

Control of High Power Factor Boost Rectifier based on Variable Structure Theory

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Abstract

This paper presents the design of Sliding Mode Control for high power factor boost rectifier. The general features are no input voltage sensing, and no use of multiplier. The proposed controller operates in continuous conduction mode. Moreover, due to the sliding mode control, fast and stable response is achieved, in spite of the large output filter. Control complexity is the same as that of standard current-mode controls. The proposed circuit is an interesting solution when high-efficiency and low-cost control circuits are required. Converter analysis, design criteria, and Simulation results are presented. Simulation is done by using mathematical model of boost rectifier in MATLAB/SIMULINK environment.

Index terms: Modeling of Boost Rectifier, CCM, VSCS, and SMC.

Introduction

With the ever-increasing demand for power quality from the utility grid, power factor correction (PFC) is becoming a basic requirement for switch-mode power supplies. High quality rectifiers also used to interface ac line and dc load and appear to the line as a resistive load, so achieving unity power factor even in presence of distorted line voltage.

An inherent limitation of these converters is their slow dynamic response. In fact, sinusoidal input current means large input power fluctuations at twice the line frequency, resulting in low-frequency output voltage ripple, which cannot be corrected by the control, otherwise input current waveform is affected. Big output filter capacitors must therefore be adopted in order to limit the voltage ripple, which

further reduces the bandwidth of the voltage control loop.

In practice, this bandwidth limitation can be overcome if some input current distortion is accepted (within the limits posed by the standards, e.g. IEC 555-2), provided that this results in smaller power fluctuation. From this point of view, sliding mode control is very powerful, since it can provide an optimal trade-off between the needs for increasing response speed and reducing input current distortion and output voltage ripple.

Control of switch-mode power supplies can be difficult, due to their intrinsic non-linearity. In fact, control should ensure system stability in any operating condition and good static and dynamic performances in terms of rejection of input voltage disturbances (audio susceptibility) and effects of load changes (output impedance). These characteristics, of course, should be maintained in spite of large input voltage, output current, and even parameter variations (robustness). In classical control approach relies on the knowledge of a linear small-signal model of the system to develop a suitable regulator. The design procedure is well known, but is generally not easy to account for the wide variation of system parameters, because of the strong dependence of small-signal model parameters on the converter operating point. This aspect becomes even more problematic in Power Factor Correction circuits, in which the operating point moves continuously, following the periodic input/output voltage variations.

A different approach, which complies with the non-linear nature of switch-mode power supplies, is represented by the sliding mode control, which is derived from the variable structure systems (VSS) theory [6]-[7]. This control technique offers several advantages:

- First, let N be the system order, system response has order $N-1$. In fact, under sliding mode only $N-1$ state variables are independent, the N -th being constrained by sliding surface.
- Second, system dynamic is very fast, since all control loops are concurrently.
- Third, stability (even for large input and output variations) and Robustness (against load, supply and parameter variations) are excellent, as for any other hysteretic control.
- Fourth, system response slightly on actual converter parameters.

The proposed sliding mode control is based on the property of quasi-steady-state operation of CCM converters. Indeed, it is an extension of the works reported in [1]-[2], in which the same property was used to simplify the control circuitry of a PWM controller.

Control Design

Control of Variable Structure Systems: The Sliding Mode Approach

VSS are systems whose physical structure is changed intentionally during the time in accordance with a preset structure control law. The instants at which the changing of the structure occur are determined by the current state of the system. From this point of view, switch-mode power supplies represent a particular class of VSS, since their

structure is periodically changed by the action of controlled switches and diodes. During the control process, the structure of the control system varies from one structure to another thus earning the name *variable structure control* (VSC).

VSC is a high-speed switching feedback control resulting in a sliding mode. For example, the gains in each feedback path switch between two values according to a rule that depends on the value of the state at each instant. The purpose of the switching control law is to drive the nonlinear plant's states trajectory on this surface for all subsequent time. This surface is called *Switching Surface*. When the plants state trajectory is "above" the surface, feedback path has one gain and a different gain if the trajectory drops "below" the surface. This surface defines the proper rule for switching. The surface is also called *Sliding Surface* because, ideally speaking, once intercepted, the switching control maintains the plant's state trajectory on the surface for all subsequent time and the plant's state trajectory along this surface. The plant dynamics restricted to this surface represent the controlled system behavior.

