'93 한국자동제어학술회의논문집 (1993. 10. 20~22)

영전류 스위칭 방식의 직렬 공진형 AC/DC 컨버터를 위한 전환모드 이산 슬라이딩 제어

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Switched Discrete Sliding Mode Control for ZCS Series Rosonant AC to DC Converter

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Abstract: A buck-boost zero current switched(ZCS) series resonant AC to DC converter for the DC output voltage regulation together with high power factor is proposed. The proposed single phase AC to DC converter 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 for this AC to DC converter is developed and an analysis for the internal operational characteristics is explored. Based on this analysis, a switched discrete sliding mode control(SDSMC) technique is investigated and its advantages over the other types of current control techniques are discussed. With the proposed control technique, the unity power factor without a current overshoot and a wide range of output voltage can be obtained.

1 Introduction

In order to obtain high quality AC power lines and EMI requirements, it is necessary for many power electronic equipments to improve the waveform quality of the AC to DC conversion. Recently, there has been a great deal of research on the wave shaping of the active line current by way of the hard switching. 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 [1-4]. To minimize these problems, the resonant converter concepts are promising, as they eliminate the switching losses to such a great extent that the switching frequency can be increased and the level of EMI reduced [5-8]. Among the afore mentioned approaches for the power factor correction and output voltage regulation, the boost type is the most popular configuration [8].

The boost converter has the advantage of a continuous inductor current in AC line that can be controlled to get a unity power factor. However, there exists an inherent disadvantage like the uncontrolled range exists when the output voltage is below the source voltage. Therefore this results in the load overcurrent as well as start-up inrush surge. Ultimately, the inductor and the isolation transformer will sturate and components will fail. Thus much higher ratings of the switching device are required for safe operation. To overcome this disadvantage, the buck operation should be employed for the improved control performance and system design optimization in the start-up transient periods and fault load condition. The buck converter has the characteristics of degrading the input power factor because the input current is chopped [9]. Although the buck converter has a disadvantage, it can limit load overcurrent and start-up inrush current. Thus, the buck operation is very useful for safe operation and to extend the output voltage range below the source peak voltage. In this paper, it is basically considered that the fast transient response without load overcurrent and start-up inrush current can be available by fully utilizing both the possible operational modes.

Recently, several papers introduced the sliding mode control(SMC) technique for the resonant power switching supply [10-15]. Among them, an interesting control technique using the sliding mode in the continuous time domain has been reported [14]. With this technique, since the full state feedback loop is analyzed simultaneously, the analysis is more complex and the existence condition of the sliding regime has not been proved analytically. It is also difficult to design a control law considering the characteristics of each state. To design the sliding mode controller in the continuous time domain, an approximate model under several assumptions has been derived and used. In this paper, the switched SMC in the discrete time domain is proposed to control the resonant current in the inner-loop. Using the

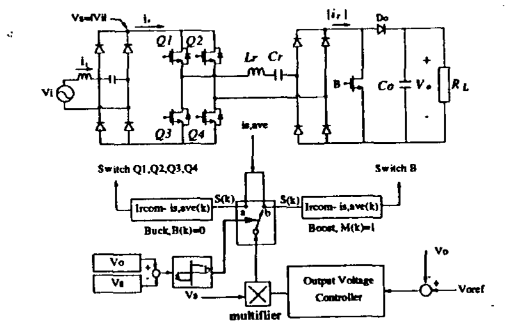
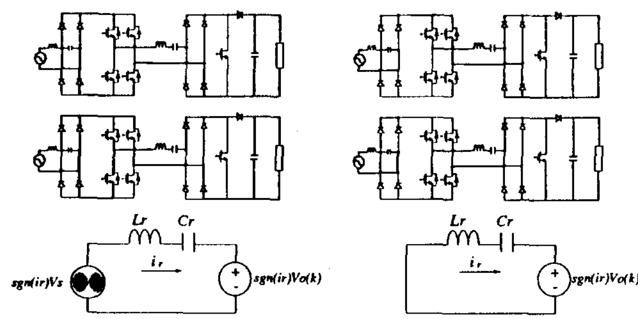
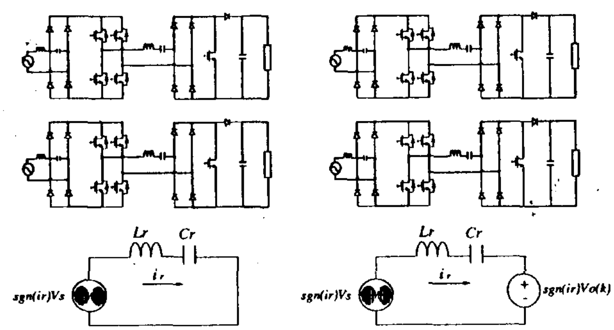


Fig. 1. Circuit Diagram of a Switched Sliding Mode Controlled ZCS Series Resonant AC to DC Converter.



