A Nonlinear Control of the Instantaneous Power in dq Synchronous Frame for PWM AC/DC Converter under Generalized Unbalanced Operating Conditions

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Abstract— **This paper proposes a new control scheme for regulating the instantaneous power for pwm ac/dc type rectifiers under generalized unbalanced operating conditions. By nullifying the oscillating components of instantaneous power at the poles of the converter instead of the front-end through solving a set of nonlinear control equations in real time, the harmonics in the output dc voltage can be eliminated more effectively under generalized unbalanced operating conditions on the ac input side. The control scheme allows the pwm rectifier to generate a dc output without substantial even-order harmonics and to maintain nearly unity power factor under generalized unbalanced operating conditions, which makes it possible to reduce the size of the dc-link capacitor and ac inductors leading to the possibility of reduced total cost. Simulation results along with experimental results under the two examples of the unbalanced operating conditions confirm the feasibility of the new control method.**

Keywords— PWM ac/dc converter, generalized unbalanced operating conditions, instantaneous power, nonlinear control

I. INTRODUCTION

The pwm ac/dc converter has been increasingly employed in recent years owing to its advanced features including sinusoidal input current at unity power factor and high quality dc output voltage with a filter capacitor of small size. These features are not necessarily achieved under the operating conditions of unbalanced input supply and unbalanced input impedances. Such a generalized unbalanced operating condition is quite common in power systems, particularly in a weak ac system. Unevenly distributed single-phase loads or nonsymmetrical transformer windings as well as faults could lead to a generalized unbalanced network. It has been shown that unbalanced input voltages or impedances result in the appearance of even harmonics at the dc output and odd harmonics in the input currents [1-3]. Therefore, pwm ac/dc converters under generalized unbalanced operating conditions necessitate the use of input/output filters of large size and thereby completely offset several advantages of the pwm ac/dc converter [2].

Stankovic and Lipo [3] and Enjeti and Choudhury [2] have proposed methods to eliminate the input-output harmonics of the boost and buck type rectifiers, respectively. These methods suffer from two disadvantages: the power factor cannot be adjusted and the methods cannot operate well under the extreme unbalanced operating conditions such as unbalanced two-phase input or single-phase systems that often appear during the faults in power systems. Rioual *et al.* [4] and Song and Nam [5] derived control schemes regulating the instantaneous power at the input of the converter in dq synchronous frame for eliminating harmonics of the boost type rectifier under unbalanced input supply. These approaches also cannot be applied under the extreme unbalanced input supply. Stankovic and Lipo [6] addressed the cases of the generalized unbalanced operating conditions. However, the proposed control algorithm is a feed-forward method in phasor notation that does not regulate the instantaneous power explicitly so that it is not suitable for implementation with the feedback controlled space vector pwm in a dq synchronous frame.

This paper proposes a new control scheme for the pwm ac/dc type rectifier under generalized unbalanced operating conditions. The positive and negative sequence dq components of the input voltages, pole voltages, and currents in the synchronous frame are employed to accurately describe the behavior of the ac/dc converter. The proposed technique nullifies the oscillating components of the instantaneous power at the poles of the converter so that harmonics in the output dc voltage can be eliminated under generalized unbalanced operating conditions in the ac input side. It is necessary to compute a set of nonlinear equations in real time for obtaining the references of the input currents in the synchronous frame in order to regulate the instantaneous power at the poles of the converter. The proposed control scheme allows the pwm rectifier to generate a dc output without substantial even-order harmonics and to maintain nearly unity power factor under generalized unbalanced operating conditions, which makes it possible to reduce the size of the dc-link capacitor and ac inductor.

Simulation and experiments were conducted under two different cases of unbalanced operating conditions. The first case is 15% unbalance in one phase input voltage, which is quite common in a weak ac system. The second unbalance case is a double sinusoidal input with opposite phase and a center tapped neutral input. This second unbalance case is one example of the extreme unbalanced operating conditions. This particular operating condition is also commonly found in residential areas where both 110Vac and 220Vac input voltages are available. Simulation and experimental results confirm the proposed control method.

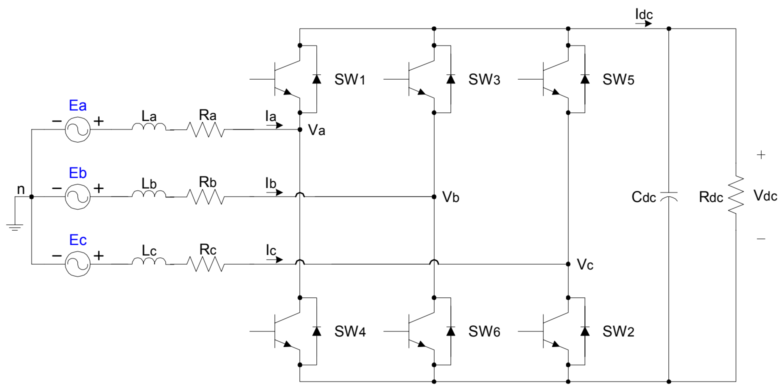

Fig. 1. PWM ac/dc converter under generalized unbalanced operating conditions.

