Design Guidelines of Circuit Parameters for Modular Multilevel Converter with H-bridge Cell

Toshiki Nakanishi Student Member, Jun-ichi Itoh Senior Member

(Manuscript received Aug. 3, 2016, revised Jan. 4, 2017)

This paper presents theoretical formulae for circuit parameter design as design guidelines in a step-down rectifier in a power system connected to a 6.6-kV AC power grid, where a modular multilevel converter (MMC) is applied. In particular, this paper focuses on the design of heat sinks, capacitors, and arm inductors. In addition, the worst case for each component design is shown as the first step of the converter design. First, the formula for the ripple current in the electrolytic capacitor is derived in order to evaluate the capacitor lifetime. Second, the formulae for the semiconductor losses are clarified on the basis of the analysis of the arm current. Third, the formula for the ripple current in the arm inductor is derived on the basis of the ripple current model of a chopper circuit. Finally, all formulae are verified by experiments with a miniature model. It is confirmed that theoretical values from formulae agree with the measured values with the errors lower than 12%.

Keywords: modular multilevel converter, high power density design, H-bridge cell, step-down rectifier

1. Introduction

Recently, DC micro-grids have been actively researched as next generation power distributions[10]. In the DC microgrid, a transformer and a rectifier are required in order to convert from 6.6-kV AC voltage to several hundred volts DC voltage[20]. However, the isolated transformer is bulky due to the operation at the utility frequency. In addition, the conventional power system for 6.6-kV AC grid also requires bulky static capacitors for power factor correction and series reactors for the reduction of the harmonic distortion of the grid current[21].

For the volume reduction of the above system, neutral-point-clamped (NPC) converters or diode-clamped converters have been actively researched in order to avoid the utilization of the bulky transformer[8]. In the three-level diode-clamped converter, the high voltage IGBTs with the voltage rating more than 6.5 kV are required. However, it is generally difficult to employ the IGBTs with the voltage rating more than 6.5 kV due to their unique specifications. Moreover, if the five-level diode-clamped converter is applied to solve the problem with the switching devices, the voltage balancing circuit is required as the auxiliary circuit in order to correct the unbalanced voltage among four capacitors in the DC link[5]. In addition, the above converters cannot achieve step-down rectification. Therefore, the specification of the NPC converter are different from the requirements of the proposed system.

On the other hand, the employment of a modular multilevel converter (MMC) with an H-bridge cell can achieve the step-down operation without the isolated transformer in the DC micro-grid[22]. The MMC has been actively researched as a next generation high voltage power converter without the bulky transformer[23][24]. In addition, the H-bridge cell type MMC also achieves the reduction of static capacitors and series reactors because the MMC controls the input power factor and reduces the harmonic distortion without large capacitors and inductors[25]. Besides, it is possible to apply 1.7-kV or 1.2-kV IGBTs with the low loss characteristic with the increase in the number of cells in the MMC. Therefore, these IGBTs can operate at the high frequency for the volume reduction.

On the other hand, in the output side of the MMC, an isolated DC-DC converter is connected to achieve the isolation of the AC power grid and the DC micro-grid[26]. Thus, the high step-down function from 6.6 kV to several hundred volts is accomplished by both the MMC and the isolated DC-DC converter. In the H-bridge cell type MMC, the degree of freedom in the design of the output DC voltage is high because the output DC voltage of the MMC is easily controlled by adjusting the DC bias in each cell output voltage[27]. Moreover, the rating of the circulating current (DC component) changes responding to the output voltage when the power rating is same. Thus, it should determine the combination of the output DC voltage and the circulating current for the volume reduction of the MMC and the isolated DC-DC converter.

In AC-DC converters using the MMC topology for the medium voltage application, (i) capacitors, (ii) heat sinks and (iii) arm inductors are the majorities of the overall volume as same as general power converters[28]. The design guideline in order to achieve each individual purpose such as the high power density or high efficiency has been considered so far[29][30]. First, a capacitor design based on the ripple voltage have been reported[31][32]. In the MMC, the capacitor voltage includes the ripple component. Consequently, the capacitance is generally determined dependent on the ripple factor as with general power converters. Moreover, the formula of the capacitor voltage included the ripple voltage has already clarified. In addition, customized film capacitors or ceramic capacitors under a fixed number of cells have also been investigated as the cell capacitor[20][21]. When the film
capacitor or the ceramic capacitor is used in the MMC, the capacitor volume is evaluated by the capacitance and the electrostatic energy (21)–(23). On the other hand, the employment of the electrolytic capacitor to the cell in the prototype or the miniature model of the MMC has also been reported (24)–(26). In general, it is commonly known that the capacitance per unit volume of the electrolytic capacitor is high compared to film capacitors or ceramic capacitors. Moreover, the required voltage rating is simply achieved by changing the number of the series connected capacitors. Besides, in terms of the lifetime, it is possible to adjust the ripple current of one capacitor by changing the number of the parallel connected capacitors. In the MMC, the capacitor voltage rating changes depending on the number of cells because the divided voltage of each cell is varied. In addition, the ripple current of the capacitor is varied by the output DC voltage of the MMC because the DC component of the arm current changes. However, the ripple current of the cell capacitor has not been thoroughly considered although the ripple current is the important factor in the design guideline for the electrolytic capacitor. In addition, the formulae of the ripple current and the ripple voltage focused on the number of cells and the output voltage of the MMC have not been also clarified.

