A PWM Method Reducing Harmonics of Two Interleaved Converters

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***Abstract*—This paper proposes a PWM method reducing harmonics of two interleaved converters at twice the switching frequency region. If harmonics at twice the switching frequency region can be reduced by the PWM method, the size of filter interfacing the grid and converter would be reduced. Based on Fourier series of phase voltage in every switching period, offset voltage for PWM minimizing harmonics at twice the switching frequency region can be calculated in real time. With the proposed PWM, the size and volume of inductive filters of two interleaved parallel converters can be shrunken conspicuously. Effectiveness of the proposed method is verified by the computer simulation and experimental results. The maximum harmonic at twice the switching frequency region has been reduced by 58% at modulation index 0.8.**

I. INTRODUCTION

Harmonic components near and above the switching frequency have been one of the major concerns in three-phase grid-connected pulse-width modulation (PWM) converter systems. Since the stringent harmonic current and voltage regulations at Point of Common Coupling (PCC) should be met, harmonic components of the current flowing into the grid should be sufficiently attenuated by adequate filters, such as just inductive filters or LCL filters [1]. In the whole power conversion system, the filters usually take up a considerable portion of the cost, volume, and weight.

On the other hand, to effectively increase the power rating with smaller power converters, parallel operation is generally accepted strategy. By using the parallel operation, system can achieve the benefits in the view point of reliability, maintenance, cost, and efficiency [2]. One of the most preferred modulation schemes in this operation mode is interleaved modulation method. Using the interleaved parallel-operation of three-phase converters, particular harmonic components in output current can be eliminated, e.g., harmonics at odd multiples of switching frequency for the parallel operation of two converters [3]-[4]. However, since the harmonics at even multiples of switching frequency remain unchanged, output filters are designed focusing on the harmonics at even multiples, especially, twice the switching frequency region. Therefore, if harmonic components in this frequency region are more diminished, the size of filters can be also reduced. Naturally, it leads to less cost and weight of overall system.

Many researches have been conducted regarding the interleaved operation. In [4], ripple cancellation effects in interleaved operation were analyzed using double Fourier series. Also, it was shown that harmonic sidebands as well as the N multiples of the switching frequency could be eliminated by interleaved operation. Meanwhile, in [5]-[7], asymmetric interleaving was introduced, where asymmetric phase shifts for each carrier were utilized. In those papers, reduced circulating current and current THD were verified. In actual filter design, however, twice the switching frequency region is the most matter of concern, rather than THD, since they have the largest harmonic component and are the lowest harmonic frequencies. In this paper, a novel method reducing the harmonics in that region is proposed.

Aforementioned, since twice the switching frequency component is dominant in interleaved parallel-operation of two converters, the purpose of the proposed PWM method is minimization of these frequency components. Based on the analysis of Fourier series coefficients of synthesized phase voltage, determination algorithm of the offset voltage, which is used as a degree of freedom in PWM, is presented, and the voltage minimizes the harmonic components at twice the switching frequency region. These components are effectively suppressed without any additional hardware. Therefore, advantages in cost, volume, and weight of the filters can be achieved. Validity of the proposed method is proved by simulation and experimental results, and the results show that the proposed method can reduce the size of the interface inductor by over 50% compared to the conventional method.

II. CONVENTIONAL INTERLEAVING OPERATION

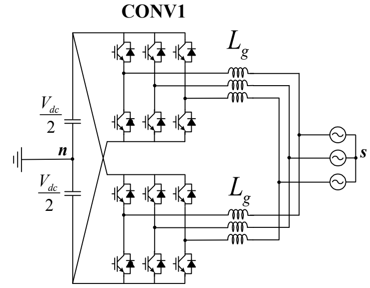


Fig. 1. Diagram of parallel operation of two converters.

