Soft Switching Space Vector PWM Inverter Using a New Quasi-Parallel Resonant DC Link

Yong C. Jung, Hyo L. Liu, Guk C. Cho, and Gyu H. Cho

Abstract— A soft switching quasi-parallel resonant dc-link (QPRDCL) inverter with improved PWM capability has been recently presented [5]. The circuit has the minimum voltage stress of the devices and provides the flexibility of selecting the on/off instants of the resonant link, resulting in improved PWM capability. In this paper, the operational principles and the detailed analysis of the QPRDCL inverter are presented for the resonant components design and the inverter control. An SVPWM with optimal vector sequence suitable for the QPRDCL inverter is also presented through the comparisons among five different modified space vector PWM (SVPWM) techniques classified by the voltage vector sequences. The performance of the selected optimal SVPWM is verified by the experimental results.

I. INTRODUCTION

T O solve the demerits of the voltage source-type hard switching inverter, such as low switching frequency, high switching loss, high EMI and acoustic noise, etc., many types of resonant dc-link inverters using various soft switching techniques have been proposed and studied intensively [1]–[5]. Among them, particularly the actively clamped resonant dc-link (ACRDCL) inverter [1] and the quasi-resonant dc-link (QRDCL) converter [4] are well-known topologies. The former, however, has high voltage stress and subharmonic problem due to discrete pulse modulation (DPM). The latter has also the restricted PWM capability owing to the fixed bus voltage notching duration, although it has minimum voltage stresses.

To overcome these problems, other types of soft switching PWM inverters have been suggested, which use a parallel resonant dc-link (PRDCL) [2], [3]. These PRDCL inverters have minimum voltage stresses and allow variable off-pulse width and on-pulse width of dc-link voltage, which greatly enhances the PWM capability. For this purpose, however, three additional switches are used.

Recently, the principle of a new soft switching quasi-parallel resonant dc-link (QPRDCL) inverter has been briefly presented by the authors [5]. It has minimum voltage stress of the

devices and improved PWM capability due to the flexible selectability of the on/off instants of the resonant link using two additional switches. In this paper, the detailed analysis of the QPRDCL operation is carried out for the design of the resonant components and the inverter control.

Based on the previous researches about the space vector modulation techniques of the hard switching inverter [6]–[12], five modified space vector PWM (SVPWM) techniques applicable to the QPRDCL inverter are also compared in this paper. They are classified by the different voltage vector sequences in a sampling period. A SVPWM method with optimal vector sequence is selected as a result of the comparisons among them, and the performance is verified by the experimental results.

II. OPERATIONS OF THE QPRDCL INVERTER

The circuit configuration of the QPRDCL inverter is illustrated in Fig. 1. The QPRDCL consists of two switching devices S_{a1} and S_{a2}, two diodes D_1 and D_2, resonant inductor L_r and resonant capacitors C_{r1} and C_{r2}. In this case, C_{r1} is the main resonant capacitor and C_{r2} is the auxiliary capacitor used to reverse the resonant inductor current i_{Lr}. Because the resonant inductor L_r is sufficiently smaller than the load inductance, the inverter with three-phase load can be replaced by current source I_o during the switching period. Fig. 2 shows the equivalent circuit of the QPRDCL inverter for explanation of the link operation. In this case, S_{INV} stands for all inverter switches. Operation modes consist of eight intervals, as shown in Fig. 3, and related waveforms are shown in Fig. 4.

At mode 0 (M_0), the output load current I_o flows through either switch S_{a1} or antiparallel diode D_{a1} of switch S_{a1} and the second auxiliary switch S_{a2} is in off state. If the switching status of the inverter needs to be changed, switch S_{a2} is turned on with zero current condition for initializing the resonant inductor current i_{Lr} (M_1). As i_{Lr} reaches the previously fixed initializing current I_i, the resonance between L_r and C_{r1} occurs by turning off S_{a1} with zero voltage condition (M_2). The main resonant capacitor voltage v_{Cr1} decreases resonantly from the dc source voltage V_d to zero, and thereafter, the resonant inductor current i_{Lr} freewheels through antiparallel diodes of inverter switches (M_3). The freewheeling duration is controllable, and thus, the link pulse position can be located at any position that is given by

