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Control Strategy for Three-Phase PWM Boost Rectifier Operating Under Different SupplyVoltage Conditions

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Abstract:*In this paper, a proposed control strategy is presented to improve the performance of the pulse width modulation (PWM) boost type rectifier when operating under different supply voltage conditions (balanced, unbalanced, and distorted three-phase supply voltages). The proposed control strategy is divided into two parts, the first part is voltage controller and the second part is current controller. In the voltage controller, Repetitive Controller (RC) is used to reduce the even order harmonics in the regulated output dc voltage so small output capacitor (filter) is used instead of large capacitor. RC also reduces the even order harmonics which appear in the reflected dc current (I_{MAX}), this leads to reduce the odd order harmonics which appear in the input currents. While in the current controller, Enhanced Phase Locked Loop (EPLL) technique is used to obtain sinusoidal and balanced three phases, to construct the reference currents, which are in phase with the fundamental supply voltages. Therefore, the supply-side power factor is kept close to unity. A proportional controller is used to give excellent tracking between the line and the reference currents. The complete system with the proposed control strategy are simulated using Matlab/Simulink. The results for the complete system using repetitive voltage controller are obtained and compared to the results of the system with the conventional voltage controller (Proportional-Integral (PI) controller connected in series with a Low Pass Filter (LPF)). The results with the repetitive controller show better response and stable operation in the steady state under different input voltage conditions, as well as in the transient response under changing the load condition.*

***Index Terms*—Enhanced Phase Locked Loop,Repetitive Controller,Three-Phase PWM Boost Rectifier, Proportional-Integral controller.**

I. INTRODUCTION

The boost type PWM rectifier has been increasingly employed in recent years since it offers the possibility of a low distortion line current withnear unity power factor for any load condition. Another advantage over traditional phase-controlled thyristor rectifiers is its capability for nearly instantaneous reversal of power flow. Unfortunately, the features of the PWM boost type rectifier are fully realized only when the supply three phase input voltages are balanced. It has been shown that unbalanced input voltages cause an abnormal second order harmonic at the dc output voltage, which reflects back to the input causing third-order harmonic current to flow. Next, the third-order harmonic current causes a fourth-order harmonic voltage on the dc bus, and so on. This results in the appearance of even harmonics at the dc output and odd harmonics in the input currents. An attempt was made to reduce low order harmonics at the input and the output of the PWM Boost Type Rectifier under unbalance input voltages [1]. The authors in [2] used two synchronous reference frames: a positive-sequence current regulated by a

proportional integral (PI) controller in a positive synchronous reference frame (SRF), and a negative sequence current regulated by a PI controller in a negative SRF. In [3], a new control scheme is proposed to minimize harmonic distortions of the input current and dc-link voltage in the converter. A hybrid digitalrepetitive current controller (RC) is used to minimize the line-side current harmonics and the dc link voltage harmonics under the distorted and unbalanced operating conditions [4]. But these strategies require a series of frame transformation and calculation which increase the complexity of implementation. New control method for input-output harmonic elimination of the pulse width modulation(PWM) boost-type rectifier under conditions of both unbalancedinput voltages and unbalanced input impedances is presented in [5]. But in this method, hysteresis current controllers are used to regulate the actual three-phase currents, these controllers produce drawbacks such as a variable switching frequency and an irregularity of the position of modulation pulses. These drawbacks provide high current ripples, acoustic noise, and difficulty in designing of input filter.

In [6], a new method for control of PWM rectifiers is presented. This method is classified in Direct Power Control (DPC) group among different control methods for PWM rectifiers which uses Model Predictive Control (MPC) and SVM. The method also usesVirtual Flux (VF) vector of the input voltage, which improves the performance of the rectifier under harmonic conditions of the networks, the method has several advantages: simple, it uses constant switching frequency, and excellent step response. Alsothis method has some disadvantages: the unbalanced condition at the input voltages is not discussed, only balanced and 5% level of 5^{th} order harmonic distortion was discussed. This level of distortion is very low compared to the level used in this paper. Another disadvantage inthis method is that it uses large input inductances as well as large output capacitor.

In this paper, the proposed method can be divided into two parts; the first part is voltage controller (Repetitive Controller) to produce the magnitude of the reference currents (I_{MAX}) without even order harmonics. The second part is current controller (Enhanced phase locked loops are used) to produce pure and balanced sine waves, these sine waves multiplied by the output of the voltage controller (I_{MAX}) to obtain sinusoidal reference currents. Finally proportional controllers are used to force the line currents to follow these sinusoidal references.

II. MODELING OF THREE-PHASE PWM BOOST RECTIFIER UNDER BALANCED INPUT VOLTAGE CONDITION

The main circuit of the three phases PWM ac to dc converter is shown in Figure1. Three-phase line voltages and the balanced input line currents are:

$$e_1=E_m\sin(\omega t) \quad (1a)$$

$$e_2=E_m\sin(\omega t-120) \quad (1b)$$

$$e_3=E_m\sin(\omega t+120) \quad (1c)$$

Assume unity power factor

$$i_1=I_m\sin(\omega t) \quad (2a)$$

$$i_2=I_m\sin(\omega t-120) \quad (2b)$$

$$i_3=I_m\sin(\omega t+120) \quad (2c)$$

Where E_m (I_m) and ω are amplitude of the phase voltage (current) and angular frequency, respectively. For phase 1: [7]

$$L\frac{di_1}{dt}+R_L i_1=V_{AD}=e_1-(V_{DN}+V_{NO}) \quad (3)$$

When switch S_1 is ON and switch S_1' is OFF, the switching functionis:

$d_1=1$ and $d_1'=0$

$$V_{DN} = i_1 R_S + V_o \quad (4)$$

Where R_S is equivalent resistance of a switching device.When switch S_1' is ON, the switching function is:

$d_1=0$ and $d_1'=1$ and

$$V_{DN}=i_1R_S \quad (5)$$

Therefore,equation (3) becomes:

$$L\frac{di_1}{dt}+R_L i_1 = e_1-[(i_1R_S+V_o)\, d_1+ (i_1R_S)\, d_1' +V_{NO}] \quad (6)$$

Because either S_1 or S_1' is conducting and only one of them is allowed to conduct in any moment, i.e. :

$d_1 + d_1'=1$ so

$$L\frac{di_1}{dt}+R_L i_1 = e_1- [i_1R_S\, d_1+V_o\, d_1+ i_1R_S\, d_1' + V_{NO}] \quad (7)$$

$$L\frac{di_1}{dt} + R_L i_1 = e_1- [i_1\, R_S\, (d_1+ d_1')+V_o\, d_1 +V_{NO}] \quad (8)$$

$$L\frac{di_1}{dt} + R_L i_1+R_S i_1 = e_1- [V_o\, d_1+V_{NO}] \quad (9)$$

$$L\frac{di_1}{dt} + R\, i_1 = e_1- [V_o\, d_1+V_{NO}] \quad (10)$$

Where $R=R_L+R_S$, the total series resistance in one phase. Similarly, for phase 2 and 3: [7]

$$L\frac{di_2}{dt} + R\, i_2 = e_2- [V_o\, d_2+V_{NO}] \quad (11)$$

$$L\frac{di_3}{dt} + R\, i_3 = e_3- [V_o\, d_3+V_{NO}] \quad (12)$$

For a three-phase system without neutral line,

$$i_1+i_2+i_3= 0 \quad (13)$$

The sum of three phase supply is

$$e_1+e_2+e_3= 0 \quad (14)$$

The voltage V_{NO} can be obtained by addingequations (10),(11),and(12)

$$L\frac{di_1}{dt}+L\frac{di_2}{dt}+L\frac{di_3}{dt}+R(i_1+i_2+i_3)=e_1+ e_2+ e_3-V_o*(d_1+d_2+d_3)+3V_{NO} \quad (15)$$

$$0 = V_o(d_1+d_2+d_3)+3V_{NO}$$

$$V_{NO} = \frac{-1}{3} V_o\sum_{k=1}^{3} d_k \quad (16)$$

Substitute equation (16) in equations (10), (11), and (12), the result will be

$$L\frac{di_1}{dt} + R\, i_1 = e_1- u_{s1} \quad (17)$$

$$L\frac{di_2}{dt} + R\, i_2 = e_2- u_{s2} \quad (18)$$

$$L\frac{di_3}{dt} + R\, i_3 = e_3 - u_{s3} \quad (19)$$

Where

$$u_{s1} = V_o\left(\frac{2d_1-(d_2+d_3)}{3}\right) \quad (20)$$

$$u_{s2} = V_o\left(\frac{2d_2-(d_1+d_3)}{3}\right) \quad (21)$$

$$u_{s3} = V_o\left(\frac{2d_3-(d_1+d_2)}{3}\right) \quad (22)$$

In the Figure1, the input line current for each phase is multiplied by switching function, then adding them to result output current i_{dc}, i.e:

$$i_{dc}= i_1d_1 + i_2d_2 + i_3d_3$$

According to Figure 1, the output current i_{dc} is equal to:

$$i_{dc}=i_c + I_o \quad (23)$$

Where i_c ($C\frac{dV_o}{dt}$) is the current of the output capacitor while I_o is the dc load current ($I_o=\frac{V_o}{R_o}$). Soanother differential equation can be written as:

$$C\frac{dV_o}{dt} = i_1\, d_1+i_2\, d_2+i_3\, d_3-\frac{V_o}{Ro} \quad (24)$$

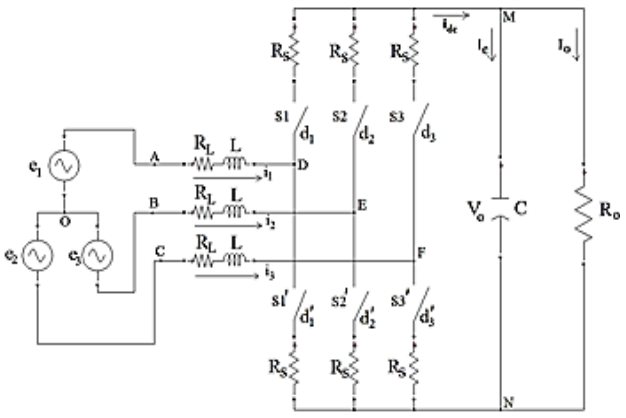

Figure1: Circuit of PWM rectifier

A block diagram of the PWM rectifier with ABC model is presented in Figure2.

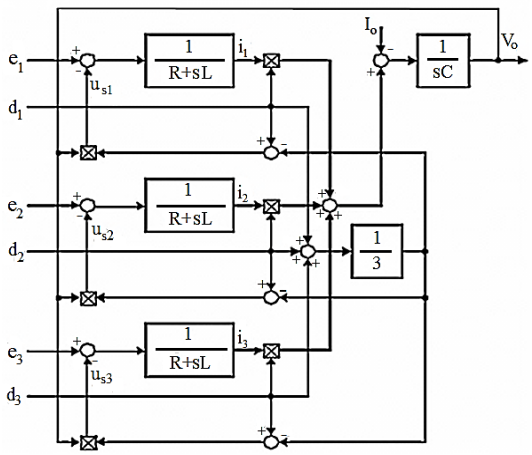

Figure2: PWM rectifier in ABC- model

III. ANALYSIS OF THREE-PHASE PWM BOOST RECTIFIER UNDER UNBALANCED INPUT VOLTAGE CONDITION

In Figure1, it is assumed that the PWM rectifier is supplied by unbalanced input voltages but balanced input impedances. The assumptions used in the following derivation are

1) The system losses are very small and can be neglected.

2) The switching functions used to represent switching action of the converter are unbalanced but contain no zero sequence.

3) Only fundamental components of switching functions and input currents are taken into account (PWM switching harmonics are not considered).

4) There is no phase difference between fundamental components of the switching functions and the input currents. Sothat the fundamental component of the switching functions can be written as

$$d_{f1} = A_1 \sin(\omega t) \quad (25)$$

$$d_{f2} = A_2 \sin(\omega t - 120) \quad (26)$$

$$d_{f3} = A_3 \sin(\omega t + 120) \quad (27)$$

And the unbalanced input currents can be written as

$$i_1 = I_{m1} \sin(\omega t) \quad (28)$$

$$i_2 = I_{m2} \sin(\omega t - 120) \quad (29)$$

$$i_3 = I_{m3} \sin(\omega t + 120) \quad (30)$$

The output dc current is

$$i_{dc} = \overline{T} . \bar{i} \quad (31)$$

The converter transfer function vector $\overline{T}$ is composed of three switching functions.[8]

$$\overline{T} = [\, d_{f1} \; d_{f2} \; d_{f3} \,] \quad (32)$$

And the current $\bar{i}$vector is

$$\bar{i} = \begin{bmatrix} i_1 \\ i_2 \\ i_3 \end{bmatrix} \quad (33)$$

The three-phase unbalanced input currents can be presented as a sum of two balanced sets of positive and negative sequence component.

$$\bar{i} = \overline{i_p} + \overline{i_n} \quad (34)$$

$$\overline{i_p} = \begin{bmatrix} I_p \sin(\omega t) \\ I_p \sin(\omega t - 120) \\ I_p \sin(\omega t + 120) \end{bmatrix} \quad (35)$$

$$\overline{i_n} = \begin{bmatrix} I_n \sin(\omega t) \\ I_n \sin(\omega t + 120) \\ I_n \sin(\omega t - 120) \end{bmatrix} \quad (36)$$

Similarly the converter transfer function can be decomposed into two balanced sets of positive and negative sequence components under unbalanced voltages that is

$$\overline{T} = \overline{T}_p + \overline{T}_n \quad (37)$$

$$\overline{T}_p = \begin{bmatrix} A_p \sin(\omega t) \\ A_p \sin(\omega t - 120) \\ A_p \sin(\omega t + 120) \end{bmatrix}^T \quad (38)$$

$$\overline{T}_n = \begin{bmatrix} A_n \sin(\omega t) \\ A_n \sin(\omega t + 120) \\ A_n \sin(\omega t - 120) \end{bmatrix}^T \quad (39)$$

Now from equation (34) to equation (39), the resultant output current in equation (31) under unbalanced input voltages becomes [8]

$$i_{dc} = (T_p + T_n)(i_p + i_n) \quad (40)$$

$$i_{dc} = T_p i_p + T_p i_n + T_n i_p + T_n i_n \quad (41)$$

Equation (41) represents the general expression for the converter output current i_{dc} under unbalanced voltages in term of positive and negative sequence components of the converter transfer function and the input currents respectively.

$$T_p i_p = \frac{3A_p I_p}{2} \quad (42)$$

$$T_n i_n = \frac{3A_n I_n}{2} \quad (43)$$

$$T_p i_n = \frac{-3A_p I_n}{2}\cos(2\omega t) \quad (44)$$

$$T_n i_p = \frac{-3A_n I_p}{2}\cos(2\omega t) \quad (45)$$

$$i_{dc} = \frac{3A_p I_p}{2} + \frac{3A_n I_n}{2} - \frac{3A_p I_n}{2}\cos(2\omega t) - \frac{3A_n I_p}{2}\cos(2\omega t) \quad (46)$$

The cross product (T_p i_n) and (T_n i_p) yield the abnormal second order harmonics components.

$$i_{dc} = I_{dc} + i_{sh} \quad (47)$$

Where

$$I_{dc} = \frac{3A_p I_p}{2} + \frac{3A_n I_n}{2} \quad (48)$$

$$i_{sh} = \frac{-3A_p I_n}{2}\cos(2\omega t) - \frac{3A_n I_p}{2}\cos(2\omega t) \quad (49)$$

$$I_o = i_{dc} - C\frac{dV_o}{dt} \quad (50)$$

$$I_o = I_{dc} + i_{sh} - C\frac{dV_o}{dt} \quad (51)$$

$$V_o = R_o I_o \quad (52)$$

Where V_o and R_o are the output dc load voltage and the load resistance

$$V_o = R_o[I_{dc} + i_{sh} - C\frac{dV_o}{dt}] \quad (53)$$

$$V_o = \underbrace{[R_o I_{dc} - R_o C\frac{dV_o}{dt}]}_{\text{constant (pure dc value } V_{dc})} + \underbrace{R_o i_{sh}}_{\text{second order harmonic}(V_{sh})} \quad (54)$$

$$V_o = V_{dc} + v_{sh} \quad (55)$$

Due to the unbalance in the input voltages, the output dc voltage contains dc term (V_{dc}) and ac term (v_{sh}). For a desired level V_{REF}, a controller is used to regulate the output dc voltage. Due to the existence of the harmonic voltage (v_{sh}) in V_o, the output current of the voltage controller (I_{MAX}) will contain dc term (I) and ac term ($I_c \cos(2\omega t)$) as shown below

$$I_{MAX} = I + I_c \cos(2\omega t) \quad (56)$$

I_{MAX} is used as magnitude for the reference currents sothat forphase1, the reference current is

$$i_{r1} = I_{MAX} \sin(\omega t)$$

$$i_{r1} = I \sin(\omega t) + I_c \cos(2\omega t) \sin(\omega t)$$

$$i_{r1}=I\sin(\omega t)-\frac{I_c}{2}(\sin(\omega t)-\sin(3\omega t) \quad (57)$$

It is implied from this analysis that the unbalance in the input voltages leads to appearance of second order harmonic (100 Hz for the 50 Hz supply) in the output dc voltage (v_{sh}), which causes third order harmonic (150 Hz for the 50 Hz supply) in the input current (see equation (57)). This interaction continues and results in the appearance of even order harmonics at the dc link voltage and odd order harmonics in the input currents.