Second, in terms of the heat sink design, the loss analysis and the breakdown of the semiconductor losses have been reported in the chopper cell type MMC (27)–(28). Besides, the evaluation of the heat sink volume in IGBTs with the voltage rating of which varies from 600 V to 6.5 kV has been also discussed (29). However, the formulae of the semiconductor losses in the H-bridge cell type MMC have not been thoroughly clarified. In the H-bridge cell type MMC, the arm current is varied by the input active power i.e. the output power because the circulating current changes. In particular, the arm current becomes AC including the DC component or DC including the AC component. Hence, it is necessary to analyze the current path in the H-bridge cell and the magnitude of each device current based on the arm current. Thus, the semiconductor loss should be clarified based on the analysis of the arm current. However, the detail analysis, the process of the formula clarification and the verification of the theoretical formula have not been considered.

Third, as the design of the arm inductor, the design guideline which the inductance is determined to be several percent of the impedance rating of the MMC has been explained (30)–(31). In addition, the design guideline which focuses on the harmonic components of the circulating current and the resonance phenomenon between the arm inductor and capacitors in cells has been also considered (32). On the other hand, as the design guideline for the inductor in general power converters, the inductance is determined based on the ripple factor of the inductor current (33)–(34). In the MMC, the ripple current changes responding to the switching frequency and the number of cells due to the change of the equivalent switching frequency. In particular, when the unipolar modulation is used in the H-bridge cell, the equivalent switching frequency is different compared to the MMC with the chopper cell. Thus, it is greatly important to consider the relationship between the ripple current of the arm inductor and the equivalent switching frequency including the number of cells as the variables. However, when the unipolar modulation is applied, the design formula of the ripple current in the arm inductor based on the relationship between the ripple current and the equivalent switching frequency has not been thoroughly reported.

The problem of the conventional design method is that the number of cells and the output voltage of the MMC are not focused in the design of the H-bridge cell type MMC. Besides, the fundamental design guideline, i.e. the worst case for the ripple component or the semiconductor loss, has not also been considered due to the above reasons. The formulae for the ripple component or the semiconductor loss have many variables in the MMC. Hence, it is difficult to find the worst case for each circuit component design. In addition, it is also difficult to acquire required values such as the minimum ripple current and the minimum semiconductor loss with circuit simulations, because there are many factors which affect the ripple component or the semiconductor loss. Therefore, it is necessary to clarify the theoretical formulae to find the worst case for each component design in the H-bridge cell type MMC.

This paper presents the theoretical formulae to find the worst case for the circuit parameter design in the H-bridge cell type MMC. The main contribution of this paper is to show the factors which decides the worst case for each component design after the clarification of the theoretical formulae focused on the number of cells and the output voltage of the MMC. It is expected that the discussion of the design for the optimum number of cells and the optimum value of the output voltage will be promoted to achieve for the minimum ripple component or the minimum semiconductor loss. This paper is a first step of the optimum design for the H-bridge cell type MMC.

In this paper, the formulae of (i) the ripple current of the cell capacitor, (ii) semiconductor loss and (iii) the ripple current of the arm inductor are derived. First, the formula of the ripple current in the cell capacitor is derived for the lifetime design of the electrolytic capacitor. It is necessary to clarify the relationship between the input power factor and the ripple current because the ripple current changes according to the input active power i.e. the output power. Second, the conduction loss and the switching loss in switching devices are derived based on the analysis of the arm current. The semiconductor losses change depending on the arm current. Thus, it is required to obtain the relationship between the arm current and each switching device loss in order to design the heat sink in the maximum point of the semiconductor loss. Third, the formula of the ripple current in the arm inductor is derived in order to determine the inductance of the arm inductor. From the relationship between the duty which changes depending on the number of cells and the ripple current, the maximum point of the ripple current is clarified. Finally, the theoretical formulae are verified by using a miniature model in order to obtain the validity of the theoretical formulae.

2. Circuit Configuration

2.1 Conventional Power System Connected to Utility Grid of 6.6 kV

Figure 1 shows a structure of the conventional power system of 200 kVA connected to the AC power grid of 6.6 kV. The model of this system is “cubicule-type high-voltage power receiving equipment” which is widely applied in Japan (40)–(42). The conventional power system has several isolated transformers in order to convert the grid voltage of 6.6 kV into the AC distribution voltage of 200 V or 100 V. Moreover, the power system also includes the static capacitors in order to correct the input power factor and the series reactor in order to reduce the harmonic distortion of the grid current. Transformers, static capacitors and series reactors are main factors which increase the system volume. In the conventional system, the component volume which is defined as the total volume of transformers, static capacitors
2.2 Front-end Converter with MMC and Isolated DC-DC Converter Figure 2 shows the configuration of the front-end converter using the H-bridge cell type MMC and the isolated DC-DC converter. Each arm of the MMC consists of the arm inductor $L_a$ and cascaded H-bridge cells in order to operate as the step-down rectifier. The MMC outputs a multi-level voltage waveform which reduces the harmonic distortions of the input current. In addition, the MMC is able to reduce the voltage rating of devices on each cell due to cascade connections of cells. Therefore, lower voltage rating devices are utilized. On the other hand, the output DC voltage of the MMC depends on the sum of the output average voltage of cells. Thus, the cell output voltage includes both the AC component and the DC component to control the input current and the output voltage. Moreover, the MMC converts the high AC voltage to the DC voltage of several hundred volts. After this step, the DC voltage of 400 V is supplied to the DC bus of the DC distribution by the isolated DC-DC converter. The isolation capability between the AC power grid and the DC micro-grid is achieved by the isolated DC-DC converter. Hence, the high step-down functions from 6.6 kV to several hundred volts is accomplished by both the MMC and the isolated DC-DC converter. As a result, the output voltage rating of the MMC can be designed with a high freedom degree. In the MMC, the circulating current (DC component) is changed responding to the output voltage of the MMC under the conditions of same power rating. Thus, a variety of the combination between the output voltage and the circulating current can be determined for the volume reduction of the MMC and the isolated DC-DC converter.