A VSC control design breaks into two phases.

- The first phase is to design or choose a switching surface so that the plant state trajectory restricted to the surface has desired dynamics.
- The second phase is to design a switched control that will drive the plant state to the switching surface and maintain into on the surface upon interception.

A lyapunov approach is used to characterize this second design phase. Here a generalized Lyapunov function, that characterizes the motion of the state trajectory to the surface, is defined in terms of the surface. For each chosen switched control structure, one chooses the "gains" so that the derivative of this lyapunov function is negative definite, thus guaranteeing motion of the state trajectory to the surface. To emphasize the important role of the Sliding mode, the control is also called *Sliding mode control*. It should be noted that a variable structure control system can be devised without a sliding mode, but such system does not possess the associated merits.

System Model, and Control Structure, and Sliding Mode

Let us consider now a class of systems with a state model nonlinear in the state vector $x(\cdot)$ and linear in the control vector $u(\cdot)$ in the form

$$x(t) \in R^m \tag{1}$$

Where $x(t) \in R^m$, and $B(x,t) \in R^{n \times m}$: further, each entry in $f(x,t)$ and $B(x,t)$ is assumed continuous with a bounded continuous derivative with respect to x .

The first phase is to designed or chooses a switch surface so that the plant state trajectory restricted to the surface has desired dynamics. According to the Sliding mode control theory [17], all state variables area sensed, and the states are multiplied by proper gains K_i and added together to form the sliding function $\sigma(x,t)$. Then, hysteretic block maintains this function to zero, so that we can define sliding surface as:

$$\sigma(x,t) = \sum_1^N K_i \cdot x_i = 0 \quad (2)$$

Where N is the system order (number of state variables)
For second order systems

$$\sigma = x_1 + \tau \cdot x_2 \quad (3)$$

Which is a linear combination of the two state variables. In the phase plane, equation $\sigma = 0$ represents a line, called sliding line, passing through the origin (Which is the final equilibrium point for the system).

We now define the following control strategy

$$\text{If } \sigma > +\beta \Rightarrow u = 0 \quad (4)$$

$$\text{If } \sigma < -\beta \Rightarrow u = 1$$

Where β define a suitable hysteresis band. In this way, the phase plane is divided in two regions separated by the sliding line, each associated to one of the two sub topologies defined by switch status u . When the system status is in P, Since we are in the region $\sigma < -\beta$, the switch is closed and the motion occurs along a phase trajectory corresponding to $u=1$. When the system status crosses the line $\sigma = +\beta$, according to (4), $u=0$ and the system status follows a phase trajectory corresponding to $u=1$. observing that the phase trajectories, in proximity of the sliding line, are directed toward the line itself, the resulting motion is made by continuous commutations around the sliding line, so that the system status is driven to the final equilibrium point.

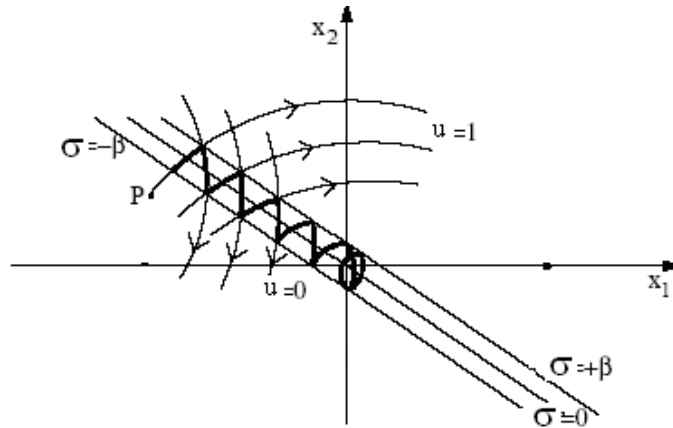


Figure 1: Sliding surface.

From this example, in the hypothesis of a suitable small value of β , two important conclusions arise

- When the system is in the sliding mode, its evolution is independent of the circuit parameters. It depends only on the sliding line chosen. In the example shown in Fig.3 the dynamic is of the first order with a time constant equal to τ .
- If N is the order of the original system, the dynamic of the controlled system in sliding mode has order $N-1$, since the state variables are constrained by the equation $\sigma = 0$.

