Modeling and Improved Predictive Current Control for Buck-Boost Series Resonant Inverter

Gun-Woo Moon, Jung-Hoon Lee, In-Cheol Baik, Kyeong-Hwa Kim, and Myung-Joong Youn

Abstract

An improved predictive current control technique for a zero current switched(ZCS) buck-boost series resonant inverter(SRI) is proposed to overcome the inherent disadvantages such as the uncontrollable large overshoot and the large current ripple. Using the proposed technique, four quadrant operations of the output voltage and current for an uninterruptible power supply(UPS) application are guaranteed and the buck-boost operation can also be obtained without an additional bidirectional switch.

I. Introduction

Recently, a great deal of researche on the high frequency DC or AC link power conversion techniques have been done due to the increasing demands for the high power density converter and inverter especially in the aerospace applications [1]-[7]. Among them, a new topology and an output voltage error estimation method of the zero current switched SRI for an UPS have been recently introduced[2]. Although the voltage ripple can be reduced, the implementation of this method is complicated due to the large computational effort required for the estimation routine and the current ripple is not sufficiently reduced. As an another approach to deal with these problems, a predictive current control technique for the buck-boost quantum series resonant converter(QSRC) has been recently attempted[3]. With this technique, however, the current control performance for the buck-boost operation is not sufficiently optimized and the uncontrollable ranges are appeared, because of being employed only the powering and freeresonance modes. Furthermore, an additional bidirectional switch across the isolation transformer is required and the controlled frequency of the pulsating current transferred to the output filter stage is randomly changed, so that an optimum output filter design is difficult. In this paper, the conventional predictive current control concept using the dynamic model is extended to the zero current switched buck-boost SRI without the constraint on the use of the operational modes and an additional bidirection switch. Furthermore, a fast transient response and near optimum operational mode patterns to reduce the current ripple are obtained and comparatively investigated with the other types of the control techniques[2]-[6].

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II. Basic Operations and Problem Statement

1. Basic Operations

The resonant power circuit of a zero current switched SRI shown in Fig.1 consists of the resonant power stage, isolation transformer, and full bridge inverter. The resonant power stage has three operational modes such as the powering, free resonance, and regenerating. The applied voltage to the resonant tank circuit is in phase with the resonant current during the powering mode operation, while the zero voltage is applied to this circuit during the freeresonance mode operation. Thus, the powering and free resonance mode can generally be used to increase and gradually decrease the resonant current respectively. On the other hand, the applied voltage is out of phase with the resonant current during the regenerating mode operation, so that the resonant current decreases faster than the freeresonance mode case [2]-[5]. Note that every operational mode changing instant is only allowed at every zero crossing instant of the resonant current.

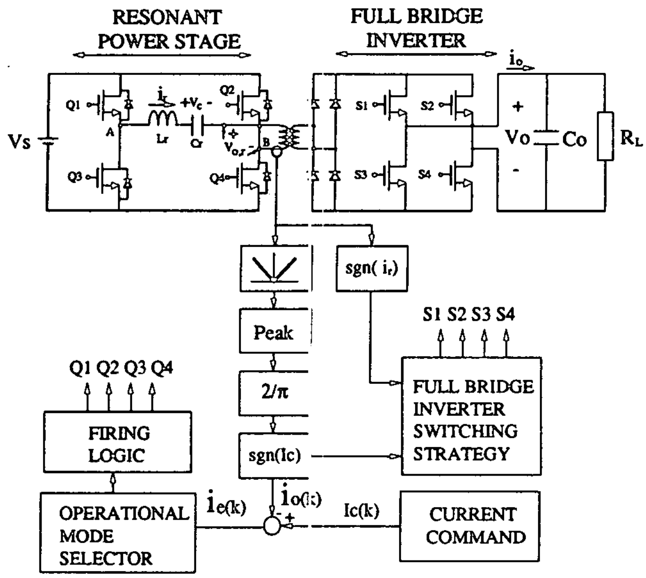


Fig. 1. Circuit Diagram of a Zero Current Switched Buck-Boost SRI.