3. Control Strategy of Step-Down MMC

3.1 Control Concept for MMC Figure 3 shows a control block diagram of the H-bridge cell type MMC (10). The control block consists of the capacitor voltage control, the arm current control and the MMC output voltage control. In addition, the capacitor voltage control is constructed by the capacitor voltage averaging control and the capacitor voltage balancing control.

3.2 Capacitor Voltage Averaging Control The capacitor voltage averaging control is employed in order to maintain the average value of all capacitor voltages in the arm (10). Therefore, the average value of all capacitor voltages has to be calculated in each arm. The output value of the PI controller is given as the command of a positive sequence component in the arm current. Moreover, the voltage command $v_C^*$ is given by (1).

$$v_C^* \geq \frac{1}{nL} \left( 2 \sqrt{\frac{2}{3}} E + V_{\text{mmc}} \right)$$

where $E$ is an root-mean-square (rms) value of the input line-to-line voltage, $V_{\text{mmc}}$ is the output DC voltage of the MMC, and $n$ is the number of cells at each leg.

3.3 Arm Current Control The arm current control is employed to control the AC component of the arm current which flows from the AC power grid. The AC component is defined as the positive sequence component. From the previous section, the command of the positive sequence component is generated by the capacitor voltage averaging control. In other words, the positive sequence component is controlled in order to maintain the average value of all capacitor voltages in each arm constant.

On the other hand, the arm current also includes the DC component (a zero sequence component). Thus, it is necessary to eliminate only the zero sequence component from the arm current in order to control the capacitor voltage.

3.4 MMC Output Voltage Control In the MMC output voltage control, the command of the output DC voltage $V_{\text{mmc}}^*$ is added into the output value of the controller in the arm current control. The zero sequence component is applied
to supply the power to the output side of the MMC. Moreover, the output DC voltage is divided by each cell. Thus, the output DC voltage of one cell is fundamentally set as $V_{\text{mmc}}/n$.

3.5 Capacitor Voltage Balancing Control The capacitor voltage balancing control corrects the unbalanced voltage which occurs among capacitors in same arm. In the conventional balancing control, it is required to design the control parameters. On the other hand, in the proposed voltage balancing control, it is not necessary to design control parameters. In particular, each cell has the DC bias of $V_{\text{mmc}}/n$ in the ideal state. In the proposed method, the dividing ratio is adjusted automatically depending on the capacitor voltage. Each cell output power is adjusted by varying the dividing ratio. Therefore, each capacitor voltage is automatically balanced by varying the dividing ratio.

4. Clarification of Ripple Voltage and Ripple Current Formulae for Capacitor Design

In this chapter, the formula of the ripple current is clarified for the design guideline of the electrolytic capacitor. In this paper, it is assumed that the electrolytic capacitor is applied as the cell capacitor. In fact, the employment of the electrolytic capacitor to the prototype MMC has been reported. It is well known that it is possible to easily vary the voltage rating and the rated ripple current by varying the numbers of series connection and the parallel connection of electrolytic capacitors. On the other hand, the ripple current which flows to the electrolytic capacitor affects the lifetime of the capacitor. Thus, the ripple current is one of the important factors for the MMC design.

4.1 Calculation of Capacitor’s Voltage Rating

The capacitor voltage in the cell depends on both the input voltage and the output voltage. The output DC voltage $V_{\text{ mmc}}$ is applied to each leg because each leg is connected to the load in parallel. In addition, the maximum value of the input phase voltage is applied to each arm. Therefore, each capacitor voltage depends on both the input voltage and the output voltage. The capacitor voltage command $v_c(t)$ is given by (1). Note that the modulation index $\lambda$ is set to 0.95 or less. In the circuit analysis of this paper, the voltage drop on the arm inductor is ignored in order to consider the fundamental operation. As an actual fact, the voltage drop on the arm inductor is sufficiently small compared to the input phase voltage and the output voltage of the MMC because the arm inductance in the MMC can be reduced.

4.2 Calculation of Ripple Voltage

In order to employ the ripple current as the evaluation index for the selection or the design of the electrolytic capacitor, it is important to understand the relationship between the ripple current and the circuit parameters of the MMC. Thus, it is necessary to clarify the formula of the ripple current which flows to the electrolytic capacitor. In this section, as a first step of the clarification for the formula of the ripple current, the formula of the capacitor voltage is clarified.

First, the cell output voltage $v_{\text{cell}}(t)$ is given by (2). The formula of (2) means that the input phase voltage is divided by each cell per arm and the output DC voltage $V_{\text{ mmc}}$ is divided by each cell per leg. On the other hand, the arm current $i_{\text{arm}}$, which includes the AC component and the DC component, is given by (3). The AC current flows from the AC grid in order to maintain the capacitor voltage. Whereas, the DC current flows into the load in order to supply the power. Moreover, the arm current $i_{\text{arm}}$ flows into each cell in same arm. Note that the sign of positive or negative is decided based on the lower arm of Fig. 2.

$$v_{\text{cell}}(t) = \frac{1}{n} \left[ \frac{2}{3} E \cos(\omega t + \phi) + V_{\text{ mmc}} \right]$$

$$i_{\text{arm}}(t) = -\frac{1}{2} \frac{2}{3} \frac{E \cos \omega t}{V_{\text{ mmc}}}$$

where $S$ is an apparent input power, $P$ is an input active power, $Q$ is an input reactive power, and $\phi$ is an input phase difference. $S$ is defined by (4).

$$S = \sqrt{P^2 + Q^2}$$

From (2) and (3), the instantaneous power of the cell is calculated. The energy which is stored in the capacitor is calculated by the integration of the instantaneous power and is given by (5).

$$W_C(t) = \int v_{\text{cell}}i_{\text{arm}} dt$$

where $W_C$ is the constant of integration, which does not change with time. Thus, the constant of integration $W_C$ is defined as the average value of the energy which is stored in the capacitor. The relationship between the capacitor voltage $v_c(t)$ and the energy $W_C(t)$ is given by (6) in terms of the electrostatic energy.

$$W_C(t) = \frac{1}{2} CV_C(t)^2$$

From (5) and (6), the capacitor voltage $v_c(t)$ is given by (7). Note that the theoretical formula is calculated by the linear approximation of Taylor expansion.

$$v_c(t) = V_C0 - \frac{1}{2} \sqrt{2 V_{\text{ mmc}} S \frac{E}{6 \omega C V_C0}} \sin \omega t$$

where the constant value $V_C0^2$ is defined by $2W_C0/C$, $V_C0$ is defined as the average value of the capacitor voltage because $W_C(t)$ does not change with time.

