소프트 스위칭 방식의 삼상 다이오드 정류기

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Three-Phase Diode Rectifier Employing Soft Switching Methods

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Abstract: Two new schemes of three-phase rectifier using softing switching methods are introduced for the input power factor correction. These three-phase rectifiers are employed the zero voltage switching for the parallel resonant and zero current switching for the series resonant AC link type rectifiers. The dynamic modeling and discontinuous integral cycle mode control technique are also presented. With the proposed circuits and control technique, the high power factor can be obtained.

1. Introduction

In order to obtain a high quality three-phase AC power lines and EMI requirments, it is necessary for many power electronic circuits to build a simple and low-cost three-phase rectifier. Recently, there has been a great deal of researches on the wave shaping of the active line current in single and three phase[1]-[7]. Among them, the PWM type three-phase schemes have been dominated in many power electronic equipments[1]. This conventinal type rectifiers have many problems such as complexity and high cost. As an effective way of overcoming the complexity, Delcho and et al [3] proposed three single-phase AC to DC converter. This topology yields unity power factor and simple control circuitry. However, it exhibits some disadvantages such that it requires complicated input synchronization logic and three-phase balance. To deal with these disadvantages, Ziogas et al [2] proposed the simple three-phase AC to DC converter with discontinuous input current control technique which draws the high quality input current wavefomrs and exhibits none of the above mentioned disadvantages. This type of approach for the line condition, however, has many problems such as high switching losses and high EMI level due to the hard switching of a boost converter [1]-[4]. To minimize these problems, the resonant converter concepts are promising as they eliminate the switching losses to a great extent such that the switching frequency can be increased and the level of EMI is reduced [5]-[7]. In this paper, two new AC link type three-phase rectifier circuits are proposed to achieve the high power factor. The proposed rectifier shown in Fig. 1(a) employ the series resonant AC link to provide the zero current switching(ZCS) condition to all devices and the input phase voltage is always discontinuous. Fig. 1(b) shows the parallel resonant AC link type rectifier to provide the zero voltage switching(ZVS) condition to all devices and the input current is always discontinuous. A dynamic model in the discrte time domain and a discontinuous integral cycle mode control(DICMC) technique for the proposed ZVS parallel resonant rectifier is developed. With the proposed circuits and control technique, the high power factor can be obtained.

2. Principles of Basic Operation

The proposed ZVS three-phase rectifier as shown in Fig. 1(b) consists of three main power conversion stages. The first stage is a three-phase rectifier consisting of an input filter, a boost inductor, and a three-phase diode rectifier. The second stage is a parallel resonant power stage with a full bridge inverter which provides the ZVS condition for all devices and desirable control performance. Finally, the third stage can be modelled as any type of the load requiring a regulated or unregulated DC bus with low frequency output filter. The resonant power stage of the proposed ZVS three-phase rectifier has two basic operational modes, namely energizing and de-energizing mode. For the convenence of the explainaiton, the equivalent single-phase circuit for the basic operational modes and input side equivalent circuit of each mode is shown in fig. 2. These modes are classified according to the direction of power flow and described as follows:

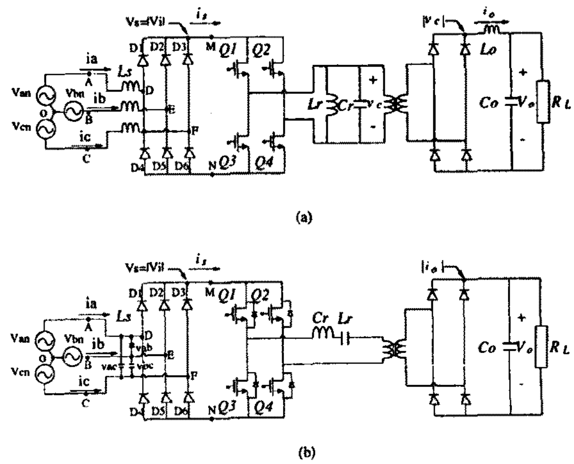


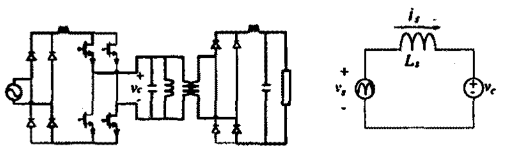
Fig. 1. Circuit Diagram of the Soft Switching Three-Phase Rectifiers.