IV. CONTROL STRATEGY

After knowing the types of the dominant harmonics in the output (second order harmonic) and in the input (third order harmonic), a control strategy is needed to reduce these harmonics. The proposed control strategy is divided into two parts, the first part is a voltage controller and the second part is a current controller. Figure 3 shows the control strategy block diagram.

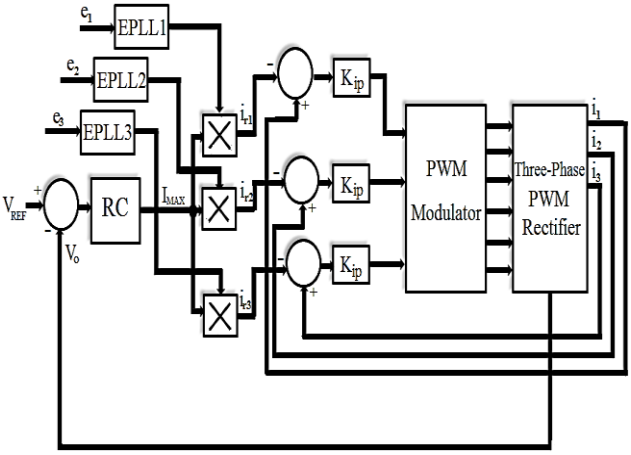

Figure3:Overall control strategy block diagram

A. Voltage Controller

The first objective of this voltage controller is to reduce the second order harmonic in the output dc voltage (less than 5% of V_o) inequation (55), and the second objective is to reduce the second order harmonics in the reflected I_{MAX} inequation (56). This leads to the reduction of the third order harmonics in the reference currents inequation (57), and reduces the total harmonic distortions (THD_s) for these references.To achieve these objectives, there are several control schemes have been proposed based on the stationary frame and the rotating frame methods. However, most of them only consider regulation of the PWM boost type rectifier under slight to medium levels of imbalance. In order to acquire these objectives under extremely unbalanced and distorted input voltages, a Repetitive Controller (RC) is used. According to the internal model principle (IMP), zero error tracking of any reference input, in steady-state, can be accomplished if a generator of the reference input is included in a stable closed-loop system. For example, a type 1 closed-loop system with an integrator ($\frac{1}{S}$) [or $\frac{z}{z-1}$ in discrete time domain], i.e., the generator of unit step function, in the loop offers tracking of a step input with zero steady state error. Repetitive control (RC) is a special case of the internal model principle in control systems with periodic signals [9]. Figure 4 shows the structure of the repetitive controller.

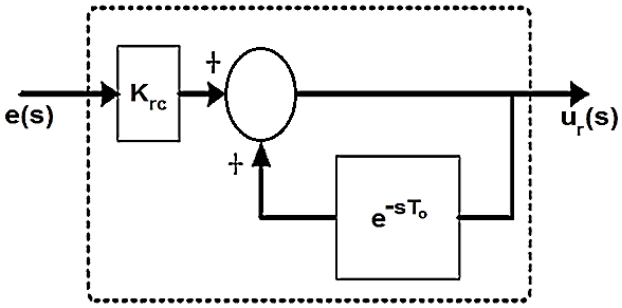

Figure 4: Repetitive controller

The transfer function of the repetitive controller is:

$$G_{rc}(s)=\frac{u_r(s)}{e(s)}=\frac{K_{rc}}{1-e^{-sT_o}} \quad (58)$$

Where K_{rc} is the gain and $T_o=\frac{1}{f_o}$, f_o is the fundamental compensation harmonic frequency.

$$G_{rc}(s) = \frac{K_{rc}}{1-e^{-sT_o}} = K_{rc}\left[\frac{1}{2} - \frac{1}{2} + \frac{1}{1-e^{-sT_o}}\right]$$

$$= K_{rc}\left[\frac{1}{2} + \frac{2-(1-e^{-sT_o})}{2(1-e^{-sT_o})}\right]$$

$$= K_{rc}\left[\frac{1}{2} + \frac{1+e^{-sT_o}}{2(1-e^{-sT_o})}\right]$$

$$= K_{rc}\left[\frac{1}{2} + \frac{1}{2}\left[\frac{1+e^{-sT_o}}{1-e^{-sT_o}}\right]\right] \quad (59)$$

According to the properties of the exponential function [10]

$$\pi\frac{e^{\pi x}+e^{-\pi x}}{e^{\pi x}-e^{-\pi x}} = x\sum_{K=-\infty}^{+\infty}\frac{1}{x^2+K^2} = \frac{1}{x} + \sum_{K=1}^{+\infty}\frac{2x}{x^2+K^2} \quad (60)$$

$$\frac{e^{\pi x}+e^{-\pi x}}{e^{\pi x}-e^{-\pi x}} = \frac{1}{\pi}\left[\frac{1}{x} + \sum_{K=1}^{+\infty}\frac{2x}{x^2+K^2}\right]$$

If $s = x\,\omega_o$, where $\omega_o = \frac{2\pi}{T_o}$, so

$$\left[\frac{1+e^{-sT_o}}{1-e^{-sT_o}}\right] = \left[\frac{1+e^{-x\frac{2\pi T_o}{T_o}}}{1-e^{-x\frac{2\pi T_o}{T_o}}}\right] = \left[\frac{1+e^{-2\pi x}}{1-e^{-2\pi x}}\right].$$

$$2\pi\left[\frac{1+e^{-2\pi x}}{1-e^{-2\pi x}}\right] = \frac{1}{x} + \sum_{K=1}^{+\infty}\frac{2x}{x^2+K^2}$$

$$\left[\frac{1+e^{-2\pi x}}{1-e^{-2\pi x}}\right] = \frac{1}{2\pi}\left[\frac{1}{x} + \frac{2x}{x^2+1} + \frac{2x}{x^2+4} + \frac{2x}{x^2+9} + \cdots\right]$$, each $x = \frac{s}{\omega_o}$ so

$$\left[\frac{1+e^{-sT_o}}{1-e^{-sT_o}}\right] = \frac{1}{T_o}\left[\frac{1}{s} + \frac{2s}{s^2+\omega_o^2} + \frac{2s}{s^2+(2\omega_o)^2} + \frac{2s}{s^2+(3\omega_o)^2} + \cdots\right]$$

$$G_{rc}(s) = K_{rc}\left[\frac{1}{2} + \frac{1}{2T_o}\left[\frac{1}{s} + \frac{2s}{s^2+\omega_o^2} + \frac{2s}{s^2+(2\omega_o)^2} + \frac{2s}{s^2+(3\omega_o)^2} + \cdots\right]\right]$$

$$G_{rc}(s) = K_p + \frac{K_I}{s} + \frac{K_r s}{s^2+\omega_o^2} + \frac{K_r s}{s^2+(2\omega_o)^2} + \frac{K_r s}{s^2+(3\omega_o)^2} + \cdots \quad (61)$$

Where $K_p = \frac{K_{rc}}{2}$, $K_I = \frac{K_{rc}}{2T_o}$, $K_r = \frac{K_{rc}}{T_o}$

Mathematically, RC is equivalent to parallel combination of proportional controller, integral controller, and many resonant controllers, as shown in Figure 5. [11]

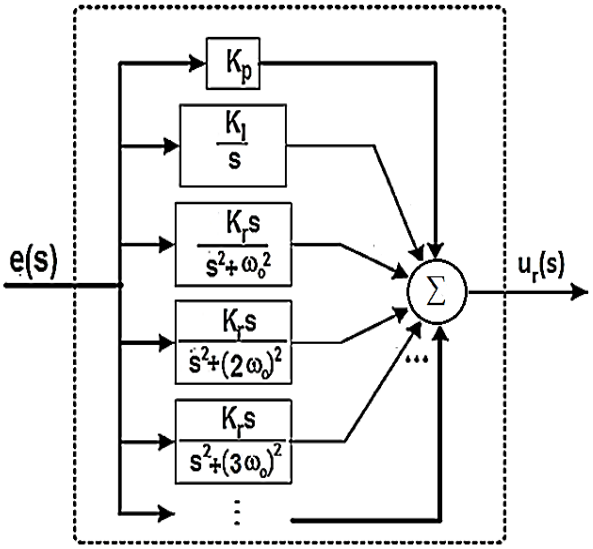

Figure 5: Equivalent form of RC controller

Hence, Figure 5 shows that the transfer function of the repetitive scheme [right-hand side of equation (61)] can be considered as a bank of infinite number of resonant filters, all connected in parallel.