The ripple voltage includes the fundamental frequency component whose frequency is same as the frequency of the input voltage source and the second-order frequency component with twice the frequency of the input frequency. Moreover, the MMC has to control the power factor and reduce the harmonic distortion in order to eliminate the static capacitors and the series reactor. Thus, it is necessary to evaluate the ripple voltage over the wide range of the input power factor.

Figure 4 shows the relationship between the input power factor and the ripple voltage. The fundamental frequency component and the second-order frequency component is drawn by (7). The second-order frequency component of the ripple voltage does not change against the change of the input power factor because the input apparent power $S$ is constant. On the other hand, the fundamental frequency component becomes maximum when the input power factor is 1.0 or $-1.0$. Thus, it is necessary to design the capacitance only when the input power factor is 1.0, i.e. the worst case.
4.3 Clarification of Ripple Current

In this section, the formula of the ripple current in the cell capacitor is clarified. In general, the ripple current value of the capacitor is shown in the datasheet of commonly-marketed electrolytic capacitors which are generally implemented into products. The capacitor should be selected in order that the ripple current value which flows to each capacitor is sufficiently small compared to the specified value on each datasheet in terms of the lifetime. Additionally, the number of the capacitors which are connected in parallel on each cell should be increased in order to meet the specification of the ripple current. Hence, it is necessary to clarify the relationship between the ripple current and the circuit parameters in order to design the lifetime.

The relationship between the capacitor current \( i_c \) and the capacitor voltage \( v_c \) is given by (8).

\[
i_c = C \frac{dv_c}{dt} \quad \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdots \cdotted
which flow to \( S_1, S_4, D_2 \) and \( D_3 \) is given by (16). In contrast, each switching device (36). The voltage drop 
\[
R = \frac{V_{\text{on}}}{n_{\text{VC}}}
\]
is the on-resistance of the device, \( V_{\text{on}} \) is the current which flows to the device, \( d_{SD} \) is the duty of each switching device (36). The voltage drop \( V_0 \) and the on-resistance \( R \) are obtained from the datasheets.

When the arm current becomes positive, by substituting the variable number \( \theta \) of (15) by the angle \( \theta_0 \) of (13), the current which flows to \( S_1, S_2, D_3 \) and \( D_4 \) is given by (16). In contrast, when the arm current becomes negative, by substituting the variable number \( \theta \) of (15) by the angle \( \theta_0 \) of (14), the current which flows to \( S_1, S_2, D_1 \) and \( D_2 \) is given by (17).

\[
i_{\text{cell,SW}} = \frac{1}{2} \sqrt{\frac{2}{3}} \frac{S}{E} \cos \theta_a + \frac{P}{3V_{\text{mmc}}} \tag{16}
\]

\[
i_{\text{cell,SW}} = \frac{1}{2} \sqrt{\frac{2}{3}} \frac{S}{E} \cos \theta_b - \frac{P}{3V_{\text{mmc}}} \tag{17}
\]

Next, each duty is given by (18), (19), (20) and (21) respectively.

\[
d_1 = \frac{1}{2} \left[ \frac{1}{n_{\text{VC}}} \left( \frac{2}{3} E \cos(\theta_a + \phi) + V_{\text{mmc}} \right) \right] \tag{18}
\]

\[
d_2 = \frac{1}{2} \left[ \frac{1}{n_{\text{VC}}} \left( \frac{2}{3} E \cos(\theta_b + \phi) + V_{\text{mmc}} \right) \right] \tag{19}
\]

\[
d_1 = 1 - d_2 \tag{20}
\]

\[
d_2 = 1 - d_1 \tag{21}
\]

where \( V_{\text{C}} \) is the capacitor voltage.

Each conduction loss of \( S_1, S_2, D_1 \) and \( D_2 \) is given by (22), (23), (24) and (25) respectively.

\[
P_{\text{loss,SW}} = \frac{V_{\text{on}}}{4\pi} \left( 1 - \frac{V_{\text{on}}}{n_{\text{VC}}} \right) \left( \frac{2}{3} E \sin \theta_0 + \frac{2}{3} V_{\text{mmc}} (\pi - \theta_0) \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
= \frac{V_{\text{on}}}{4\pi} \left( 1 - \frac{V_{\text{on}}}{n_{\text{VC}}} \right) \left( \frac{2}{3} E \sin \theta_0 + \frac{2}{3} V_{\text{mmc}} (\pi - \theta_0) \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]

\[
+ \frac{1}{n_{\text{VC}}} \left( \frac{P}{3} (2(\pi - \theta_0) - \sin 2\theta_0) \right) \frac{2}{3} \sqrt{\frac{2}{3}} \frac{PS}{E} \cos \phi \sin \theta_0 \right)
\]
5.3 Calculation of Switching Loss and Recovery Loss  

The switching loss $P_{SW}$ and the recovery loss $P_{Rec}$ are given by (26) [28]–[31]:

$$P_{SW(Rec)} = V_C \frac{1}{2} \left[ 1 - \frac{2}{3} E \right] \sin \theta_0 + \frac{P}{3 V_{\text{mmc}}} (\pi - \theta_0) \frac{W_{\text{on}} + W_{\text{off}}}{V_{\text{dcl}} I_{\text{dcl}}} f_c$$  

(26)

where $w$ is the loss energy which is described in the datasheet, $V_{\text{dcl}}$ and $I_{\text{dcl}}$ are the voltage value and the current value when $w$ is measured, $f_c$ is the carrier frequency ($\times 10^{6}$ Hz).

