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Harmonic Reduction Using Voltage Source Converter Based Active Power Filter with One Cycle Control

G.Madhavi Latha
M.Tech Student,
Dept. of Electrical and Electronics Engineering,
Vignan's Institute of Engineering for Women, VIEW,
Visakhapatnam, India.

K. Kusal Kumar
Assistant Professor,
Dept. of Electrical and Electronics Engineering,
Vignan's Institute of Engineering for Women, VIEW,
Visakhapatnam, India.

Abstract

Distribution line consists of various loads which are both linear and non linear ,due to these non linear loads a harmonic current is developed at the source current which leads to harmful disturbances, aging effect, poor power factor, lower efficiency. Employing active power filter is one of the viable solutions to eliminate power line harmonic currents generated by non linear loads and to improve the power factor. APF is operated in dual boost convertor mode with constant switching frequency by using one cycle control method. The advantages of going for this control is it requires only source current and voltage as reference. And flip flops are used to generate the pulse signal to the convertor which is connected in parallel with the disturbution system. The inverter will provide phase shifted current. This phase shifted current is helpful in injecting and cancelling out the harmonic current and to achieve three phase unit power factor for the current to and from the powergrid A DC capacitor is designed which acts as a source to the converter. In order to reduce the error and to produce firing signal, PI controller is used which is extended using a new adaptive fuzzy based one cycle controller.

Index Terms—***Nonlinear loads, Active Power Filter (APF), Dual -Boost converter, One-Cycle control, fuzzy logic controller.***

INTRODUCTION

When a three phase source is connected to a non linear load generally current harmonics are introduced. These current harmonics result in problems such as Low efficiency, Power system voltage fluctuations, Low power factor. One of the viable solutions to eliminate harmonics are power factor correction techniques such as passive power filters and active power filters.. Passive filters are easy to design, simple structure and inexpensive but passive filters have many disadvantages, such as Resonance, fixed compensation character, Large size, possible overload. To overcome the above disadvantages due to Passive Filters, Active Power Filters (APFs) have been proposed as a current-harmonic compensator. The Active Power Filter is connected in shunt of injecting harmonic current into the ac system, of the same amplitude but of reverse phase to that of the load current harmonics.

This will further result in unity power factor and sinusoidal line currents in the input power system. In this case, a minor portion of the energy is processed, which may result in overall higher power processing capability and overall higher energy efficiency[1]. These types of approaches are applicable for low-power (less than 5kVA) to high-power applications (around 100kVA). A three-phase shunt APF is generally comprised of a three-phase bridge converter and control circuitry. Most of the earlier used control approaches require to sense the load current and calculate its harmonics and reactive components in order to produce the reference for controlling the current of the bridge converter[2]. Those control methods require fast and accurate calculation; therefore, high-performance A/D converters and a high-speed digital microprocessors are required, which yields high cost, low stability and complexity. So here introducing a promising solution based on One-Cycle control.

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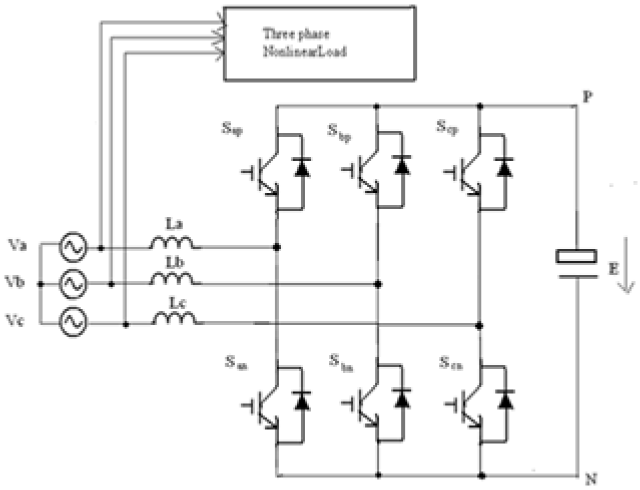

Fig.1 Power stage of the three-phase APF

The one-cycle control method does not require the use of multipliers in the control loop and the need for calculating current reference. The control circuitry is simple and reliable. In pulse width-modulation (PWM) active power filter, all the switches are switched with switching frequency; hence, the switching losses are relatively much more than that of the vector based active power filters. In this paper, a three-phase shunt filter with six-switch bridge voltage-source converter is investigated.