(a) Powering Mode : M(k)=1 (b) Free-wheeling Mode : M(k)=0

Fig. 2. Equivalent Circuits for Buck Operation



(a) Switching Status : B(k)=1 (b) Switching Status : B(k)=0

Fig. 3. Equivalent Circuits for Boost Operation

proposed switched SMC technique, a unity power factor, a better control performance of the DC output voltage regulation with a wide range, and reduced load overcurrent and start-up inrush current are obtained.

2 Basic Operation and Problem Statement

2. 1 Principles of Basic Operation

The basic configuration of a ZCS series resonant AC to DC converter shown in Fig.1 consists of a diode bridge rectifier with a high frequency input filter, a full bridge resonant power stage, a single switch boost chopper B, and an output filter stage. The resonant power stage has two useful operational modes, the powering and free-wheeling modes. As shown in Fig. 1, a chopper switch B is added to obtain a boost operation. Note that the changing instants of the operational modes and the switching status of B are only allowed at the zero crossing instants of the resonant current.

2. 1. 1 Buck Operation

If a chopper switch B is open, the proposed converter can be classified as a buck cônverter. In this case, the amplitude of the resonant current can be controlled according to the operational mode pattern of the resonant power stage. The DC voltage transfer ratio, D, can be determined discretely as follows:

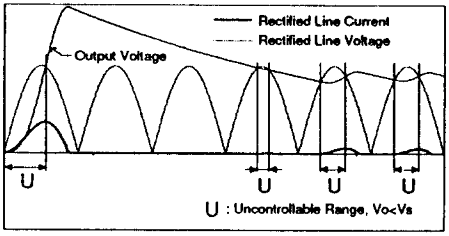
$$D(p,f)=\frac{p}{p+f}<1 \tag{1}$$

where p and f denote the number of powering and free-wheeling modes, respectively. Since the DC voltage transfer ratio is less than the unity, the buck type operation can be possible with the off status of the chopper switch B. The detailed operational principles and an equivalent circuit for each mode are shown in Fig. 2. As can be seen in Fig. 2(a), the applied rectified line voltage V_r to the resonant thank circuit is in phase with the resonant current during the powering mode. Thus, the resonant current is increased by using the powering mode and is equal to the input current. On the other hand, Fig. 2(b) shows a detailed operational principles and an equivalent circuit for the free-wheeling mode. As can be seen in this figure, the zero voltage is applied to the resonant thank circuit. Thus, the amplitude of the resonant current is decreased and the input current is zero. The characteristics of these basic operational modes are summarized in Table 1.

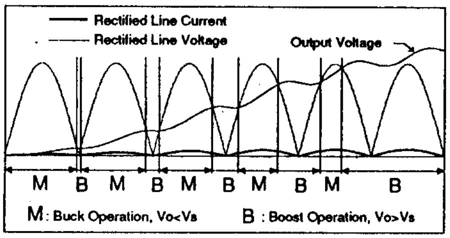
2. 1. 2 Boost Operation

The operation of the ZCS series resonant AC to DC converter with the boost characteristics can be easily obtained by using the chopper switch B. The switch pairs Q1, Q4 and Q2, Q3 are always turned on and off alternatively at zero crossing points of the resonant current. It is noted that the rectified line voltage V_r to the resonant thank circuit is always in phase with the resonant current. Thus, the resonant power stage is continuously energized by the rectified line voltage V_r and the input current is equal to the resonant current. The DC voltage transfer ratio, D, can be determined discretely as follows:

$$D(k,b)=1+\frac{b}{k}>1 \tag{2}$$



(a) Boost Type



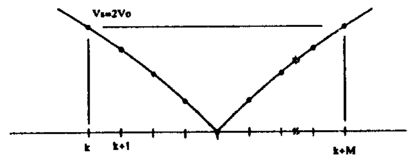
(b) Switched Sliding Mode Control Technique