In parallel operation of two converters shown in Fig. 1, several harmonics resulted from each converter can be canceled out during the summation of converter currents by shifting the phase of one carrier of PWM by $180°$ from the other, which is so called an interleaved operation. If the sinusoidal-PWM (SPWM) is adopted as a modulation scheme in each converter, then an A-phase pole voltage of converter 1 ($0°$ phase shift), i.e., v_{an1}, can be expressed as (1). Similarly, that of converter 2 ($180°$ phase shift), i.e., v_{an2}, can be also described as (2) [7].

$$v_{an1}(t)=\frac{V_{dc}}{2}M\cos(\omega_1 t+\varphi_0)+\sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty}C(m,n)\cos[m(\omega_c t+\varphi_{c1})+n(\omega_1 t+\varphi_0)]. \tag{1}$$

$$v_{an2}(t)=\frac{V_{dc}}{2}M\cos(\omega_1 t+\varphi_0)+\sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty}C(m,n)\cos[m(\omega_c t+\varphi_{c2})+n(\omega_1 t+\varphi_0)]. \tag{2}$$

where $C(m,n)=\frac{2V_{dc}}{\pi m}\cdot J_n(m\frac{\pi}{2}M)\cdot\sin(m+n)\frac{\pi}{2}$, M is the modulation index, m and n are the multiplier of the switching and fundamental frequency, respectively. Also, V_{dc} indicates the dc-link voltage, $J_n(x)$, the Bessel function of the first kind, ω_1, the fundamental frequency, ω_c, the switching frequency, φ_{c1} and φ_{c2} are the interleaving angle of converter 1 and 2, respectively, and φ_0 is the initial angle of the reference voltage for A-phase. Therefore, the initial angles for B- and C-phases are $\varphi_0-\frac{2}{3}\pi$ and $\varphi_0+\frac{2}{3}\pi$, respectively. And, as aforementioned, φ_{c1} is $0°$ and φ_{c2} is $180°$.

Meanwhile, the line current is determined by (3) where the line current is the summation of each converter current. In (3), $v_{g,as}$ indicates the A-phase line voltage and it is assumed to have only a fundamental component [5].

$$\frac{d}{dt}i_{as}(t)=\sum_{k=1}^{2}\frac{d}{dt}i_{ak}(t)=\sum_{k=1}^{2}\frac{(v_{ank}(t)-v_{sn}(t))-v_{g,as}(t)}{L_g}. \tag{3}$$

Also, L_g means the inductance value between converter and AC source, and v_{sn} in (4) is defined as common mode voltage which can be calculated from (1) and (2).

$$v_{sn}=\frac{1}{2}\sum_{k=1}^{2}\frac{v_{ank}(t)+v_{bnk}(t)+v_{cnk}(t)}{3}=\sum_{m=1}^{\infty}\sum_{n=-\infty}^{\infty}C(2m,3n)\cos[2m\omega_c t+3n(\omega_1 t+\varphi_0)]. \tag{4}$$

From (1)-(4), it can be noticed that the line current consists of fundamental $(=i_{as,f})$ and harmonic $(=i_{as,h})$ components. As the harmonic components come from $(v_{ank}(t)-v_{sn}(t))$ term in (3), they can be extracted as (5). Then, by integrating (5), harmonic components of $i_{as}(t)$ can be derived as (6).

$$\frac{d}{dt}i_{as,h}(t)=\sum_{m=1}^{\infty}\sum_{\substack{n=-\infty\\ n\neq 0,\pm3,\ldots}}^{\infty}\frac{2C(2m,n)\cos[2m\omega_c t+n(\omega_1 t+\varphi_0)]}{L_g}. \tag{5}$$

$$i_{as,h}(t)=\sum_{m=1}^{\infty}\sum_{\substack{n=-\infty\\ n\neq 0,\pm3,\ldots}}^{\infty}\frac{2C(2m,n)}{(2m\omega_c+n\omega_1)L_g}\sin[2m\omega_c t+n(\omega_1 t+\varphi_0)]. \tag{6}$$

From (6), it is worth of noting that all harmonics at odd multiples of switching frequency are eliminated by $180°$ interleaved parallel-operation of two converters. Meanwhile, the harmonics at even multiples remain the same with non-interleaved case since the difference between $m\varphi_{c1}$ and $m\varphi_{c2}$ is multiple of 2π. Therefore, in filter design, only even multiple components of switching frequency would be considered in the interleaved parallel-operation of two converters.