B. Current Controller

The objectives of the current controller are: to construct the reference currents by using Phase Locked Loop (PLL) technique, and to achieve good tracking between the actual and the

reference currents by employing proportional controller only. The components of the current controller will be discussed in details in the following sections.

- **Enhanced Phase Locked Loop (EPLL):**

The EPLL enhances the standard PLL by removing its main drawback that is the presence of double-frequency errors. It achieves this task by means of estimating the amplitude of the input signal. Thus, in addition to removing the ripples, the EPLL provides an estimate of the input signal magnitude and a filtered version of the input signal. This makes the EPLL function as a filter and as a controller too. The block diagram of the EPLL is shown in Figure 6. The EPLL comprises a PLL (shown in the box on the bottom of Figure 6) and also a branch that generates a signal y that is the filtered version of the input signal u. Thus, Y estimates the peak value of the input signal, and ϕ estimates its phase angle. The frequency is estimated at ω. The signal S is unity sinusoidal signal in phase with the input signal, and this represents a stable synchronizing reference. [12]

Assume (u=U sin θ, where θ=ωt) and (y = Ysinϕ). Obviously, when (Y = U and ϕ= θ), the EPLL is in a steady situation, and the error signal e = (u – y) is zero. If this steady situation is stable, then it means that the EPLL approaches the correct solution. For (u = U sin θ) and (y = Y sinϕ), the error signal is (e = u – y) = (U sin θ – Y sin ϕ).The output of phase detector (PD) (the multiplier in PLL) in Figure 6 is equal to

$$z = e\cos\phi = (U\sin\theta - Y\sin\phi)\cos\phi$$

$$z = \frac{U}{2}\sin(\theta-\phi) + \underbrace{\frac{U}{2}\sin(\theta+\phi) - \frac{Y}{2}\sin 2\phi}_{\text{high-frequency}} \qquad (62)$$

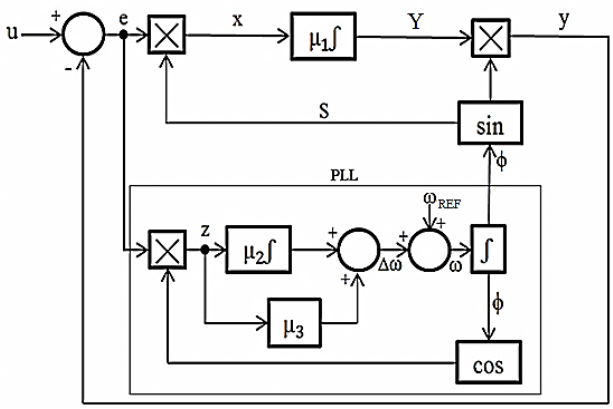

Figure 6: EPLL block diagram

Assuming that the steady situation (i.e., Y = U and ϕ = θ) is stable, the high-frequency term approaches zero as the system approaches to the steady situation. This means that the high- or double-frequency term keeps being removed from the loop and the frequency and phase angle estimations will carry no double-frequency ripple as they approach their steady values. The output of the top multiplier in Figure 6 is equal to

$$x = e\sin\phi = (U\sin\theta - Y\sin\phi)\sin\phi$$

$$x = \frac{U}{2}\cos(\theta-\phi) - \underbrace{\frac{Y}{2}\cos(2\phi) - \frac{U}{2}\cos(\theta+\phi)}_{\text{high-frequency}} \qquad (63)$$

As the system approaches the steady condition, the high frequency term approaches zero. Therefore, there will be no double-frequency ripple on the estimated peak value.The selection of the parameters in Figure 6 (μ1, μ2, and μ3) depends on the following observation:

- Increasing the value of μ_1 will increase the speed of estimating of the magnitude. However, it creates oscillations in the response. There is a tradeoff between speed and accuracy (or smoothness).
- Decreasing μ_1, μ_2, and μ_3 yield an estimation of the peak, and phase which is insensitive/robust to theundesirable variations and noise in the input signal.

Finally, the output of the EPLL(S in Figure 6) will be pure unit sinusoidal and synchronized with the voltage in each phase, therefore the reference currents are:

$$i_{r1} = I_{MAX}\sin(\omega t) \quad (64)$$

$$i_{r2} = I_{MAX}\sin(\omega t - 120) \quad (65)$$

$$i_{r3} = I_{MAX}\sin(\omega t + 120) \quad (66)$$

- **Proportional Controller:**

Theblock diagram of the current control loop shown in Figure 7, gains and time constants associated with various elements of the this block diagram are as follow

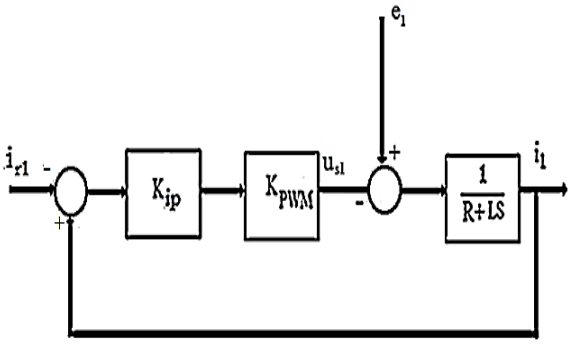

Figure 7: Block diagram of the current control loop.

e_1 source voltage
u_{s1} converter input voltage
K_{ip} gain of the P controller
K_{PWM} gain of the PWM block
T_{RL} time constant of the plant = $\frac{L}{R}$
K_{RL} gain of the plant = $\frac{1}{R}$
i_1 line current
i_{r1} reference current

Forward transfer function = $\frac{K_{ip}K_{PWM}K_{RL}}{1+ST_{RL}}$

The closed loop transfer function is

$$= \frac{\frac{K_{ip}K_{PWM}K_{RL}}{1+ST_{RL}}}{1+\frac{K_{ip}K_{PWM}K_{RL}}{1+ST_{RL}}}$$

$$= \frac{\frac{K_{ip}K_{PWM}K_{RL}}{1+ST_{RL}}}{\frac{1+ST_{RL}+K_{ip}K_{PWM}K_{RL}}{1+ST_{RL}}}$$

$$= \frac{K_{ip}K_{PWM}K_{RL}}{ST_{RL}+K_{ip}K_{PWM}K_{RL}+1}$$

$K_{ip}K_{PWM}K_{RL} >> 1$, so the closed loop transfer function becomes

$$\frac{i_1}{i_{r1}} = \frac{1}{1+T_eS} \quad (67)$$

Where $T_e = \frac{T_{RL}}{K_{ip}K_{PWM}K_{RL}}$

V. SIMULATION RESULTS

The operation of the three-phase PWM boost-type rectifier under severe unbalanced and distorted operating conditions has been simulated in MATLAB Simulink by using SimPowerSystems toolbox. Five different cases have been selected to verify feasibility and performance of the new control method (with RC). The converter operates at near unity power factor with a stable behavior in spite of level of imbalance and distortion of the input conditions. The main electrical parameters of the power circuit and control data are given in Table 1.