Each switching loss in $S_1$ and $S_2$ is given by (27) and (28) respectively. Similarly, each recovery loss in $D_1$ and $D_2$ is given by (29) and (30) respectively.

$$P_{B,S1SW} = \frac{V_C}{\pi} \left[ \frac{1}{2} \sqrt{\frac{2}{3}} \frac{E}{E} \sin \theta_0 + \frac{P}{3 V_{\text{mmc}}} (\pi - \theta_0) \frac{W_{\text{on}} + W_{\text{off}}}{V_{\text{dcl}} I_{\text{dcl}}} f_c \right]$$  

(27)

$$P_{B,S2SW} = \frac{V_C}{\pi} \left[ \frac{1}{2} \sqrt{\frac{2}{3}} \frac{E}{E} \sin \theta_0 - \frac{P}{3 V_{\text{mmc}}} \theta_0 \frac{W_{\text{on}} + W_{\text{off}}}{V_{\text{dcl}} I_{\text{dcl}}} f_c \right]$$  

(28)

$$P_{B,D1Rec} = \frac{V_C}{\pi} \left[ \frac{1}{2} \sqrt{\frac{2}{3}} \frac{E}{E} \sin \theta_0 - \frac{P}{3 V_{\text{mmc}}} (\pi - \theta_0) \frac{W_{\text{on}} + W_{\text{off}}}{V_{\text{dcl}} I_{\text{dcl}}} f_c \right]$$  

(29)

$$P_{B,D2Rec} = \frac{V_C}{\pi} \left[ \frac{1}{2} \sqrt{\frac{2}{3}} \frac{E}{E} \sin \theta_0 + \frac{P}{3 V_{\text{mmc}}} (\pi - \theta_0) \frac{W_{\text{on}} + W_{\text{off}}}{V_{\text{dcl}} I_{\text{dcl}}} f_c \right]$$  

(30)

5.4 Relationship between Input Power Factor and Semiconductor Losses  

The relationship between the input power factor and semiconductor losses of the H-bridge cell is clarified in order to design the heat sink in the worst case. The formulae of the semiconductor losses include the semiconductor loss because the switching device at the unity power factor. The theoretical values of the semiconductor loss and the simulation are compared. Fig. 7 shows the comparison result of the theoretical formula of the semiconductor loss and the simulation. Note that the input apparent power $S$ is constant. The theoretical values of the formulae agree with the simulation values in the maximum error of 1.0% or less even though the arm current changes from AC to DC in the input power factor of 0.22. Moreover, the losses of $S_1$ and $D_1$ increase as raising of the input power factor. In contrast, the losses of $S_2$ and $D_2$ decrease. The reason why the semiconductor losses change is because the DC component in the arm current changes depending on the input power factor. When the input power factor is zero, the semiconductor loss is generated by only the AC component because the active input power, i.e. the output power is zero. The semiconductor loss which is generated by the DC component increases gradually when the input power factor moves toward 1.0. Thus, the losses of $S_1$ and $D_2$ increase as increasing of the DC component. In contrast, the losses of $S_2$ and $D_1$ decrease to zero because the arm current does not flow to $S_2$ and $D_1$. As a result, the total loss becomes maximum when the input power factor is 1.0.

In conclusion, as the worst case for the semiconductor loss and the design guideline of the heat sink in the MMC, it is necessary to design the heat sink when the input power factor is 1.0. Besides, it is also necessary to focus on the switching device where the loss generates because the semiconductor loss occurs in the particular switching device at the unity power factor.

6. Clarification of Ripple Current Formula for Arm Inductor Design  

In this chapter, the formula of the ripple current in the arm inductor is clarified for the design of the arm inductor based on the ripple factor of the inductor current. In particular, the ripple current changes drastically as the feature of the multilevel converter when the number of cells or the switching frequency changes. Therefore, both the number of cells and the switching frequency should be included in the formula of the ripple current as the variable. First, the relationship between the ripple current and the number of cells is clarified when the number of cells or the switching frequency changes. Second, the derived relationship, the formula of the ripple current is clarified based on a chopper circuit. Finally, the theoretical value with the formula and the simulation result are compared.

6.1 Modulation for H-bridge Cell  

Figure 8 shows the diagram of the unipolar modulation for the H-bridge cell (ex. $v_{Ar1}$). In Fig. 5, the leg of the H-bridge cell is constructed by $S_1$ and $S_2$. The other leg is constructed by $S_3$ and $S_4$. The switching patterns of $S_1$ and $S_2$ are determined by the voltage command $v_{Ar1}$. Moreover, the switching patterns of $S_3$ and $S_4$ are determined by the inverted voltage command $-v_{Ar1}$. In the H-bridge type circuit such as the grid-tied inverter with the unipolar modulation, the waveform of the inverter output voltage is three-level.

In addition, the phase-shifted triangular carrier is applied.
where \( N \) is the number of levels of the arm output voltage. The fundamental component is 12 kHz, whereas the frequencies of the second-order component and the third-order component are 24 kHz and 36 kHz.

In the H-bridge cell, with the employment of the unipolar modulation, the output voltage of the cell has twice the frequency of the carrier frequency. The equivalent switching frequency of the cell total output voltage \( v_{Br} \) in each arm \( f_{e,\text{cell}} \) is given by (33).

\[
f_{e,\text{cell}} = \frac{1}{2} \cdot 2f_c = nf_c, \quad n = \ldots \quad (33)
\]

From (33), under the condition which the number of cells is 8 and the carrier frequency is 1.5 kHz, the equivalent switching frequency \( f_{e,\text{cell}} \) is 12 kHz. It is confirmed that the theoretical value by (33) is equal to the result from the frequency analysis.