III. Proposed Method

As stated above, harmonics at even multiples of switching frequency are not affected by the interleaved operation. Only the magnitude is varied according to the PWM methods. Therefore, only way to reduce the even multiple harmonic components is synthesizing the phase voltage with less those components with fixed interleaving angle, which is $180°$. However, since the harmonics above the switching frequency region are expressed as a complicated form including Bessel functions in fundamental period, it is tricky to handle and reduce them.

In this part, new PWM method to attenuate harmonic components around twice the switching frequency is proposed with the introduction of a novel concept dealing the harmonics above switching frequency. Since the components at twice the switching frequency are dominant in interleaved parallel-operation of two converters, it is essential to minimize them in order to reduce the size of inductive filters.

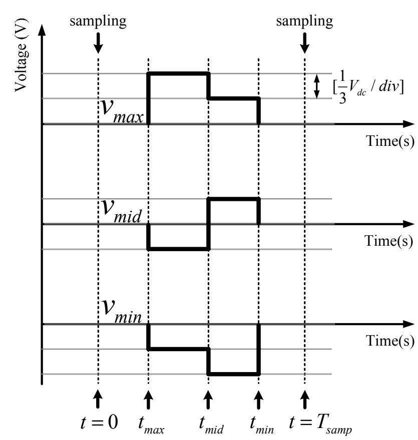


Fig. 2. Phase voltages for the three-phase PWM converter(On sequence).

Fig. 2 shows one example of the applied phase voltages in two-level three-leg converter without dead time effect. In the proposed method, these phase voltages are expressed by Fourier series with switching period T_{sw} (=$2T_{samp}$) as (7), under the assumption that the phase voltage reference is constant during a switching period and, therefore, the synthesized phase voltages are even.

$$v_{xs}(t) = a_0 + \sum_{n=1}^{\infty} a_n \cos\frac{n\pi t}{T_{samp}} \ (0 \le t \le T_{sw}). \tag{7}$$

where $a_0 = \frac{2}{T_{samp}}\int_0^{T_{samp}} v_{xs}(t)dt$, $a_n = \frac{2}{T_{samp}}\int_0^{T_{samp}} v_{xs}(t)\cos\frac{n\pi t}{T_{samp}}dt$.

In the representation of Fourier series with the period T_{sw}, sideband harmonics of n multiples of switching frequency do not appear and only the sum of them is shown.

Among the phase voltage references, v_{max}^* is defined as the reference with a maximum value, v_{mid}^* with a medium value, and v_{min}^* with a minimum value. From these definitions, on/off time of switching device for on/off sequences by triangular carrier based PWM can be calculated as in (8) [8]. For on sequence case such as shown in Fig. 2, t_{max}, t_{mid}, and t_{min} indicate the on time of maximum, medium, and minimum pole voltages, respectively, and off time for off sequence case. Therefore, $a_{2,max}$ means the coefficient of Fourier series for v_{max} at n=2 during the switching period. $a_{2,mid}$ and $a_{2,min}$ are also similarly defined. Then, $a_{2,max}$, $a_{2,mid}$, and $a_{2,min}$ in (9) can be derived from (7)-(8).

Additionally, the phase voltage in a fundamental period, $T_b = 2\pi / \omega_1$, can be expressed by (10).

$$v_{xs}(t) = a_0(t) + \sum_{n=1}^{\infty} a_n(t) \cos\frac{n\pi t}{T_{samp}}, \ (0 \le t \le T_b). \tag{10}$$

Eq. (10) is the representation of (7) with extended time range from a switching period to a fundamental period. There are fundamental component, $a_0(t)$, and ripple components, $\sum_{n=1}^{\infty} a_n(t)\cos\frac{n\pi t}{T_{samp}}$. In addition, Ripple components can be divided into $1f_{sw}$, $2f_{sw}$, … , nf_{sw} region ripple. For example, $a_2(t)\cos(2\pi\times 2f_{sw}\times t)$ represents the net expression for twice the switching frequency components including some sideband harmonics, which can be expressed as (11) with the base period, T_b , where $a_{2,k}$ means the coefficient of $2f_{sw}+kf_b$ frequency harmonic.