A. Energizing Mode : The switches, Q1, Q4 and Q2, Q3 pairs are turned on and off alternatively in synchronization with the zero crossing points of the resonant capacitor voltage. The current i_s applied to the resonant tank circuit is in phase with the resonant capacitor voltage. Thus, the resonant tank circuit is energized by the source current i_s and delivers it to the load through the output filter as shown in Fig. 2(a). In other words, the applied voltage across the input inductor is $v_s - |v_c|$, since the reflected resonant capacitor voltage to the input side is always positive during this mode. Thus, the energizing mode can be used to decrease the line current as shown in Fig. 3.

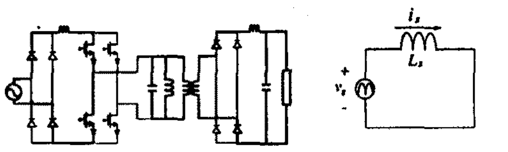
B. De-energizing Mode : De-energizing modes are defined when Q1, Q3 or Q2, Q4 pair is turned on at the zero crossing points of the resonant capacitor voltage. During this mode, the resonant capacitor voltage is de-energized by the output current i_o. Since the reflected resonant capacitor voltage to the input side section is zero, the applied voltage across the input inductor is v_s (Fig. 2(b)). As shown in Fig. 3, the line current is increased like the boosting action in a conventional AC to DC boost converter.

Since the mode changing instants are allowed only at the zero crossing points of the resonant capacitor voltage, interval length of the mode changing should be an interger multiple of half-resonant frequency. The duty cycle of the operational modes is varied for load variation. During the de-energizing mode, all three input AC phases become shorted through inductors L_{smsa}, L_{smsb}, L_{smsc}, the six rectifier diodes and the switches Q1 and Q3 (or Q2 and Q4). Consequently, the three input currents begin simultaneously to increase at a rate proportional to the instantaneous values of their respective phase voltages. Moreover the specific peak current values during each de-energzing mode are proportional to their input phase voltages during the same de-energizing mode.

Furthermore since the input current pulses always begin at zero, it means that their avrage valuea also vary sinusoidally. Consequently, all three input AC currents consist of the fundamental(60Hz) component, a half-resonant frequency component, and an unwanted band centered at around the integer multiple of half-resonant frequency. Since the resonant frequency and its integer multiple frequency can be in the order of several tens or hundreds of kHz, filtering out of the unwanted input current harmonics becomes a relatively easy task.



(a) Energizing Mode



(b) De-energizing Mode

Fig. 2. Equivalent Single-Phase Basic Operational Modes and Input Side Section Equivalent Circuit of Each Mode.

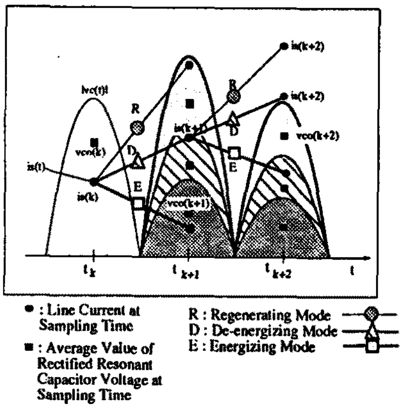


Fig. 3. Typical Waveforms with respect to Basic Operational Modes.

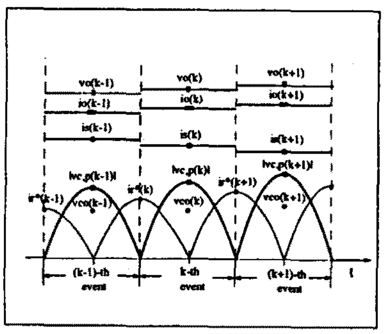


Fig. 4. Definition of the Discrete State Variables.

3. Dynamic Modeling and Discintinuous Integral Cycle Mode Control

In order to obtain a simple and effective analytical tool for the analysis, a dynamic model is developed. The dynamic modeling in a discrete time domain is carried out under the following assumption :

1) All components are ideal.

2) The line current i_s, output current i_o, and output voltage v_o are constant throughout a half cycle operation by the low ripple approximation, because the source inductor L_s and output inductor L_o are sufficiently larger than the resonant tank circuit L_r in the practical case.

3) The turns ratio of the isolation transformer is 1.

4) The line voltage is an ideal sinewave.