PARALLEL CONNECTED DUAL BOOST Converter

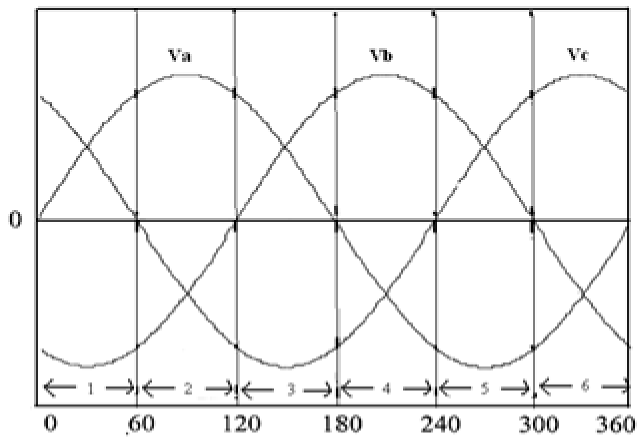

Fig.2 Normalized three-phase grid voltage waveforms

Principle of operation

The waveforms of the three-phase voltages Va, Vb and Vc of the grid are shown in Fig..2 During the each 60^o region in Fig.2, the voltage-source converter in Fig.1 can remain decoupled as a parallel-connected dual-boost converter.

Here the total 360^o region is divided into six regions as shown in Fig.2. In every region two voltages remain either positive or negative, one voltage is either positive or negative, and depending on this condition the total 360^o region is divided into six regions. In each respective region the voltage-source converter is decoupled as a dual-boost converter as explained in the fyrther section. In the region ($0\sim60^o$), the phase voltage V_b is the smallest. In this case, switch S_{bp} is kept off and switch S_{bn} is kept on throughout the complete 60^o region, while switches in the extra two branches such as S_{an},S_{ap} and S_{cn},S_{cp} are controlled complementally (with small dead time in between) at the switching frequency. For example, during respective switching cycle, if switch S_{an} is OFF, switch S_{ap} will be ON and vice versa. Here line frequency is much lesser than the switching frequency. Here Line frequency is 60 HZ and Switching frequency is 50 KHZ.

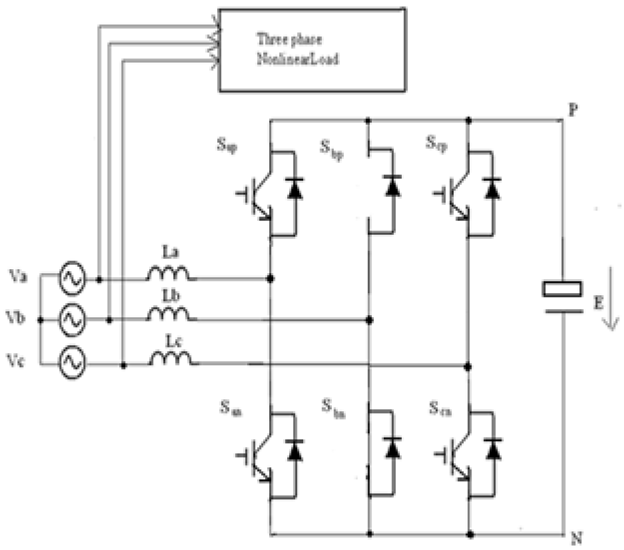

Fig.3 Power stage of the three-phase APF during $0\sim60^o$ regions.