Table 1 : Electrical parameters of power circuit and control data

Switching Frequency(F_S)	10000Hz	Output dc voltage V_o	300V
Resistance of reactor(R_L)	0.01 Ω	Nominal input voltage (V)	120V
Inductance of reactors (L)	5 mH	supply frequency	50 Hz
dc-bus capacitor (C)	480 μF	K_{ip}	133
Load resistance R_o	100 Ω	μ_1, μ_2, μ_3	18,0.1,0.1

The cases are:

CASE 1

In this case, the operation of the three-phase PWM boost rectifier is simulated under balanced input voltages condition. Figure 8 shows three-phase input voltages condition of this case, where $Em_1=Em_2=Em_3=120V$. Figure 9 show the steady-state three-phase input currents and Figure 10 shows the output dc voltage when using RC.

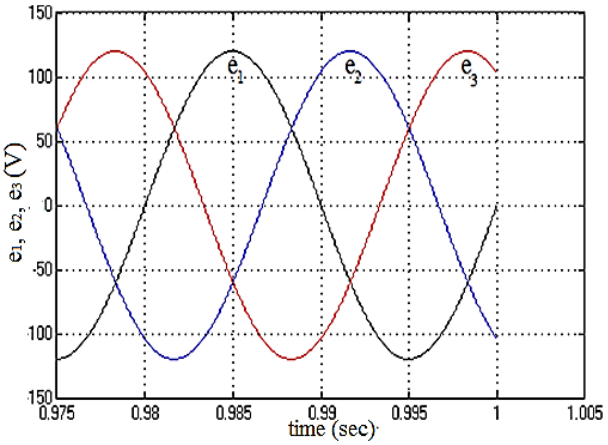

Figure 8: Input voltages of case1

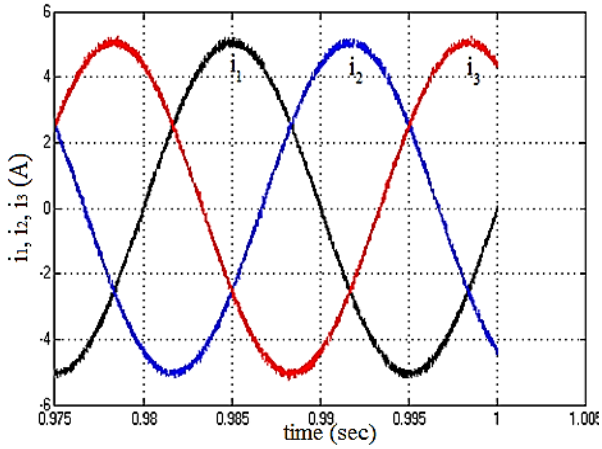

Figure 9: Input line currents when using repetitive controller (RC) in case1

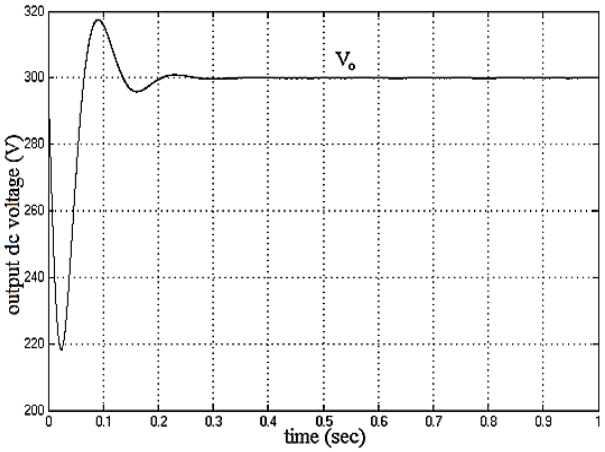

Figure 10: Output dc voltage when using RC in case1

CASE 2

If the input voltages are $Em_1=190V$, $Em_2=120V$, and $Em_3=70V$ as shown in Figure 11, then the input currents obtained when using RC are shown in Figure 12. The negative sequence current in Figure 12 is 0.02705A, and the third order harmonic currents in each phase are 0.01A, 0A, and 0.01A, while the THD_s are 1.96%, 1.98%, and 2.04% respectively.

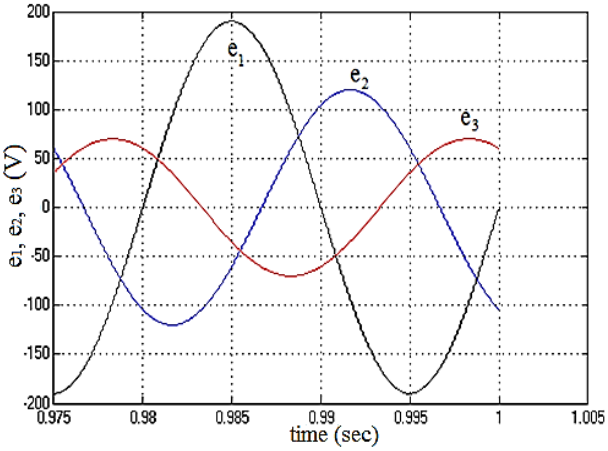

Figure 11: Unbalanced input voltages of case2

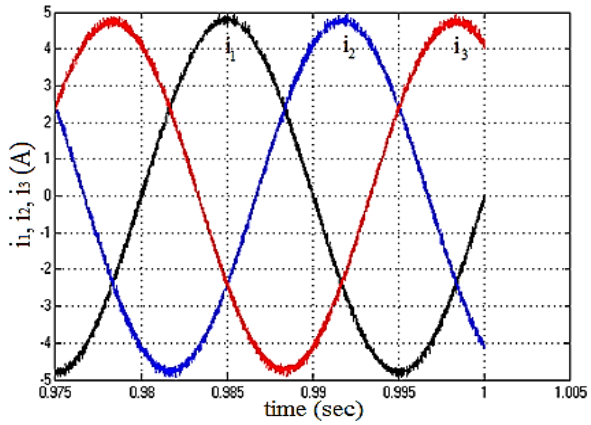

Figure 12: Input currents obtained when using RC in case2

Figure 13 shows the output dc voltage of this case, the second order harmonic voltage is equal to (5.62V peak-to-peak (P.P)) when RC is used.The power factor obtained for this case is 0.9874.

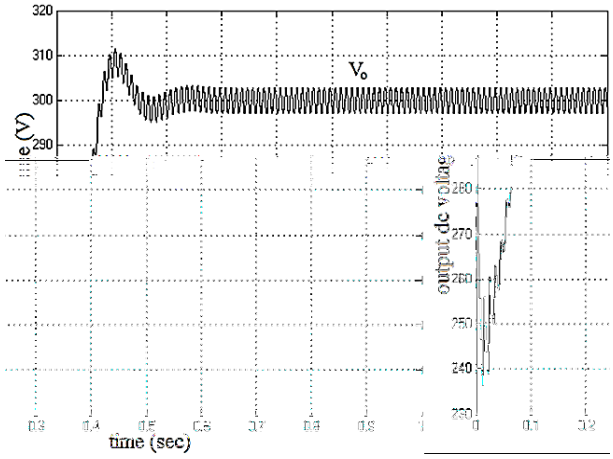

Figure 13: Output dc voltage obtain when using RC in case2

CASE 3:

In this case, a 5th order harmonic of 25% distorts unbalanced input voltages (Em_1=157V, Em_2=120V, and Em_3=85V). Figure 14 shows the input voltages of this case, while the input currents obtained when using RC in this case are shown in Figure 15. Figure 16 shows the output dc voltage.

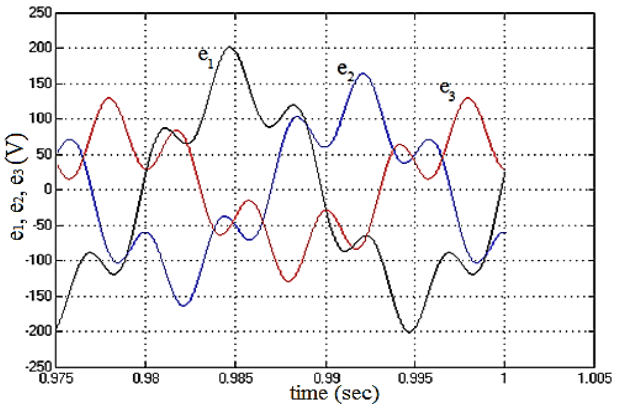

Figure 14: Input voltages of case3

The THD_s for the input currents are 1.94%, 2.03%, and 2.15% respectively, while the 3rd order harmonics are 0A, 0.01A, and 0.01A. The 5th order harmonics in these currents are 0.02A, 0.03A, and 0.04A respectively.The second order harmonic in the output dc voltage is (3.54V P.P). The power factor for this case is 0.912.

