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Multilevel PWM Sinusoidal Inverter

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Abstract

This paper discusses two types of multilevel *PWM* sinusoidal inverter: one employs the multiplexing through addition of two-step switched *PWM* using different-phase sawtooth wave carriers and the other uses the multiplexing of compound *PWM*, an expanded version of the former. It also gives the configurations of these inverters and compares their characteristics.

In connection with the configurations and characteristics of the inverters, we also explained *PWM* waveform sequences with the use of single carriers obtained through their respective modulation methods, clarified the frequency spectrum distribution of the multiplex *PWM* waveform made through the addition of the above-mentioned sequence by means of double Fourier series expansions. It was found that the lowest angular frequency of the unnecessary elements of higher harmonics at N different-phase carriers is $N\omega_c - \omega_s$ and for both types, the output voltage level becomes N times the applied *DC* voltage and that while for both types the maximum value of higher harmonics is reduced as N increases, the reduction is larger with the compound *PWM* type than with the two-step switched *PWM* type.

In examining the possibility of a multiplexed inverter for commercial use, we devised a main inverter circuit with consideration to the need for a larger capacity two-step switched *PWM* type inverter. The output waveform measured with the prototype inverter we made shows that very satisfactory sinusoidal waves obtain with the main inverter circuit. Our experiment suggests that the proposed type of inverter could be useful as a *CVCF* power source or *UPS* as an *OA* equipment power supply. For the commercial use of the compound *PWM* type, it is expected that if a single-chip control signal generator is developed for use in combination with the main inverter circuit, an inverter with much improved characteristics would be realized.

Key words ; *PWM* sinusoidal inverter, Two-step switched *PWM* method, Compound *PWM* method, Frequency spectrum distribution, Double Fourier series expansion, Multilevel *PWM* sinusoidal inverter, Large power inverter, Uninterruptible power supply (*UPS*), Constant voltage constant frequency (*CVCF*) power source.

1. Introduction

With the development and spreading use of *OA* and *FA* machines, there are growing needs for improved output power waveforms with, for example, uninterrupted power supply (*UPS*) and also for higher levels of power and performance for motor control with industrial equipment.

In the past, the level of power has been increased by multiplexing the output of square-wave inverters with phase differences, and output waveforms of power supply have been improved through the use of inverters that allow unnecessary higher harmonics to be moved to the high frequency area as sideband waves of carriers for modulation. But the former involves unnecessary harmonic elements in relatively low frequency areas, making it difficult to reduce output wave distortion, while the latter requires increasing carrier frequencies to move unnecessary harmonics to higher frequency areas, causing the switching loss to increase.[(1)~(3)]

This paper will propose one type of multilevel *PWM* sinusoidal inverter designed to attain power as well as decrease unnecessary higher harmonic elements to improve the distortion of power supply output waveforms and explain its configuration and characteristics.

The multilevel *PWM* inverter uses the phase difference of unnecessary higher harmonic elements contained in the *PWM* waveforms which are compared and sorted out with carriers having different phases to cancel lower-level harmonics through multiplexing.[(8)(11)] As a result, the output waveform of the inverter follows the reference sinusoidal wave signals in an orderly manner; unnecessary harmonics elements, with reduced amplitude, are distributed in the high-frequency area; and the dynamic range between the reference signal and lower-limit unnecessary higher harmonic frequency can be expanded. This means that satisfactory sinusoidal waves can be output simply by adding a filter with a high cut-off frequency and a low level of attenuation. It is also possible to attain a larger capacity because the switching frequency can be kept lower as compared with the use of a single *PWM* inverter.[(9)]

In this paper, we will look at the basic principle of *PWM* multiplexing, theoretically analyze the characteristics of the two-step switched *PWM* method using sawtooth waves as a carrier by applying double Fourier series expansions, discuss them with experimental data obtained by using a prototype inverter, and then show the configuration and characteristics of an inverter multiplexed through the addition of *PWM* waves obtained individually with carriers having different phases. Then, we will discuss how a multilevel *PWM* sinusoidal inverter can be realized through a compound *PWM* method, which is an expanded version of the two-step switched *PWM* method. Then, we will make a comparison between the two types of inverter in terms of characteristics and the possibility of commercial applications.

2. Basic principle of *PWM* multiplexing

Fig. 1 gives a configuration diagram of a multilevel *PWM* inverter. To keep track of the

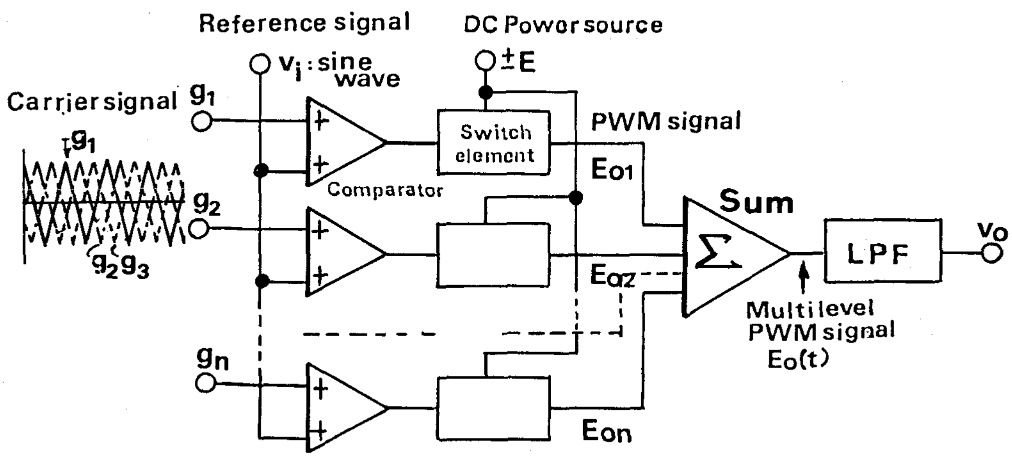

Fig. 1 Basic configuration of multilevel PWM sinusoidal inverter.

reference sinusoidal wave signal V_i, we obtain a sequence of *PWM* waveforms (E_{0i} (i=1, 2, . . . , n)) by chopping the *DC* power supply voltage *E*, approximate V_i with the multilevel *PWM* waveform $E_0(t)$ obtained by adding the sequence of *PWM* waveforms, and take out a low-distortion inverter output V_0 by eliminating unnecessary higher harmonic elements contained in $E_0(t)$ by means of a low-path filter (*LPF*).