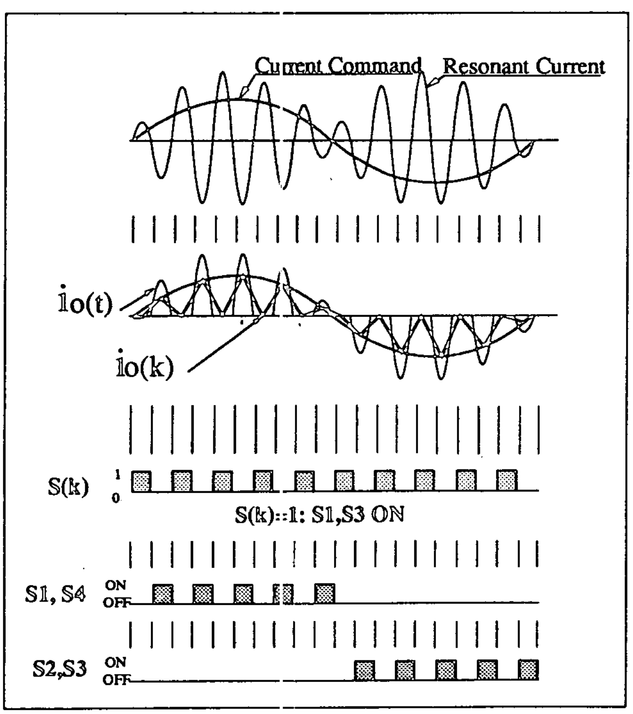


Fig. 2. Switching Strategy for the Full Bridge Inverter.

It is a common constraint of all the ZCS resonant power conversion schemes. By changing the operational mode of resonant power stage, the amplitude of the resonant current can be controlled. This high frequency resonant current is transferred to the isolation transformer, and then controlled by a full bridge inverter to produce a desired output frequency. Note that the switches of a full bridge inverter are also turned on and off in accordance with the zero crossing instants of the resonant current. With the operational mode of resonant power stage, the boost operation can be obtained using the switching strategy for the full bridge inverter. A typical waveform of the full bridge inverter controlled resonant current and a switching strategy for a full bridge inverter are shown in Fig.2. As can be seen in this figure, the full bridge inverter switches S1 and S3 are both on to get a boost operation if the sign of the resonant current is positive. On the other hand, if the sign of the resonant current is negative, the switches S1 and S4 are closed during the interval that the polarity of resonant current is same as that of current command, while the control signals for S2 and S3 are the complement of that for the switches pair S1, S4. With this kind of a switching strategy, the size of isolation transformer can be effectively reduced[2]. The applied source voltage (V_{AB}) and reflected output voltage ($V_{o,r}$) to the tank circuit are investigated with respect to the current command, resonant current, operational modes, and control signal for the full bridge inverter and summarized in Table 1.

Table 1. The Reflected Output Voltage and Applied Source Voltage to Tank Circuit for the Proposed Full Bridge Inverter Switching Strategy.

Operational Conditions		Full Bridge Inverter Switches				Reflected Output Voltage	Applied Voltage To Tank Circuit	
		S_1	S_2	S_3	S_4		Mode	V_{AB}
$i_r>0$		ON	OFF	ON	OFF	0	P	V_s
							F	0
							R	$-V_s$
$i_r<0$	$I_c>0$	ON	OFF	OFF	ON	V_0	P	$-V_s$
							F	0
							R	V_s
	$I_c<0$	OFF	ON	ON	OFF	$-V_0$	P	$-V_s$
							F	0
							R	V_s
Generalized Expressions $V_{AB}=\text{sgn}(i_r)M(k)V_s$ M(k)=1:Powering(P) M(k)=0: Free-Resonance(F) M(k)=-1: Regenerating(R)					$v_{o,r}=(1-S(k))\text{sgn}(ir)\text{sgn}(Ic)vo$ S(k)=1:S1, S3 ON S(k)=0:S1, S4 or S2, S3 ON			