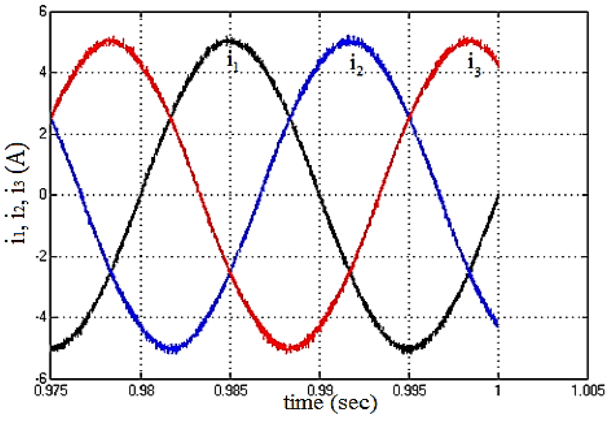

Figure 15: Input line currents obtained when using RC in case3.

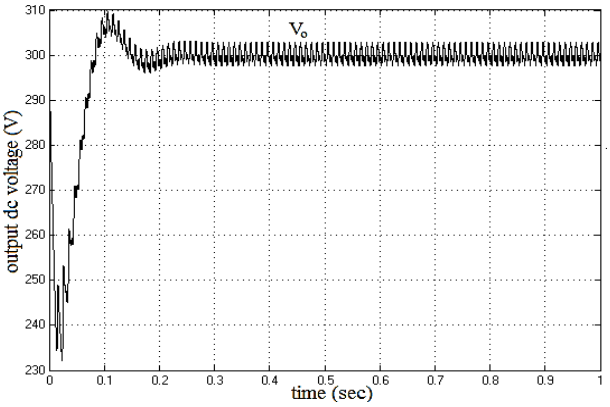

Figure 16: Output dc voltage obtained when using RC in case3.

CASE 4:

In this case, a 7th order harmonic of 25% distorts unbalanced input voltages (Em_1=157V,Em_2=120V,and Em_3=85V). Figure 17 shows the waveforms of the input voltages of this case. Figure 18 shows the input line currents, while Figure 19 shows the output dc voltage.

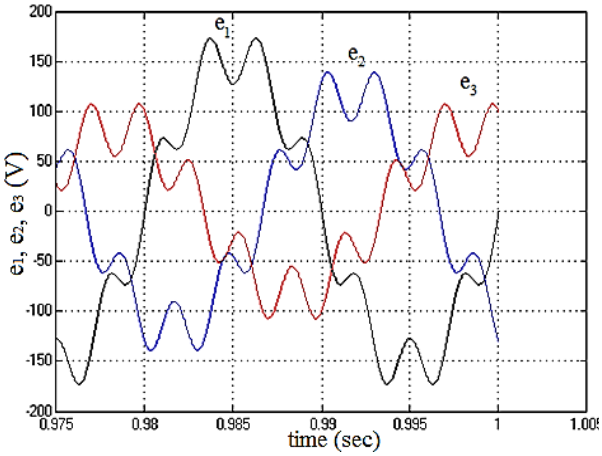

Figure 17: Waveforms of the input voltages of case4

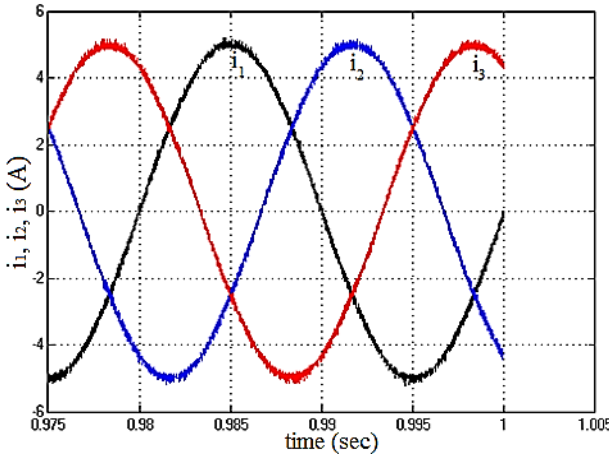

Figure 18: Input currents obtained when using RC in case4

The THD_s for the input currents in Figure 18 are 1.97%, 1.97%, and 2% for the three phases respectively, while the 7th order harmonic is 0.02 A for each phase. The obtained power factor for this case is 0.9573.

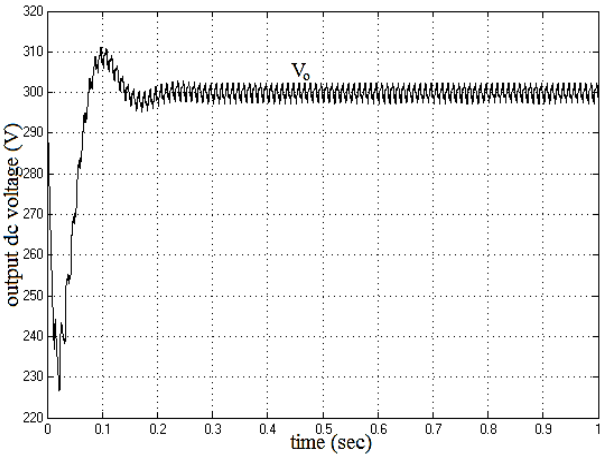

Figure 19: Output dc voltage obtain when using RC in case4

CASE 5:

In this case, unbalanced input voltages Em_1=157V, Em_2=120V, and Em_3=85V are considered, these voltages are distorted by a 5th harmonic of 20% and a 7th harmonic of 20%. Figure 20 shows the waveforms of the input voltages.

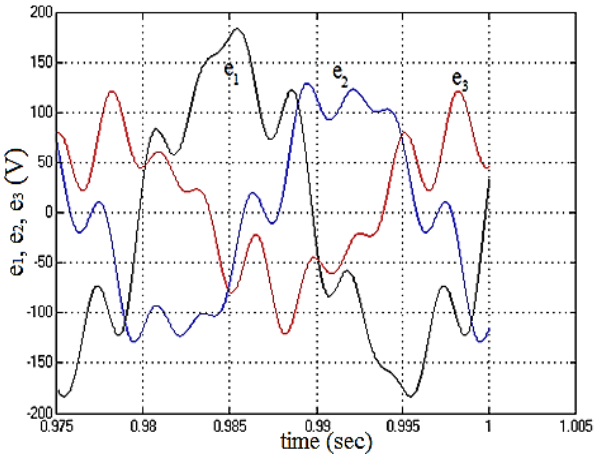

Figure 20: Waveforms of the input voltages of case5

Figure 21 shows the input line currents obtained when using RC. The THD_s for these currents are 1.99%, 1.89%, and 2.07% respectively, while the third order harmonic in each phase current is 0A. Figure 22 shows the output dc voltage when using RC, the second order harmonic is (5V P.P.). The power factor obtained for this case is 0.9036.