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Characteristics of the proposed converter

For the dual-boost converter shown below in Fig.4 or 5, four different switching states are possible for the two switches T_p and T_n . The four switching states and inductor voltages are shown in Table 1. Where

$$\begin{bmatrix} V_p^* \\ V_n^* \\ V_t^* \end{bmatrix} = \begin{bmatrix} \frac{2}{3} & -\frac{1}{3} \\ -\frac{1}{3} & \frac{2}{3} \\ \frac{1}{3} & \frac{1}{3} \end{bmatrix} \cdot \begin{bmatrix} V_p \\ V_n \end{bmatrix} \quad \text{-----------}(1)$$

For a three-phase APF with a constant switching frequency, only two switching sequence possibilities occur, i.e., I, II, IV (condition $d_p > d_n$ where d_p , d_n are the duty ratios of the switches , T_p, T_n respectively) or I, III, IV (condition $d_p < d_n$) during the every switching cycle, if trailing-edge modulation is achieved.

State	T_p	T_n	$\overline{T}_p$	$\overline{T}_n$	V_{Lp}	V_{Ln}	V_{Lt}
I	ON	ON	OFF	OFF	V_p^*	V_n^*	V_t^*
II	ON	OFF	OFF	ON	$V_p^*+\frac{1}{3}\cdot E$	$V_n^*-\frac{2}{3}\cdot E$	$V_t^*-\frac{1}{3}\cdot E$
III	OFF	ON	ON	OFF	$V_p^*-\frac{2}{3}\cdot E$	$V_n^*+\frac{1}{3}\cdot E$	$V_t^*-\frac{1}{3}\cdot E$
IV	OFF	OFF	ON	ON	$V_p^*-\frac{1}{3}\cdot E$	$V_n^*-\frac{1}{3}\cdot E$	$V_t^*-\frac{2}{3}\cdot E$

The voltage waveforms across the inductors L_p, L_n, L_t are shown in Fig.6 .For the first switching sequence ($d_n<d_p$). Based on the assumption that line frequency is much lesser than the switching frequency, the inductor voltage-second balance is approximately applicable, that is

$$\begin{aligned} &V_p^* d_n + \left(V_p^* + \tfrac{1}{3}E\right).\left(d_p - d_n\right) + \left(V_p^* - \tfrac{1}{3}E\right).\left(1 - d_p\right) = 0 \\ &V_n^* d_n + \left(V_n^* - \tfrac{2}{3}E\right).\left(d_p - d_n\right) + \left(V_n^* - \tfrac{1}{3}E\right).\left(1 - d_p\right) = 0 \\ &V_t^* d_n + \left(V_t^* - \tfrac{1}{3}E\right).\left(d_p - d_n\right) + \left(V_t^* - \tfrac{2}{3}E\right).\left(1 - d_p\right) = 0 \end{aligned} \quad \text{-----}(2)$$

The following equation is valid for a symmetrical three-phase system:

$$V_p^* + V_n^* - V_t^* = 0 \quad \text{---------} \quad (3)$$

From (2) and (3) with additional simplification we will get

$$\begin{bmatrix} (1-d_p) \\ (1-d_n) \end{bmatrix} = \begin{bmatrix} 2 & 1 \\ 1 & 2 \end{bmatrix} \cdot \begin{bmatrix} \frac{V_p^*}{E} \\ \frac{V_n^*}{E} \end{bmatrix} \quad \text{-------------} \ (4)$$

It has been verified that this equation is valid for the other switching sequence I, III, and IV ($d_p<d_n$) also.

Equation (4) gives an inherent relationship between the duty cycle and the input, output voltage for the shunt-connected dual-boost converter.