Let $V_i=E_i \sin \omega_s t$ be the reference sinusoidal wave signal, and the $E_0(t)$ be the non-addition *PWM* waveform by carrier g_1 with an angular frequency of ω_c ($\omega_c > \omega_s$). Then, this can be expressed by a double Fourier series as follows:

$$E_0(t) = \sum_{m,\, n=-\infty}^{\infty}\sum C(m,\ n)\, e^{j(m\omega_s + n\omega_c)} \tag{1}$$

where, $C(m,\ n)$ is Fourier coefficents

The principle of this multiplex *PWM* inverter is to eliminate unnecessary higher harmonics through the *LPF* and obtain only the V_i signals as output.

Let $C_1(m,\ n)$ be the Fourier series where the reference sinusoidal wave signal and carrier are shifted by θ and, ϕ, respectively. Then, it becomes as follows:

$$C_1(m,\ n) = e^{j(m\theta + n\phi)} \cdot C(m,\ n) \tag{2}$$

From expression (2), the following expression can express Fourier coefficient $C_N(m,\ n)$ obtained when adding the sequence of *PWM* waveforms E_{01}, E_{02}, . . . , E_{0n} each chopped by carriers g_1, g_2, . . . , g_N, each of which are shifted by $(2\pi n/N) \cdot C_{k-1}$ (k=1, 2, . . . , N):

$$\begin{aligned} C_N(m,\ n) &= \left[e^{j\frac{2\pi n}{N}} + e^{j\frac{2\pi n}{N}\cdot 2} + \cdots + e^{j\frac{2\pi n}{N}(N-1)}\right] \\ &\quad \cdot C(m,\ n) \\ &= Z \cdot C(m,\ n) \end{aligned} \tag{3}$$

where, the phase of g_i (i=1, 2, . . . , N) is $(2\pi n/N)(k-1)$ (k=1, 2, . . . , N).

In this expression, Z equals 0 or N depending on whether n/N is a non-integer or an integer.

It follows from this that the frequency spectrum of $E_0(t)$ distributes with only the carrier element of the n-fold frequency and its $N\omega_c$ and sideband element being multiplied by n, eliminating the other harmonic elements. As a result, multiplexing makes it possible to move unnecessary higher harmonic elements to the high frequency area and reduce the amplitude of the harmonic elements when a single carrier is used. At the same time, it is possible to multiplex the output voltage onto the *DC* power supply voltage *E* through addition of the *PWM* waveform sequence E_{0i} (*NE*), leading to improved output voltage waveforms and increased power.[6]

3. Configuration and characteristics of sawtooth-carrier multilevel *PWM* inverter

In the multiplexing of *PWM* using multiple carriers with different phases, the Fourier coefficient $C_N(m,\ n)$ of the multilevel *PWM* waveform $E_0(t)$ that approximates the reference signal waveform is given by expression (3). Decreasing this value involves selecting the carrier signal wave capable of degenerating the Fourier coefficient $C(m,\ n)$ during non-multiplexing, that is, in single-carrier *PWM* operations, and examining this *PWM* method.

Carrier signals can be triangle waves, sawtooth waves or special-pattern waves (e.g., cosine waves and lissajous waves). Research has been done for the commercial application of multiplexing using triangle waves. The author, too, showed that the use of sawtooth waves, rather than triangle waves, is suitable for this type of multiplexing. In this paper, therefore, we will examine two types of multilevel *PWM* sinusoidal inverter—one using the two-step switched *PWM* method employing sawtooth waves as carriers, and the other using the composite *PWM* method combining the two-stage switching *PWM* method and *DC* power supply voltage level switching. We will then look at their basic configurations and characteristics.

3-1 Multiplexing of two-step switched *PWM*[7][11]

Fig. 2 shows the principle of multiplexing the two-step switched *PWM* method with the phase difference of sawtooth wave carriers g_1

and g_2 set at 180°. Fig. 3 is an example of its circuit configuration.

As can be known from Fig. 2, this modulation method is a polarity inversion single-pole single-edge modulation, which we call two-step switched *PWM*), to cause the modulation operation to be switched from the front edge to the rear edge (or front edge) at the positive or negative peak of the reference sinusoidal voltage and also cause the polarity of the carrier to be switched at the zero point. The switching between the positive and negative gradient of a sawtooth wave is done by selecting the four

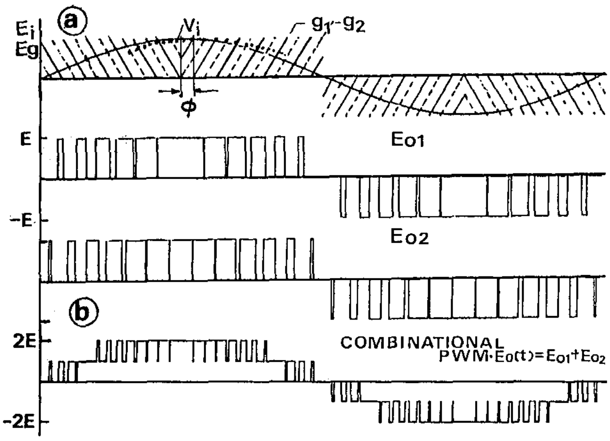

Fig. 2 Operation principle of 2-step switched modulation type multilevel PWM (N=2).

types of sawtooth wave element, shown in Fig. 3. Specific bias (Carrier), by driving the analog switch by means of the signals that detect the positive and negative peaks and zero point of V_i.

By comparing V_i and the different-phase carriers of g_1 and g_2 as shown in Fig. 2 (a), it is possible to generate the *PWM* wave sequences E_{01} and E_{02} with the respective carriers. Addition of these results in the multilevel *PWM* waveform $E_0(t)$ shown in Fig. 2 (b). The Fourier coefficient $C_{N=2}(m,\ n)$ that gives unnecessary higher harmonic elements of $E_0(t)$ becomes as follows:

$$C_{N=2}(m,\ n)=e^{j\frac{2\pi n}{2}}\cdot C_1(m,\ n) \tag{4}$$

The principle of this multiplex *PWM* inverter is to eliminate unnecessary higher harmonics through the *LPF* and obtain only the V_i signals as output.