II. Model of PWM AC/DC Rectifier under Generalized Unbalanced Operating Conditions

It is possible to represent an unbalanced three-phase input voltage without a zero sequence as the orthogonal sum of positive and negative sequence. E^p_{dq} and E^n_{dq} are transformed space vectors in the positive and negative synchronous rotating frame respectively. $E^p_a(t)$ is a positive sequence component of $E_a(t)$ while $E^n_a(t)$ is a negative sequence component of $E_a(t)$. $E^p_a(t)$ and $E^n_a(t)$ are summed to represent $E_a(t)$.

$$E_{dqs} = e^{j\omega t} E^p_{dq} + e^{-j\omega t} E^n_{dq} = \frac{2}{3}(E_a + a\, E_b + a^2 E_c) \quad (1)$$

$$E^p_{dq} = \frac{2}{3}(E^p_a + a\, E^p_b + a^2 E^p_c)\, e^{-j\omega t} = E^p_d + j\, E^p_q \quad (2)$$

$$E^n_{dq} = \frac{2}{3}(E^n_a + a\, E^n_b + a^2 E^n_c)\, e^{j\omega t} = E^n_d + j\, E^n_q \quad (3)$$

where: $a = e^{j(2\pi/3)}$.

I^p_{dq}, I^n_{dq}, V^p_{dq}, and V^n_{dq} can be defined in the same manner as in E^p_{dq} and E^n_{dq} . The conventional electrical equations on the ac side of the pwm ac/dc converter are shown in (4) and (5).

$$E^p_{dq} = V^p_{dq} + L\frac{d}{dt} I^p_{dq} + j\omega L\, I^p_{dq} + R\, I^p_{dq} \quad (4)$$

$$E^n_{dq} = V^n_{dq} + L\frac{d}{dt} I^n_{dq} - j\omega L\, I^n_{dq} + R\, I^n_{dq} \quad (5)$$

Arranging the real and imaginary terms of (4) and (5) results in the following four real equations.

$$E^p_d = V^p_d + L\frac{d}{dt} I^p_d - \omega L\, I^p_q + R\, I^p_d \quad (6)$$

$$E^p_q = V^p_q + L\frac{d}{dt} I^p_q + \omega L\, I^p_d + R\, I^p_q \quad (7)$$

$$E^n_d = V^n_d + L\frac{d}{dt} I^n_d + \omega L\, I^n_q + R\, I^n_d \quad (8)$$

$$E^n_q = V^n_q + L\frac{d}{dt} I^n_q - \omega L\, I^n_d + R\, I^n_q \quad (9)$$

The input complex power of the converter is given in (10). At most six real and imaginary terms are found in S_{in} considering only the first harmonics of the input voltage and current.

$$S_{in} = E_{dqs}\, I^*_{dqs}$$
$$= \frac{3}{2} (e^{j\omega t} E^p_{dq} + e^{-j\omega t} E^n_{dq}) (e^{j\omega t} I^p_{dq} + e^{-j\omega t} I^n_{dq})^* \quad (10)$$

$$S_{in} = (P^{in}_o + P^{in}_{c2} cos(2\omega t) + P^{in}_{s2} sin(2\omega t))$$
$$+ j(Q^{in}_o + Q^{in}_{c2} cos(2\omega t) + Q^{in}_{s2} sin(2\omega t)) \quad (11)$$

In order to cancel 120-Hz ripple in the dc output, P^{in}_{s2} and P^{in}_{c2} are set to zero [4-5]. The average reactive power exchanged between the utility source and the converter becomes zero by nullifying Q^{in}_o leading to unity power factor [4-5]. These conditions are incorporated into the following matrix equation (12). The positive and negative dq-components of the currents in the synchronous frame are regulated to the reference values obtained from (13).