$$\sum_{k=-\infty}^{\infty} a_{2,k}\cos 2\pi(2f_{sw}+kf_b)t, \ (k \ne 2m, k \ne 3n). \tag{11}$$

Then, the RMS value of this component in a fundamental period can be represented as (12).

$$\begin{aligned} &\sqrt{\frac{1}{T_b}\int_0^{T_b} a_2(t)^2 \cos^2(2\pi\times 2f_{sw}\times t)dt} \\ &= \sqrt{\frac{1}{2T_b}\int_0^{T_b} a_2(t)^2 dt} \\ &= \sqrt{\frac{1}{T_b}(a_{2,1}{}^2 + a_{2,5}{}^2 + a_{2,7}{}^2 + \cdots)} \end{aligned} \ . \tag{12}$$

First equation of (12) is the RMS value of $a_2(t)\cos(2\pi\times 2f_{sw}\times t)$ in a fundamental period and can be reduced to second equation. Third equation of (12) is also the RMS value of (11) in a fundamental period. From (12), relationship between $a_2(t)$ and sideband harmonics is briefly explained. In a fundamental period, $a_2(t)$ is changing in every sampling time and $v_{xs}(t)$ would be one of the three phase voltage references, i.e., v_{max}, v_{mid}, and v_{min}. Thus, $a_2(t)$ has $a_{2,max}$, $a_{2,mid}$, and $a_{2,min}$ periodically. Consequently, if $a_{2,max}{}^2 + a_{2,mid}{}^2 + a_{2,min}{}^2$ is minimized in every sampling time, then the RMS value of $2f_{sw}$ frequency band ripple could be also minimized in a fundamental period.

To summary, if the coefficient of twice the switching frequency is minimized in every switching period, then the dominant components among sideband harmonics which occupy most of the harmonic sum, would be nearly minimized.

In the proposed method, therefore, v_{offset} for PWM is selected to minimize the square sum of $a_{2,max}$, $a_{2,mid}$, and $a_{2,min}$ in a switching period. If $F(v_{offset})$ in (13) is defined as the square sum of three coefficients, v_{offset} should be determined to satisfy two conditions in (14), which correspond to minimum $F(v_{offset})$.

$$t_{\max} = T_{samp}\times(0.5 - \frac{v_{max}{}^* + v_{offset}}{V_{dc}}), \quad t_{mid} = T_{samp}\times(0.5 - \frac{v_{mid}{}^* + v_{offset}}{V_{dc}}), \quad t_{\min} = T_{samp}\times(0.5 - \frac{v_{min}{}^* + v_{offset}}{V_{dc}}). \tag{8}$$

$$\begin{aligned} a_{2,max} &= \frac{V_{dc}}{3\pi}[\sin\frac{2\pi}{V_{dc}}(v_{mid}{}^* + v_{offset}) + \sin\frac{2\pi}{V_{dc}}(v_{min}{}^* + v_{offset}) - 2\sin\frac{2\pi}{V_{dc}}(v_{max}{}^* + v_{offset})] \\ a_{2,mid} &= \frac{V_{dc}}{3\pi}[\sin\frac{2\pi}{V_{dc}}(v_{min}{}^* + v_{offset}) + \sin\frac{2\pi}{V_{dc}}(v_{max}{}^* + v_{offset}) - 2\sin\frac{2\pi}{V_{dc}}(v_{mid}{}^* + v_{offset})]. \\ a_{2,min} &= \frac{V_{dc}}{3\pi}[\sin\frac{2\pi}{V_{dc}}(v_{max}{}^* + v_{offset}) + \sin\frac{2\pi}{V_{dc}}(v_{mid}{}^* + v_{offset}) - 2\sin\frac{2\pi}{V_{dc}}(v_{min}{}^* + v_{offset})] \end{aligned} \tag{9}$$