Fig. 1.   The circuit configuration of the quasi-parallel resonant dc-link (QPRDCL) inverter.

the PWM controller. In this mode, all inverter switches are turned on under zero voltage. Some time after, switch S_{a2} is turned off under zero voltage condition, and the resonant inductor current i_{Lr} is reversed by the resonance between L_r and C_{r2} (M_4). When this mode is completed, the inductor current i_{Lr} freewheels again through the inverter switches (M_5). When the total zero voltage duration, $t_5 - t_2$, equals a proper value which is precalculated by the PWM controller, the switches of the inverter selected according to the modulation strategy are turned off under zero voltage condition. A new resonance between L_r and C_{r1} occurs, and the main capacitor voltage v_{Cr1} increases up to the dc source voltage V_d (M_6). The residual current flows through the diode D_{a1} and lastly becomes zero (M_7). In this period, switch S_{a1} can be turned on under zero voltage condition. One switching cycle of the QPRDCL is completed at the end of this mode. The mode analysis of the QPRDCL operation is carried out in the Appendix. The equations for the resonant components are derived, and the required times of each mode ($T_1 \sim T_7$) to control the resonant link operation are also obtained.

III. CHARACTERISTICS AND DESIGN OF THE NEW INVERTER

Fig. 5 shows the voltage waveforms of the auxiliary switches, S_{a1} and S_{a2}, for one link operation. S_{a1} has always the minimum voltage stress. However, the voltage stress of S_{a2} can be minimum only if the following equation derived from (A12) is satisfied:

$$Z_{r2} \leq \frac{V_d}{I_{p,\,\mathrm{max}}}   \tag{1}$$

where $I_{p,\,\mathrm{max}}$ is the maximum value of the resonant inductor peak current represented by (A8). If $I_{p,\,\mathrm{max}}$ is determined, the auxiliary resonant capacitor C_{r2} can be selected using this equation. $I_{p,\,\mathrm{max}}$ and the first resonant components, L_r and C_{r1}, can be easily calculated by the following two equations

Fig. 2.   Equivalent circuit of the QPRDCL inverter for explanation of the link operation.

derived from (A8) and (A18), like [2]:

$$I_p \geq \frac{V_d}{Z_{r1}} + I_{on}   \tag{2}$$

$$I_i \geq \left[\left(\frac{V_d}{Z_{r1}} + I_o n + I_o \right)^2 - \left(\frac{V_d}{Z_{r1}} \right)^2 \right]^{1/2} - I_o.   \tag{3}$$

In this way, the minimum voltage stresses for all of the devices can be achieved. On the other hand, the current stresses of S_{a1} and S_{a2} are 1.5 p.u. in cases where IGBT devices are used.

There are two freewheeling modes, M_3 and M_5, in the QPRDCL operation. These two interval periods contribute to form the variable zero vector. In practical case, there are some declines of the resonant inductor current in these modes mainly caused by the conducting voltages of the switching devices in the freewheeling patches. These should be considered to choose the initializing current I_i. For the simplicity of the link operation control, M_3 is fixed to minimum duration and M_5 is varied to get a zero vector in the experiment.

IV. MODIFIED SVPWM TECHNIQUES FOR THE QPRDCL INVERTER

The space vector PWM (SVPWM) method is widely used, because the maximum output voltage is 15% larger than the

Fig. 3. The mode diagrams of the QPRDCL operation.

one obtained by sinusoidal PWM (SPWM), and the harmonic characteristics are better than the others maintaining lower switching frequency [6], [7]. However, in order to apply the SVPWM to the QPRDCL inverter, the sequence of the voltage vectors selected in a sampling period should be modified.