Moreover, the above result implies that the frequency of the ripple current on each arm inductor is determined by the equivalent switching frequency \( f_{e,\text{cell}} \).

From the above result, in following considerations of the ripple current, only the lower arm in Fig. 9 is focused on.

### 6.3 Relationship between Duty of Switching Device and Ripple Current in Chopper Circuit

The relationship between the duty and the ripple current is clarified based on the general power converter as the fundamental consideration.

Figure 11 shows the circuit diagram of a chopper circuit and the pattern which shows the relationship between the duty and the ripple current (38). In the chopper circuit, the duty of \( SW_1 \) is defined as \( D_{in} \). In contrast, the duty of \( SW_2 \) is defined as \( (1 - D_{in}) \). In Fig. 11(b), the ripple current in the chopper circuit \( \Delta i_{L,\text{ch}} \) is varied by \( D_{in} \).

The peak-to-peak value of the ripple current in the chopper circuit \( \Delta i_{L,\text{pp},\text{ch}} \) is given by (34) (39).

\[
\Delta i_{L,\text{pp},\text{ch}} = \frac{V_{in,\text{ch}}}{L} T_{on} = \frac{V_{in,\text{ch}}}{L} D_{in} T, \quad \ldots \quad (34)
\]

where \( V_{in,\text{ch}} \) is the input voltage, \( L \) is the inductance, \( T_{on} \) is the on-time of \( SW_1 \), and \( T \) is the switching cycle.

The boost ratio \( \alpha \) is defined by (35) (39).

\[
\alpha = \frac{V_{out,\text{ch}}}{V_{in,\text{ch}}} = 1 - D_{in}, \quad \ldots \quad (35)
\]

where \( V_{out,\text{ch}} \) is the output voltage of the chopper circuit.

From (35), the relationship between the input voltage \( V_{in,\text{ch}} \) and the output voltage \( V_{out,\text{ch}} \) is given by (36).

\[
V_{in,\text{ch}} = V_{out,\text{ch}} (1 - D_{in}), \quad \ldots \quad (36)
\]
By substituting (36) into (34), the peak-to-peak value $\Delta I_{pp, ch}$ and the maximum value $\Delta I_{ml, ch}$ is given by (37) and (38) respectively. Note that the average value of the inductor current is set to zero in order to calculate only the ripple component.

$$\Delta I_{pp, ch} = \frac{V_{in, ch}}{L} D_{ch} T$$

$$\Delta I_{ml, ch} = \frac{\Delta I_{pp, ch}}{2} = \frac{V_{in, ch}}{2L} D_{ch} (1 - D_{ch})$$

### 6.4 Definition of Duty in Each Multilevel-Voltage Step

For clarification of ripple current formula In this section, the ripple current in the arm inductor is clarified. From (37) and (38), the relationship between the duty of the switching device and the ripple current of the inductor in the chopper circuit is clarified. In the same way, it is possible to clarify the formula of the ripple current in the MMC with the duty of the multilevel voltage. The general duty is given by (18), (19), (20) and (21). However, these duties cannot express the variation of the pulse width in each step of the multilevel voltage. As discussed above, the ripple current is varied by the variation of the pulse width. Thus, it is necessary to define a duty with which the variation of the pulse width in each step of the multilevel voltage waveform is considered.

First, the basic level in each step of the multilevel voltage is employed in order to consider the variation of the pulse width in each step of the multilevel voltage.

Figure 12(a) shows the multilevel waveform of the cell total output voltage $v_{ch}$, the command waveform of the cell total output voltage and the basic level voltage. The multilevel waveform is obtained, whereas the basic level in each step of the multilevel voltage is also obtained as the lower limit of each step. In other words, the MMC outputs the pulse voltage based on the basic level in each step.

Figure 12(b) shows the command duty of the cell total output voltage $V_{ch}$, as well as the duty of the basic level voltage. It is possible to calculate the duty of the basic level voltage $d_{mlb}$ by (39).

$$d_{mlb} = \frac{1}{n}(N_{ml} - 1) \quad N_{ml} = 1, 2, 3 \cdots n \cdots$$

where $N_{ml}$ is the index which is varied from 1 to $n$.

For example, in $n = 8$, the step of the multilevel voltage varies periodically when $d_{mlb}$ is 0.125, 0.250 or 0.375. From (39) and Fig. 12, it is understood that the variation of the duty in each step such as 0.125, 0.250 is determined solely by the number of cells $n$.

Figure 12(c) shows the duty $d_{mlv}$ in each step of the cell total output voltage. The duty $d_{mlv}$ varies from 0.0 to 1.0 in each step. As the derivation of the duty $d_{mlv}$, first, the difference between the command voltage and the basic level voltage is calculated. Then, the difference is divided by the capacitor voltage. As a result, the duty $d_{mlv}$ is defined by (40).

$$d_{mlv} = \frac{|N_{ml} - 1|}{n}$$

where $D$ is the duty of the cell output voltage such as (18). Note that it is possible to replace $D$ to $d_{ch}$. Moreover, $N_{ml}$ should be modified according to the number of voltage levels when the number of voltage levels changes depending on the modulation.