Fig. 4 Analysis for the uncontrollable Range and Switched Operation

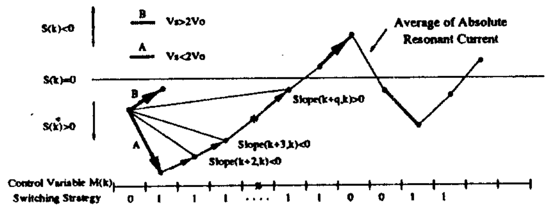
where k and b denote the number of off and on status of the chopper switch B, respectively. Since the DC voltage transfer ratio is more than the unity, the boost type operation can be possible with the continuous powering mode. Using the continuous powering mode, the amplitude of the current can be rapidly increased with a maximum slop when a switch B is closed. If a switch B is open, as can be seen in Fig. 3, 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. In steady state, the value of the output voltage is generally much larger than that of the source voltage in the boost operation. Thus, the amplitude of the resonant current is decreased when the chopper switch B is open. The characteristics of these switching status of B are summarized in Table 2.

2. 2 Problem Statement

The boost type ZCS series resonant AC to DC converter has many advantages over the conventional boost type PWM converters, such as the zero switching loss, smaller filter, and lower EMI level, etc. Furthermore, it shows a much smaller steady state peak current level than a buck type ZCS series resonant AC to DC converter. However, the boost topology can control and limit current only as long as the output voltage is greater than the source voltage. If a load fault condition is occurred, the output voltage will be dragged down below the peak vaule of the AC line voltage. When this occurs, the current will rise rapidly and



(a)



(b)

Fig. 5. Typical Response of Buck Operation

without limit. Since the current is above the desirable level, the chopper switch B is held off by control circuit. However it can't control to reduce the current. Therefore the boost-type converter loses control. Ultimately, the inductor and the isolation transformer will sturate and components will fail and much higher ratings of the switching devices are required for safe operation. Furthermore, since the output voltage is zero during start-up transients, the start-up inrush current and output voltage overshoot are also occurred. This results in the afore mentioned problems. Thus the constraint on the use of the operational mode in boost type AC to DC converter causes the uncontrollable range during load fault condition and start-up transient period. In this paper, it is considered that a wide range of output voltage without a large current overshoot can be available by fully utilizing both the possible operational modes, buck mode and boost mode. The internal operational characteristics should be

Table 1. Characteristics of the Basic Operational Modes

Basic Operational Mode		Applied Source Voltage to Tank Circuit	Input Power	Output Power
Mode	Notation ($M(k)$)			
Powering	1	$sgn(i_r(t))v_s(t)$	$\|i_r(t)\| v_s(t)$	$\|i_r(t)\| v_o(t)$
Free-wheeling	0	0	0	$\|i_r(t)\| v_o(t)$

Table 2. Characteristics of the Switching Status of B

Switching Status of B		Reflected Output Voltage to Tank Circuit	Input Power	Output Power
Status	Notation ($B(k)$)			
ON	1	0	$\|i_r(t)\| v_s(t)$	0
OFF	0	$sgn(i_r(t))v_o(t)$	$\|i_r(t)\| v_s(t)$	$\|i_r(t)\| v_o(t)$

analyzed using an exact dynamic model and operational modes, and also the switching signal for B should be carefully determined to give the high power factor and output voltage control performance.

3 Dynamic Modeling

In order to obtain a simple and effective analytical tool for the analysis, a dynamic model in the discrete time domain is developed. Since both the changes of the operational modes and switching status of B can be effected at any control instant, the control variables representing these control inputs for the k-th time event are defined as $M(k)$ and $B(k)$, respectively. Note that $M(k)$ has the values of 1 and 0 denoting the powering and free-wheeling modes for the k-th time event. Also $B(k)$ has the values of 1 and 0 expressing the on and off status of B for the k-th time event, respectively. Based on the basic operational principles, the governing equations for the ZCS buck-boost AC to DC converter can be derived as follows:

$$sgn(i_r(t))M(k)v_s(t) = L_r\frac{di_r(t)}{dt} + v_c(t) + (1-B(k))sgn(i_r(t))v_o(t) \quad (3)$$

$$C_r\frac{dv_c(t)}{dt} = i_r(t) \quad (4)$$

$$(1-B(k))\,|\,i_r(t)\,| = C_o\frac{dv_o(t)}{dt} + \frac{1}{R_L}v_o(t) \quad (5)$$