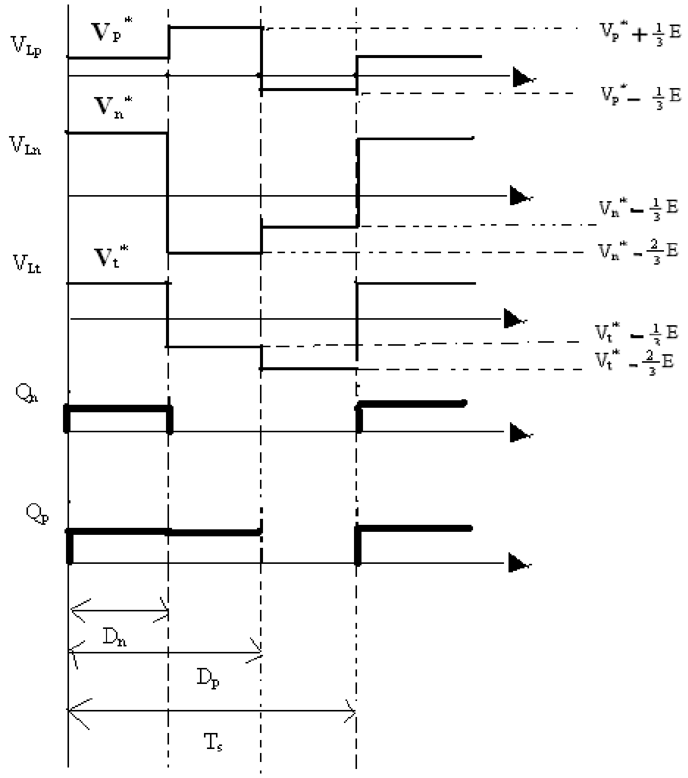

Fig.4 Inductor voltage waveforms for the converter during the condition $d_p > d_n$

In Fig.6 V_{Lp}, V_{Ln}, and V_{Lt} represents the voltage across inductors V_p , V_n , and V_t , accordingly. Q_p and Q_n are the driving signals for switches T_p and T_n respectively.

PROPOSED ONE-CYCLE CONTROLLER FOR THREE-PHASE APF

For the unity-power-factor of the three-phase APF, the control goal is to force the grid line current in each phase to follow the corresponding sinusoidal phase

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voltage, i.e.,

$$\left.\begin{aligned} V_a &= Re \cdot i_a \\ V_b &= Re \cdot i_b \\ V_c &= Re \cdot i_c \end{aligned}\right\} \text{---------------- (5)}$$

where Re is the emulated resistance that reflects real power of the load. This control goal can be achieved by controlling the equivalent currents i_p and i_n to follow the voltages V_p^* and V_n^* The control goal equation of the three-phase APF can be rewritten as

$$\left.\begin{aligned} V_p^* &= Re \cdot i_p \\ V_n^* &= Re \cdot i_n \end{aligned}\right\} \text{------------- (6)}$$

Substituting (6) into (4) and considering the switch is ON for the whole 60° region, it is achieved that

$$\left.\begin{bmatrix} (1-d_p) \\ (1-d_n) \end{bmatrix} = \frac{Re}{E\,Rs} \cdot Rs \cdot \begin{bmatrix} 2 & 1 \\ 1 & 2 \end{bmatrix} \cdot \begin{bmatrix} i_p \\ i_n \end{bmatrix}\right\}_{d_t=1} \text{-------------- (7)}$$

Define

$$Vm = \frac{E\,Rs}{Re} \text{ ------------- (8)}$$

where the signal V_m can be generated from output voltage feedback compensator which is used to regulate the output voltage of the capacitor E of the voltage source converter consequently to the load level; Rs is the equivalent current sensing resistance and it is fixed constant. Merging of the two equations 7,8 and the control key equation is obtained as

$$\left.Vm \cdot \begin{bmatrix} (1-d_p) \\ (1-d_n) \end{bmatrix} = Rs \cdot \begin{bmatrix} 2 & 1 \\ 1 & 2 \end{bmatrix} \cdot \begin{bmatrix} i_p \\ i_n \end{bmatrix}\right\}_{d_t=1} \text{- (9)}$$

The above equation indicates that three-phase power factor can be derived by controlling the duty ratios of switches so that first-order polynomial equation (9) is satisfied. This can be recognized by the one-cycle control core as shown in Fig.7. The operation waveforms are shown in Fig.8.