$$\frac{2}{3}\begin{bmatrix} P^{in}_o \\ Q^{in}_o \\ P^{in}_{s2} \\ P^{in}_{c2} \end{bmatrix} = \begin{bmatrix} \frac{2}{3} P^{in}_o \\ 0 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} E^p_d & E^p_q & E^n_d & E^n_q \\ E^p_q & -E^p_d & E^n_q & -E^n_d \\ E^n_q & -E^n_d & -E^p_q & E^p_d \\ E^n_d & E^n_q & E^p_d & E^p_q \end{bmatrix} \begin{bmatrix} I^p_d \\ I^p_q \\ I^n_d \\ I^n_q \end{bmatrix} \quad (12)$$

$$[\, I_{dq}\,] = [\, E_{dq}\,]^{-1} [\, S_{in}\,] \quad (13)$$

In a second example case of unbalanced operating conditions introduced in this paper, this approach cannot be applied because the matrix of [E_{dq}] becomes singular. Since,

$$E^p_d = -E^n_d,\ E^p_q = -E^n_q \quad (14)$$

The determinant of $[\,E_{dq}\,] = (E_d^p)^2 + (E_q^p)^2 - (E_d^n)^2 - (E_q^n)^2$

$$= 0 \qquad (15)$$

This singularity comes from the fact that the first and the fourth rows of the matrix $[\,E_{dq}\,]$ are linearly dependent. This physically means that it is impossible to make the oscillating component of the instantaneous input power, P_{c2}^{in}, zero while having nonzero average input power, P_o^{in}.

III. Control Strategy

The complex power in the dq reference frame is conserved not only in the whole but also in each term by term as expected from the orthogonality among average, cosine, and sine terms in real and imaginary parts of the complex power: P_o^{in}, P_{c2}^{in}, P_{s2}^{in}, Q_o^{in}, Q_{c2}^{in}, Q_{s2}^{in}. Therefore, if the terms of P_{s2}^{in} and P_{c2}^{in} are compensated in the inductors then P_{s2}^{out} and P_{c2}^{out} calculated at the three poles of the converter vanish resulting in dc output without 120-Hz ripple. This can also be achieved by regulating P_{s2}^{out} and P_{c2}^{out} to zero directly. Even-order harmonic contents in a dc output voltage are directly caused by oscillatory active power components applied through three-pole ac/dc converters. The output complex power, S_{out}, is described as the following:

$$S_{out} = V_{dqs}\, I_{dqs}^{*}$$

$$= \frac{3}{2}\left(e^{j\omega t}\, V_{dq}^{p} + e^{-j\omega t}\, V_{dq}^{n} \right)\left(e^{j\omega t}\, I_{dq}^{p} + e^{-j\omega t}\, I_{dq}^{n} \right)^{*} \qquad (16)$$

Therefore,

$$Real\,[S_{out}] = P_o^{out} + P_{c2}^{out}\cos(2\omega t) + P_{s2}^{out}\sin(2\omega t) \qquad (17)$$

where:

$$\frac{2}{3} P_{c2}^{out} = V_d^n I_d^p + V_q^n I_q^p + V_d^p I_d^n + V_q^p I_q^n \qquad (18)$$

$$\frac{2}{3} P_{s2}^{out} = V_q^n I_d^p - V_d^n I_q^p - V_q^p I_d^n + V_d^p I_q^n \qquad (19)$$

Under the assumptions of steady state condition and zero input resistances, (6), (7), (8), and (9) can be simplified to (20), (21), (22) and (23), respectively.

$$V_d^p = E_d^p + \omega L\, I_q^p \qquad (20)$$

$$V_q^p = E_q^p - \omega L\, I_d^p \qquad (21)$$

$$V_d^n = E_d^n - \omega L\, I_q^n \qquad (22)$$

$$V_q^n = E_q^n + \omega L\, I_d^n \qquad (23)$$

Applying the above four equations into (18) and (19) yields (24) and (25), respectively.

$$\frac{2}{3} P_{c2}^{out} = E_d^n I_d^p - \omega L\, I_q^n I_d^p + E_q^n I_q^p + \omega L\, I_d^n I_q^p + E_d^p I_d^n + \omega L\, I_q^p I_d^n + E_q^p I_q^n - \omega L\, I_d^p I_q^n \qquad (24)$$

$$\frac{2}{3} P_{s2}^{out} = E_q^n I_d^p + \omega L\, I_d^n I_d^p - E_d^n I_q^p + \omega L\, I_q^n I_q^p - E_q^p I_d^n + \omega L\, I_d^p I_d^n + E_d^p I_q^n + \omega L\, I_q^p I_q^n \qquad (25)$$

In order to cancel 120-Hz ripple in a dc output voltage, P_{c2}^{out} in (24) and P_{s2}^{out} in (25) should be set to zero.