Let ϕ be the phase angle due to the shift of the carrier switching operation at the peak. Then, the Fourier coefficient $C_1(m,\ n)$ of *PWM* waveform sequence E_{01} with carrier g_1 is given in the following expression:

$$C_1(m,\ n)=\frac{E}{2\pi^2 n}\{1-(-1)^m\}\Big[\frac{(-1)^{\frac{m-1}{2}}}{m}\sin m\phi$$
$$-\sum_{k=0}^{\pm\infty}J_{2k}(2n\pi M)\frac{(-1)^{\frac{m-2k-1}{2}}}{m-2k}\sin$$

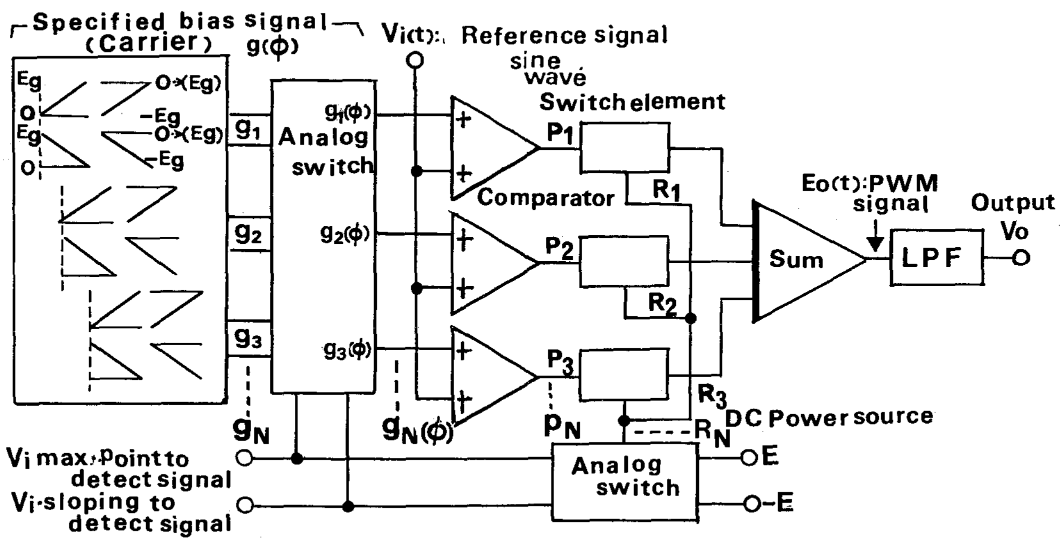

Fig. 3 Basic circuit construction of 2-step switched modulation type multilevel PWM sinusoidal inverter.

$$(m-2k)\phi\Big] + j\frac{E}{2\pi^2 n}\{1-(-1)^m\}$$
$$\Big[\frac{(-1)^{\frac{m-1}{2}}}{m}\cos m\phi - \frac{\pi}{2}\boldsymbol{J}_m(2n\pi M)$$
$$-\sum_{k=0}^{\pm\infty}\boldsymbol{J}_{2k}(2n\pi M)\frac{(-1)^{\frac{m-2k-1}{2}}}{m-2k}$$
$$\cos(m-2k)\phi\Big]\ ;\ \text{for } m \text{ odd} \qquad (5)$$

where, $M=E_i/E_g \leq 1$ and $\boldsymbol{J}_m(X)$ is Bessel function of m-th order.

The frequency spectrum becomes as shown in Fig. 4 (a). With the addition of *PWM* waveform sequence E_{02} resulting from carrier g_2 with a 180° phase delay from g_1, the frequency spectrum of multiplex *PWM* waveform $E_0(t)$ has a distribution of more than $2\omega_c$ sideband elements as shown in Fig. 4 (b). The unnecessary higher harmonic elements shift to the high frequency area through the multiplexing. The measured results showed a satisfactory level of agreement with the theoretical values.

The frequency spectrum of E_{01} shows that only the sideband wave elements for ω_c, $2\omega_c$, . . . , $n\omega_c$ distribute in an asymmetrical manner. The angular frequency of the lower-limit unnecessary higher harmonics is $\omega_c - 3\omega_s$ and the upper limit of the angular frequency of the reference sinusoidal wave becomes $\omega_{smax} = \omega_c/4$. The multiplexing of $N=2$ has the following features: the elements of $2\omega_c$, $4\omega_c$, . . . do not appear; the angular frequency of the lower-limit unnecessary higher harmonics is $2\omega_c - \omega_s$ shifts to the higher frequency area; the maximum value of unnecessary higher harmonic elements with respect to the reference signal elements can be degenerated to −18dB; the interval between the reference signal frequency and lower-limit unnecessary harmonic elements, that is, the frequency allowance, can be expanded; and the amplitude of unnecessary higher harmonic elements can be suppressed.

The displacement angle ϕ at the time of modulation switching operation has no influence on the frequency spectrum distribution if $\phi = \pm 15°$. It is practically easy to make a setting within this range. No consideration is, therefore, required with respect to

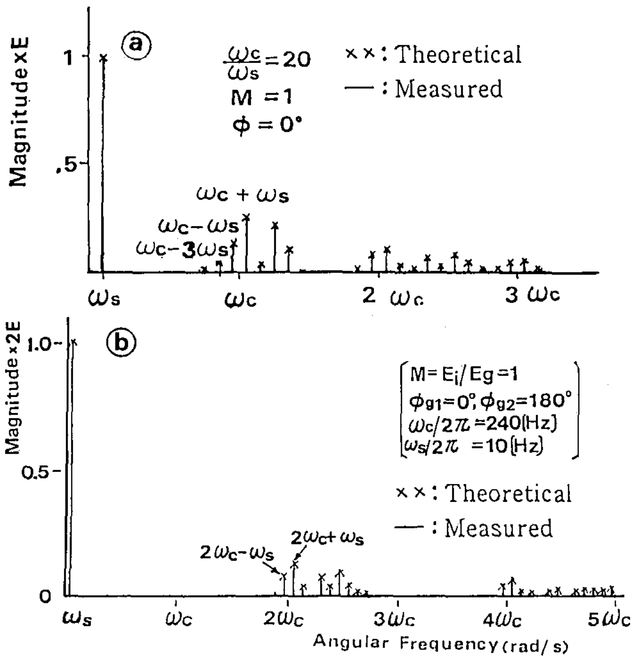

Fig. 4 ⓐ : 2-step switched PWM frequency spectrum.
ⓑ : Multilevel PWM frequency spectrum (N=2).