The left hand side of (3) implies the actual applied rectified line voltage to the resonant tank circuit with respect to the operational mode in the buck operation, and the third term of the right hand side denotes the output voltage reflected to the resonant power stage in case of the boost operation. Equation (4) refers to the dynamic behavior between the resonant capacitor voltage and resonant inductor current. Equation (5) expresses the output equation. Because the line voltage and output voltage during the half resonant interval can be assumed as constants by the low ripple approximation [17], the absolute resonant current can be derived from (3) and (5) as follows:

$$|\,i_r(t)\,| = \frac{v_c^*(k) + M(k)v_s(k) - (1-B(k))v_o(k)}{Z}\sin\{\omega_r(t - kT/2)\}\ , \quad for\ \frac{kT}{2} \le t < \frac{(k+1)T}{2} \quad (6)$$

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 (3)-(5) and (6), the following equation can also be easily obtained as

$$v_c^*(k+1) = \frac{1}{C_rZ}\int_{kT/2}^{(k+1)T/2}\{v_c^*(k) + M(k)v_s(k) - (1-B(k))v_o(k)\}\sin\left\{\omega_r\left(t-\frac{kT}{2}\right)\right\}dt - v_c^*(k) \quad (7)$$

$$v_o(k+1) = \frac{1}{C_oZ}\int_{kT/2}^{(k+1)T/2}\{v_c^*(k) + M(k)v_s(k) - (1-B(k))v_o(k)\}\sin\left\{\omega_r\left(t-\frac{kT}{2}\right)\right\}dt + v_o(k) - \frac{1}{C_o}\int_{kT/2}^{(k+1)T/2}\frac{v_o(k)}{R_L}dt. \quad (8)$$

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 (6) as

$$i_{s,ave}(k) \equiv \frac{2}{\pi}|\,i_{r,p}(t)\,| = \frac{2}{\pi}\frac{v_c^*(k) + M(k)v_s(k) - (1-B(k))v_o(k)}{Z}, \quad for\ \frac{kT}{2} \le t < \frac{(k+1)T}{2} \quad (9)$$

where $|\,i_{r,p}(t)\,|$ denotes the absolute peak value of the resonant current. Solving (7) and (8) gives

$$v_c^*(k+1) = v_c^*(k) + 2M(k)v_s(k) - 2(1-B(k))v_o(k) \quad (10)$$

$$v_o(k+1) = \gamma(1-B(k))v_c^*(k) + M(k)v_s(k) + (1-B(k))v_o(k) + v_o(k) - \gamma^*v_o(k) \quad (11)$$

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 (9) by $(k+1)$ as

$$i_{s,ave}(k+1) = \frac{2}{\pi}\frac{v_c^*(k+1) + M(k+1)v_s(k+1) - (1-B(k+1))v_o(k+1)}{Z}. \quad (12)$$

Using (9) through (12), the following discrete state equation for a ZCS series resonant AC to DC converter can be derived as

$$\begin{pmatrix} i_{s,ave}(k+1) \\ v_o(k+1)\end{pmatrix} = \begin{pmatrix} 1-\gamma(1-B(k))(1-B(k+1)) & -\frac{2}{\pi Z}(1-B(k)) + (1-\gamma^*)(1-B(k+1)) \\ \frac{\pi Z\gamma}{2}(1-B(k)) & 1-\gamma^* \end{pmatrix} \begin{pmatrix} i_{s,ave}(k) \\ v_o(k)\end{pmatrix} + \begin{pmatrix} \frac{4v_s(k)}{\pi Z} \\ 0\end{pmatrix} M^*(k+1). \quad (13)$$

Since C_r is chosen as much smaller than C_o, γ and γ^* become much smaller than the unity. Therefore, the equation (13) can be simplified as follows:

$$\begin{pmatrix} i_{s,ave}(k+1) \\ v_o(k+1)\end{pmatrix} = \begin{pmatrix} 1 & -\frac{4}{\pi Z}(1-B^*(k+1)) \\ \frac{\pi Z\gamma}{2}(1-B(k)) & 1-\gamma^* \end{pmatrix} \begin{pmatrix} i_{s,ave}(k) \\ v_o(k)\end{pmatrix} + \begin{pmatrix} \frac{4v_s(k)}{\pi Z} \\ 0\end{pmatrix} M^*(k+1) \quad (14)$$