### 6.5 Clarification of Ripple Current Formula

In order to clarify the formula of the ripple current in the arm inductor, each parameter of (38) is replaced respectively as following:

$$V_{out, ch} = v_C \quad T = \frac{1}{nf}, \quad D_{ch} = d_{mlv} \quad L = L_a \cdots$$

The maximum value of the ripple current $\Delta I_{ml}$ is given by (42).

$$\Delta I_{ml} = \frac{V_c}{2nfL_a} d_{mlv} (1 - d_{mlv})$$

In (42), the maximum value of the ripple current is varied by the duty $d_{mlv}$ which changes frequently. Therefore, the formula of (42) shows the envelope of the ripple current which traces the maximum value of the ripple current. The peak value of the ripple current is given by differentiating
Design Guidelines for Modular Multilevel Converter (Toshiki Nakanishi et al.)

From (43), when \( d_{\text{mlvs}} \) is 0.5, the formula reaches the extremum. Thus, \( \Delta i_{Lm} \) reaches the peak value when \( d_{\text{mlvs}} \) is 0.5.

Figure 13 shows the multilevel voltage waveform of the cell total output voltage in one arm, the duty of the cell total output voltage in each step and the waveforms of the ripple current, envelope of the ripple current and reversed envelope of the ripple current. It is confirmed that the envelope of the ripple current calculated by (42) traces the maximum value of the ripple current. Additionally, the ripple current has the peak value when \( d_{\text{mlvs}} \) is 0.5.

In conclusion, as the worst case for the design of the arm inductor, it is necessary to determine the inductance when the duty \( d_{\text{mlvs}} \) in each step of the cell total output voltage is 0.5. In addition, it is necessary to focus on the number of voltage levels because the number of voltage levels changes depending on the modulation.

7. Experimental Result in Miniature Model

In this chapter, the theoretical formulae are verified by the experiment in the miniature model. Note that it is assumed that the unity power factor is obtained over entire load range where the theoretical values are calculated. Thus, the apparent power \( S \) in each formula is replaced by the output power \( P \). In addition, the output voltage \( V_{\text{mmc}} \) is calculated from the output power and the load resistor in the calculation of theoretical values.

7.1 Verifications of Fundamental Operation for Step-down Rectifier using MMC

Table 1 shows experimental conditions. The miniature model is constructed by four cells per leg. Additionally, as fundamental experiments, the resistance of 5.3 \( \Omega \) is employed as the load of the MMC without a smoothing capacitor.

Figure 14 shows the waveforms of the input phase voltage, the input current of R-phase and the output DC voltage. First, from the waveforms of the input phase voltage and the input current, it is confirmed that the unity power factor is obtained in the input stage. Moreover, the total harmonic distortion (THD) of the input current is 3.1% when the normalized impedance \( \%Z \) of the inductor is 6.1%. Second, the waveform of the output DC voltage in the lower side of Fig. 14 shows that the step-down rectifier converts from the input voltage of 200 V into the output DC voltage of 75 V, which is maintained at constant. Moreover, it is possible to determine the output DC voltage freely because the output voltage is controlled by varying the DC offset of the cell output voltage. As a result, the proposed rectifier of the MMC achieves the step-down rectification.

Figure 15 shows the waveforms of the arm voltage which is the summation of the output voltage of all cells in each arm and the waveform of the line voltage between R-phase and S-phase in the upper side. First, the arm voltage of five levels is obtained because one of the H-bridge cells with the unipolar modulation obtains three-level voltage and the arm has two cells. In addition, the waveforms are not symmetry in the positive side and the negative side because the cell output voltage includes the DC component.

Besides, the waveform of the line voltage between R-phase and S-phase in the upper side is nine levels. However, the multilevel voltage waveform becomes non-uniform.
Fig. 14. Waveforms of input phase voltage, input phase current and output voltage

Fig. 15. Waveforms of arm voltage which is the summation of output voltage of all cells in each arm and line voltage between R-phase and S-phase in upper side

Fig. 16. Waveforms of all capacitor voltage in R-phase

reason is that the arm voltage is not symmetry in the positive side and the negative side because the command voltage of the cell includes the DC bias with the MMC output voltage control. However, the voltage fluctuation is still smaller compared to the employments of the chopper cells and the H-bridge cells with bipolar modulation. Thus, the MMC achieves the size reduction of the arm inductors.

Figure 16 shows the waveforms of all cell capacitor voltages which are connected to the R-phase leg. The cell capacitor voltage is controlled according to the capacitor voltage command \( v_{c}^{*} \). As a result, the proposed step-down rectifier maintains the capacitor voltage of each H-bridge cell to the voltage command of 130 V. Therefore, the MMC also achieves the capacitor voltage control. In addition, the maximum voltage error between the voltage command of the cell capacitor and the measured voltage is 2% or less.

7.2 Verifications for Theoretical Formula of Ripple Current in Capacitor

In this section, the verification result for the theoretical formula of the capacitor ripple current is shown. In order to evaluate the ripple current of the capacitor, the frequency analysis of the capacitor voltage is conducted because it is difficult to measure the ripple current of the capacitor directly in the miniature model. However, because the ripple voltage is caused by the ripple current, the validity of the formula is still verified when the measured ripple voltage and the theoretical value by (7) are proved to be same. In particular, the fundamental frequency component and the second-order frequency component of the ripple voltage which are given by (7) and the measured value are compared.

Figure 17 shows the comparison result of the theoretical value and the measured value of the fundamental frequency component in the ripple voltage. From the result, measured values agree with theoretical values with the small error. In particular, the maximum error between the theoretical value and the measured value is 4.0%.

Figure 18 shows the comparison result of theoretical values and measured values of the second-order frequency component. The difference between the theoretical value and the measured value is large. The maximum error is 49.0%. The cause of the large error is the mismatch of the input active power and the output power due to the semiconductor loss and the inductor loss.