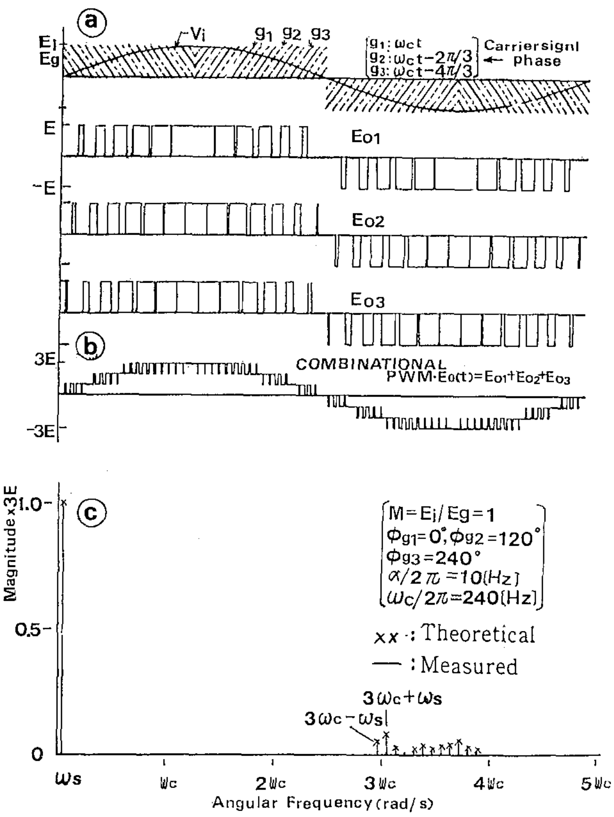

Fig. 5 2-step switched modulation type multilevel PWM operation and its frequency spectrum (N=3).

Fig. 5 shows the principle of multiplexing ($N=3$) of two-step switched *PWM* with the phases of g_2 and g_3 set at 120° and 240° against g_1, respectively (Fig. 5 (a)), and the frequency spectrum of $E_0(t)$. Fig. 6 shows the frequency spectrum of multiplex ($N=4$) *PWM* using a carrier with the phase delays of g_2, g_3, and g_4 from g_1 being 90°, 180°, and 270°, respectively.

It is known from these figures that the lower limit of unnecessary higher harmonic angular frequency is $3\omega_c-\omega_s$ for $N=3$ and $4\omega_c-\omega_s$ for $N=4$. It moves to the high frequency area as N increases. Also, as compared with the reference signal amplitude, the maximum amplitude of unnecessary higher harmonic elements can be reduced to −21.8dB for $N=3$, and −24.8dB for $N=4$. For the configuration of the inverter, the *LPF* can be made such that the cut-off angular frequency ω_{cut} and attenuation level are $3\omega_c/4$ with 14dB/oct for $N=3$, and ω_c with 12dB/oct, respectively. This can be realized easily by increasing N. The output voltage level of the inverter is $3E$ for $N=3$ and $4E$ for $N=4$. Multiplexing has the effect of enhanced utilization of the *DC* power voltage E. This also makes it possible to obtain a large-capacity power supply and an improved output waveform.

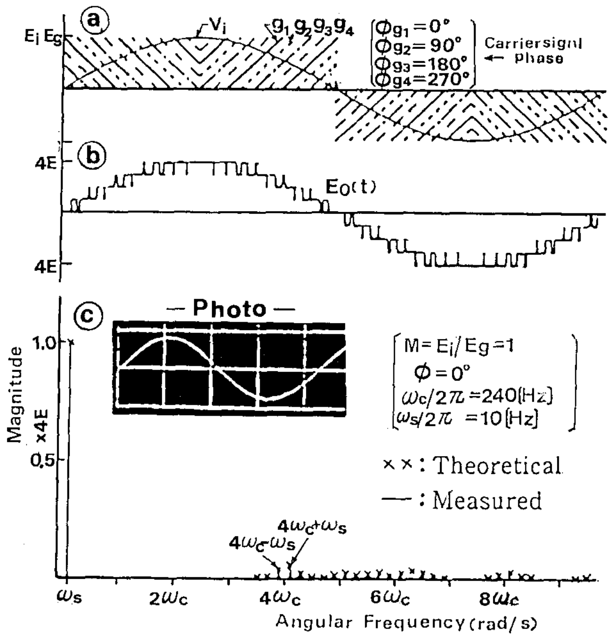

Fig. 6 2-step switched modulation type multilevel PWM inverter operation and its frequency spectrum (N=4).
(Photo; Experimental results due to multilevel inverter output voltage (V_0) waveform LPF: $\omega_{cut}/2\pi=100$[Hz], 12dB/oct.)

The photo given in Fig. 6 (c) shows an example of an output waveform of the prototype inverter. It is a sinusoidal wave with distortion being as low as 0.3%.

3–2 Multiplexing of compound *PWM*

Fig. 7 gives the basic principle of compound *PWM* incorporating the carrier level switching of the two-step switched *PWM* method plus the switching of *DC* power supply voltage levels. Fig. 8 shows an example circuit configuration of a multilevel inverter of this type.

This modulation method was devised for the purpose of degenerating the unnecessary higher harmonic elements contained in the *PWM* signals of the two-step switched *PWM* method while expanding the upper frequency limit of the reference signal.[4][5]

As shown in Fig. 7 (a) and Fig. 8, Specified bias signal (Carrier), this modulation method provides eight types of $g(\phi)$ for selection depending on the reference signal. That is, it is controlled by means of logical operations on the gradient, and positive and negative peaks, of the reference sinusoidal signal, as well as each 1/2-level ($\pm E_i/2$) detection signal of the amplitude voltage. In the interval where the amplitude of the reference sinusoidal wave V_i is greater than $E_i/2$, the *DC* power supply voltage $E/2$ is chopped off. The same operation takes place in the interval where V_i is negative. In this manner, the *PWM* waveform $E_0(t)$ is generated.

Fig. 8 shows the basic circuit of an inverter that adds compound *PWM* waveform sequence E_{01}, E_{02}, and E_{03} using different-phase carriers

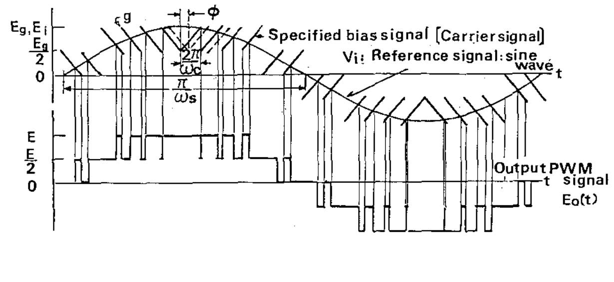

Fig. 7 Operation principles of compound PWM sinusoidal inverter.

