통신용 전력변환장치를 위한 새로운 영전류 스위칭 방식의 고 역률 정류기

문 건우 , 정 영석 , 김 만고 o , 윤 명중
한국과학기술원 전기 및 전자 공학과 o 한국 통신

A New Zero-Current Switched High Power Factor Rectifier for Power Conversion System for Telecommunication

Gun Woo Moon, Young Seok Jung, Marn Go Kim ° and Myung Joong Youn

Department of Electrical Engeering KAIST *o Korea Telecommunication*

Abstract

A new Zero-Current Switched(ZCS) High Power Factor Rectifier for the power factor correction is proposed. The proposed single phase rectifier enables a zero-current switching operation of all the power devices allowing the circuit to operate at high switching frequencies and high power levels. A dynamic model and a predictive current control technique for the ZCS-High Power Factor Recitfier(HPFR) are proposed. With the proposed dynamic model, an analysis for the internal operational characteristics is explored. With the proposed control technique, the unity power factor.

I. Introduction

In order to comply with the future regulations on frequency distortions of main AC power lines and EMI requirements, it is necessary for many power electronic circuits to improve the waveform quality of the AC to DC conversion. Recently, there has been a great deal of researches on the wave shaping of the active line current and a boost converter is believed to be a preferred power circuit configuration. This type of approach for the line condition, however, has many problems such as the 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 [4]-[7]. However, several papers employing the resonant converter concept use the variable switching frequency to shape a line current [7] and do not concern the discrete time domain modeling[9] of the AC to DC converter. In this paper, a new Zero-Current Switched High Power Factor Rectifier(ZCS-HPFR) is proposed to achieve the unity power factor and wide range of output voltages. The proposed rectifier shown in Fig.1 employs the series resonant AC-link to provide the ZCS condition to all devices. A dynamic model in the discrete time domain for the proposed rectifier is developed and an analysis for the internal characteristics is presented. Using the proposed circuitry, unity power factor and wide output voltage ranges are obtained.

II. Principles of Basic Operation

The basic configuration of ZCS-High Power Factor Rectifier is shown in Fig.1. The proposed rectifier consists of a diode bridge rectifier with the push-pull switches *S1* and *S2*, an isolated resonant power stage, a single switch boost chopper *Q*, and an output filter stage. A transformer with a center tapped primary is inserted to achieve electrical isolation and extends the ranges of the output voltage to lower level of the input voltage according to its turn ratio. The push-pull switches *S1* and *S2* are always turned on and off alternatively in synchronization with the zero crossing points of the resonant current. It is noted that the rectified line voltage V_s to the resonant tank circuit is always in phase with the resonant current. Thus, the resonant power stage is continuously energized by the rectified line voltage V_s. The push-pull switches used to produce a resonant frequency modulated rectified line voltage at the input of the resonant power stage. A typical waveforms of the push-pull switches controlled

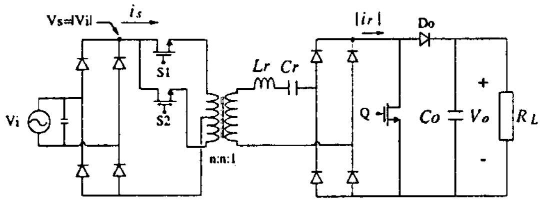


Fig. 1. Circuit Diagram of a Zero-Current Switched High Power Factor Rectifier.

rectified line voltage, current and a switching strategy for the push-pull switches are shown in Fig. 2. As can be seen in this figure, a switch *S1* is closed(*S2* is open) during the resonant half cycle, while the control signal for *S2* is high(*S1* is open) during the next resonant half cycle. With this kind of a switching strategy, the size of isolation transformer can be effectively reduced[10]. With the proposed switching strategy for the push-pull switches, the operation of the ZCS-High Power Factor Rectifier for a unity power factor can be obtained using a single switch boost chopper, *Q*, where the switching is also occurred at zero crossing point of the resonant current. Using the previously stated switching strategy of *S1* and *S2* , the amplitude of the resonant current can be rapidly increased with a maximum slope when a switch *Q* is closed. If *Q* is open, the output stage is connected in series with the resonant tank circuit and hence the stored energy in the resonant power stage is transferred to the output stage. Since the resonant power stage is always energized by the switching strategy for the push-pull switches as that of the hard-switched boost converter, a boost characteristics of a ZCS-HPFR can be obtained. A characteristic waveforms with respect to switching status of a boost chopper *Q* is shown in Fig. 3.