There is growing interest in definitions and interpretations of notions of reactive power in polyphase systems, particularly with nonsinusoidal waveforms and unbalanced networks [7-8]. The definition of average reactive power in (11) appears to have inconsistency with the classical notion of power factor especially in the case of extreme unbalanced operating conditions such as a center tapped split single-phase input supply. The more relevant description of average reactive power in synchronous rotating frame is derived on the basis of the following classical definition of reactive power in sinusoidal steady state condition:

$$Q_o^{in} = E_{arms}\, I_{arms} \sin\theta_{e\text{-}i}^{a} + E_{brms}\, I_{brms} \sin\theta_{e\text{-}i}^{b} + E_{crms}\, I_{crms} \sin\theta_{e\text{-}i}^{c} \qquad (26)$$

where E_{arms} and I_{arms} are the rms values of $E_a(t)$ and $I_a(t)$. $\theta_{e\text{-}i}^{a}$ is the phase angle between $E_a(t)$ and $I_a(t)$. After applying a few trigonometric identities together with the Park Transformation, equation (26) can be simplified to (27).

$$Q_o^{in} = E_q^p I_d^p - E_d^p I_q^p - E_q^n I_d^n + E_d^n I_q^n \qquad (27)$$

The equations of (13) and (27) are linear while the equations of (24) and (25) are nonlinear with respect to the four unknown input current components; I_d^p, I_q^p, I_d^n, and I_q^n. Therefore, the following set of nonlinear equations needs to be solved in real time to obtain the references of input currents.

$$F(X) = \begin{bmatrix} f_1(I_d^p, I_q^p, I_d^n, I_q^n) \\ f_2(I_d^p, I_q^p, I_d^n, I_q^n) \\ f_3(I_d^p, I_q^p, I_d^n, I_q^n) \\ f_4(I_d^p, I_q^p, I_d^n, I_q^n) \end{bmatrix} = 0 \qquad (28)$$

where f_1, f_2, f_3, and f_4 represent (13), (27), (24), and (25) with the appropriate conditions on P_o^{in}, Q_o^{in}, P_{c2}^{out}, and P_{s2}^{out}.

A control block in charge of obtaining the positive and negative dq-components from the unbalanced three-phase variables is considered to be important function block for a regulation of input currents in a positive and negative synchronous frame. Notch filters or low pass filters have been extensively used to obtain the positive and negative dq-components in synchronous frame [5].

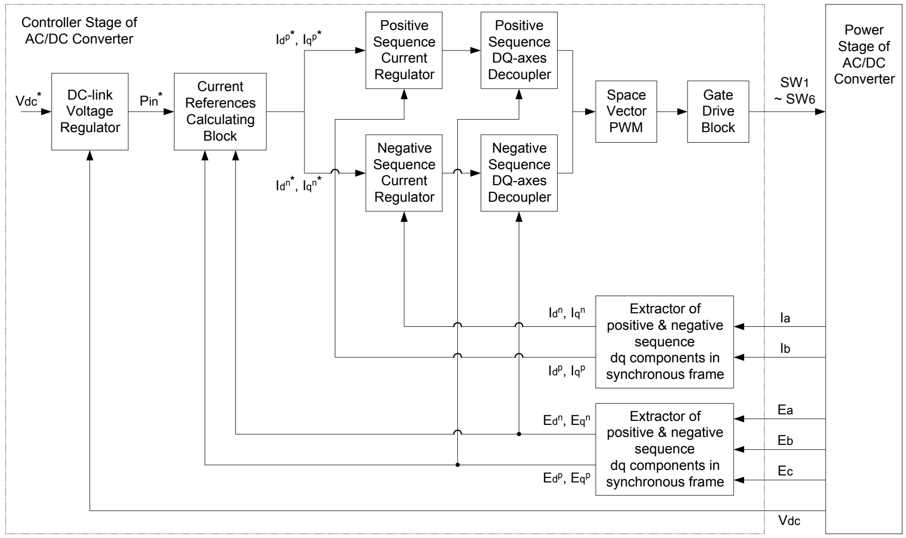

Fig. 2. System block diagram.

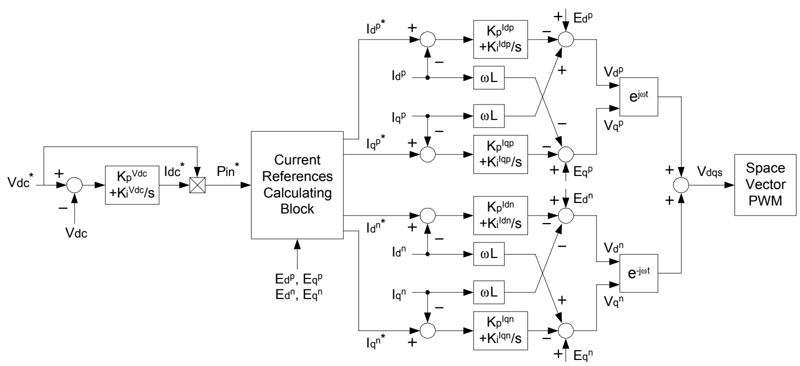

Fig. 3. Detailed control block diagram.