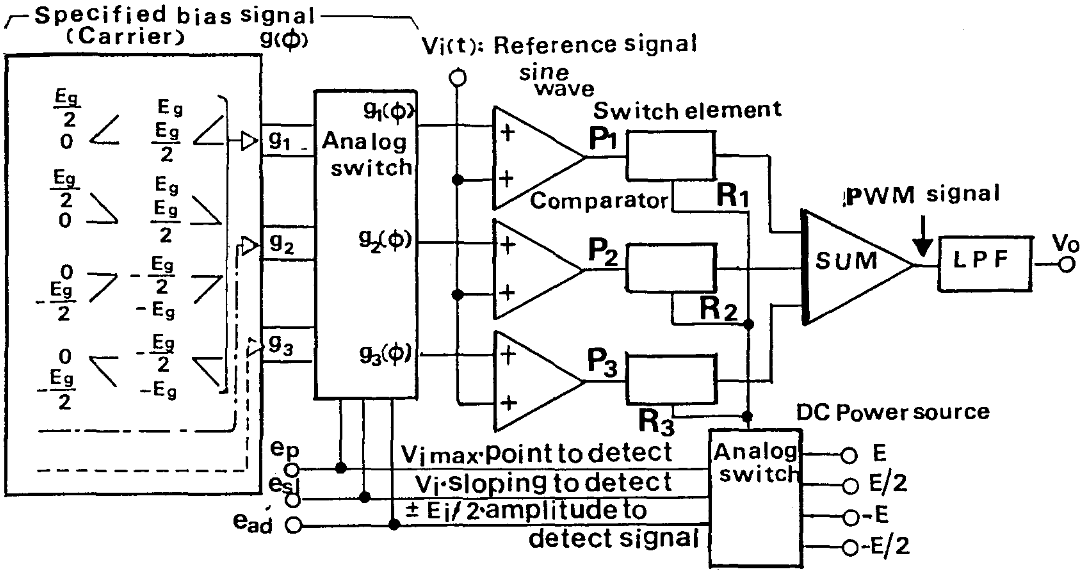

Fig. 8 Basic circuit construction of compound modulation type multilevel PWM sinusoidal inverter (N=3).

g_1, g_2 and g_3, eliminates the unnecessary higher harmonic elements from the multiplex *PWM* signal, and takes out the output sinusoidal wave V_0.[11]

As with the case of the multiplexing of two-stage switching *PWM*, the Fourier series $C_1(m, n)$ of the compound *PWM* waveform sequence E_{01} with a single carrier $g_1(\phi)$ (ϕ will hereafter be omitted) is given in the following formula, where ϕ is the displacement phase angle from the peak of the reference sinusoidal wave at the switching operation:

$$C_1(m, n) = \frac{E}{2n\pi^2}\Big[\frac{1}{m}(-1)^{\frac{m-1}{2}}\sin(m\phi) + \sum_{k=-\infty}^{\infty} \boldsymbol{J}_{2k}(4n\pi M) \frac{1}{m-2k}(-1)^{\frac{m-2k-1}{2}}\sin(m-2k)\phi + j\frac{1}{m}(-1)^{\frac{m-1}{2}}\cos(m\phi) - j\frac{\pi}{2}\boldsymbol{J}_m(4n\pi M) - j\sum_{k=-\infty}^{\infty}\boldsymbol{J}_{2k}(4n\pi M) \frac{1}{m-2k}(-1)^{\frac{m-2k-1}{2}} \cos(m-2k)\phi\Big] \quad (6)$$

where, m is odd, $M=E_i/E_g \leq 1$, and $\boldsymbol{J}_m(X)$ is Bessel function of m-th order.

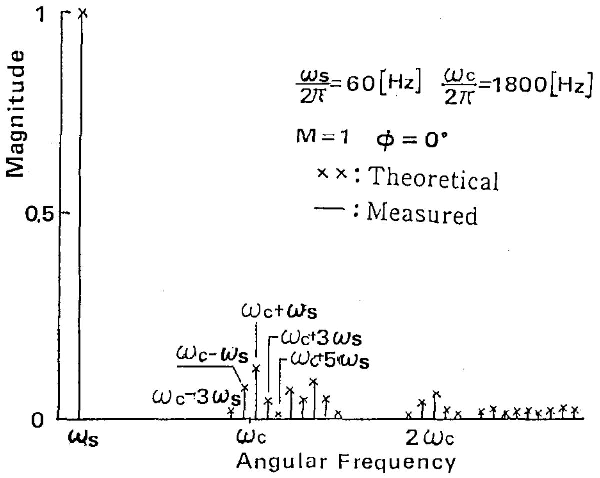

Fig. 9 Compound PWM frequency spectrum.

Fig. 9 gives the frequency spectrum distribution of E_{01}. If ϕ is between $\phi=\pm 13$ degrees, the distribution will not change: As with the case of two-step switched *PWM*, the lower limit angular frequency is $\omega_c - 3\omega_s$, but the maximum amplitude of unnecessary higher harmonic elements is reduced to 12.7% (−18dB) of the reference signal amplitude and the attenuation characteristic of the *LPF* is 14dB/oct, making it possible to obtain a sinusoidal wave output with the upper limit of ω_s reaching $\omega_c/4$ (distortion rate of 0.3% or less, $M=1$).

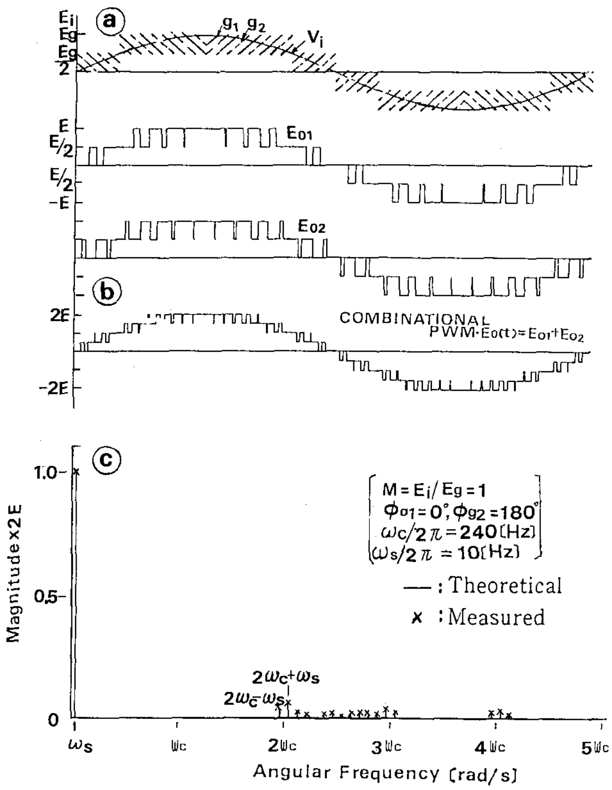

Fig. 10 Compound modulation type multilevel PWM operation and its frequency spectrum (N=2).