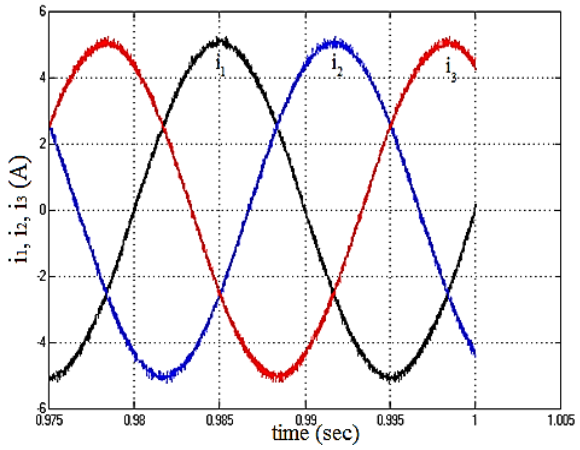

Figure 21: Input line currents obtained when using RC in case 5

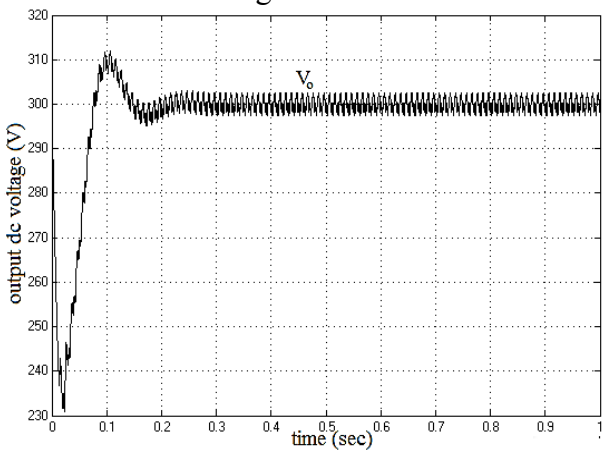

Figure 22: Output dc voltage obtained when using RC in case5

For all cases, it is noticed that the output dc voltage requires 140ms (settling time) to be stable at the beginning of the operation within a variation of 6%. Figure 23 shows the dc link voltage and the ac line current with using RC when the load is changed from 100 Ω to 50 Ω at 0.6 sec and back to 100 Ω at 0.8 sec from starting time.In increasing the load,it is clear that an oscillation in the dc link voltage is occurred within 8% (decrease in the output dc voltage), and it is damped after 85 msec from the instant of increasing the load. The ac line current (increased from 5A to 10A) also reaches to the steady state (stable at 10A) in two cycles, around 40ms. When the load returns to 100 Ω, the oscillation in the output dc voltage is within 9% (increase in the output dc voltage) and this oscillation is damped after 130 msec.

The ac input line current reaches to the steady state (stable at 5A) in six cycles, around 120 ms.

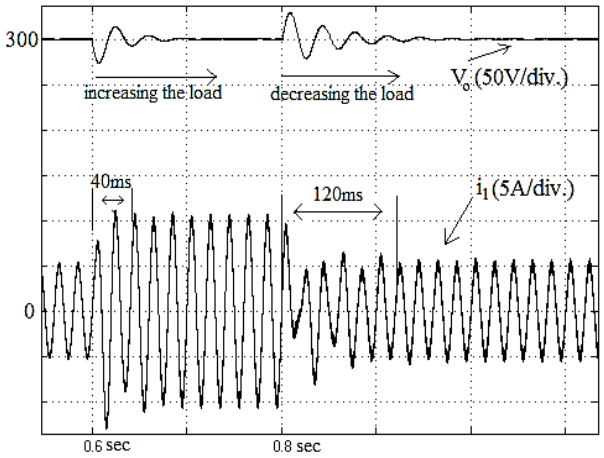

Figure 23: Simulation results of the output dc voltage and ac line current when using RC under changing the load

All the results of the output dc voltage that are mentioned above show excellent voltage regulation at steady state and transient conditions. The output dc voltage tracks the reference dc voltage with small ripple (less than 6V P.P) at steady state. While the oscillation in the output dc voltage is no more than 6% for transient condition (step change in load). Now, the conventional control method (Proportional Integral (PI) controller with a Low Pass Filter (LPF)) is compared with the RC under supplying unbalanced input voltages (case2) and under changing the load. The cutoff frequency of LPF is 50 Hz and the parameters of the PI controller are K_{vp}=1 and K_{vi}=66, where K_{vp} is the proportional gain and K_{vi} is the integral gain of the voltage controller.

Figure 24 shows the dc current I_{MAX} when the input supply voltages are Em_1=190V, Em_2=120V, and Em_3=70V as shown in Figure 11 (case2).

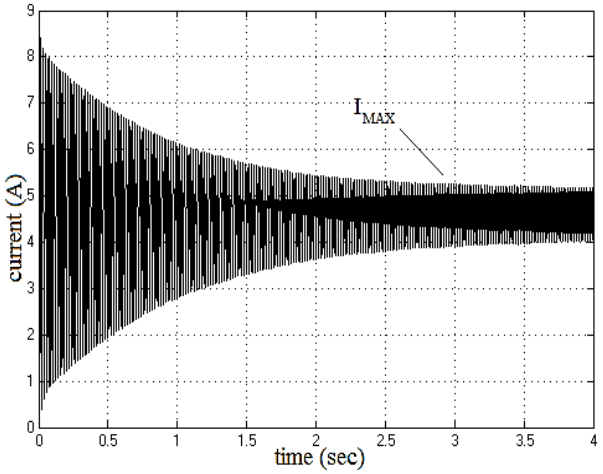

Figure 24: I_{MAX} obtained when using the conventional method

The second order harmonic in this current is 0.5A (1A P.P), this leads for appearing a third order harmonic in the input currents and makes THDs for these currents are higher than 5%. Figure 25 shows the waveforms of the three-phase input currents with using the conventional controlled method and when the input voltages in Figure 11 are used.

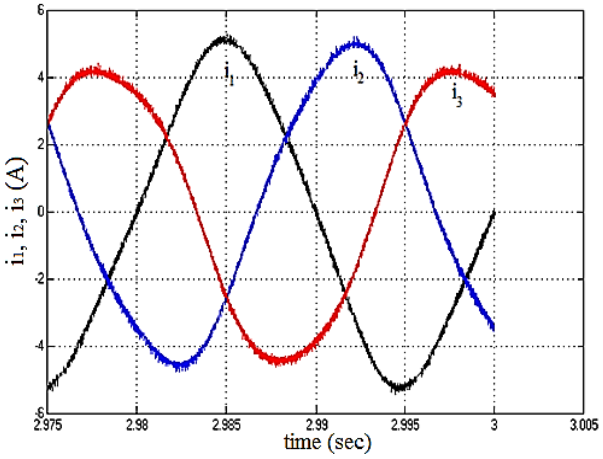

Figure 25: Input currents obtained when using the conventional method and under unbalanced supply voltage level in case2

Figure 26 shows the output dc voltage V_o, the second order harmonic in this voltage is 8.6V (P.P). This Figure shows that the output dc voltage enters the steady state after 2.6 sec from starting time of operation.

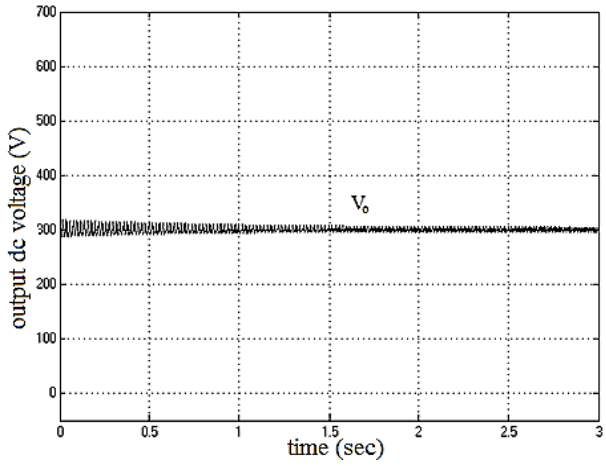

Figure26: Output dc voltage obtained when using the conventional method

Figure 27 shows the ac line current and the output dc voltage when using LPF with PI controller under load change condition. So that when the load is increased at 0.6 sec from starting time, the ac line current requires six cycles (120ms) to reach the steady state. An oscillation occurs at the output dc voltage and this oscillation is damped after 130 msec from the instant of increasing the load. The load returns to 100 Ω at 0.8 sec from starting the operation, this causes an oscillation in the output dc voltage but this oscillation is damped after 160 ms. While the ac line current requires ten cycles to be stable, around 200 ms.

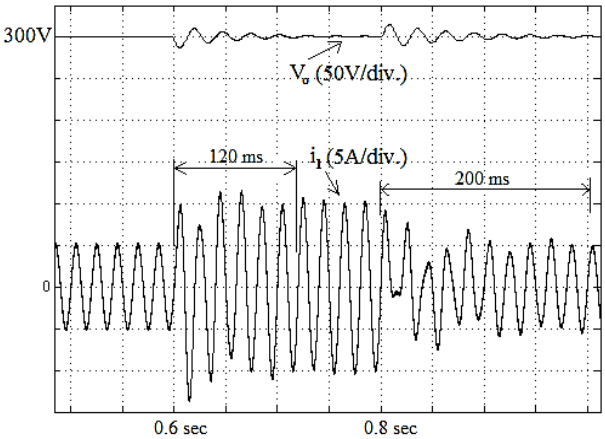

Figure 27: Simulation results of dc voltage and ac line current when using LPF and PI controller under changing the load

Table 2 shows the ripple in the output dc voltage, the negative sequence current, and the third order harmonic in the input line currents when RC and LPF with PI controller are used. Table 3 shows THDs when using the conventional method and Table 4 shows THDs when using the proposed method.