2. *Problem Statement*

The aim of the current mode control is for the full bridge inverter controlled resonant current to follow the desired current command exactly. In order to effectively control the resonant current, the internal operational characteristics should be analyzed in detail. For this analysis, a simple and exact dynamic model is required. A dynamic model in the discrete time domain presented in [3] is very useful for ZCS buck-boost series resonant inverter. However, since the resonant current is always rectified in this scheme, this model is valid only for the first quadrant operations of the output current and voltage. Thus, a modeling technique presented in [3] should be extended to the four quadrant operations of a proposed inverter. To improve the current control performance and guarantee the four quadrant buck-boost operation of the output voltage and current, every operational mode and the switching strategy for a full bridge inverter should be selected properly at every control instant. Since the regenerating mode is not utilized in the conventional predictive control technique for the buck-boost QSRC[3], the large current overshoot, current ripple, and offset current are appeared. On the other hand, the uncontrollable overshoot is not found in the bang bang type current control technique[4]. By employing the regenerating mode in a control, both the positive and negative current slopes can be available for the four quadrant operation. However, the steady state current ripple is larger than that of the proposed control technique. To overcome these disadvantages, the internal operational characteristics should be analyzed including all kinds of operational modes and the switching strategy for the full bridge inverter should also be carefully determined to get the minimized current ripple, reduced offset current, and guarantee the four quadrant operations.

III. Modeling and Improved Control Technique

In order to obtain an analytical tool for analysis, a sampled data dynamic model is developed in the following. Since the output of a QSRI is controlled by the amplitude modulated discrete current pulses, the state variables are defined in the discrete time domain as shown in Fig.3. Fig. 4 shows the equivalent circuits for the proposed inverter. Furthermore, since the changes of the basic operational modes and switching signal for the full bridge inverter can be occurred at any control instant, the control variable representing the operational mode, *M(k),* and switching signal for the full bridge inverter , *S(k),* of the kth time event are also defined in terms of discrete time index respectively. It is assumed that all components are ideal and the turn ratio of the isolation transformer is one. From Table 1 and Fig. 4, the differential equations for the resonant power stage can be easily derived as follows:

$$V_{AB} = L_r \frac{di_r(t)}{dt} + v_c(t) + v_{o,r}(t), \text{ for } \frac{kT}{2} \le t < \frac{(k+1)T}{2} \quad \text{(1-a)}$$

where V_{AB} and $v_{o,r}(t)$ denoting the applied voltage and reflected output voltage to the tank circuit respectively can be expressed as

$$V_{AB} = \text{sgn}(i_r(t)) M(k) V_s \quad \text{(1-b)}$$

$$v_{o,r}(t) = (1-S(k))\ \text{sgn}(I_c(k))\ \text{sgn}(i_r(t)) v_o(t) \quad \text{(1-c)}$$

The sgn(*x*) is defined as

$$\text{sgn}(x) = \begin{cases} 1 & \text{if } x \ge 0 \\ 0 & \text{if } x < 0 \end{cases} \quad \text{(1-d)}$$

$$C_r \frac{dv_c(t)}{dt} = i_r(t). \quad (2)$$

Similarly, the output equation can also be expressed as

$$i_o(t) = (1-S(k))\ \text{sgn}(I_c(k))|i_r(t)| = C_o \frac{dv_o(t)}{dt} + \frac{1}{R_L} v_o(t)$$

$$\text{for } \frac{kT}{w} \le t < \frac{(k+1)T}{2} \quad (3)$$

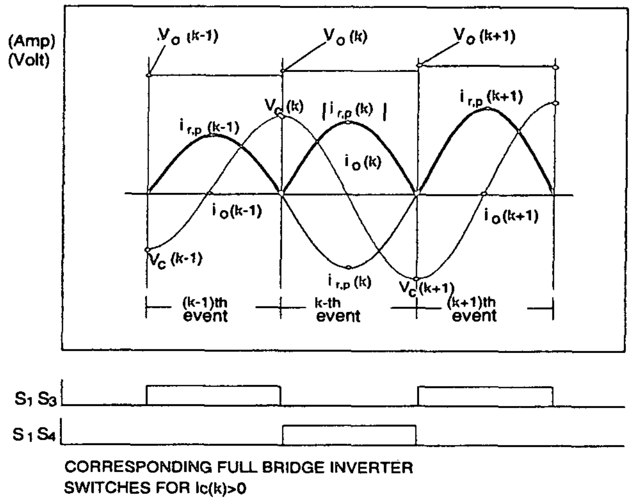
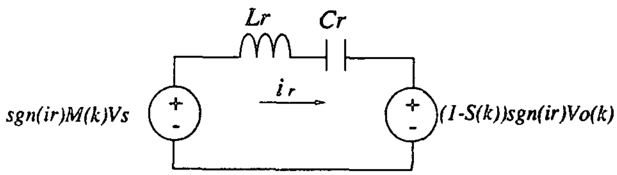
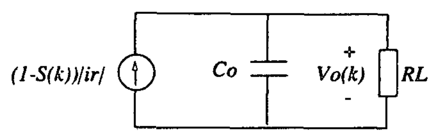


Fig. 3. Definition of the Discrete State Variable.