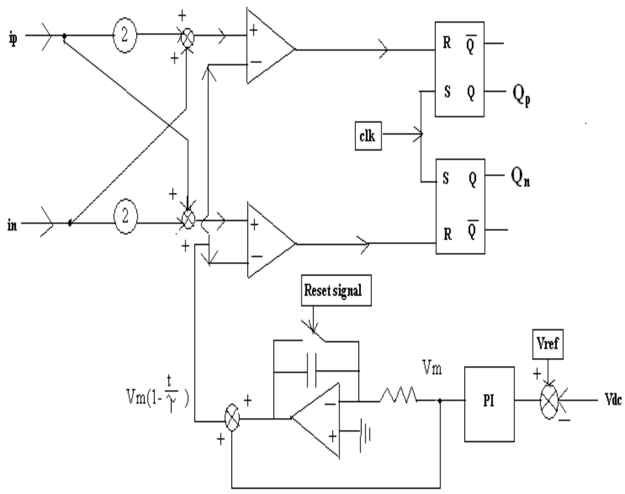

Fig.5 One- cycle control logic

In the beginning of each cycle, the clock pulse sets the two flip-flops. The currents i_n and i_p from the current selection logic is linearly merged to form an input to each of the two comparators. At further input of the two comparators is the value of Vm and the subtracted integrated value of V_m. Signal $V_m(1\text{-}t/T_s)$ is compared with $R_s(\ i_p+2\ i_n)$ in the lower comparator and is compared with $R_s(2\ i_p+\ i_n)$ in the upper comparator as shown in Fig.7. When the two inputs of a comparator happen as shown in Fig.8, the comparator changes its state, which resets the corresponding flip-flop. As an outcome, the correspondent switch is turned off. Therefore, the duty ratios d_p and d_n are determined for the respective switch in each switching cycle.

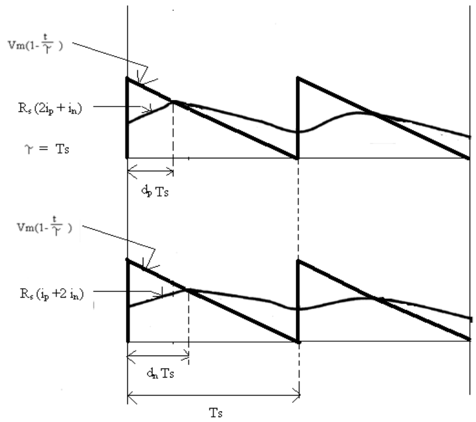

Fig.6 Operational waveforms of One-cycle controlled APF controller.

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The One-Cycle Control Approach has the below Features

Three-phase unity-power-factor and low total harmonic distortion (THD) are obtained by one integrator with reset as well as several logic and linear components. It is reliable and simple.

- Only ac mains current and voltage zero-crossing points are sensed. No sensors for load current and APF inductor current are required.
- Not required to calculate the reference for APF inductor current so that complicated digital computation is removed.
- No need of multipliers.
- Constant switching frequency, which is desirable for industrial applications, is obtained.
- For the three-phase converter, only two switches are operated at higher frequency, and switching losses are reduced compared to PWM converter.

DESIGN CONSIDERATIONS

Dc-Link Capacitor Design

The output dc-link capacitor of voltage source converter is obtained by the output voltage ripple. The equation is given by

$$c \geq \frac{P_o}{2 * f_{line} * (V^2_{omax} - V^2_{omin})} \quad \text{----------(10)}$$

Suppose the power is 7000 W for example; with 2% ripple, APF and output voltage is 400 V. The line frequency is 60 Hz. The capacitance is calculated as 4800µF.

Selection of APF Inductance

The concept of the recommended control is using one-cycle control to apply the control key equation as follows:

$$R_s . i_{eq} = V_m . (1-d) \quad \text{----------------(11)}$$

Where

$$i_{eq} = (2i_p + i_n) \quad (OR) \quad (i_p + 2\, i_n)$$

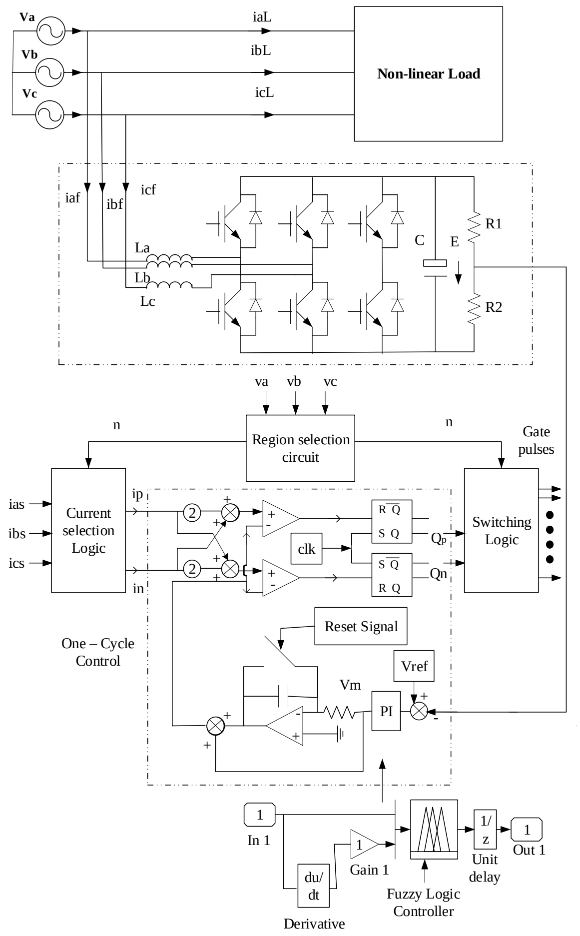

Fig.7 Block diagram of One-Cycle controlled APF with Fuzzy controllers and PI

The operation waveforms are shown in Fig.9. Similar to the peak current model control, there is convergence condition.

The stability condition is given by

$$m_c \geq \frac{(m_2 - m_1)}{2} \quad \text{------------} \ \mathbf{(12)}$$

where m_2 is the OFF slope of the input current and m_1 is the ON slope of the input current; m_c is an equivalent slope of the carrier signal, which is implemented by integrator with reset. Taking into consideration that the load current is low frequency and the influence of load current can be ignored, we only consider the inductor current in the stability analysis, we have

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Where

$$m_1 = R_s \cdot \frac{V_g}{L}$$

$$m_2 = R_s \cdot \frac{V_o - V_g}{L} \quad \text{------------ } \mathbf{(13)}$$

$$m_c = \frac{V_m}{\Upsilon} = \frac{V_m}{T_s}$$

$$\Upsilon = T_s$$

Substitution of (13) into (12) yields the convergence condition

$$V_m \geq \frac{R_s \cdot T_s}{2L} \cdot \left(V_o - 2 \left| V_g \right| \right) \quad \text{----------- (14)}$$

$$\geq \frac{R_s \cdot T_s}{2L} \cdot \left(V_o - 2 \cdot V_{grms} \left| \sin(wt) \right| \right)$$

the condition of convergence is dependent on the angular angle of input voltage and the V_m and wt , which is related to the output power and input voltage. When the convergency condition is satisfied partly, the system will still be stable.

According to (14), convergence condition for region $0^o \sim 360^o$ is given by

$$V_m \geq \frac{R_s \cdot T_s}{2L} \cdot V_o \quad \text{----------- } \mathbf{(15)}$$

however

$$V_m = \frac{V_o R_s}{R_e} \quad \text{------------}\mathbf{(16)}$$

V_m is related to input voltage and output power through (16) . It can be rewritten as

$$V_m = \frac{P_o \cdot R_s \cdot V_o}{\eta \cdot V_{grms}^2} \quad \text{------------}\mathbf{(17)}$$

where η is the estimated efficiency.

Merging of the above equations yields

$$L \geq \frac{1}{2} \cdot \eta \cdot T_s \cdot \frac{V_{grms}^2}{P_o} \quad \text{---------------}\mathbf{(18)}$$

The above equation was used to conclude the size of inductor. At full load and maximum input voltage condition, the system must be completely stable and then the inductor can be selected by

$$L \geq \frac{1}{2} \cdot \eta \cdot T_s \cdot \frac{\max\left(V_{grms}^2 \right)}{\max\left(P_o \right)} \quad \text{--------}\mathbf{(19)}$$

For η=90% , Ts=20µs, $\max(P_o)$=7000W, $\max(V_{grms})$=170 v then the minimum inductance is derived as L = 250µH.