Firstly, for the case of continuous input current and phase a,

$$V_{an} = L_s \frac{di_a}{dt} + v_{DN} + v_{NO} \tag{1}$$

when diode D1 is on and diode D4 is off for $0 < \omega t < \pi$, the diode function is

$$d_1 = 1, \qquad \bar{d}_1 = 0 \tag{2}$$

and v_{DN} is reflected resonant capacitor voltage with respect to the diode action and operational mode to the input side section as follows:

$$v_{DN} = d_1 \cdot sgn(v_c) M(k) v_c(t). \tag{3}$$

Thus, (1) becomes

$$L_s \frac{di_a}{dt} = V_{an} - d_1 \cdot sgn(v_c) M(k) v_c(t) - v_{NO}. \tag{4}$$

Similary, for phase b and c:

$$L_s \frac{di_b}{dt} = V_{bn} - d_2 \cdot sgn(v_c) M(k) v_c(t) - v_{NO}. \tag{5}$$

$$L_s \frac{di_c}{dt} = V_{cn} - d_3 \cdot sgn(v_c) M(k) v_c(t) - v_{NO}. \tag{6}$$

For a three-phase system without neutral line,

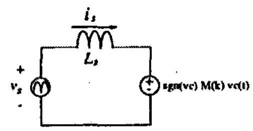
$$i_a + i_b + i_c = 0 \tag{7}$$

$$V_{an} + V_{bn} + V_{cn} = 0 \tag{8}$$

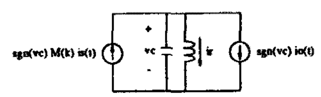
The voltage v_{NO} can be obtained by summing (4) through (6):

$$v_{NO} = -\frac{1}{3} sgn(v_c) M(k) v_c(t) \sum_{k=1}^{3} d_k. \tag{9}$$

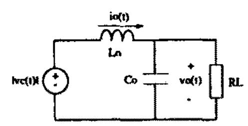
The discrete state variables are defined as shown in Fig. 4. Fig. 5 shows the equivalent circuit of the ZVS three-phase rectifier where the control variable $M(k)$ represents the operational mode of the k-th time event. Based on the basic operational principles previously stated and equivalent circuits, the governing equations for a ZVS three-phase rectifier can be obtained as follows:



(a) Input Side Section



(b) Power Stage Section



(c) Output Side Section

Fig. 5. Equivalent Single-Phase Equivalent Circuit for the ZVS Three-Phase Rectifier.

$$v_{an}(t)=L_s\frac{di_a(t)}{dt}+sgn(v_c(t))D_1\,|\,v_c(t)\,|\,M(k) \tag{10}$$

$$v_{bn}(t)=L_s\frac{di_b(t)}{dt}+sgn(v_c(t))D_2\,|\,v_c(t)\,|\,M(k) \tag{11}$$

$$v_{cn}(t)=L_s\frac{di_c(t)}{dt}+sgn(v_c(t))D_3\,|\,v_c(t)\,|\,M(k) \tag{12}$$

$$M(k)\cdot(d_1\cdot i_a+d_2\cdot i_b+d_3\cdot i_c)=i_r(t)+C_r\frac{dv_c(t)}{dt}+sgn(v_c(t))i_o(t) \tag{13}$$

$$v_c(t)=L_r\frac{di_r(t)}{dt} \tag{14}$$

$$|\,v_c(t)\,|=L_o\frac{di_o(t)}{dt}+v_o(t) \tag{15}$$

$$i_o(t)=C_o\frac{dv_o(t)}{dt}+\frac{1}{R_L}v_o(t) \tag{16}$$

where

$$D_1=d_1-\frac{1}{3}(d_1+d_2+d_3)$$

$$D_2=d_2-\frac{1}{3}(d_1+d_2+d_3)$$

$$D_3=d_3-\frac{1}{3}(d_1+d_2+d_3)$$

Since $L_s, L_o \gg L_r$, the absolute resonant capacitor voltage can be derived from (13) as follows :

$$|\,v_c(t)\,|=Z\cdot\{i_r^*(k)+M(k)\cdot(d_1\cdot i_a+d_2\cdot i_b+d_3\cdot i_c)-i_o(k)\}\cdot\sin\{\omega_r(t-kT/2)\}\quad,\quad for\,\frac{kT}{2}\le t<\frac{(k+1)T}{2} \tag{17}$$

where $Z=\sqrt{L_r/C_r}, T=2\pi\sqrt{L_rC_r}, \omega_r=1/\sqrt{L_rC_r}$, and $i_r^*(k)$ is defined as the absolute value of $i_r(k)$. If a new discrete state variable $v_{co}(k)$ is used to represent the average value of the absolute resonant capacitor voltage during the k-th time event, the following equation can be obtained from (17) as

$$v_{co}(k)\equiv\frac{2}{\pi}\,|\,v_{c,p}(t)\,|,\qquad for\,\frac{kT}{2}\le t<\frac{(k+1)T}{2}$$

$$=\frac{2Z}{\pi}\{i_r^*(k)+M(k)\cdot(d_1\cdot i_a+d_2\cdot i_b+d_3\cdot i_c)-i_o(k)\} \tag{18}$$

where $|\,v_{c,p}(t)\,|$ denotes the absolute peak value of the resonant capacitor voltage. Solving above equations gives the following resultant equations.