where $B^*(k+1)$ and $M^*(k+1)$ are defined, respectively, as

$$B^*(k+1) = \{B(k) + B(k+1)\}/2 \quad (15)$$

4 Switched Discrete Sliding Mode Control Technique and Analysis

In this section, a switched discrete sliding mode control(SDSMC) technique is developed and its advantages over the other types of current control techniques are comparatively discussed. It is noted that the main control objectives for the AC to DC converter are to shape the line current into a sinewave keeping in phase with the line voltage without large current overshoot and to obtain a desirable control performance for the output voltage regulation. The boost type converter is often used to satisfy the above control objectives due to the advantage of continuous line current. However, there exists an inherent disadvantage such that the uncontrolled range exists during the start-up transient period and load fault condition. Fig. 4(a) shows the typical start-up transient response for the boost operation. As can be seen in this figure, the uncontrollable ranges exist in the range of $v_o(k)$ less than $v_s(k)$. Since $M^*(k+1)$ is always set to the unity in a boost ZCS series resonant AC to DC converter, the current slope is always positive for any value of $B^*(k+1)$ in the range of $v_o(k)$ less than $v_s(k)$. Hence, the resonant current can not be properly controlled until $v_o(k)$ becomes greater than $v_s(k)$, and so start-up inrush current appears during the transient period. To deal with this design problem, it is considered in this paper that a buck operational mode should be properly used in the range of $v_o(k)$ less than $v_s(k)$ to eliminate the uncontrollable range, and then a boost operation is cascaded above range. This basic strategy for the elimination of the uncontrollable ranges are shown in Fig. 4(b). By properly choosing the operational mode patterns below $v_s(k)$, an sliding mode control technique for the buck ZCS series resonant AC to DC converter is employed. This control concept is further extended to a boost operation for the range of $v_o(k)$ greater than $v_s(k)$ in order to obtain the unity power factor.

4. 1 Buck Operation

4. 1. 1 Switching Strategy

The switching strategy based on the sliding mode control(SMC) in case of the buck operation is as follows:

$$M(k+1) = 1 \quad for \quad S(k) > 0 \tag{21}$$

$$M(k+1) = 0 \quad for \quad S(k) < 0. \tag{22}$$

The switching surface consisting of the average value of the resonant current and its command I_{rcom} in the k-th time event is as follows:

$$S(k) = I_{rcom} - i_{s,ave}(k). \tag{23}$$

When a quasi sliding mode exists on $S(k)=0$, the average motion of the system response is determined by the smooth control function called the equivalent operational mode. In the discrete time domain, the equivalent operational mode $M^*_{eq}(k+1)$ is obtained from the forward difference of the switching surface as

Table 3. Current Slope Analysis for the Buck Operation

$M^*(k+1)$	$Slope(k,k+1)$	Positive Current Slope Condition	Negative Current Slope Condition
0	$\frac{1}{L_{eq}}[-v_o(k)]$	$-v_o(k) > 0$ None	$-v_o(k) < 0$ Always
0.5	$\frac{1}{L_{eq}}\left[\frac{1}{2}v_s(k) - v_o(k)\right]$	$v_s(k) > 2v_o(k)$	$v_s(k) < 2v_o(k)$
1	$\frac{1}{L_{eq}}[v_s(k) - v_o(k)]$	$v_s(k) > v_o(k)$ Always	$v_s(k) < v_o(k)$ None

$$S(k+1) - S(k) = 0. \tag{24}$$

From (17), (23), and (24), the equivalent operational mode is derived as

$$M^*_{eq}(k+1) = \frac{v_o(k)}{v_s(k)}. \tag{25}$$

The equation (25) shows that the equivalent operational mode coincides with the duty ratio between the input voltage and the output voltage, and satisfies the following equation in case of the buck type converter as

$$0 < M^*_{eq}(k+1) < 1. \tag{26}$$

This equation implies that the proper operational mode can always be generated in case the output voltage is less than the rectified line voltage. If we use the equivalent operational mode to the equation (14), the dynamics of the resonant current is eliminated. In other words, the dynamic relation between the output voltage and the average value of the resonant current command can be easily obtained using (20) as

$$v_o(k+1) = (1-\gamma^*)v_o(k) + \frac{\pi Z\gamma}{2}I_{rcom}. \tag{27}$$

Since (27) is a first-order model, the various types of the output-loop voltage controller can be easily designed.