However, the use of these filters typically introduces measurement delay or phase delay leading to inefficient transient characteristics. In this paper, a new algorithm of calculating the positive and negative dq-components of the input voltages in synchronous frames($E_d^p, E_d^n, E_q^p, E_q^n$) from the measured *abc* input voltages($E_a(t)$, $E_b(t)$, $E_c(t)$) is proposed. The principle of this algorithm is explained by the following equations.

$$E_a^p(t) = \frac{1}{3}\left(E_a(t) + E_b(t+\frac{1}{3}T) + E_c(t+\frac{2}{3}T) \right) \tag{29}$$

$$E_a^n(t) = \frac{1}{3}\left(E_a(t) + E_b(t+\frac{2}{3}T) + E_c(t+\frac{1}{3}T) \right) \tag{30}$$

In (29) and (30), T is the fundamental period of input voltage which is 16.67 ms in this paper. The control block that performs this function is named as "*Extractor of positive & negative sequence dq components in synchronous frame*". The detailed block diagram of this control algorithm is shown in Fig. 4. This control block is also used to obtain I_d^p, I_q^p, I_d^n, and I_q^n from the measured input currents($I_a(t)$, $I_b(t)$). The newly proposed algorithm generates the outputs within at most 2/3 of input period time under any type of unbalanced inputs.

IV. Simulation & Experimental Results

The proposed control scheme was simulated using SABER and Simulink. The *Broyden method* was employed to solve the nonlinear control equations. The steady state performance of the pwm controlled rectifier in Fig. 1 with the proposed control circuits in Fig. 2 and Fig. 3 was evaluated using a laboratory prototype pwm rectifier. The complete control was implemented in a DSP processor TI TMS320F240 with both a sampling rate and a switching frequency of 10 kHz.

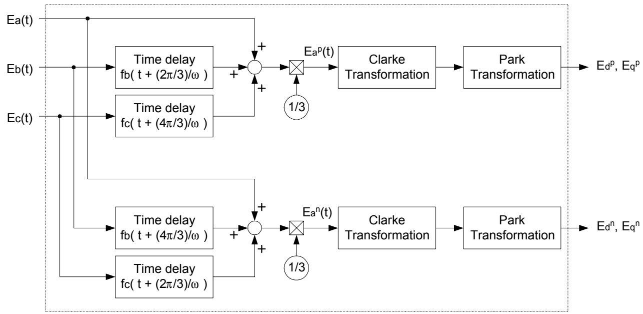

Fig. 4. Extractor of positive & negative sequence dq components in synchronous frame.

To verify the feasibility of the proposed control scheme under the various generalized unbalanced operating conditions, computer simulations and experiments were conducted under two different unbalanced operating conditions.

A. Case 1: 15% unbalance in one phase input voltage

This kind of unbalanced input condition is quite common in a weak ac system. The system parameters of this condition are summarized in Table I.

TABLE I

PARAMETERS USED IN THE SIMULATION & EXPERIMENT FOR CASE 1

Parameter	Value	Parameter	Value
L_a, L_b, L_c	1.6 mH	R_a, R_b, R_c	0.2 Ω
E_a	140 sin(ωt) V	V_{dc}*	400 V
E_b	140 sin(ωt–120°) V	C_{dc}	100 uF
E_c	119 sin(ωt+120°) V	R_{dc}	100 Ω

Fig 5 illustrates the converter output dc voltage, the phase-a input current, and the phase-a input voltage for the conventional single PI-controller in positive sequence only without having any unbalance compensation control scheme. Fig 6 shows the waveforms for the dual controller with the proposed control scheme. Notice the reduction of the output dc voltage ripple in Fig. 6 compared to that of Fig. 5. The detailed frequency spectrum of the output dc voltage and the input current is shown in Fig. 7. The three different frequency spectra(balanced case for the single controller, unbalanced case for the single controller, and unbalanced case for the dual controller) are overlapped to better compare the magnitudes of harmonics. Frequency spectrum of the balanced case for the single controller is employed as an ideal result for the purpose of comparing. As for the frequency spectrum of the output dc voltage, second and forth-order harmonics are significant as expected from the existence of an oscillating component of instantaneous active power at twice input frequency. In contrary, the input current has significant harmonics at third and fifth-order of input frequency. It is noted from Fig. 7 that the dual controller with a proposed control scheme could smooth out the dc output ripple and reduce the odd-order harmonic contents in the input current under the unbalanced input supply.

The experimental waveforms for the single controller are shown in Fig. 8. The experimental results for the dual controller are also shown in Fig. 9. The three different frequency spectra are presented in Fig. 10 for the same three cases as in Fig. 7. THD(Total Harmonic distortion) was obtained from the experimental results for each case and summarized in Table II. The experimental results in Fig. 8, 9, and 10 along with the THD data in Table II confirm the simulation results and verify the fact that the application of the proposed compensation control algorithm leads to a considerable reduction in the harmonic amplitude of the dc output voltage and the input currents.