Note that the switching frequency is determined by the amplitude of the hysteresis band β . The potentialities of this control technique in the application to switch-mode power supplies are now evident: it exploits the intrinsic non-linear nature of these converters and it is able to provide dynamic behaviors that are different from that of the substructures composing the system, and correspond to that of a reduced system.

Switching surface design

In the simple case of the second-order systems considered in the previous section, sliding mode control design requires only selection of parameter τ . Selection, must be done in order to ensure the following three constraints The *hitting condition*, which requires that the system trajectories cross the sliding line irrespective of their starting point in the phase plane; The *existence condition*, which requires that the system trajectories near the sliding line (in both regions) are directed toward the line itself; The *stability condition* of the system motion on the sliding line (i.e. the motion must be toward the equilibrium point).

Existence Condition

The existence condition requires that the phase trajectories are directed toward the sliding surface in a small volume around the surface itself. It is achieved by defining the appropriate Lyapunov function $V(x,t,\sigma)$. For signal input systems it is ordinarily convenient to choose a Lyapunov function of the form $V(x,t,\sigma) = 0.5\sigma^2(x)$. To determine the gains necessary to drive the system state to the surface $\sigma(t) = 0$, they may be choose so that

$$\dot{V}(x,t,\sigma) = 0.5 \frac{d\sigma^2}{dt} = \sigma(x) \frac{d\sigma(x)}{dt} = \sigma \cdot \dot{\sigma} < 0 \quad (5)$$

Thus sliding mode does exist on a discontinuity surface whenever the distances to this surface and the velocity of its hange $\dot{\sigma}$ are of opposite signs, i.e. when

$$\lim_{x \rightarrow +0} \dot{\sigma} > 0 \quad \text{and} \quad \lim_{x \rightarrow +0} \dot{\sigma} < 0$$

Stability Condition

Switching surface design is predicted upon knowledge of the system behavior in a sliding mode. This behavior depends on the parameters of the switching surface. In

any case, achieving a switching-surface design requires analytically specifying the motion of the state trajectory in a sliding mode. The so called method of equivalent control is essential to this specification.

Equivalent Control

Equivalent control constitute an equivalent input which, when exciting the system, produces the motion of the system on the sliding surface whenever the initial state is on the surface. Suppose at t_1 the plant's state trajectory intercepts the switching surface, and a sliding mode exists. The existence of a sliding mode implies that, for all $t \geq t_1$, $\sigma(x(t), t) = 0$, and hence $\dot{\sigma}(x(t), t) = 0$. Using the chain rule, we define the equivalent control u_{eq} for system of the form eq.3.70 as the input satisfying

$$\dot{\sigma} = \frac{\partial \sigma}{\partial t} + \frac{\partial \sigma}{\partial x} \dot{x} = \frac{\partial \sigma}{\partial t} + \frac{\partial \sigma}{\partial x} f(x, t) + \frac{\partial \sigma}{\partial x} B(x, t) u_{eq} = 0 \quad (6)$$

Assume that the matrix product $\frac{\partial \sigma}{\partial x} B(x, t)$ is nonsingular for all t and x , and one can compute u_{eq} as

$$u_{eq} = - \left[\frac{\partial \sigma}{\partial x} B(x, t) \right]^{-1} \left(\frac{\partial \sigma}{\partial t} + \frac{\partial \sigma}{\partial x} f(x, t) \right) \quad (7)$$

Therefore, given that $\sigma(x(t_1), t_1) = 0$, then for all $t \geq t_1$, the dynamics of the system on the switching surface will satisfy

$$\dot{x}(t) = \left[I - B(x, t) \left[\frac{\partial \sigma}{\partial x} B(x, t) \right]^{-1} \frac{\partial \sigma}{\partial x} \right] f(x, t) - B(x, t) \left[\frac{\partial \sigma}{\partial x} B(x, t) \right]^{-1} \frac{\partial \sigma}{\partial t} \quad (8)$$

This equation represents the equivalent system dynamics on the sliding surface. The driving term is present when some form of tracking or regulation is required of the controlled system, e.g. when

$$\sigma(x, t) = \sum_1^N K_i \cdot x_i + r(t) = 0 \quad (9)$$

With $r(t)$ serving as a “reference” signal.