III. Dynamic Modeling and Anaysis

In order to obtain a simple and effective analytical tool for the analysis, a dynamic model in discrete time domain is developed. The discrete state variables are defined as in Fig. 4. Fig. 5 shows the equivalent circuit of the ZCS-HPFR. To obtain a dynamic model for a ZCS-HPFR, the control variable for the *k*-th time event representing the status of the switch, *Q*, is defined as *S(k)*. Based on the basic operational principles previously stated, the governing equations for the *k*-th time event can be derived as follows:

$$\frac{1}{n}sgn(i_r(t))v_s(t) = L_r\frac{di_r(t)}{dt} + v_c(t) + (1 - S(k))sgn(i_r(t))v_o(t) \quad (1a)$$

$$C_r\frac{dv_c(t)}{dt} = i_r(t) \quad (1b)$$

$$(1 - S(k))\,|\,i_r(t)\,| = C_o\frac{dv_o(t)}{dt} + \frac{1}{R_L}v_o(t) \quad (1c)$$

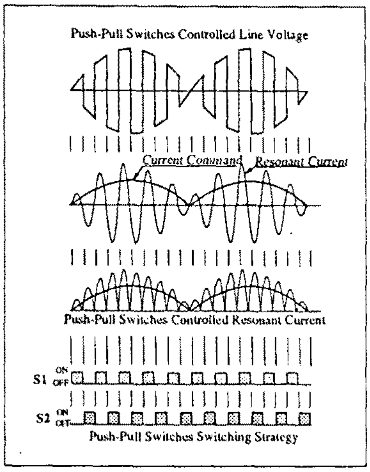


Fig. 2. Switching Strategy for Push-Pull Switches

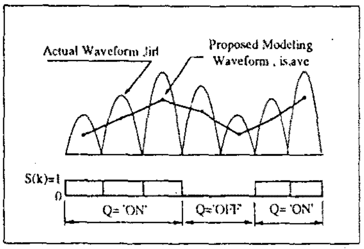


Fig. 3. Characteristic Waveforms with respect to Switching Status of Q

where $S(k)$ has the values 1 and 0 expressing the on and off status of the switch Q, respectively. Since the push-pull switches $S1$ and $S2$ are always turned on and off alternatively, the left hand side of (1a) implies the actual applied voltage to the resonant tank circuit with respect to the turns ratio of an isolation transformer and the direction of the resonant current. The third term of the right-hand side of (1a), $(1-S(k))sgn(i_r(t))v_o(t)$, denotes the output voltage reflected to the resonant tank circuit with respect to the control variable. Equation (1b) implies the dynamic relation between the resonant capacitor voltage and inductor current, and (1c) express the output equation. Because the line voltage and output voltage during the half resonant interval can be assumed as a constant by the low ripple approximation [8], the absolute resonant current can be derived from (1a) and (1b) as follows:

$$|i_r(t)| = \frac{v_c^*(k)+v_s(k)-(1-S(k))v_o(k)}{Z}\sin\{\omega_r(t-kT/2)\} \quad ,$$

$$for\ \frac{kT}{2} =< t < \frac{(k+1)T}{2} \qquad (2)$$

where $Z=\sqrt{L_r/C_r}$, $T=2\pi\sqrt{L_rC_r}$, $\omega_r=1/\sqrt{L_rC_r}$, and $v_c^*(k)$ is defined as the absolute value of $v_c(k)$. From the governing equations (1a)-(1c) and (2), the following equation can also be easily obtained :

$$v_c^*(k+1) = \frac{1}{C_rZ}\int_{kT/2}^{(k+1)T/2}\{v_c^*(k)+v_s(k)-(1-S(k))v_o(k)\}\cdot$$

$$\sin\left\{\omega_r\left(t-\frac{kT}{2}\right)\right\}dt - v_c^*(k) \qquad (3)$$

$$v_o(k+1) = \frac{1}{C_oZ}\int_{kT/2}^{(k+1)T/2}\{v_c^*(k)+v_s(k)-(1-S(k))v_o(k)\}$$

$$\sin\left\{\omega_r\left(t-\frac{kT}{2}\right)\right\}dt + v_o(k) - \frac{1}{C_o}\int_{kT/2}^{(k+1)T/2}\frac{v_o(k)}{R_L}dt. \qquad (4)$$