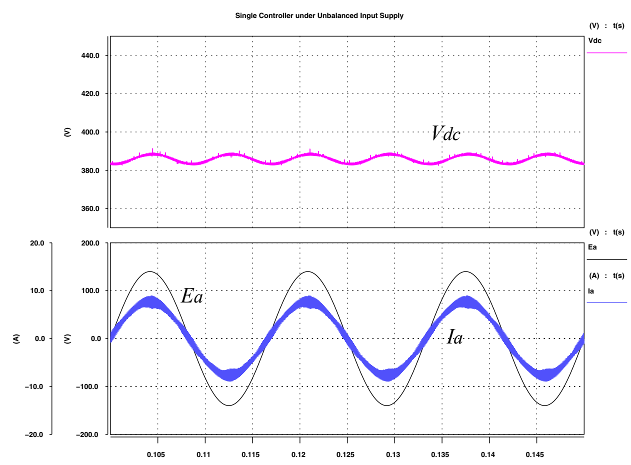

Fig. 5. Simulation waveforms for the single controller under unbalance.

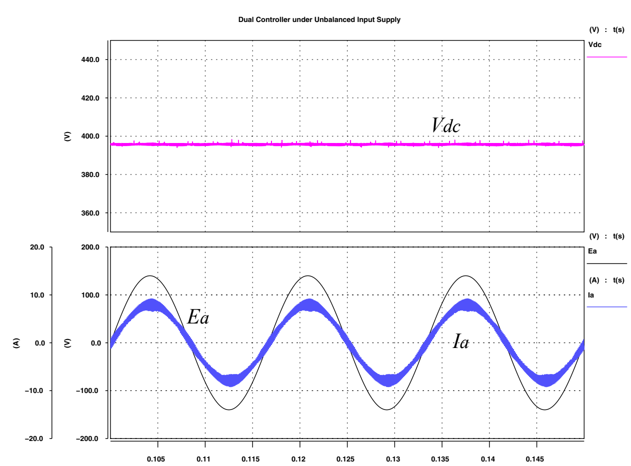

Fig. 6. Simulation waveforms for the dual controller under unbalance.

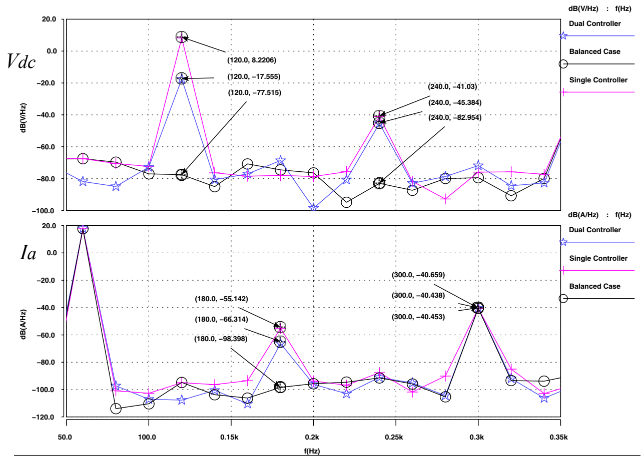

Fig. 7. Frequency spectrum of the output dc voltage and the input current from the simulation results.

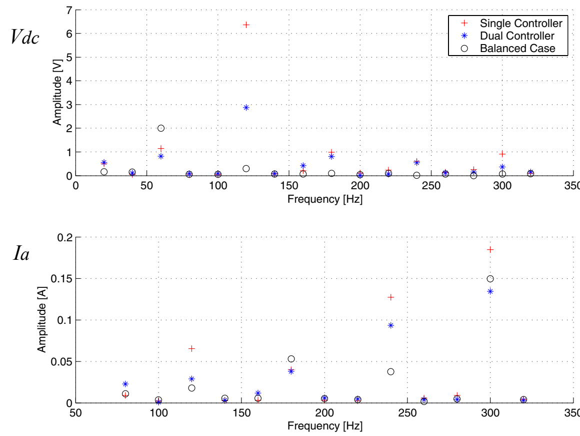

Fig. 10. Frequency spectrum of the output dc voltage and the input current from the experimental results.

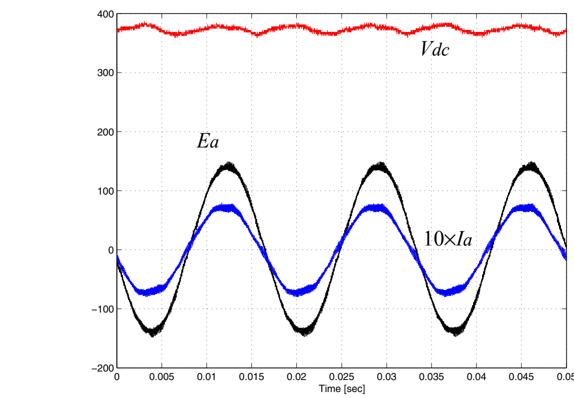

Fig. 8. Experimental waveforms for the single controller under unbalance.