(a) Power Stage Section



(b) Output Side Section

Fig. 4. Equivalent Circuits of the ZCS-Buck-Boost Series Resonant Inverter.

Note that T denotes the resonant cycle, $i_o(t)$ represents the full bridge inverter controlled resonant current, and the setting of *M(k)* as 1, 0, and 1 implies the powering, freeresonance, and regenerating mode for the kth time event respectively. Since the output voltage during the half resonant interval can be regarded as constant by the low ripple approximation[3], the absolute resonant current can be easily derived from (1) and (2) as follows:

$$|i_r(t)| = \frac{v_c^*(k) + V_s M(k) - (1-S(k))\,\mathrm{sgn}(I_c(k))v_o(k)}{Z}\sin\{\omega_r(t-kT/2)\},$$

$$\text{for } \frac{kT}{2} \le t < (k+1)\frac{T}{2} \tag{4}$$

where $Z=\sqrt{L_r/C_r}$, $T=2\pi\sqrt{L_rC_r}$, $\omega=1/\sqrt{L_rC_r}$, and $v_c^*(k)$ is defined as the absolute value of $v_c(k)$. Define a new discrete $i_{r,avg}(k)$ representing the average value of the resonant current and $i_o(k)$ representing the average value of the full bridge inverter controlled resonant current during the kth time event, then the following equation can be obtained from (4) as

$$i_{r,avg}(k) \equiv \frac{2}{\pi}\,\mathrm{sgn}(I_c(k))\{|i_r(t)|\}_{peak},$$

$$i_o(k) \equiv \frac{2}{\pi}(1-S(k))\,\mathrm{sgn}(I_c(k))\{|i_r(t)|\}_{peak}, \text{ for } \frac{kT}{2} \le t < \frac{(k+1)T}{2}$$

$$= \frac{2}{\pi}\,\mathrm{sgn}(I_c(k))\frac{v_c^* + V_s M(k) - (1-S(k))v_o(k)}{Z}. \tag{5}$$

From (2),(3) and (5), the following equation can also be derived as

$$v_c^*(k+1) = \frac{1}{C_rZ}\int_{kT/2}^{(k+1)T/2}\{v_c^*(k)+V_sM(k)-(1-S(k))\,\mathrm{sgn}(I_c(k))v_o(k)\}$$

$$\sin\left\{\omega_r\left(t-\frac{kT}{2}\right)\right\}dt - v_c^*(k) \tag{6}$$

$$v_o(k+1) = \frac{1}{C_rZ}\int_{kT/2}^{(k+1)T/2}\mathrm{sgn}(I_c(k))\{v_c^*(k)+V_sM(k)-(1-S(k))\,\mathrm{sgn}(I_c(k))v_o\}$$

$$\sin\left\{\omega_r\left(t-\frac{kT}{2}\right)\right\}dt - v_c^*(k) - \frac{1}{C_o}\int_{kT/2}^{(k+1)T/2}\frac{v_o(k)}{R_L}dt \tag{7}$$

Solving these equations gives rise to

$$v_c^*(k+1) = v_c^*(k) - 2(1-S(k))\,\mathrm{sgn}(I_c(k))v_o(k) + 2V_sM(k) \tag{8}$$

$$v_o(k+1) = \mathrm{sgn}(I_c(k))\{\gamma v_c^*(k) + \gamma V_sM(k)\} + (1-S(k))(1-\gamma-\gamma^*)v_o(k) \tag{9}$$

where $\gamma = 2C_r/C_o$ and $\gamma^* = \pi Z\gamma/(2R_L)$. The average value of the full bridge inverter controlled resonant current for the (k+1)th time event can be obtained by simply replacing all the time index k of equation (5) by k+1 as follows:

$$i_o(k+1) = \frac{2}{\pi Z}\{v_c^*(k+1)+V_sM(k+1)-(1-S(k))\,\mathrm{sgn}(I_c(k))v_o(k+1)\} \tag{10}$$