$$\begin{aligned} &v_{offset} = \frac{V_{dc}}{4\pi}\times\left\{\tan^{-1}\left(\frac{k_1\sin 2\pi\frac{v_{\min}{}^*}{V_{dc}} + k_2\sin 2\pi\frac{v_{\max}{}^*}{V_{dc}} - k_3\sin 2\pi\frac{v_{\max}{}^* + v_{\min}{}^*}{V_{dc}}}{k_1\cos 2\pi\frac{v_{\min}{}^*}{V_{dc}} + k_2\cos 2\pi\frac{v_{\max}{}^*}{V_{dc}} + k_3\cos 2\pi\frac{v_{\max}{}^* + v_{\min}{}^*}{V_{dc}}}\right) + n\pi\right\}, \text{ where } \begin{aligned} k_1 &= \sin^2\frac{(2v_{\max}{}^* + v_{\min}{}^*)\pi}{V_{dc}} \\ k_2 &= \sin^2\frac{(v_{\max}{}^* + 2v_{\min}{}^*)\pi}{V_{dc}} \\ k_3 &= \sin^2\frac{(v_{\max}{}^* - v_{\min}{}^*)\pi}{V_{dc}} \end{aligned} . \\ &n = 0, \pm 1, \pm 2, \cdots \end{aligned} \tag{15}$$

$$F(v_{offset}) \equiv a_{2,max}^{\;\;2} + a_{2,mid}^{\;\;2} + a_{2,min}^{\;\;2}$$
$$= \frac{V_{dc}^{\;2}}{3\pi^2} \times \left[\begin{array}{l} \left\{ \sin\frac{2\pi}{V_{dc}}(v_{\max}^{\;*} + v_{offset}) - \sin\frac{2\pi}{V_{dc}}(v_{\text{mid}}^{\;*} + v_{offset}) \right\}^2 \\ + \left\{ \sin\frac{2\pi}{V_{dc}}(v_{\text{mid}}^{\;*} + v_{offset}) - \sin\frac{2\pi}{V_{dc}}(v_{\min}^{\;*} + v_{offset}) \right\}^2 \\ + \left\{ \sin\frac{2\pi}{V_{dc}}(v_{\min}^{\;*} + v_{offset}) - \sin\frac{2\pi}{V_{dc}}(v_{\max}^{\;*} + v_{offset}) \right\}^2 \end{array} \right]. \quad (13)$$

$$\frac{d}{dv_{offset}} F(v_{offset}) = 0 \quad \text{and} \quad \frac{d^2}{dv_{offset}^{\;\;2}} F(v_{offset}) > 0. \quad (14)$$

From (14), solution can be derived analytically as (15). However, the selected v_{offset} should meet the condition of (16) which is the boundary conditions for carrier based PWM.

$$-\frac{V_{dc}}{2} - v_{\min}^{\;*} < v_{offset} < \frac{V_{dc}}{2} - v_{\max}^{\;*}. \quad (16)$$

If the boundary condition in (16) is satisfied, then the solution can be found. However, if not, v_{offset} should be re-determined for $F(v_{offset})$ to be minimum at $F(\frac{V_{dc}}{2} - v_{\max}^{\;*})$ or at $F(-\frac{V_{dc}}{2} - v_{\min}^{\;*})$, because $F(v_{offset})$ has no local minimum in the region of (16). In this way, the pole voltage references reducing the dominant harmonics at twice the switching frequency region can be determined in every sampling period.