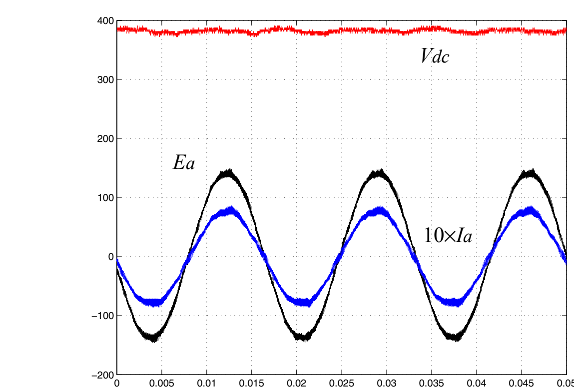

Fig. 9. Experimental waveforms for the dual controller under unbalance.

TABLE II

TOTAL HARMONIC DISTORTION OF EXPERIMENT RESULTS FOR CASE 1

Output dc voltage(V_{dc})		Input ac current(I_a)	
Single controller	1.79 %	Single controller	3.53 %
Dual controller	0.88 %	Dual controller	2.67 %
Balanced Case	0.55 %	Balanced Case	2.44 %

B. Case 2: a double sinusoidal input with opposite phase and a center tapped neutral input

Although this kind of unbalanced operating condition is not as common as that of Case 1, a double sinusoidal input with opposite phase and a center tapped neutral input is considered to be one of most extreme unbalanced operating conditions. Therefore, it is an appropriate example to test the performance capability of a proposed control algorithm under severe unbalanced operating conditions. The system parameters of this condition are summarized in Table III.

TABLE III

PARAMETERS USED IN THE SIMULATION & EXPERIMENT FOR CASE 2

Parameter	Value	Parameter	Value
L_a, L_b	1.9 mH	R_a, R_b	0.2 Ω
L_c	5.4 mH	R_c	0.35 Ω
E_a	144 sin(ωt) V	V_{dc}*	400 V
E_b	144 sin(ωt–180°) V	C_{dc}	470 uF
E_c	0 V	R_{dc}	320 Ω

The active and reactive power components at the front input of the converter, across the inductors, and at the poles of the converter are calculated and shown in Fig. 11, Fig. 12, and Fig. 13, respectively. It is noted from these three figures that the complex power in the dq reference frame is conserved not only in the whole but also in each term by term as expected from the orthogonality among average, cosine, and sine terms in real and imaginary parts of the complex power. The output instantaneous active power, $P^{out}(t)$ in Fig. 13 is almost flat compared to the $P^{in}(t)$ in Fig. 11. This also explains the fact that P^{out}_{s2} and P^{out}_{c2} have to be regulated to zero in order to achieve a constant output instantaneous active power. The input currents are unbalanced, as expected, to cancel the pulsating component of instantaneous active power coming from the input supply.

Experimental results are shown in Fig. 15 verifying the simulation results. E_{in} and I_{in} are the primary side input voltage and current of the center tapped input transformer. As noted from Fig. 14 and 15, E_{in} and I_{in} are nearly in-phase. This means that the input power factor looking at the primary side of the input transformer is also almost unity. This result of almost unity power factor is consistent with the condition of setting the average input reactive power(Q^{in}_o) to zero in (27). Smooth DC bus voltage can be observed despite the severe unbalanced operating conditions. It is noted that V_{dc} is regulated within $\pm 5\ V$ around the nominal output voltage using C_{dc} of 470 uF. These preferred results are accomplished by regulating the positive and negative dq-components of the input currents to the references of the currents calculated in (28). The positive and negative dq-components of the input voltages in the synchronous frame($E^p_d, E^n_d, E^p_q, E^n_q$) can be obtained from the measured *abc* input voltages($E_a(t)$, $E_b(t)$, $E_c(t)$) through the *Extractor of positive & negative sequence dq components in synchronous frame* as shown in Fig. 4.

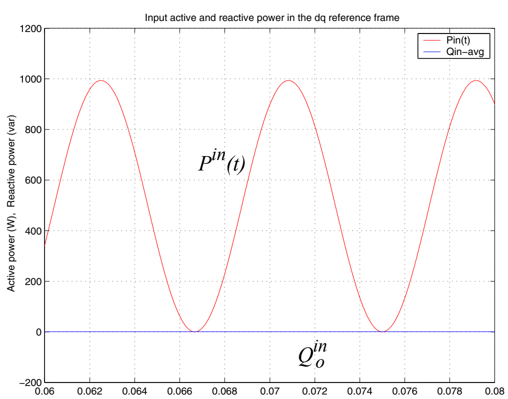

Fig. 11. Input active and reactive power in the dq synchronous reference frame from the simulation results.