4. 1. 2 Analysis for Sliding Regime

The existence condition for a sliding mode in the continuous time domain is

$$\lim_{s \to 0} s\dot{s} < 0. \tag{28}$$

It is clear that the condition (28) which assures the sliding motion on the switching surface in a continuous system is no longer applicable to the proposed discrete time domain model. Thus, an intuitively obvious definition of sliding motion in the discrete time domain was proposed in [16]. For quasi sliding, this is given as follows:

$$|S(k+1)| < |S(k)|. \tag{29}$$

If(29) is decomposed into two inequalities, it becomes

$$(S(k+1)-S(k))sgn(S(k))<0 \quad (30)$$

$$(S(k+1)+S(k))sgn(S(k))>0. \quad (31)$$

The equation (30) assures the quasi sliding motion on the switching surface and the convergence of the state trajectories on the switching surface is assured by (31). In the following discussion, the detailed analysis for the sliding condition (30) is examined for the proposed buck type AC to DC converter which is modeled in the discrete form. Using the current slope equation (18), (30) can be expressed as

$$(S(k+1)-S(k))sgn(S(k))=\frac{4}{\pi Z}sgn(S(k))\cdot[M^{*}(k+1)v_{s}(k)-v_{o}(k)]$$

$$=\frac{4}{\pi Z}\left(\frac{T}{2}\right)sgn(S(k))\cdot Slope(k,k+1)<0. \quad (32)$$

As can be seen in (32), the slope of the average resonant current must be positive in order to satisfy the sliding condition in the regions of $S(k)>0$, that is, $Slope(k,k+1)>0$. On the other hand, the negative current slope is needed to assure the sliding condition in the regions of $S(k)<0$, that is, $Slope(k,k+1)<0$. It is obviously implied that the state is moved toward the switching surface $S(k)=0$. In other words, the average values of the resonant current is directly followed by the current command. Note that the current slope is controlled by the moving average value of the operational mode, as can be seen in eqn (17). Since three values are possible for $M^{*}(k+1)$, three kinds of the condition are available for both the positive and negative current slopes. The possible conditions for the current slope with respect to $M^{*}(k+1)$ are summarized in Table 3. However, the condition $M^{*}(k+1)=0$ is does not occur because $M(k+1)$ is always the unity by the switching strategy (21) in the regions of $S(k)>0$. Similarly, the case of $M^{*}(k+1)=1$ for the negative slope condition is not available because $M(k+1)$ always has a zero value by the switching strategy (22) in the regions of $S(k)<0$. The conditions, $v_{s}(k)>v_{o}(k)$ and $-v_{o}(k)<0$, when $M^{*}(k+1)=1$ and $M^{*}(k+1)=0$, respectively, are always satisfied in the buck operation. Thus, the only interesting condition is confined to the case of $M^{*}(k+1)=0.5$ which comes from the change of the operational modes. For example, if $M(k)=0$ for the positive slope condition, the next operational mode is set to the unity, $M(k+1)=1$, by the switching strategy (21), and if $M(k)=1$, the next operational mode has the zero value, $M(k+1)=0$, by the switching strategy (22). In this case, $M^{*}(k+1)=0.5$, the only negative current slope is available under the condition of $v_{s}(k)<2v_{o}(k)$ as can be seen from the Table 3. Therefore, the sliding regime is assured in the region of $S(k)<0$ in which only the negative current slope is needed to confirm the sliding condition (30). However, the sliding condition is violated in the region of $S(k)>0$ in which the positive current slope is needed to confirm the sliding condition. Although the sliding condition is relaxed during the transient periods of the operational mode under the conditions of $v_{s}(k)<2v_{o}(k)$ for $S(k)>0$ and $v_{s}(k)>2v_{o}(k)$ for $S(k)<0$, the sliding condition is satisfied at least after the finite sampling time. In other words, the prediction of the sliding surface has the negative values which is equivalent to the positive current slope, or v_{s} is larger than $2v_{o}$ because v_{s} is a rectified sinusoidal time varying waveform after the finite sampling time for $S(k)>0$. This analysis is summarized in the following Lemma 1.

> Lemma 1(Modified Sliding Condition) : For the proposed buck type AC to DC converter(17), the sliding condition for the existence of a quasi sliding regime is satisfied at least after the finite N sampling time. This can be expressed as follows :
>
> $$[S(k+N)-S(k)]sgn(S(k))<0 \quad for \quad \forall k. \quad (33)$$