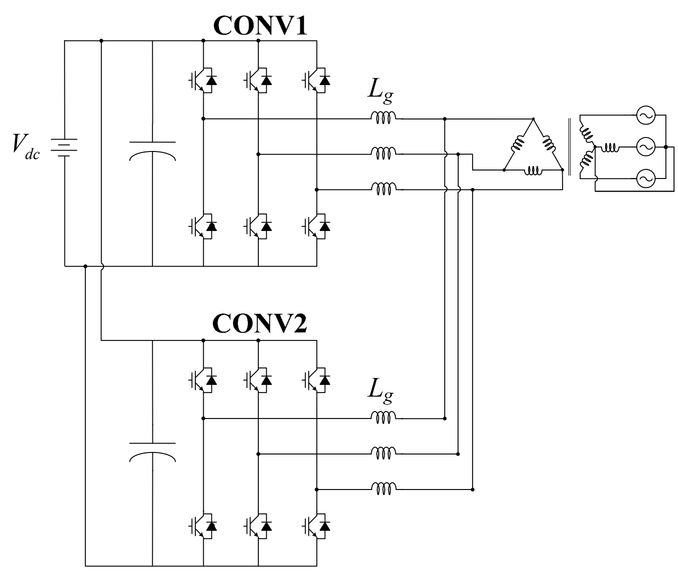


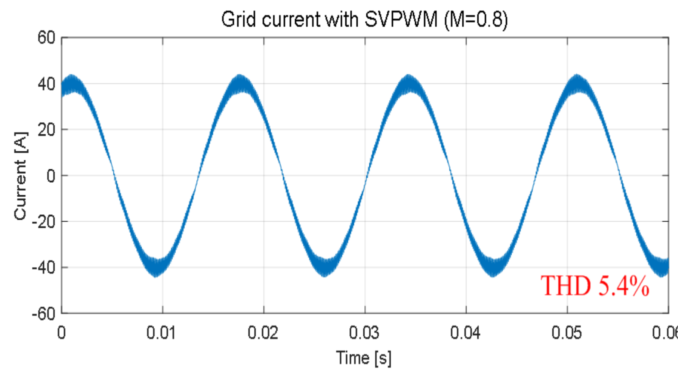
Fig. 3. Circuit diagram of system under experimental verification test.

IV. Experimental Results

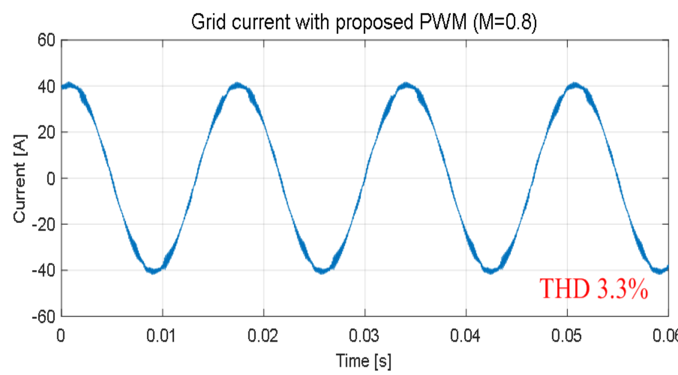
Experimental system for verification of the proposed method has been configured as shown in Fig. 3 and parameters of the system are listed in Table. 1. For each converter, rated current is 20 A_{peak} and a delta-wye connected transformer is placed between the system and grid. Grid side voltage is 220 V_{rms} and transformed to 110 V_{rms} in converter side for the limitation of available dc-link voltage. Two

Table. 1. Parameters of system.

Parameter	Test values
Converter power	5 [kVA]
Line voltage (RMS)	63.5 [V_{rms}]
Line current (peak)	40 [A_{peak}]
Fundamental frequency	60 [Hz]
Switching frequency	5.04 [kHz]
Sampling frequency	10.08 [kHz]
L_g	1 [mH] (0.167 pu)
V_{dc} (M=0.8)	240 [V]



(a) SVPWM (M=0.8)

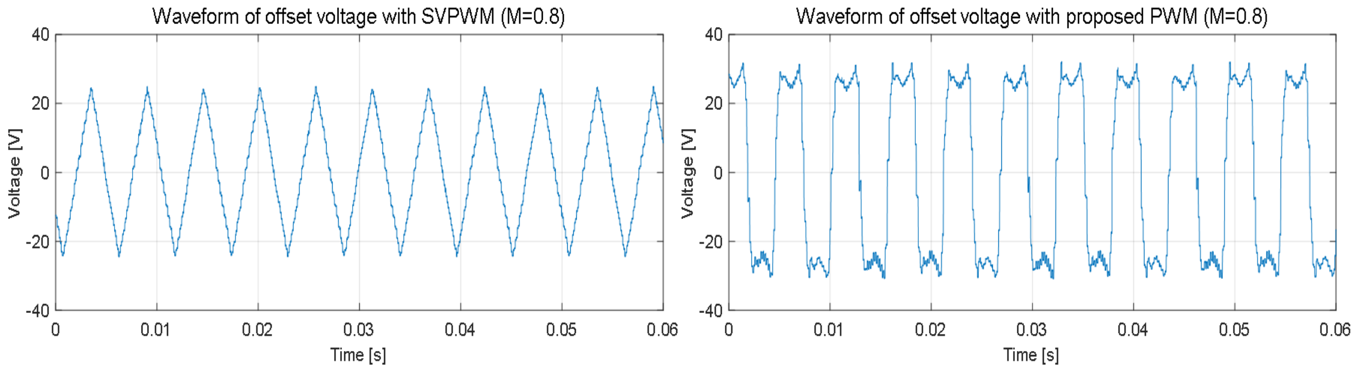


(b) Proposed PWM (M=0.8)