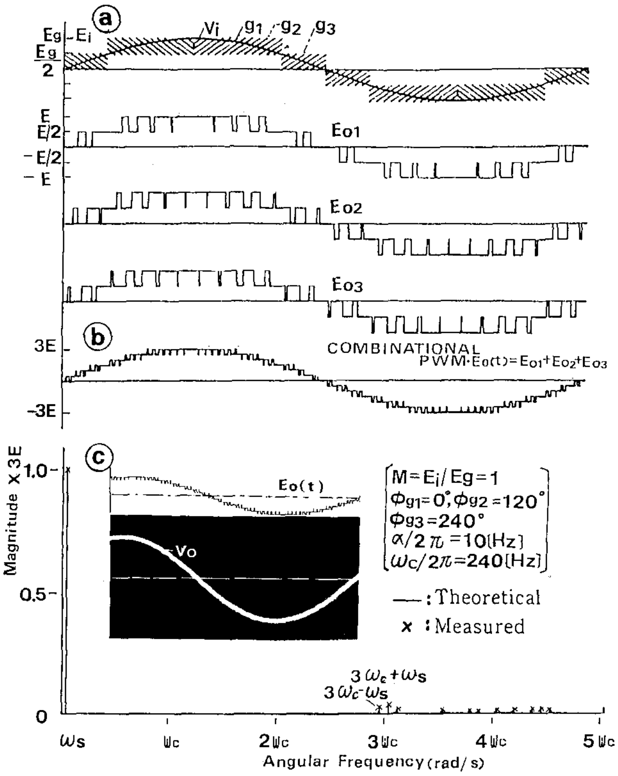

Fig. 11 Compound modulation type multilevel PWM sinusoidal inverter operation and its frequency spectrum (N=3).

(Photo; Experimental results
upper is multilevel PWM signal waveform ($E_0(t)$).
lower is output voltage waveform (V_0)
LPF: $\omega_{cut}/2\pi=100$[Hz], 12dB/oct.)

With the frequency spectrum of multilevel *PWM* waveform $E_0(t)$ with the addition of the compound *PWM* waveform sequence E_{02} using carrier g_2 with a 180-degree phase difference from carrier g_1 (Fig. 10 (a) and (b)), the maximum level of unnecessary higher-harmonic frequency amplitudes is suppressed to half (-26.6dB) that obtained when no multiplexing is done (Fig. 10 (c)). The lower limit of angular frequency is $2\omega_c-\omega_s$ and at this frequency, the amplitude of unnecessary higher harmonics can be reduced to 6% of the reference signal amplitude. Also, the level of the *LPF* can be 12dB/oct or less, making its configuration easy. The level of the sinusoidal output voltage is $2E$.

Fig. 11 is the frequency spectrum of $E_0(t)$ of multiplexed composite *PWM* with the phases of carriers g_1, g_2 and g_3 set at 0, 120 and 240 degrees, respectively (that is, $N=3$): The lower limit of the angular frequency of unnecessary higher harmonics becomes $3\omega_c-\omega_s$, and its maximum amplitude can be suppressed to about 3.5% to 4.05% (-28 to -30dB) of the reference signal amplitude. That is to say, the multiplexing of compound *PWM* gives an edge over the multiplexing of two-step switched *PWM* in terms of the degeneracy of unnecessary higher harmonic amplitude. This means that 10dB/oct is adequate for the attenuation characteristic of the *LPF*, making its configuration easy. The output waveform example given in Fig. 11 (c) was a waveform actually measured with the prototype circuit shown in Fig. 8. It shows a satisfactory output sinusoidal wave. With this type of inverter, the voltage level of the output sinusoidal wave is $3E$, and the upper limit of the angular frequency of the reference signal is $3\omega_c/4$, up 25% over the level with $N=2$.

With this method, too, as discussed above, a tracking-type (following the reference signal) multilevel *PWM* sinusoidal inverter can be realized by increasing the number of different-phase carriers and adding thus-obtained composite *PWM* waveform sequences. From the viewpoint of practical use, however, it seems desirable to limit the number of different-phase carriers to three to four for circuit configuration purposes.

3-3 Comparison between two types

Table 1 gives a comparison between the multiplexing of two-step switched *PWM* and the multiplexing of compound *PWM* in terms of the maximum level of the higher harmonic amplitude of *PWM* waveform and the angular frequency of lower-limit unnecessary higher harmonics appearing on the lowest frequency side as classified by the number (N) of different-phase carriers used.

Table. 1 Comparison of modulated method
[C(1, 0)=0dB]

N	heading	maximum harmonic component (dB)		earest component to ω_s (rad/s)	
	modulated type	2-step switched	compound	2-step switched	compound
2		-18.0	-26.6	$2\omega_c-\omega_s$	
3		-21.8	-30.03	$3\omega_c-\omega_s$	
4		-24.5	-33.7	$4\omega_c-\omega_s$	

With this, both types of multiplexing have the same lower limit of unnecessary higher harmonics, which is $N\omega_c-\omega_s$ ($N=1, 2, \ldots, 4$). In terms of the maximum level of higher harmonics elements, however, the multiplexing of compound *PWM* is about 32% ($N=2$), and 27% ($N=3$ and 4) lower than the multiplexing of two-step switched *PWM*. This means that the former requires a less demanding, easier-to-make, *LPF* than the latter does. With respect to the configuration of inverter circuity, however, the compound *PWM* method is more complex than the two-step switched *PWM* method because the former involves the level switching control of *DC* power supply voltage in the process of modulation. There is a possibility of using triangular wave carriers instead of sawtooth wave carriers. In this case, the lower limit of the angular frequency of unnecessary higher harmonics is $2\omega_c-5\omega_s$ for $N=2$, $3\omega_c-6\omega_s$ for $N=3$, and $4\omega_c-7\omega_s$ for $N=4$. The more the level of multiplexing, the more the number of carrier sideband elements on the side of the reference signal side. Besides, the maximum level of higher harmonic elements is 4% to 5% higher with triangular carriers than with sawtooth carriers. That is, the proposed two types are superior in terms of the degeneration of unnecessary higher harmonics. Irrespective of the type of carrier used, multiplexing has an effect of making the output voltage level N times the *DC* power supply voltage E.