Substituting (8) and (9) into (10) and rearranging with (5), the following equation can be obtained as

$$i_o(k+1) = \{\mathrm{sgn}(I_c(k))\,\mathrm{sgn}(I_c(k+1))-\gamma\}i_o(k)$$

$$-\frac{2}{\pi Z}(1-S(k))\{\mathrm{sgn}(I_c(k))\,\mathrm{sgn}(I_c(k+1))+1-\gamma^*\}v_o(k)$$

$$+\frac{2V_s}{\pi Z}\,\mathrm{sgn}(I_c(k+1))\{M(k)+M(k+1)\} \tag{11}$$

Similarly from (9), $v_o(k+1)$ can also be expressed in terms of $i_o(k)$ and $v_o(k)$ as follows:

$$v_o(k+1) = \frac{\pi Z\gamma}{2}(1-S(k))i_o(k) + (1-\gamma^*)v_o(k) \tag{12}$$

In order to interpret the internal operations of a proposed inverter, the dynamic model is simplified under the following assumptions: $\gamma \ll 1$ and $\gamma^* \ll 1$. This assumptions are resonable one since C_o is generally chosen as much larger than C_r for low output voltage ripple. In the range that $I_c(k)$ and $I_c(k+1)$ have the same polarity, from (11) and (12), the discrete state equation can be expressed as follows:

$$\begin{pmatrix} i_o(k+1) \\ v_o(k+1)\end{pmatrix} = \begin{vmatrix} 1 & -\frac{4}{\pi Z}(1-S^*(k+1)) \\ \frac{\pi Z\gamma}{2}(1-S(k)) & 1 \end{vmatrix}\begin{pmatrix} i_o(k) \\ v_o(k)\end{pmatrix}$$

$$+\begin{vmatrix}\frac{\mathrm{sgn}(I_c(k+1))4V_s}{\pi Z} \\ 0\end{vmatrix}M^*(k+1) \tag{13}$$

$$S^*(k+1) = \frac{S(k)+S(k+1)}{2},\ M^*(k+1) = \frac{M(k)+M(k+1)}{2} \tag{14}$$

The current dynamics are directly controlled by $S^*(k+1)$ and $M^*(k+1)$. The possible values for these temporary variables are examined and summarized in Table 2 and 3. In the conventional optimal predictive current control technique for the buck-boost QSRC, the dynamic model is obtained by equating $S^*(k+1)$ to $M^*(k+1)$ for all k.

Table 2. The Proposed Values of with respect to the Operational Modes.

M*(k+1)	M(k)	M(k+1)	Comments
1	P	P	P:Powering (M(k)=1) F:Free-resonance (M(k)=0) R:Regenerating (M(k)=-1)
0.5	P	F	
	F	P	
0	P	R	
	F	F	
	R	P	
-0.5	F	R	
	R	F	
-1	R	R	

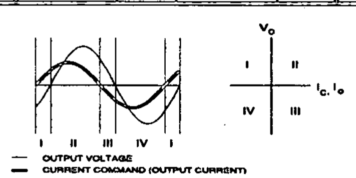
In this case, $S^*(k+1)$ becomes the only control variable and this has the possible value set of *{0, 0.5, 1}*. Since the sign of the control variable, $S^*(k+1)$, is always positive, the current slope can be positive or negative for any value $S^*(k+1)$ if $v_c^*(k)$ and $i_o(k)$ have the different polarities like the regions I and III in Table 4. Hence, the resonant current can not be properly controlled unless $v_c^*(k)$ and $i_o(k)$ have the same polarities which result in a large current overshoot. To deal with this design problem, all kinds of the operational modes should be properly used to eliminate the uncontrollable range. As can be seen in Table 1, the full bridge inverter switches S1 and S3 are both on if the sign of the resonant current is positive. On the other hand, if the sign of the resonant current is negative, the switches pair *S1, S4* or *S2, S3* is alternately turned on and off with respect to the current command. An ideal control input which can nullify the present current error at the next time event can be expressed as $S^*(k+1)$ $S^*(k+1)$

$$M^*_{ideal}(k+1) = \frac{\mathrm{sgn}\,(I_c(k+1))}{V_s}\left(\frac{\pi Z}{4}\,i_e(k) + (1-S^*(k+1))\,v_o(k)\right) \quad (15)$$

Table 3. The Possible Values of S*(k+1) with respect to the Switching Signal for the Full Bridge Inverter.