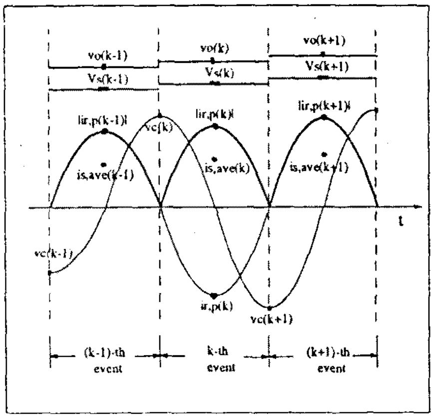
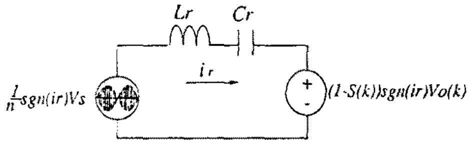
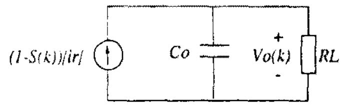


Fig. 4. Definitions of the Discrete State Variables



(a) Power Stage Section



(b) Output Side Section

Fig. 5. Equivalent Circuits of the ZCS-High Power Factor Rectifier

If a new discrete state variable is defined as $i_{s,ave}(k)$ representing the average value of the absolute resonant current during the k-th time event, the following equation can be obtained from (2) as

$$i_{s,ave}(k) \equiv \frac{2}{\pi}|i_{r,p}(t)|\,, \qquad for\ \frac{kT}{2} =< t < \frac{(k+1)T}{2}$$

$$= \frac{2}{\pi}\frac{v_c^*(k)+v_s(k)-(1-S(k))v_o(k)}{Z} \qquad (5)$$

where $|i_{r,p}(t)|$ denotes the absolute peak value of the resonant current. Solving (3) and (4) gives

$$v_c^*(k+1) = v_c^*(k)+2v_s(k)-2(1-S(k))v_o(k) \qquad (6)$$

$$v_o(k+1) = \gamma(1-S(k))v_c^*(k)+v_s(k)+(1-S(k))v_o(k)$$

$$+v_o(k)-\gamma^* v_o(k) \qquad (7)$$

where $\gamma=2C_r/C_o$ and $\gamma^*=\pi Z\gamma/(2R_L)$. The average value of the absolute resonant current for the $(k+1)$-th time event can also be obtained by simply replacing the time index k of (5) by $(k+1)$ as

$$i_{s,ave}(k+1) = \frac{2}{\pi}\frac{v_c^*(k+1)+v_s(k+1)-(1-S(k+1))v_o(k+1)}{Z}. \qquad (8)$$

Using (5) through (8), the following discrete state equation for ZCS-HPFR can be derived as

$$\begin{pmatrix} i_{s,ave}(k+1) \\ v_o(k+1) \end{pmatrix} = \begin{pmatrix} 1-\gamma(1-S(k))(1-S(k+1)) & -\frac{2n}{\pi Z}(1-S(k))+(1-\gamma^*)(1-S(k+1)) \\ \frac{\pi Z\gamma}{2n}(1-S(k)) & 1-\gamma^* \end{pmatrix}$$

$$\begin{pmatrix} i_{s,ave}(k) \\ v_o(k) \end{pmatrix} + \begin{pmatrix} \frac{4v_s(k)}{\pi Z} \\ 0 \end{pmatrix}. \quad (9)$$

Since C_r is chosen as much smaller than C_o, γ and γ^* become much smaller than the unity. Therefore, the equation (9) can be simplified as follows:

$$\begin{pmatrix} i_{s,ave}(k+1) \\ v_o(k+1) \end{pmatrix} = \begin{pmatrix} 1 & -\frac{4n}{\pi Z}(1-S^*(k+1)) \\ \frac{\pi Z \gamma}{2n}(1-S(k)) & 1-\gamma^* \end{pmatrix} \begin{pmatrix} i_{s,ave}(k) \\ v_o(k) \end{pmatrix}$$

$$+ \begin{pmatrix} \frac{4v_s(k)}{\pi Z} \\ 0 \end{pmatrix} \quad (10)$$

where $S^*(k+1) = \{S(k)+S(k+1)\}/2$. Note that the average values of the line current $i_{s,ave}(k+1)$ are directly controlled by $S^*(k+1)$. The slope of the average absolute resonant current between the k-th and $(k+1)$-th time events, $S(k,k+1)$, can be easily derived from (10) as follow:

$$S(k,k+1) = \frac{i_{s,ave}(k+1) - i_{s,ave}(k)}{(T/2)}$$

$$= \frac{8}{\pi Z T}\{v_s(k) - (1 - S^*(k+1))v_o(k)\}$$

$$= \frac{1}{L_{eq}}\{v_s(k) - (1 - S^*(k+1))v_o(k)\},$$

$$L_{eq} = \left(\frac{\pi}{2}\right)^2 L_r. \quad (11)$$

IV. Simulation and Conclusions

In this section, the typical controlled responses of the proposed circuitry are simulated. The parameters used in simulation are as follows:

$$v_s(t) = |\sqrt{2}\,100\sin(120\pi t)|, \quad L_r = 100muH, \quad C_o = 500\mu F,$$

$$R_L = 50\Omega, \quad n = 3, \quad f_r = 100kHz, \quad I_{COM} = v_s(t)/3.$$

Thus, the proposed modeling and predictive current control technique are very useful for the analysis and performance improvement of a ZCS-High Power Factor Rectifier. Fig. 6 shows the simulated waveforms of the line voltage and line current and output voltage for the proposed circuit. It can be said that the line current waveforms show the sinusoidal waveforms keeping in phase with the line voltage. By using the proposed high power factor rectifier, greatly reduced line current harmonics and unity power factor can be obtained. Thus the proposed circuit is very useful for the power factor correction of the power supply in the telecommunication equipments.

Acknowledgement : The author should like to acknowledge the financial support from Korea Telecom Research Center.

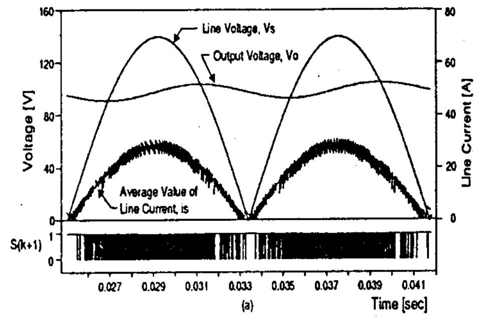


Fig.6 Simulation Results for the Bang Bang Current Control Techniques

References

[1] S.Manias,P.D.Ziogas,"An SMR Topology with Suppressed DC Link Components and Predictive Line Current Wave Shaping," IEEE Industry Application Society Annual Meeting,1986,pp.630-639.

[2] Kwang-Hwa Liu,Yung-Lin L, " Current Waveform Distortion in Power Factor Correction Circuits Employing Discontinous-Mode Boost Converters," IEEE Power Elect. Spec.Conf.,1989, pp. 825-829.

[3] M. Kazerani, P. D. Ziogas, and G. Joos," A Novel Current Waveshaping Technique for Solid-State Input Power Factor Conditioners," IEEE Trans. Ind. Elect., vol. 38, no. 1, pp. 72-78, Feb. 1991.

[4] L. Dixion," High Power Factor Switching Preregulator Design Optimization," Unitrode Switching Regulated Power Supply Design Seminar Manual, 1990.

[5] J.Sebastian,J.Uceda,J.A.Cobos,P.Gil, " Using Zero-Current-Switched Quasi- resonant Converters as Power Factor Preregulator," IEEE Industry Elect. Conf., 1991, pp.225-230.

[6] R.W.D.Doncker, G.Venkataramanan," A New Single Phase AC to DC Zero Voltage Soft Switching Converter," IEEE Power Elect.Spec.Conf.,1990, pp. 206-212.

[7] M. H. Kheraluwala, R. L. Steigerwald, and R. Gurumoorthy," Performance Characterization of a High Power Factor Power Supply with a Single Power Stage," in Conf. Rec. 1992. Annu. Meet. IEEE Ind. Appl. Soc., pp. 659-665.

[8] R. J. King and T. A. Stuart," A Large-Signal Dynamic Simulation for the Series Resonant Converter," IEEE Trans. Aerosp. Electron. Syst. vol. AES-19, no. 6, pp. 859-870, 1983.

[9] J. H. Ko, S. S. Hong, T. Y. Ann, and M. J. Youn," Dynamic Modeling and Current Control Technique for Quantum Series Resonant Converter with Non-periodic Integral Cycle Mode," Int. J. Electron., 1991, pp. 885-897.

[10] S. G. Trabert and R. W. Erickson," Steady State Anaysis of the Duty Cycle Controlled Series Resonant Converter," IEEE Power Electron. Spec. Conf. Rec., pp. 545-546, 1987.