The voltage vector **V** represented by the instantaneous space vectors is defined by

$$V = V_a + aV_b + a^2V_c,$$

$$a = \exp\left(\frac{j2\pi}{3}\right). \tag{4}$$

There are seven space vectors represented on d–q domain normalized by $\frac{2}{3}{}^*V_d$, which are depicted in Fig. 6. They are six nonzero vectors $V_1 \sim V_6$ and one zero vector V_0 or

V_7. Fig. 7 shows the three possible trajectories in which U-locus is the generated flux vector defined as a time-integral function of **V**, and U*-locus represents the reference flux vector. These are well-analyzed as regards the harmonics in the previous work [8]. The scheme of Fig. 7(a) is superior to the others when the modulation index is around unity, but the scheme of Fig. 7(b) has an advantage over a wide modulation index range. Therefore, the five modified SVPWM techniques available to the QPRDCL inverter are presented, excluding the scheme of Fig. 7(c), which is illustrated in Fig. 8. Types I and II are the scheme of Fig. 7(b), and the others are the scheme of Fig. 7(a) with different zero vector durations, respectively. In this figure, $V_{0,\,\mathrm{min}}$ means the zero vector with minimum time duration $t_{0,\,\mathrm{min}}$ which is represented by (A7), (A14), and

Fig. 4.   The operational waveforms of the resonant components.

Fig. 5.   Voltage waveforms of the auxiliary switches S_{a1} and S_{a2}.

(A20) as follows:

$$t_{0,\min} = \frac{T_{2,\max} + T_{6,\max}}{2} + T_4. \qquad (5)$$

V_{1_2} implies the V_1 vector with one half of the calculated time duration. Similarly, V_{0_2} implies the V_0 vector with half duration. As mentioned previous chapter, the zero voltage period in the link operation can be utilized as the zero voltage vector V_0 in the SVPWM method.

V. COMPARISONS AMONG THE MODIFIED SVPWM TECHNIQUES

To compare among the modified SVPWM techniques, the three criteria are used as follows.

1) Performance index, PI

$$PI \equiv \int |U^* - U|^2 \, dt. \qquad (6)$$

2) Distortion factor, DF

$$DF \equiv \frac{1}{V_1} \left[\sum_{k=2}^{\infty} \left(\frac{V_{k-th}}{k} \right)^2 \right]^{1/2} \times 100 \, [\%]. \qquad (7)$$

Fig. 6.   Voltage space vectors of the conventional SVPWM inverter.

3) Total harmonic distortion, THD

$$THD \equiv \frac{1}{V_1} \left[\sum_{k=2}^{\infty} (V_{k-th})^2 \right]^{1/2} \times 100 \, [\%] \qquad (8)$$

where V_1 and V_{k-th} denote the fundamental and the kth harmonic voltages, respectively. Thus, the optimal SVPWM technique should be satisfied to minimize the above criteria for overall modulation index range. In this case, the modulation index m is defined as

$$m \equiv \frac{|\mathbf{V}_o|}{\frac{2}{3} \cdot V_d} \qquad (9)$$

where $\mathbf{V}_o$ is the output voltage vector and V_d is the dc-link voltage.

As can be seen in Fig. 8, all types have the same sampling angle ωT_s, except for Type II. To have the same flux locus, the sampling angle of Type II should be reduced to one half of the others. Therefore, the comparison is carried out in such a case that $\Delta\theta_1 = \Delta\theta_3 = \Delta\theta_4 = \Delta\theta_5 = 2 \cdot \Delta\theta_2$ equal $10°$.

The performance index PI's according to the modulation index m are shown in Fig. 9. Since the performance index PI is defined by a time-integral function of flux error vector square, PI should be minimized to reduce the torque ripple caused by the flux error. Types I and II are superior to the others with regard to PI. The flux deviation of Type III is the largest of all.

The distortion factor DF is a criterion, which shows the quantity of the containing lower-order harmonics. Although the distortion factors of output voltages show similar trends, as shown in Fig. 10, Type II is better than any others in most of the range. Also, Fig. 11 shows the total harmonic distortion according to the modulation index. In this case, Types II and III have better characteristics over wide modulation index range. Synthetically speaking, Type II is the optimal SVPWM method available for the QPRDCL inverter.

VI. EXPERIMENTAL RESULTS

In order to verify the operations and the validity of the optimal SVPWM method selected in the previous chapter, the

Fig. 7. Three possible trajectories of space vector PWM method.

Fig. 8. Five modified SVPWM techniques with different voltage vector sequences.

prototype QPRDCL inverter is built and tested. The switching devices S_{a1}, S_{a2}, $S_1 \sim S_6$ are module type 2MBI50L-060 IGBT's and the diodes D_1, D_2 are DSEI2X61-10B. Other circuit parameters are $L_r = 20\,\mu H$, $C_{r1} = 45$ nF and $C_{r2} = 205$ nF.