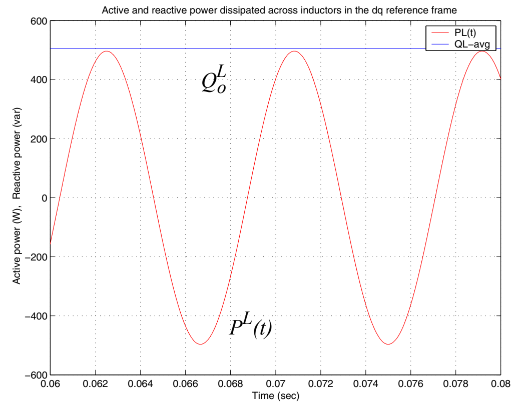

Fig. 12. Active and reactive power dissipated across the inductors in the dq synchronous reference frame from the simulation results.

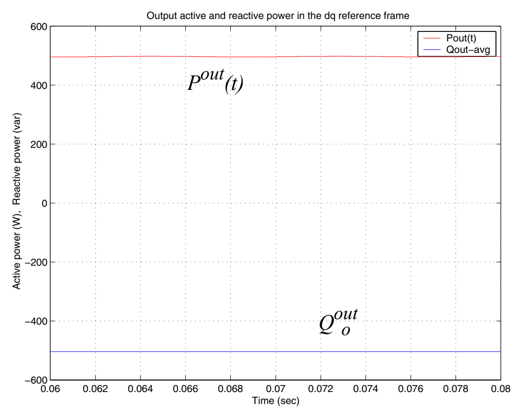

Fig. 13. Output active and reactive power in the dq synchronous reference frame from the simulation results.

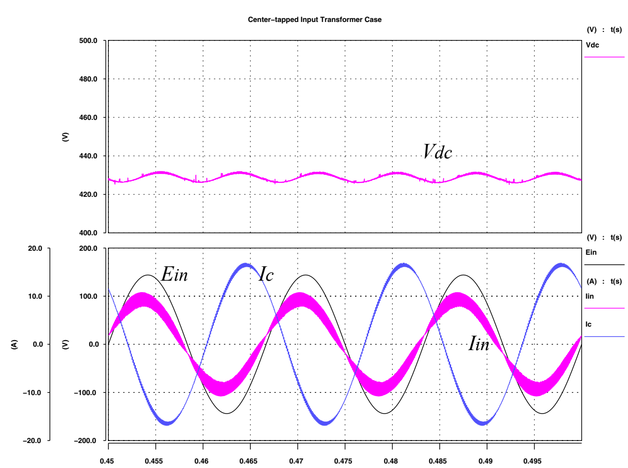

Fig. 14. Simulation waveforms for the dual controller under the input supply from a center tapped transformer.

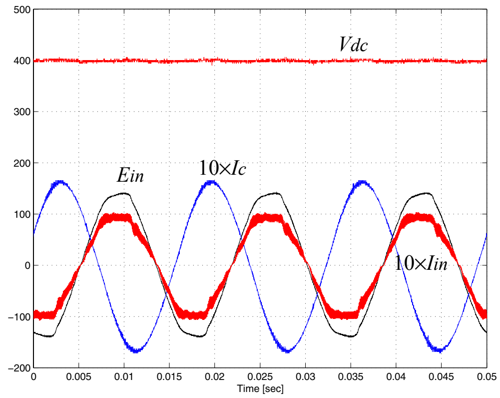

Fig. 15. Experimental waveforms for the dual controller under the input supply from a center tapped transformer.

V. Conclusion

This paper proposes a new control method for a line side connected pwm ac/dc converter operating under generalized unbalanced operating conditions. By nullifying the oscillating components of the instantaneous active power at the poles of the converter instead of the front-end, the harmonics in an output dc voltage can be effectively eliminated even under generalized unbalanced operating conditions in ac input sides. A proposed control scheme needs to solve a set of nonlinear control equations with respect to the input currents in the real time. A nonlinear equation solving algorithm called Broyden method was selected and successfully implemented in the current reference calculating block. An extreme unbalanced operating condition of a transformer sinusoidal secondary voltage with center tapped neutral and the unbalanced impedances was chosen as one of two example conditions for the simulation and the experiment in this paper. The dc output without substantial even-order harmonics and nearly unity power factor in ac side of the rectifier make it possible to reduce the size of the output dc-link capacitor and ac side inductors leading to the possibility of reduced total system cost.

Simulation results along with experimental results using a laboratory prototype converter confirm the proposed control method.

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