Proof : By using the switching strategy (21)-(22), and current slope equation (18), the p-step prediction of the sliding surface in the discrete time domain is given by

$$[S(k+p)-S(k)]sgn(S(k))=\frac{1}{L_{eq}^{p}}\{p\,v_{o}(k)-\frac{1}{2}v_{s}(k)\}sgn(S(k)) \quad (34)$$

where L_{eq}^{p} is the equivalent inductance with the p-step prediction in eqn (18). If $p\rightarrow N$ is taken, the modified sliding condition for the proposed buck type converter can be obtained. The N-step prediction of the sliding surface has the negative value or the magnitude of the rectified source voltage in the $(k+N)$-th time event $v_{s}(k+N)$ is larger than that of the $2v_{o}(k)$ in case of $S(k)>0$. Similarly, the N-step prediction of the sliding surface has the positive value, or the magnitude of the rectified source voltage in the $(k+N)$-th time event $v_{s}(k+N)$ is smaller than that of the $2v_{o}(k)$ in case of $S(k)<0$. Note that the integer number N which implies the finite sampling time is defined as follows:

$$N=\min\{M,q\}. \quad (35)$$

In eqn (35), the positive integer q represents the minimum value of the number of the finite sampling time when either the positive current slope appears in case of $S(k)>0$ or the negative current slope is available in case of $S(k)<0$ by the switching strategy as shown in Fig. 5(b). This is mathematically expressed as

$$q=\min\{i\mid[S(k+i)-S(k)]sgn(S(k))<0\}. \quad (36)$$

The integer q depends on the switching strategy which makes the current slope have the desirable polarity after the q sampling time. On the other hand, since the line voltage is a sinusoidal time varying voltage, the magnitude of the rectified line voltage in $(k+M)$-th time event is larger than that of the rectified line voltage in the k-th time event which is equal to $2v_o(k)$ in case of $S(k)>0$ and smaller when $S(k)<0$. From eqn (34), the positive current slope appears when $v_s(k+M)>2v_o(k)$ in case of $S(k)>0$ and the negative current slope is available when $v_s(k+j)<2v_o(k)$ in case of $S(k)<0$ by the variation of $v_s(k)$ as shown in Fig. 5(a). This can be mathematically expressed as

$$M=\min\{j \mid [v_s(k)-v_s(k+j)]sgn(S(k))<0\}. \qquad (37)$$

After a modified quasi sliding mode takes place, the sliding surface variation during the N sampling time is bounded as follows:

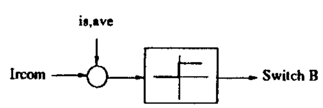
$$|S(k+N)-S(k)| \le \frac{1}{L_{eq}^N}\{Nv_o(k)-M^*(k+1)v_s(k)\}|_{max}$$

$$=\frac{1}{L_{eq}^N}Nv_o(k). \qquad (38)$$

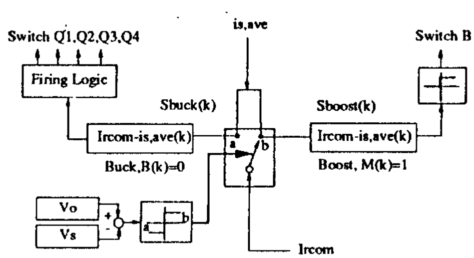
The eqn (38) can be derived by setting $M^*(k+1)$ to 0. Therefore, a modified quasi sliding motion for the proposed converter is not divergent.

5 Controller Implementation and Simulation

The advantages of the proposed control technique are comparatively illustrated with the bang bang type of the typical current control techniques. The block diagrams of two current control techniques for the purpose of comparison are shown in Fig. 6. The major advantage of Fig. 6(a) is simple implementation since only the boost operation is used. However, the control performances are generally



(a) Bang Bang Boost Type



(b) Switched Sliding Mode Controller

Fig. 6 Block Diagram of the two kinds of control techniques

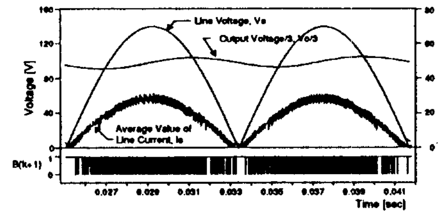
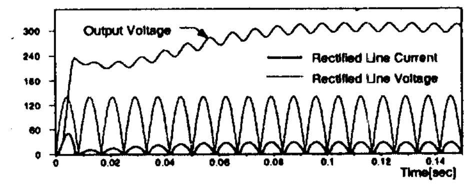
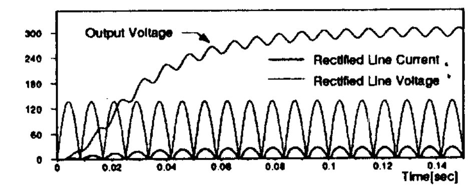