Fig. 12 shows the control block diagram for the SVPWM inverter. Mode 1 duration T_1 can be calculated by predicting the next output load current I_{on} using the output currents sensed. The three vector durations, t_0, t_1, and t_2, calculated by SVPWM strategy are loaded to the gating signal generator. In this case, the gating signal generator is implemented by using one programmable timer, two buffers, and eight gate drivers like as [3].

Fig. 13 shows the main resonant capacitor voltage v_{Cr1}, the resonant inductor current i_{Lr} and the auxiliary resonant capacitor voltage v_{Cr2}, respectively. As can be seen in this figure, they are coincided with the previous explanations. There are some droops, however, in the resonant inductor current during

the freewheeling period, which should be considered in the link operation.

To confirm the comparisons in the previous chapter, Types II and III are selected and tested in this chapter. As mentioned before, the sampling angle is determined to compare the cases under the same flux locus so that $\Delta\theta_3 = 2 \cdot \Delta\theta_2 = 10°$. Fig. 14 shows the flux locus of Type II, and the locus of Type III is also similar although not shown here. The line-to-line voltage and the line current of Type II are shown in Fig. 15 and the current harmonic spectrum is shown in Fig. 16. For comparison, the line-to-line voltage, line current, and current harmonic spectrum of Type III are shown in Figs. 17 and 18, respectively. In this case, a 5[hp] induction motor is used as a load by driving the motor with modulation index $m = 0.8$ at the output frequency f_o of 60 [Hz]. As can be seen in these figures, the lower-order current harmonics of Type II is smaller than the ones of Type III. Type II is also superior to Type III with respect to THD and DF of line current.

Fig. 9. Performance index PI's according to the modulation index m.

Fig. 11. Total harmonic distortion THD's according to the modulation index m.

Fig. 10. Distortion factor DF's according to the modulation index m.

VII. CONCLUSION

The operational principle and mode analysis of the QPRDCL inverter [5] are presented. Using this analysis, the resonant components are designed for testing the QPRDCL inverter. To apply the SVPWM technique to the QPRDCL inverter, some modifications of vector sequences are carried out. Five different modified SVPWM techniques available to the QPRDCL inverter are presented and compared. For comparison, three criteria are used by defining performance index (PI), distortion factor (DF), and total harmonic distortion (THD). As a result, Type II is chosen to be the optimal SVPWM method available for the QPRDCL inverter, which is verified by the experimental results with the motor load.

APPENDIX
MODE ANALYSIS OF QPRDCL OPERATION

1) Mode 0: $(t_7 \sim t_0,\ S_{a1}\ :\ \text{on},\ S_{a2}\ :\ \text{off},\ S_{INV}\ :\ \text{off})$

$$v_{Cr1}(t) = V_d,$$
$$v_{Cr2}(t) = 0,$$
$$i_{Lr}(t) = 0. \tag{A1}$$

2) Mode 1: $(t_0 \sim t_1,\ S_{a1}\ :\ \text{on},\ S_{a2}\ :\ \text{on},\ S_{INV}\ :\ \text{off})$

$$v_{Cr1}(t) = V_d$$
$$v_{Cr2}(t) = 0. \tag{A2}$$
$$i_{Lr}(t) = \frac{V_d}{Lr}\,t$$
$$i_{Lr}(T_1) = I_i \tag{A3}$$
$$T_1 = \frac{Lr I_i}{V_d}. \tag{A4}$$

3) Mode 2: $(t_1 \sim t_2,\ S_{a1}\ :\ \text{off},\ S_{a2}\ :\ \text{on},\ S_{INV}\ :\ \text{off})$

$$v_{Cr1}(t) = V_d \cos(\omega_{r1}t) - (I_i + I_o)Z_{r1}\sin(\omega_{r1}t)$$
$$v_{Cr2}(t) = 0 \tag{A5}$$
$$i_{Lr}(t) = \frac{V_d}{Z_{r1}}\sin(\omega_{r1}t) + (I_i + I_o)\cos(\omega_{r1}t) - I_o \tag{A6}$$

where $\omega_{r1} = 1/\sqrt{LrCr1}$, $Z_{r1} = \sqrt{Lr/Cr1}$

$$T_2 = \frac{1}{\omega_{r1}}\tan^{-1}\left(\frac{\dfrac{V_d}{Z_{r1}}}{I_i + I_o}\right) \tag{A7}$$

$$I_p = i_{Lr}(T_2)$$
$$= \left[(I_i + I_o)^2 + \left(\frac{V_d}{Z_{r1}}\right)^2\right]^{1/2} - I_o. \tag{A8}$$