Physical Meaning of the Equivalent Control

A real control always includes a slow component to which a high rate component maybe added. So decompose the control structure as

$$u(x, t) = u_{eq}(x, t) + u_N(x, t) \quad (10)$$

Where u_{eq} is only valid on the sliding surface and u_N assures the existence of a sliding mode. And u_N is defined as

$$u_N(x,t) = \text{sgn}(\sigma) \quad (11)$$

Where

$$\text{sgn}(\sigma) = \frac{\sigma}{|\sigma|}$$

Since a control plant is a dynamic object, its behavior is largely determined by the slow component while its response to the high rate component is negligible. On the other hand, the equivalent control method demands a substitution of the real control in the motion Eq (10) with a continuous function $u_{eq}(x,t)$, which does not contain any high rate component. The equivalent control equals the slow component of the real control i.e. average control value and may be measured by a first order linear filter provided its time constant is small enough as compared with the slow component, yet large enough to filter out the high rate component and approximately matched with the boundary layer width.

Sliding Mode Control of High Power Factor Boost Rectifier

In this work a new variation of Sliding mode control of High Power Factor Boost rectifier based on Quasi-Steady-State operation of CCM converter is proposed.

System Model

Taking into account that the cut-off frequency of the filter is much higher the line frequency, the resulting close loop dynamic behavior of the boost Power Factor Correction rectifier can be approximated by the following relations:

$$i_i = i_L \text{sgn}(V_{in}) \quad (11)$$

$$L \cdot \frac{di}{dt} = |V_{in}| - V_o \cdot u_{eq} \quad (12)$$

$$C \cdot \frac{dV_o}{dt} = i \cdot u_{eq} - \frac{V_o}{R} \quad (13)$$

Where u_{eq} is the average values of the control variable u ($u=0$ when switch is on and $u=1$ when switch is off).

Controller Design

The input current of a Active Power Factor Correction must follow the shape the line voltage as a resistor, and simultaneously, the average value of the output voltage ($\langle V_o \rangle$) must be regulated to a reference value ($V_o \text{ ref}$). both control goals be achieved if the input current satisfies the following relation.

$$i_L = \frac{|V_{in}|}{R_{eq}} \quad (14)$$

Where i_L is the inductor current of the converter and R_{eq} is the emulated resistor the values of which are related to the input power and the output voltage. The value of R_{eq} must be controlled to make the average value $\langle V_o \rangle$ follow a reference value $V_{o\ ref}$.

The control scheme consists of both an outer control loop, which regulates the value of the emulated resistor, and an inner control loop, which is used to make the input current follow a reference value $\frac{|V_{in}|}{R_{eq}}$.

To implement the relation of Eq. (15), avoiding the sensing of the line voltage and the use of an analogue multiplier, the quasi-steady-state approach is considered. According to the approach, the ratio between the line and the output voltage can be approximated by the voltage conversion ratio of a DC/DC boost converter. Therefore, the following relationship can be used in the control design

$$\frac{V_o}{|V_{in}|} \approx \frac{1}{\langle u \rangle} \quad (16)$$

Where $\langle u \rangle$ is the average value of the control variable u during a switching period ($u=0$ when switch is on and $u=1$ when switch is off).

Substituting Eq.(16) for Eq.(15) yields a new expression of the control goal is

$$i_L \approx \frac{V_o \cdot \langle u \rangle}{R_{eq}} \quad (17)$$

Taking into the account that i_L and $\frac{V_o}{R_{eq}}$ are nearly constant during a switching period, the relation eq. can be approximated by

$$\langle i_L - K.u \rangle \approx 0 \quad (18)$$

$$H_{LP}(s).(i_L - k.u) \approx 0 \quad (19)$$

Where K is defined as $\frac{V_o}{R_{eq}}$, and it is the new control variable which is used to regulate the output voltage. In order to achieve the unity-power-factor, the expression (18) must be satisfied when the sliding regime is reached. The expression of this required sliding surface can be deduced by identifying the invariance condition $\sigma = 0$ with Eq.(18).

$$\sigma = H_{LP}(s).(i_L - k.u) \quad (20)$$

Note that this sliding surface will satisfy the transversal condition if $H_{LP}(s)$ is the

first order low pass filter ($H_{LP}(s)=s_0/s+s_0$), it can be used to calculating the average value of $i_L - k.u$ during a switching period. The control law associated with this surface is deduced by using the reaching condition, $\sigma.\dot{\sigma} < 0$ [7], which yields

$$u = \begin{cases} 1 & \text{.....when } \sigma > 0 \\ 0 & \text{.....when } \sigma < 0 \end{cases}$$

Moreover, a proportional-integral control of K has been used to make the output voltage be regulated to a reference value $V_{o\text{ref}}$.

$$K = K_p(V_{o\text{ref}} - V_o) + K_I \int_0^t (V_{o\text{ref}} - V_o) d\tau \quad (21)$$

Fig.2 shows the block diagram of proposed controller. In order to implement the proposed sliding surface, a first order low-pas filter, an analog multiplexer, and a proportional integral controller are required. Finally, the state of the switch is determined by a hysteresis comparator, according to the control law deduced above.