Fig. 7 Steady State Response for Boost and SDSMC Techniques

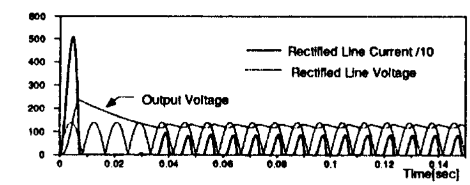


(a) Boost Type

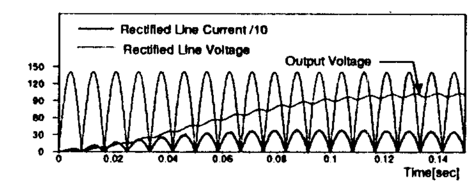


(b) Switched Sliding Mode Control Technique

Fig. 8 Start-up Transient Responses in case of Vo,ref=300V.



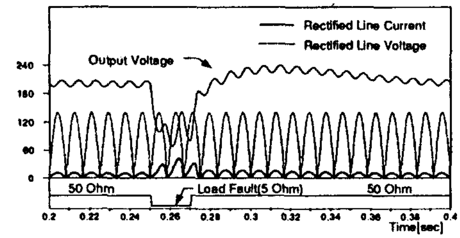
(a) Boost Type



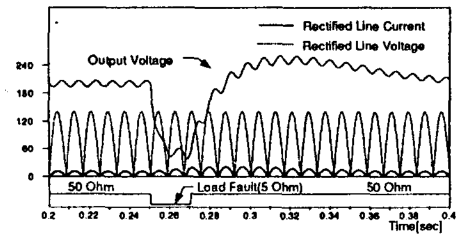
(b) Switched Sliding Mode Control Technique

Fig. 9 Start-up Transient Responses in case of Vo,ref=100V.

shown to be undesirable during the start-up transient periods and load fault condition. Fig. 6(b) shows the proposed switched sliding mode current controller. As can be seen in this figure, the switching technique of the buck and boost operation is employed to improve the control performance. With these strategies, the steady state response for the arbitrary current command are simulated. Since only the boost operation can be available in the steady state, the



(a) Boost Type



(b) Switched Sliding Mode Control Technique

Fig. 10 Load Fault Transient Responses in case of Vo,ref=200V.

responses are the same for the two kinds of control techniques. Fig. 7 shows the waveform of the line voltage, line current, and output voltage. It can be said that the line current show the sinusoidal waveforms keeping in phase with the line voltage. Thus, the unity power factor can be obtained for both control techniques. For the purpose of comparison, the rectified line voltage, line current, and output voltage of a boost type ZCS series resonant AC to DC converter using a bang bang type current control technique are simulated as shown in Fig. 8(a). It shows, however, the large current overshoot and undesirable output voltage response. As can be expected, these problems are effectively overcome using the proposed switched sliding mode control as shown in Fig. 8(b). Since the reduced current overshoot can be obtained, the system design optimization with respect to the ratings of the switching device is possible. Further, it shows a better output voltage transient response and the buck-boost operation can also be available. The other advantage of the proposed switched SMC technique is that the controllable output voltage ranges are wider than the conventional boost type. Fig. 9 shows the simulated waveforms when the output voltage reference is 100*V* which is less than the peak value of the absolute line voltage. As can be seen in Fig. 9(a), the start-up inrush current appears during the start-up transient and the uncontrollable regions are shown in steady state. This result in a poor power factor in the range of the lower output voltage with the undesirable output voltage regulation. However, the responses of the proposed switched SMC are desirable in the power factor correction and output voltage regulation with wide ranges as can be seen in Fig. 9(b). If a load fault condition is occurred, the output voltage will be dragged down below the peak vaule of the AC line voltage. As can be seen in Fig. 10(a), the current will rise rapidly and without limit. Therefore the inductor and the isolation transformer will sturate and components will fail and much higher ratings of the switching devices are required for safe operation. However these problems are eliminated by proposed scheme as can be seen in Fig. 10(a).

6 Conclusion

A buck-boost zero current switched(ZCS) series resonant AC to DC converter for the DC output voltage regulation together with high power factor is proposed. The proposed single phase AC to DC converter 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 for this AC to DC converter is developed and an analysis for the internal operational characteristics is explored. Based on this analysis, a switched discrete sliding mode control(SDSMC) technique is investigated and its advantages over the other types of current control techniques are discussed. With the proposed control technique, the unity power factor and the DC output voltage regulation in wide ranges without a inrush current in start-up transient and load fault condition can be obtained.

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