4) Mode 3: $(t_2 \sim t_3,\ S_{a1}\ :\ \text{off},\ S_{a2}\ :\ \text{on},\ S_{INV}\ :\ \text{on})$

$$v_{Cr1}(t) = 0$$
$$v_{Cr2}(t) = 0 \tag{A9}$$
$$i_{Lr}(t) = I_p \tag{A10}$$
$$T_3 = \text{variable.} \tag{A11}$$

5) Mode 4: $(t_3 \sim t_4,\ S_{a1}\ :\ \text{off},\ S_{a2}\ :\ \text{off},\ S_{INV}\ :\ \text{on})$

$$v_{Cr1}(t) = 0$$
$$v_{Cr2}(t) = I_p Z_{r2}\sin(\omega_{r2}t) \tag{A12}$$
$$i_{Lr}(t) = I_p \cos(\omega_{r2}t) \tag{A13}$$

Fig. 12. Control block diagram for the SVPWM inverter.

T_1 : mode 1 duration
t_0 : zero vector duration
t_1 : first vector duration
t_2 : second vector duration

(a)

(b)

Fig. 13. Link operations of the QPRDCL inverter. (a) Main resonant capacitor voltage (50 V/div) and resonant inductor current (10 A/div). (b) auxiliary resonant capacitor voltage (50 V/div) and resonant inductor current (10 A/div).

Fig. 14. Flux locus of Type II.

Fig. 15. Experimental waveforms of Type II with motor load. (upper: line-to-line voltage 100 V/div, lower: line current 1 A/div).

Fig. 16. Line current harmonic spectrum of Type II.

Fig. 18. Line current harmonic spectrum of Type II.

Fig. 17. Experimental waveforms of Type III with motor load. (upper: line-to-line voltage 100 V/div, lower: line current 1 A/div).

where $\omega_{r2} = 1/\sqrt{LrCr2}$, $Z_{r2} = \sqrt{Lr/Cr2}$.

$$T_4 = \frac{\pi}{\omega_{r2}}. \tag{A14}$$

6) Mode 5: ($t_4 \sim t_5$, S_{a1} : off, S_{a2} : off, S_{INV} : on)

$$v_{Cr1}(t) = 0$$
$$v_{Cr2}(t) = 0 \tag{A15}$$
$$i_{Lr}(t) = -I_p \tag{A16}$$
$$T_5 = \text{variable}. \tag{A17}$$

7) Mode 6: ($t_5 \sim t_6$, S_{a1} : off, S_{a2} : off, S_{INV} : off)

$$v_{Cr1}(t) = Z_{r1}(I_p - I_{on}) \sin(\omega_{r1}t)$$
$$v_{Cr2}(t) = 0 \tag{A18}$$
$$i_{Lr}(t) = (I_p - I_{on}) \cos(\omega_{r1}t) + I_{on} \tag{A19}$$

where I_{on} denotes the next load current.

$$T_6 = \frac{1}{\omega_{r1}} \sin^{-1} \left(\frac{\dfrac{V_d}{Z_{r1}}}{I_p - I_{on}} \right) \tag{A20}$$

$$I_r = i_{Lr}(T_6)$$
$$= \left[(I_p - I_{on})^2 - \left(\frac{V_d}{Z_{r1}} \right)^2 \right]^{1/2} + I_{on}. \tag{A21}$$

8) Mode 7: ($t_6 \sim t_7$, S_{a1} : on, S_{a2} : off, S_{INV} : off)

$$v_{Cr1}(t) = V_d$$
$$v_{Cr2}(t) = 0 \tag{A22}$$
$$i_{Lr}(t) = -\frac{V_d}{Lr} t + I_r \tag{A23}$$
$$T_7 = \frac{Lr I_r}{V_d} \tag{A24}$$

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