Fuzzy Logic Controllers

Introduction to Fuzzy Logic:

Fuzzy Logic is a very powerful method of reasoning where input I is not precise and mathematical models are unavailable.

The products based on fuzzy became highly competitive due to Low power consumption, better performance, high reliability and inexpensive. Fuzzy interference system is the process of formulating the mapping from a given input to output. TheFuzzy interference process involves logical operations, membership functions and IF THEN rules.This process involves three steps.The primary step is fuzzification in which the input is a value or a waveform.

The inputs are crisp numbers limited to a specific range. The next step is the rule base which is based on the given input membership function.Most probably IF THEN rules are used. The final step is defuzzification in which fuzzy based rules are converted into defuzzified value.

This Finally obtained value is the controller signal to the flipflop.The performance of the fuzzy controller and PI controllers are compared.

Simulation Results

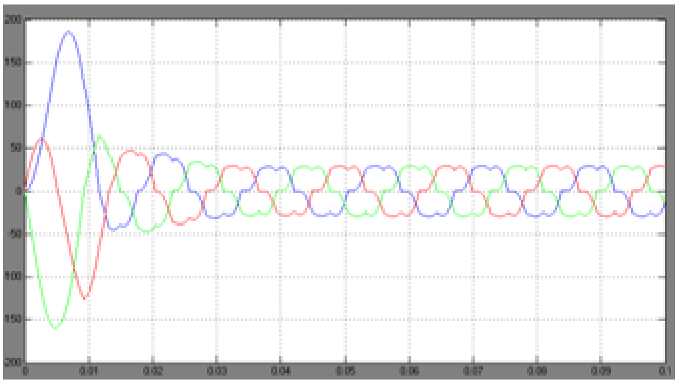

Fig.8 Source current without shunt filter

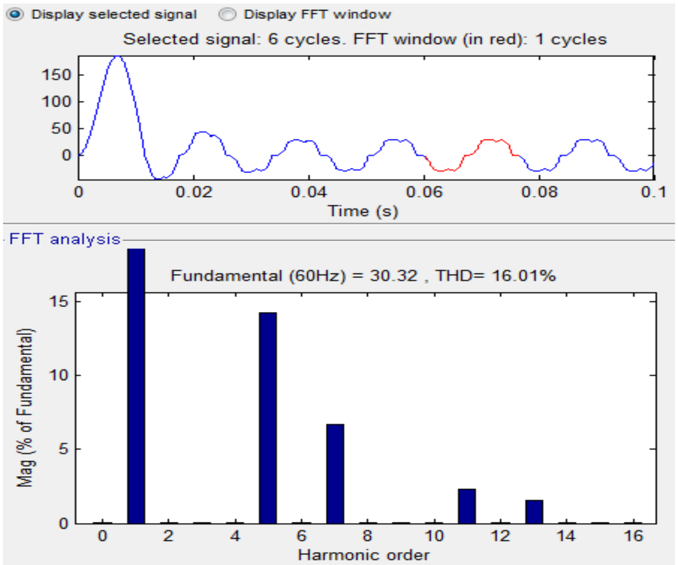

Fig.9 Total Harmonic Distortion without shunt filter