As discussed above, both types of multilevel *PWM* inverter are characterized by the fact that higher harmonic elements can be degenerated, the output waveform can be improved and the output voltage can be enhanced to NE. But all these characteristics assume $M=1$. In discussing the inverter's controllability, changes in the frequency spectrum for $M\neq1$. From the viewpoint of switching loss, it is desirable to make the value of ω_c/ω_s as small as possible. It is, however, expected that decreasing this value would increase the distortion of the output waveform.

Let us therefore define the harmonic distribution coefficient (distortion ratio) d of a multiplex *PWM* waveform by the following formula:

$$d=\frac{\sqrt{\sum_{k=2}^{\infty}(V_k/k)^2}}{V_1}\times 100[\%] \tag{7}$$

where, V_1 is $C(1, 0)$, V_k is the sum of all $C(m, n)$ for $m+n=k$.

Now let us have a look at the d characteristic with respect to changes in the value of M with ω_c/ω_s as a parameter.

Fig. 12 gives the computed values of d obtained by changing M with the value of ω_c/ω_s set at 10, 20 and 30 for the multilevel *PWM* sinusoidal inverter shown in Fig. 6. When the inverter was used in the neighborhood of $M=1$, the value remained almost unchanged at the minimum level of 0.3%, permitting making the value of ω_c/ω_s low. If ω_c/ω_s is set equal to or greater than 20, the d value for $M=0.4$ to 1.05 is almost constant within 0.5%. If the tolerance of d is increased to 1%, it is possible to make E_i variable within the range of $M=0.2$ and 1.07.

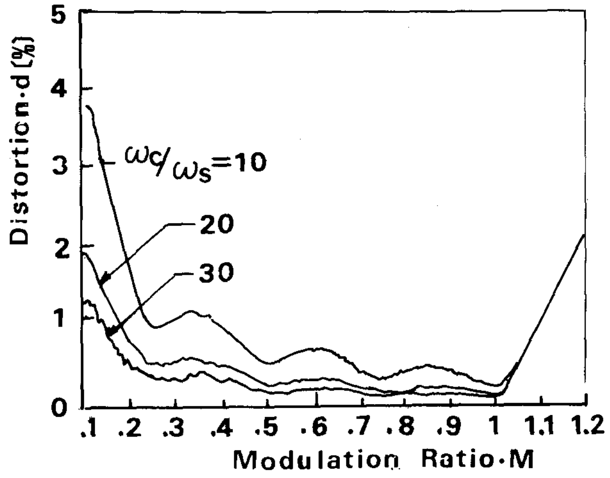

Fig. 12 Distortion of PWM inverters waveform (2-step switched modulation type, N=4).

This means that the inverter is capable of changing the output voltage by about 80% for control purposes. To obtain a well-formed sinusoidal output from this multilevel *PWM* waveform requires adding an *LPF*. Since an *LPF* with a high cut-off frequency and a low attenuation characteristic is desirable, the ideal inverter operation should take place at around $M=1$. It can therefore be said that this inverter can be better used for *CVCF* and *UPS*, although it can be used for control purposes as well.

It was found that the multiplexing of compound *PWM* gives a lower distortion of *PWM* waveform than with the multiplexing ($N=4$) of two-step switched *PWM* shown in Fig. 12, producing a very satisfactory output waveform.

These inverters can be realized by means of the basic circuit configurations shown in Figs. 3 and 8 if a small capacity will do. Making a large capacity inverter, however, involves discussing how to incorporate a power device in the *DC* power supply voltage cut-off switching element and configure an adder, that is, how to realize the main circuit of the inverter.

Fig. 13 is a sample configuration of the main inverter circuit of Fig. 3 for an increased capacity: Power transistors are used for the switching devices of T_{11} to T_{14}, and T_{21} to T_{24} which are driven by control signals P_1 to P_4. Auxiliary circuits comprising T_{11}' to T_{14}' and T_{21}' to T_{22}' are added for the polarity changing involved in the switching operations. When $N=2$, two-step switched *PWM* waveform sequences E_{01} and E_{02} are added through a current-balance reactor L_1 and a sinusoidal output voltage is obtained through an *LPF* that eliminates the unnecessary high harmonics from the multiplex *PWM* waveform at terminal P_0.[9]

When $N=4$, E_{03} and E_{04} at terminals Q_1 and Q_2 are added by means of L_2. The multiplex *PWM* waveform obtained by adding the multiplex *PWM* waveform at terminal Q_0 to E_{01} and E_{02} at terminal P_0 is further added by means of current-balance reactor L_3. In this manner, like the case of $N=2$, a multiplex *PWM* waveform is obtained at terminal R_0 to obtain a sinusoidal output through the *LPF*. While additional main switching elements are required, since they are connected in parallel, the capacity of each individual element can be reduced. The

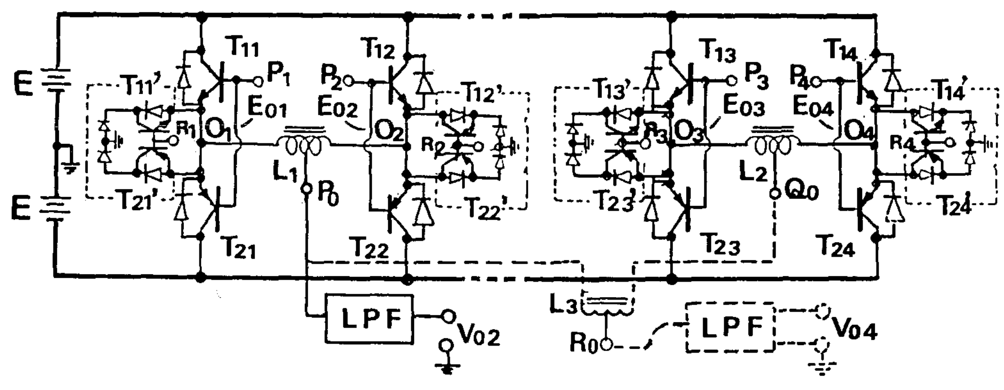

Fig. 13 Circuit configuration of main multilevel inverter (2-step switched modulation type).