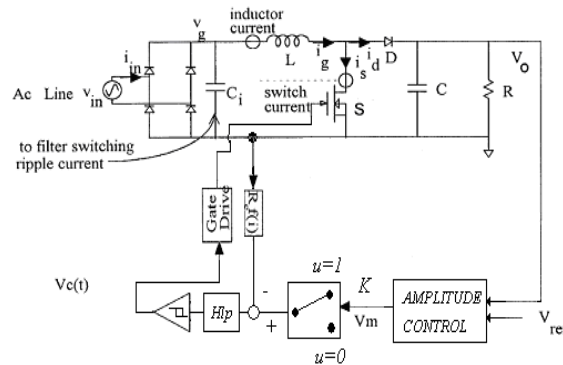


Figure 2: Block Diagram of proposed controller.

Simulation Results

A Sliding mode controller based on variable structure theory has been simulated. Fig.3 shows the simulation diagram of proposed controller. Simulation is done by using MATLAB/SIMULINK with the help of mathematical model of boost converter. Mathematical model of boost converter is developed by using state space averaging technique. The output voltage of the rectifier is regulated at 400V. The simulation waveforms of the input current (i_{in}), input voltage (v_{in}), at different load conditions are shown in Fig.4.(a) and (b).The harmonic spectrum of the corresponding input current is shown in Fig.4. (c) and (d). it can be seen that the PF of the boost rectifier is almost unity. The total harmonic distortion is less. The output voltage waveform is shown in Fig.4.(e) and (f) at $R=350\Omega$ and step change from 250Ω to 1000Ω at 0.5sec respectively. It can be seen that Peak over shoot in out put voltage under the transient

condition is around 2.5% and the output voltage is regulated with low ripple factor under steady-state condition.

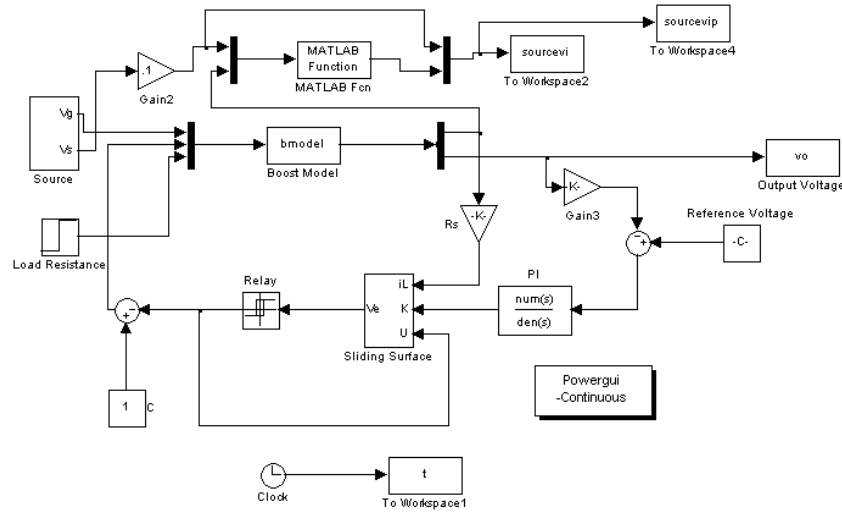


Figure 3: Simulation diagram of proposed controller.