Fig. 4. Line current of each PWM method(Experiment).

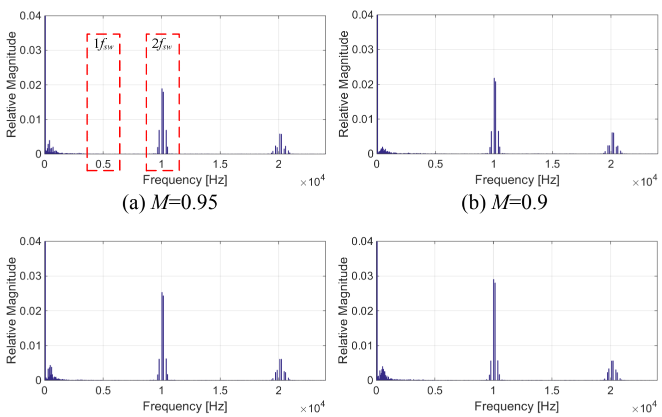
converters sharing the identical dc-link are connected in parallel and operated shifting the phase of one carrier by 180°. Also, experiments were conducted according to modulation index, M, which is changed by adjusting dc-link voltage.

In this part, the comparison results between Space-Vector PWM (SVPWM), which is generally used in various applications, and the proposed PWM are presented. Fig. 4 shows the line current waveforms when different PWM methods are adopted in the same condition. Namely, SVPWM is applied in Fig. 4(a), and the proposed PWM in Fig. 4(b). From Fig. 4, it can be seen that the harmonics have been conspicuously reduced and THD also has been reduced from 5.4% to 3.3% by simply changing PWM method.



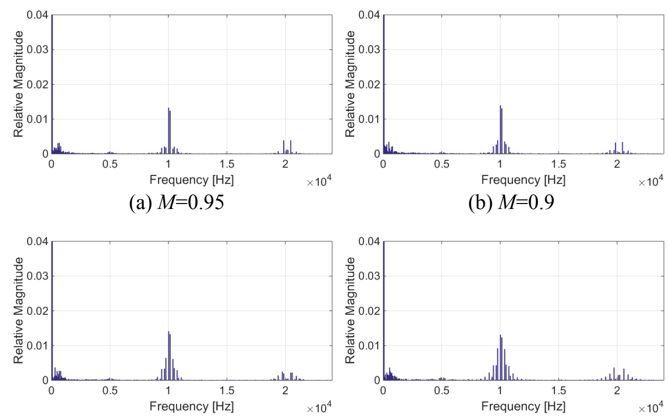
(a) SVPWM (M=0.8)   (b) Proposed PWM (M=0.8)

Fig. 5. Offset voltage of each PWM method(Experiment).



(a) M=0.95   (b) M=0.9
(c) M=0.85   (d) M=0.8

Fig. 6. Harmonic spectrum of line current under SVPWM(Simulation).



(a) M=0.95   (b) M=0.9
(c) M=0.85   (d) M=0.8

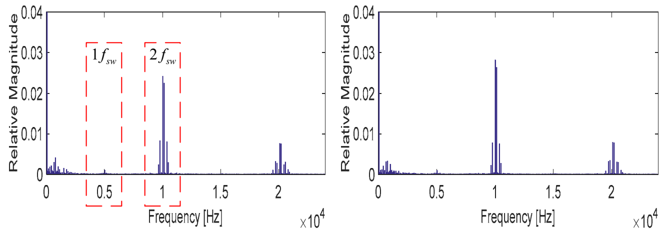
Fig. 7. Harmonic spectrum of line current under the proposed PWM(Simulation).