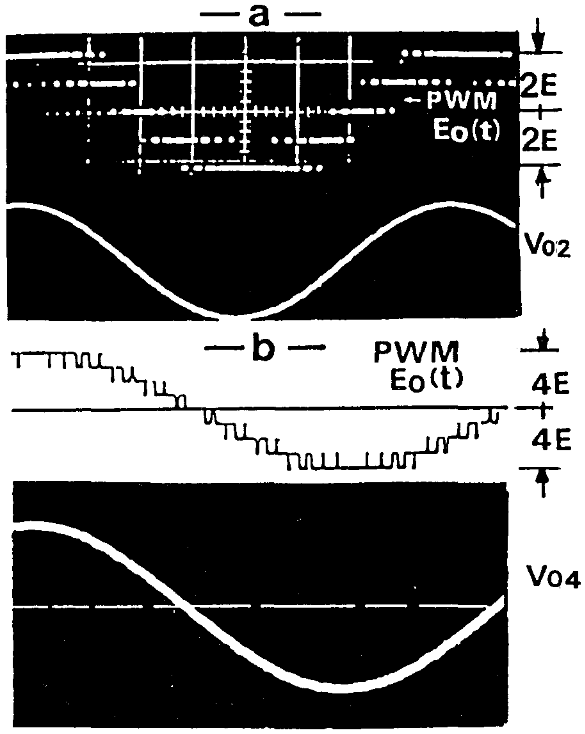

Fig. 14 An experimental results
a ; N=2, {upper is multilevel PWM signal ($E_0(t)$) waveform. lower is output voltage (V_{02}) waveform.
b ; N=4, {upper is multilevel PWM signal ($E_0(t)$) waveform. lower is output voltage (V_{04}) waveform.
($\omega_s/2\pi=60$[Hz], $\omega_c/2\pi=1200$[Hz]
LPF: $\omega_{cut}/2\pi=200$[Hz], 12dB/oct)

circuit is therefore suitable for a large-capacity inverter. While the circuit uses two more reactors than with the case of $N=2$, they can be substantially reduced in size. The *DC* voltage applied to this circuit can be reduced by half while the frequency can be doubled. Besides, the maximum time of voltage application is as short as one fourth the cycle of the carrier sawtooth wave (1/2 for $N=2$). Therefore, the size of reactors used in an inverter of the same capacity as with $N=2$ can be reduced to about one fourth.

Fig. 14 gives sample output waveforms actually measured with the prototype main inverter circuit controlled by means of two-step switched *PWM* signals obtained at terminals P_1 to P_4 and R_1 to R_4 using different-phase sawtooth carriers given in Fig. 3.

Fig. 14 (a) shows a multiplex *PWM* waveform $E_0(t)$ for $N=2$ and an output sinusoidal voltage waveform (V_{02} taken through an *LPF* with a cut-off frequency of 100Hz and an attenuation characteristic of 12dB/oct. Fig. 14 (b) is a multiplex waveform for $N=4$ and an output sinusoidal voltage waveform V_{04} obtained by using the same *LPF*. In both cases, the frequency of the output sinusoidal voltage was 60Hz, and the effective voltage and current were 100V and 10A. The distortion of the waveform was as satisfactory as 0.5% for $N=2$ and 0.3% or less for $N=4$. The *DC* power supply voltage for $N=4$ was half that for $N=2$. The *LPF* was capable of reducing the waveform distortion to 0.3% or less even with the use of the inductance (L) element alone. This is useful for a power supply for *OA* equipment.

We will not discuss the commercial multiplexing circuity of compound *PWM* here because efforts are now being made to develop its main inverter circuit and *PWM* control signal generator as integrated in a single-chip computer.

4. Conclusion

In this paper we examined two types of DC-AC converting multilevel *PWM* sinusoidal inverter designed to provide an improved output waveform and increased capacity: one is realized through the multiplexing of two-step switched *PWM* using different-phase sawtooth carriers and the other through the multiplexing of compound *PWM*, or an expanded version of the former. We also had a comparative look at their configurations, features and characteristics.

We established the logical basis for the multiplexing of two-step switched *PWM* by giving theoretical expressions to obtain the frequency spectrum of multiplex *PWM* waveforms through the addition of *PWM* waveform sequences obtained each of the different-phase carriers used. We also obtained the frequency spectrum of two-step switched *PWM* waveform sequences with sawtooth carriers by applying double Fourier series expansions and checked this theoretical analysis with measured data for validity. Then, we examined frequency spectra for multiplex *PWM* by varying the number (N) of different-phase carriers from 2 to 4. It was found that with multiplexing, the lower limit of unnecessary higher harmonics shifted to the high frequency area $N\omega_c-\omega_s$, resulting in an expanded dynamic range. At the same time, the upper limit of higher harmonics was suppressed, showing the effectiveness of multiplexing for the applied *DC* voltage level.

We also examined the multiplexing of compound *PWM* with focus on compound *PWM* waveform sequences and the frequency spectrum of the multiplex *PWM* waveform obtained through the addition of these sequences. As a result, it was found that with this method, the lower limit of the angular frequency of unnecessary higher harmonics was the same as with the two-step switched *PWM* method, but the maximum value of higher harmonics could be much more reduced. This in turn would make *LPF* configuration easier and permit obtaining a less-distorted sinusoidal output wave.

We made a comparison of the two types with respect to the characteristics of unnecessary higher harmonics and obtained a useful guideline for making these multiplex inverters. Furthermore, we examined a main inverter circuit for the purpose of attaining an increased capacity with the two-step switched *PWM* method. It was found that since this uses the parallel connection of power switching elements, it is possible to reduce the capacity of each individual element, permitting making an inverter with an increased capacity. It was found that while three current-balance reactors are required for the configuration of $N=4$, they can be made substantially smaller in size than those used for the configuration of $N=2$. Our prototype multilevel *PWM* sinusoidal inverter using this main inverter circuit generated low-distortion, satisfactory output waveforms, showing that it would be useful as a *CVCF* inverter or *UPS* for *OA* equipment. Efforts are now being made to develop a main inverter circuit, together with a *PWM* control signal generator, for the multiplexing of composite *PWM* as integrated in a single-chip computer. The author plans to present a separate report on this.

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