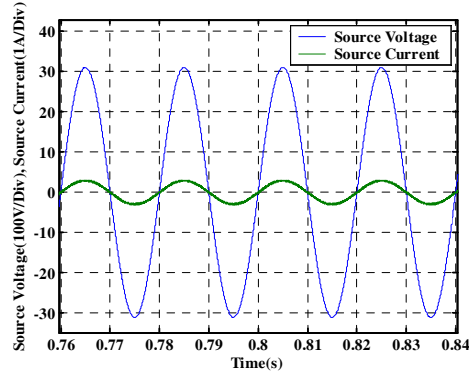


Figure 4 (a)

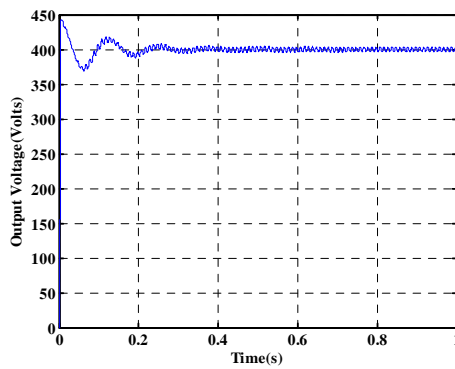


Figure 4(b)

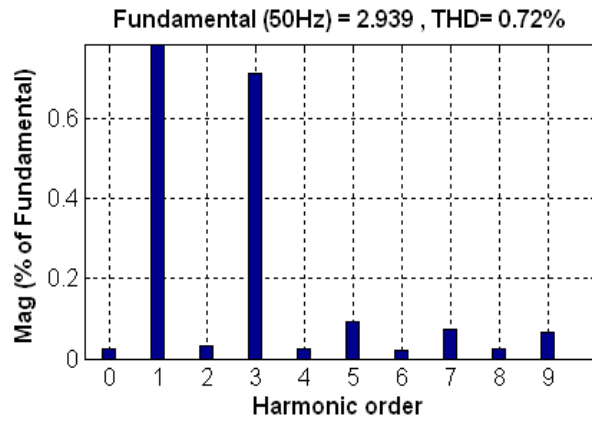


Figure 4(c)

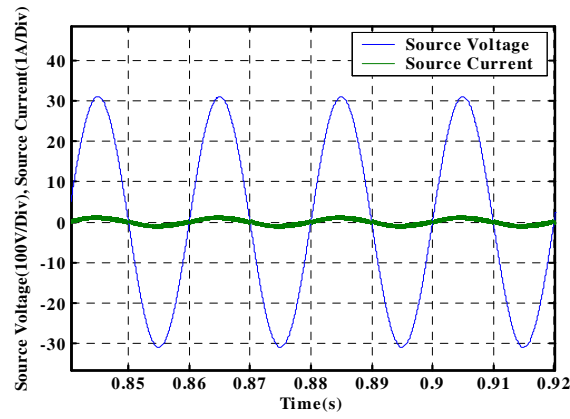


Figure 4(d)

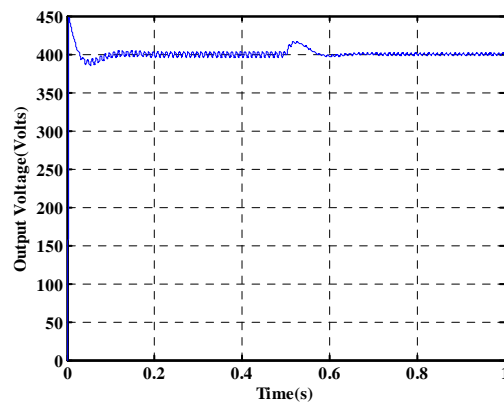


Figure 4(e)

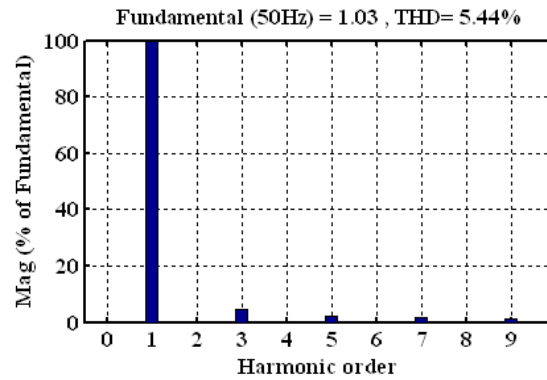


Figure 4(f)

Figure 4: Simulation results: (a),(b) are input current and voltage,(c),(d) are THD in input current, (e) and (f) are output voltage at $R=350\Omega$ and $R=250\Omega-1000\Omega$ at 0.5sec respectively.

Conclusion

This paper proposes a new controller for a high power factor boost rectifier. The proposed scheme operates in continuous conduction mode and eliminates the multiplier and the input voltage sensor in the multiplier approach with penalty of slight distortion of the input current.

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