$$i_a(k+1)=i_a(k)-\delta_{Ls}M(k)D_1v_{co}(k)+\delta_{Ls}v_{an}(k) \tag{19}$$

$$i_b(k+1)=i_b(k)-\delta_{Ls}M(k)D_2v_{co}(k)+\delta_{Ls}v_{bn}(k) \tag{20}$$

$$i_c(k+1)=i_c(k)-\delta_{Ls}M(k)D_3v_{co}(k)+\delta_{Ls}v_{cn}(k) \tag{21}$$

$$\begin{aligned}v_{co}(k+1)=&\frac{4Z}{\pi}M^*(k+1)(d_1\cdot i_a+d_2\cdot i_b+d_3\cdot i_c)\\&+[1-\gamma_o-\gamma_sM(k+1)M(k)(d_1\cdot D_1+d_2\cdot D_2+d_3\cdot D_3)v_{co}(k)\\&-\frac{4Z}{\pi}i_o(k)+\gamma_ov_o(k)\\&+\gamma_sM(k+1)(d_1\cdot V_{an}+d_2\cdot V_{bn}+d_3\cdot V_{cn})\end{aligned} \tag{22}$$

$$i_o(k+1)=\delta_{Lo}v_{co}(k)+i_o(k)-\delta_{Lo}V_o(k) \tag{23}$$

$$v_o(k+1)=\delta_{Co}i_o(k)+(1-\delta_{RL})v_{co}(k) \tag{24}$$

where $\gamma_s=2L_r/L_s$, $\gamma_o=2L_r/L_o$, $\delta_{Ls}=T/2L_s$, $\delta_{Lo}=T/2L_o$, $\delta_{Co}=T/2C_o$, and $\delta_{RL}=T/2R_LC_o$. For the case of the discontinuous input current , the $i_a=i_b=i_c=0$ and the eqation (22) through (24) is same for the discontinuous case.

The output voltage of the proposed rectifier can be controlled by proper selection of the basic operational modes. Since the mode changing instants are allowed only at the zero crossing points of the resonant capacitor voltage, interval length of the mode changing should be an interger multiple of half-resonant frequency to control the discontinuous input current. In this sense, it is named as an DICMC. The duty cycle of the operational modes is varied for load variation.

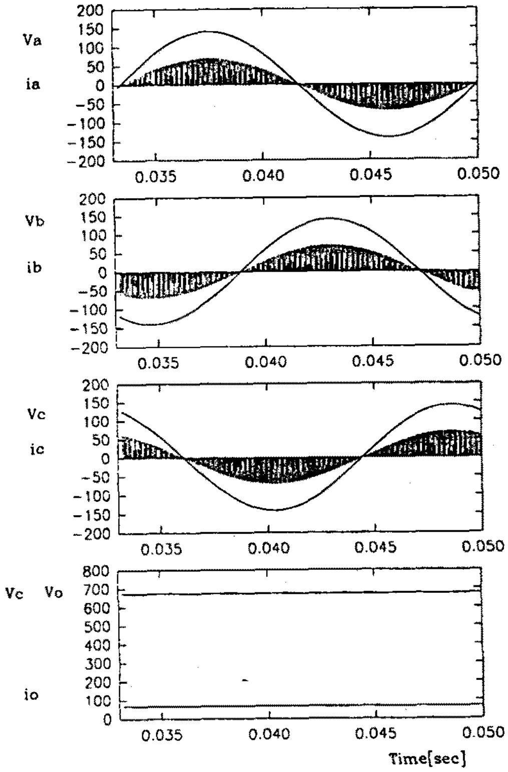


Fig. 6. Computer Simulation Results Using the Proposed Dynamic Modeling.

4. Simulation Results and Conclusions

Two new classes of the soft switched three-phase rectifiers are introduced here which perform the function of high power factor. The zero-voltage or zero-current switching operation of the rectifier allows a substantial reduction in size and weight as compared to the conventional power factor techniques. The dynamic modeling and discontinuous integral cycle mode control technique are also presented. Due to the many advantages including the high efficiency, ZVS(or ZCS), and low EMI, the proposed three-phase rectifiers can be useful for the power factor correction circuit.

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