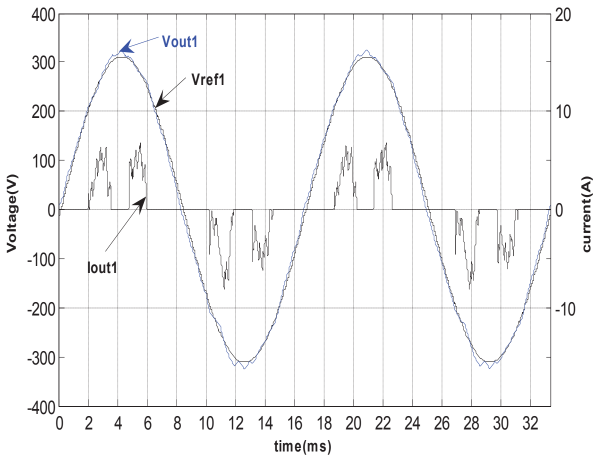

Fig. 12. Output voltage and current of the first simple phase for a non linear load

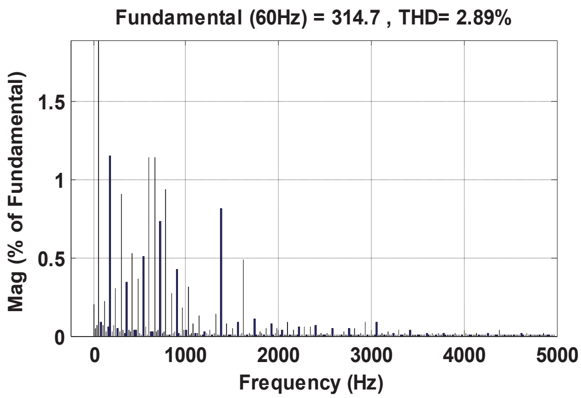

Fig. 13. Spectral analysis of the output voltage for a non linear load.

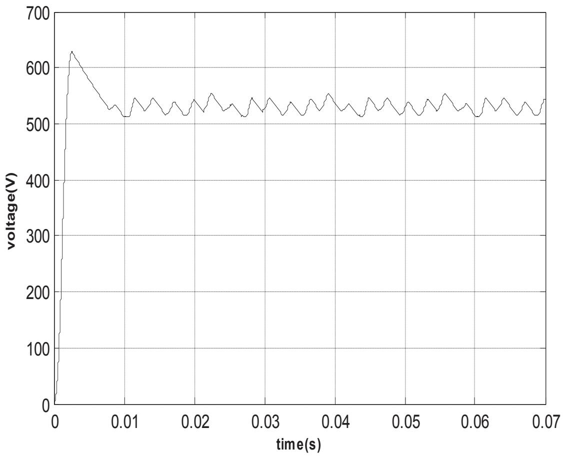

Fig. 14. Output dc voltage of the diode bridge rectifiers with RC load.

7. Conclusions

A Linear Quadratic Regulator was successfully developed for a three phase UPS application. The linear quadratic regulator gains are calculated by minimizing a cost function which can be changed by the designer by modification of the weighting factors. Therefore, it is possible to reduce the control efforts in tracking the sinusoidal reference. The RLS estimator identifies the plant parameters which are used to compute LQR gains periodically. The discrete control law has shown good performances to linear and nonlinear loads when operated at low switching frequency (1500 Hz). These characteristics make this scheme suitable to be used in high power applications as well as to be implemented through a low cost microcontroller.

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