S*(k+1)	S(k)	S(k+1)	Comments
1	1	1	S(k)=1:S1,S3
0.5	1	0	ON
	0	1	S(k)=0:S1,S4
0	0	0	or S2,S3 ON

Table 4. The Signs of the Current Slopes with respect to Four Quadrant Operational of the Output Voltage and Current for a Current Control Technique Using the Powering and Free-Resonance Mode.

Current Selector Type	Possibl Values of $M^*(k+1)$	The Signs of Current Slopes with respect to Four Quadrant Operations I	II	III	IV	Comments
Powering, Free- Resonance	1	PCS	NCS	NCS	PCS	PCS : Positive Current Slope NCS : Negative Current Slope
	0.5	PCS for $v_o < V_s/2$, NCS for $v_o > V_s/2$	NCS	PCS for $v_o > V_s/2$, NCS for $v_o < V_s/2$	PCS	
	0	NCS	NCS	PCS	PCS	
Definitions of Four Quadrants						

where $i_e(k) \equiv I_c(k) - i_{r,avg}(k)$. In eq. (15), $S^*(k+1)$ is directly governed by the proposed full bridge inverter switching strategy. *S(k)* is unity in case of $\mathrm{sgn}\,(i_r(k)) = 1$. However, the next control input S(k+1) is zero because the sign of the resonant current is negative. Thus, $S^*(k+1)$ can be taken as 0.5 in every operation. This switching strategy for the full bridge inverter confirm the buck-boost operation and make it easy to design the current controller. Thus (15) is rewritten as follows:

$$M^*_{ideal}(k+1) = \frac{\mathrm{sgn}\,(I_C(k+1))}{V_s}\left(\frac{\pi Z}{4}\,i_e(k) + 0.5v_o(k)\right) \quad (16)$$

which contents the only one control variable. Since the available five values for $M^*(k+1)$ are *{ 1, 0.5, 0, 0.5,1}*, the next step is to find one of the practically possible value of $M^*(k+1)$ which is nearest to the ideal control input. This can be mathematically expressed as

$$M^*(k+1) = P_j^* = [\ P_j \in \{-1,-0.5,0,0.5,1\} |\, \min_{j=1\cdots5} \{p_j - M^*_{ideal}(k+1)\}] \quad (17)$$

Finally, the actual operational mode of the resonant power stage for (k+1)th time event can be easily determined using (14) as follows:

$$M(k+1) = \begin{cases} 1, & \text{if } \ 2p_j^* - M(k) > 1 \\ 2p_j^* - M(k) & \text{if } \ -1 \le 2p_j^* - M(k) \le 1 \\ -1 & \text{if } \ 2p_j^* - M(k) < 1 \end{cases} \quad (18)$$

With this controller, the current control performance can be remarkably improved since the current slopes are directly controlled by an optimal manner.

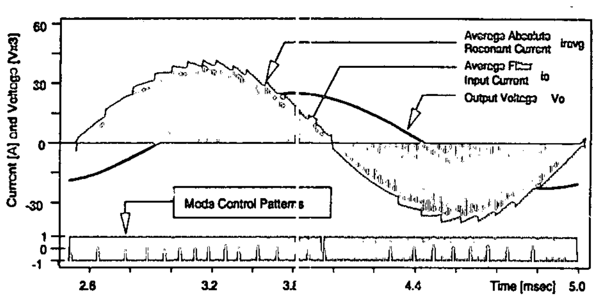
IV. Simulations and Discussions

In this section, the responses of the proposed current control technique are comparatively shown with the other types current control techniques. The parameters used in these simulations are given as follows:

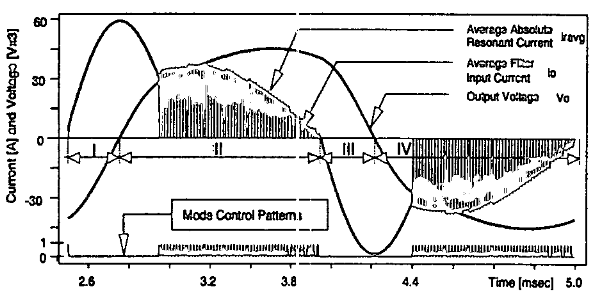
$$V_s = 50V, L_r = 200\mu H, C_o = 100\mu F, R_L = 10\Omega,$$

$$f_r = 100kHZ,\ currentCommand = I_c = 36\sin(2\pi 400t).$$

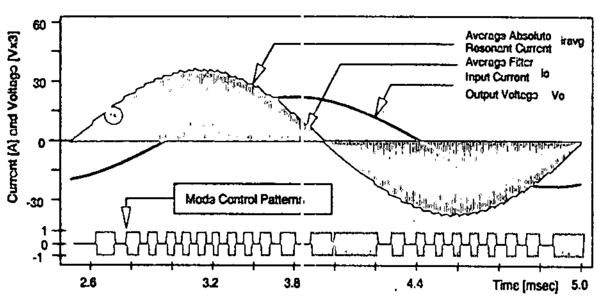
Fig.5(a) shows the current and output voltage responses for the bang bang type of the previous studies[4] and the corresponding mode control patterns. These control patterns show how the internal current slopes are utilized. However, the large current ripple and offset current are shown in this figure. As an effective way of dealing with these problems, the conventional predictive current control concept [3] is investigated for the buck-boost SRI as shown in Fig. 5(b). With this technique, the offset current can be reduced and the relatively small current ripple can be obtained. However, the large current overshoot due to the uncontrollable range is appeared. As can be expected, these problems are effectively solved by employing the proposed predictive control strategy for the buckboost SRI as shown in Fig. 5(c). The minimized current and output voltage ripples with the reduced offset current can be obtained, and four quadrant buck-boost operations without an additional bidirectional switch are guaranteed. Furthermore, the controlled frequency of the pulsating output current transferred to the output filter stage becomes constant as a resonant frequency, which results in an optimum output filter design. The proposed control technique is expected to be widely applicable for the improvement of the current control performance in many applications.



(a)



(b)



(c)

Fig. 5. Responses (a) Bang Bang Type (b) Conventional Optimal Predictive Type (c) Proposed Type.

V. Concluding Remarks

In this paper, a dynamic modeling and an improved current control technique for a ZCS buck-boost series resonant inverter are newly proposed to obtain the improved control performance without much complicating the implementation. Dynamic modeling in the discrete time domain is developed. Using this model, the internal operational characteristics affecting to the current control performances are investigated and the possibility of a remarkable current ripple reduction can be found. Based on this analysis, an improved predictive current control technique for a zero current switched(ZCS) buck-boost series resonant inverter(SRI) is proposed to overcome the inherent disadvantages such as the uncontrollable large overshoot and the large current ripple. Using the proposed technique, four quadrant operation of the output voltage and current for an uninterruptible power supply(UPS) application is guaranteed and the buck-boost operation can also be obtained without an additional bidirectional switch. Furthermore, an offset current can also be remarkably reduced. The proposed technique is expected to be widely applied to the other types of the ZCS resonant power conversion schemes.

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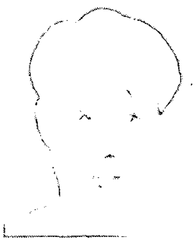

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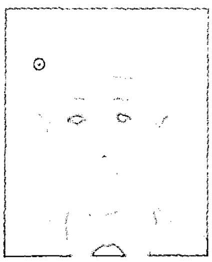

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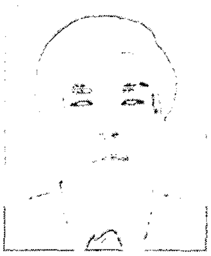

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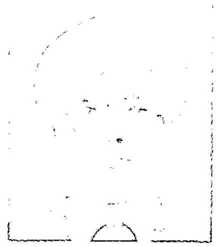

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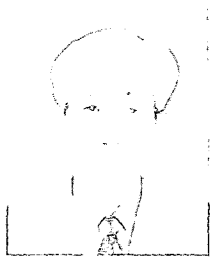

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