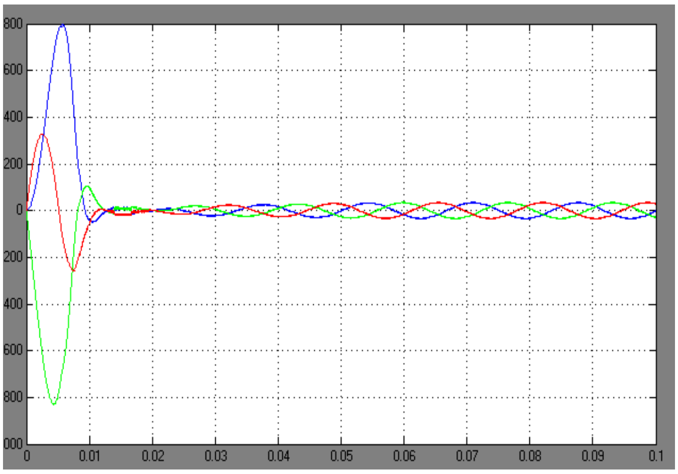

Fig.10Source current with pi one cycle control

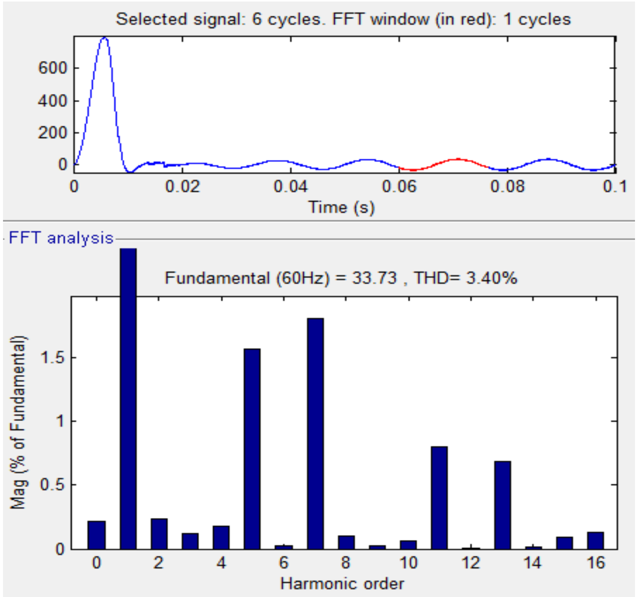

Fig.11 Total Harmonic Distortion with pi one cycle controller

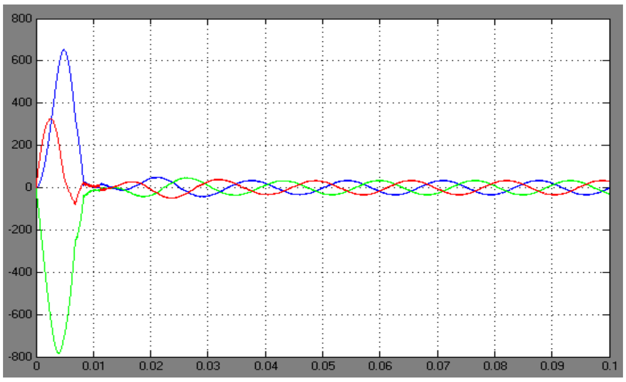

Fig.12 Source current with fuzzy one cycle control

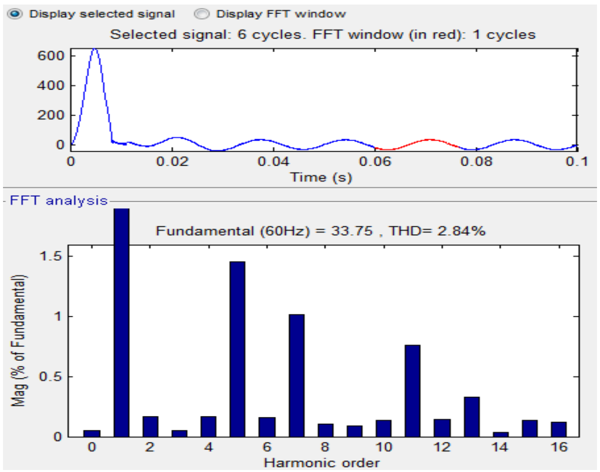

Fig.13 THD with fuzzy one cycle controller

Conclusion

A three phase shunt harmonic filter with one cycle fuzzy controller has been designed in this paper. This new controlling method is determined with more reliability and fast harmonic compensating capacity. The main advantage of this method is it senses only source current and voltage but not the load current and voltage. The intensive computation is removed as there is no need of calculating the reference filter inductor current. When a non linear load is connected to a three phase source some harmonics are introduced. In order to overcome these harmonics in the source, a pi one cycle control based filter and fuzzy one cycle based filter are connected in parallel with the load. The total harmonic distortion without shunt filter is observed to be 16.1%, with PI one cycle control is 3.40%,with fuzzy one cycle control is 2.84%.

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