Table 2: Third order harmonics, output ripple, and negative sequence current obtained for both conventional and RC methods when the level voltages in case 2 are used

	RC			Conventional method		
	i_1	i_2	i_3	i_1	i_2	i_3
3rd order(A)	0.01	0	0.01	0.25	0.25	0.25
Ripple(P.P)	5.62 V			8.6 V		
Negative sequence current (A)	0.026			0.271		

Table 3 : THDs obtained when using the conventional method

Conventional method			
	THD% of phase1	THD% of phase2	THD% of phase3
CASE1	1.78	1.75	1.83
CASE2	6	6.13	6.25
CASE3	3.57	2.93	3.27
CASE4	3.64	3.18	3.91
CASE5	4.11	3.58	3.98

Table 4 : THDs obtained when using the proposed method

Proposed method			
	THD% of phase1	THD% of phase2	THD% of phase3
CASE1	1.93	1.93	2.05
CASE2	1.96	2.01	1.98
CASE3	1.94	2.03	2.15
CASE4	1.97	1.97	2
CASE5	1.99	1.89	2.07

It can be seen from these tables that when using the proposed method (RC), the THD_s in the input line currents, the negative sequence currents, and the second order harmonic in the output dc voltage are less, in comparing to their values when using the conventional method (LPF with PI controller).

Figure 28 shows the output dc voltage when using the proposed method. In this Figure, the reference output dc voltage is changed from 300V to 400V at 0.3 sec from starting time and from 400V to 350V at 0.6 sec.

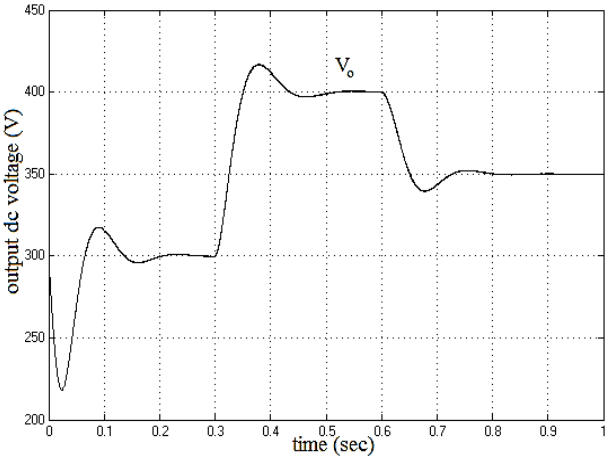

Figure 28: changing the output reference dc voltage when using the proposed method

Figure 29 shows the changing of the reference output dc voltage when using the conventional method.

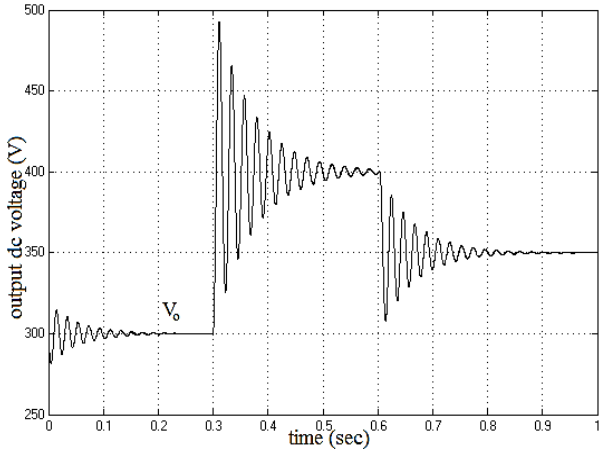

Figure 29: changing the output reference dc voltage when using the conventional method

From Figures 28 and 29, the oscillation in the output dc voltage within 6% when using the proposed method, while 28% when using the conventional method. Therefore, the proposed method produces excellent performance with using small value of the output capacitor and producing excellent input-output harmonic elimination. To improve the performance of the conventional method, the value of the output capacitor must be larger.

Finally, Figure 30 and Figure 31show three-phase input line currents and the output dc voltage with using RC when inductor (3mH) is added to the resistive load which its value is 100Ω under supplying unbalancedinputvoltages in case 2.

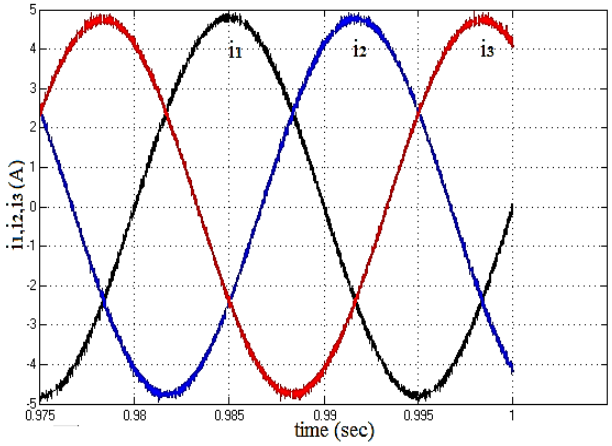

Figure 30: Input currents obtained when using the proposed method and under unbalanced supply voltage level in case2

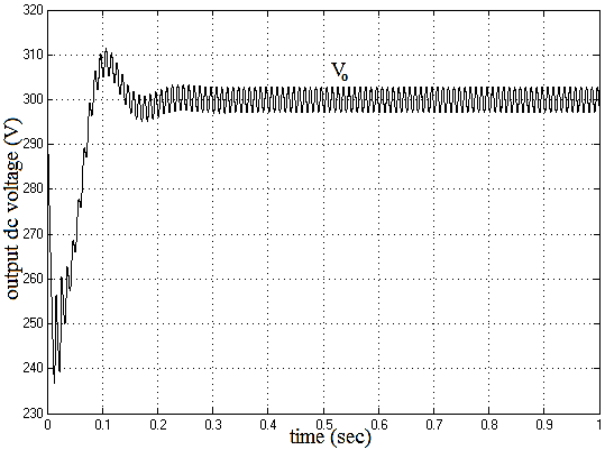

Figure 31: Output dc voltage obtained when using RC

THDs for the input line currents in Figure 30 are 1.99%, 2%, and 2.05%. While the power factor is 0.9915 and the second order harmonic in the output dc voltage in Figure 31 is 5.64V. So the performance of PWM rectifier when using RC is the same with and without adding inductor (3mH) to the resistive load (100Ω).

VI. Conclusions

This paper proposed a new control strategy (RC is used in the voltage controller, and EPLL with proportional controller are used in the current controller) on a three-phase PWM boost rectifier, to achieve input-output harmonic reduction under different input supply voltage conditions. It is concluded from the analysis of the PWM rectifier that when the input supply voltages are unbalanced a second order harmonic appears at the output dc voltage and this result in a third order harmonic in the input currents. The performance of reduction these harmonics for both the proposed method and the conventional method is evaluated. The proposed control method (RC) uses small output capacitor, this makes the system is smaller in size, improves the transient response under the disturbances and decreases the cost of the system. In addition, the proposed method produces less second order harmonic in the reflected

dc current I_{MAX}, this reduces the third order harmonic in the input currents and improves the THD_s (less than 5%), less third order harmonic in the input currents results in reducing the second order harmonic in the output dc voltage. While if the conventional method uses small output capacitor, I_{MAX} has higher second order harmonic and this makes the THD higher than 5% and more harmonics will exist in the output dc voltage. Simulation results in this paper are discussed in the transient response under changing the load condition and in the steady state response under different input voltage conditions. In the steady state response, five different cases were performed to verify the feasibility of PWM rectifier with the proposed control strategy.The advantages of using the proposed method are:

1- Balanced three-phase sinusoidal input currents.
2- THD_s are less than 5% for each phase.
3- The power factor is kept close to unity.
4- The output dc voltage can be regulated at any desired level.

The transient response of the input line current under changing the load when using the proposed control method requires two cycles (40 ms) to be stable when the load is increased from 100 Ω to 50 Ω and requires five cycles (100 ms) when the load back to 100 Ω. While with the conventional method, the ac line current requires six cycles (120 ms) to be stable when the load is increased and ten cycles (200 ms) when the load back to 100 Ω. Finally, the results in the steady state response and in the transient response are better when using the proposed control method.

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