On the other hand, in Fig. 5, waveforms of the offset voltage according to each PWM method are presented. When SVPWM is applied, as shown in Fig. 5(a), sharp and triangular-shape v_{offset} with 3rd harmonic is generated. From the offset voltage, it can be imagined that the peak point of reference voltage is lowered, extending the linear synthesis range [10]. Meanwhile, v_{offset} in Fig. 5(b), has the similar form with the offset voltage of DPWM. This comes from the obtained v_{offset} which cannot satisfy the condition of (16) near the peak point of reference voltage. In this case, therefore, v_{offset} is determined by the boundary conditions, i.e., $v_{offset} = -V_{dc} / 2 - v_{\min}^*$ or $v_{offset} = V_{dc} / 2 - v_{\max}^*$, and become similar with that of DPWM.

Additionally, it can be noted that there are asymmetries in 3rd harmonic component in both offset voltages. They might be caused by several reasons including slight unbalances of inductances and switching devices, and discretization effects in the experimental system.

Fig. 6 and Fig. 7 show the simulation results of the harmonic spectrum of line current under various modulation index conditions when SVPWM and the proposed PWM are applied, respectively. Harmonics at odd multiples of switching frequency region are almost the same in both PWM methods since those harmonics are canceled out by interleaved operation. However, harmonics at twice the switching frequency region are much different. The proposed method reduces the maximum magnitude of the harmonic at target frequency region compared with SVPWM case. Moreover, reduction ratio increases as M is getting smaller. In the case of M=0.8 which is shown in Fig. 6(d) and Fig. 7(d), maximum harmonic at twice the switching frequency region has been reduced by 56%.

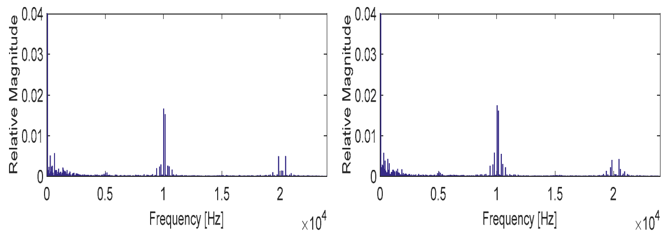
Similarly, Fig. 8 and Fig. 9 show the experimental results in the same condition with the previous cases. Some lower order harmonics near fundamental frequency are slightly increased than the simulation results, which is caused by nonlinear effect such as dead time and the voltage drop by switching elements. It would be simply eliminated by proper compensation methods [11]. Nevertheless, it can be noticed from the results that maximum harmonic at twice the switching frequency region has been reduced by 58% by only changing the PWM method. Therefore, it can be said that the proposed PWM method would satisfy the grid-code only using a smaller interface inductor by 58% from the inductor with the conventional SVPWM case.



(a) M=0.95 (b) M=0.9

(c) M=0.85 (d) M=0.8

Fig. 8. Harmonic spectrum of line current under SVPWM(Experiment).



(a) M=0.95 (b) M=0.9

(c) M=0.85 (d) M=0.8

Fig. 9. Harmonic spectrum of line current under the proposed PWM(Experiment).

V. CONCLUSION

In this paper, for paralleled interleaved two converters connected to the grid, a new PWM scheme reducing harmonic components at twice the switching frequency region in grid side has been proposed. Based on Fourier series analyses, v_{offset} that minimizes the square-sum of Fourier series coefficients at twice the switching frequency was calculated in every sampling period. Using the proposed PWM method which can be simply implemented in the software, each converter synthesizes the phase voltages reducing the harmonics at twice switching frequency region, where the interleaved operation cannot cancel out the harmonics in the region. Therefore, the proposed method would effectively suppress the harmonics, which is the utmost concern in inductive filter design. Proposed method has been confirmed by computer simulation and experiment results. Peak value of harmonics at twice the switching frequency region of the proposed PWM method was reduced by over 50% compared with those of SVPWM method in 0.8 modulation index. By only changing PWM method in the software, the harmonics can be reduced and it would result in the conspicuous shrink of the size and cost of the filter inductor between converters and the grid.

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