Figure 19 shows the comparison result of theoretical values and measured values of the second-order frequency component based on the input active power. In particular, the input active power is assigned to the power \( P \) in the
theoretical formula of (7). From the result, the maximum error is reduced to 10.3% compared to the results from Fig. 18. Hence, this result clarifies the reason of the high error between the theoretical value and the measured value of the second-order frequency component in Fig. 18, which is the mismatch of the input active power and the output power due to the semiconductor loss and the inductor loss. In the practical converter, every losses are minimized in order to obtain the high efficiency. Thus, the error in the second-order frequency component based on the output power is small because of the low loss. On the other hand, in the above evaluation, when the output power is low, the error between the theoretical value and the measured value of the second-order frequency component may be large because the unity power factor is not obtained. However, the ripple voltage including the second-order frequency component should be designed in the worst case when the ripple voltage is maximum in the rated output power. Hence, the effect of the above error is small for the design.

In conclusion, the validity of the formula for the capacitor voltage ripple is verified. Besides, this result confirms that the capacitor ripple current can be calculated exactly by the formula.

7.3 Verifications for Theoretical Formula of Ripple Current in Arm Inductor

In this section, the verification result of the ripple current in the arm inductor is shown.

Figure 20 shows the expanded waveforms of the lower arm voltage in the R-phase and the arm current. The arm voltage is the total output voltage of two cells. In addition, the expanded waveform focuses on the output voltage of which the duty in each step of the cell total output voltage $d_{\text{mod}}$ is 0.5, i.e. when the ripple current becomes maximum. As shown in Fig. 20, the measured peak-to-peak value of the ripple current is 118 mA. The theoretical value of the twice value by (42) is 124 mA because the peak-to-peak value is twice value of the maximum value. As a result, the error between the theoretical value and the measured value is 4.9%. Note that the inductance used in (42) is measured at the frequency of 32 kHz because 32 kHz is the equivalent frequency of one arm in the miniature model. Moreover, the unipolar modulation is employed.

Figure 21 shows the comparison result of theoretical values and measured values of the ripple current in the arm inductor dependent on capacitor voltage. The measured input power is defined as the no-load loss. Next, the input power, the output power and the inductor loss are measured same time when the MMC operates in state where the converter connects to the load. In general converters, the switching device has the parasitic capacitor and the current flows between the parasitic capacitor and the switching device when the switching device turns on. In this state, the measured input power is defined as the no-load loss. Thus, the inductance decreases due to the increase of the core temperature. In the experiment, semiconductor losses, inductor losses and the no-load loss are considered. The no-load loss is measured when the MMC operates in state where the converter does not connects to the load. In general converters, the switching device has the parasitic capacitor and the current flows between the parasitic capacitor and the switching device when the switching device turns on. In this state, the measured input power is defined as the no-load loss. Next, the input power, the output power and the inductor loss are measured same time when the MMC operates in state where the converter connects to the load. By subtracting the output power, the inductor loss and the no-load loss from the input power, the semiconductor loss is obtained.

Table 2 shows the circuit parameters and the loss parameters in the verification for formulae of semiconductor losses. The loss parameters relating to the conduction loss are obtained from the datasheet of the used device, whereas the loss parameters relating to the switching loss and the recovery loss are obtained from the measured values in the switching test.
Figure 22 shows the comparison result of measured values and theoretical values. The theoretical value is defined as the total value of the conduction loss, the switching loss and the recovery loss in all cells. The maximum error is 3.4%. On the other hand, in all range, measured values are larger than theoretical values. The reason of this error is the loss in the equivalent series resistance (ESR) of the capacitor and the wiring loss which is not considered in the measurement. Thus, the semiconductor loss which is measured in the experiment includes the above losses which are not considered. As a result, measured values are larger than theoretical values. Besides, the error between the measured value and the theoretical value increases with the increase of the output power. This reason is that both the loss in the capacitor and the wiring loss increase with the increase of the output power because the ripple current of the capacitor and the arm current increases.

On the other hand, the switching device has the tolerance in the loss characteristic. This tolerance may affect the errors between calculated and measured losses. However, the MMC consists of many switching devices. As a result, the effect of the difference between the nominal value and the practical value of each switching device is small for overall system by averaging tolerance of the switching device. In addition, the relationship between the input power factor and the semiconductor loss is not changed in principle. Thus, it is possible to design the heat sink as same as general power converters after considering the worst case or the margin even when the tolerance in the loss characteristic exists.

Figure 23 shows the breakdown of the losses in the MMC. The no-load loss is constant against the change of the output power. On the other hand, the semiconductor loss and the inductor loss increase with the increase of the output power because the arm current increases.

8. Conclusion

This paper presented the theoretical formula focused on the number of cells and the output DC voltage for the circuit parameter design in the H-bridge cell type Modular Multilevel Converter (MMC) in order to show the design guideline. Moreover, the worst case for each component design was also clarified as following:

(i) As the worst case in the cell capacitor design, the ripple component was maximum when the input power factor was 1.0. In particular, it was necessary to design the rated ripple current in this point when the electrolytic capacitor was applied.

(ii) As the worst case in the heat sink design, the semiconductor loss was maximum when the input power factor was 1.0.

(iii) As the worst case in the arm inductor design, the ripple current was maximum when the duty in each step of the cell total output voltage was 0.5.

In addition, theoretical formulae were verified in the miniature model of the MMC. As a result, the following results were obtained.

(a) As the verification of the capacitor ripple voltage, the measurement resulted in the maximum error of 4.0% between theoretical values and measured values. Because the ripple voltage was caused by the ripple current, it is possible to calculate exactly the ripple current of the electrolytic capacitor with the formula.

(b) As the verification of the ripple current in the arm inductor, the measurement resulted in the maximum error of 11.8% between theoretical values and measured values. The error was caused by the change of the core characteristic.

(c) As the verification for the formulae of semiconductor losses, the measurement resulted in the maximum error of 3.4% between the theoretical value and the measured value. Hence, it is possible to calculate the semiconductor loss for design of the heat sink.

It is expected that these results promote the discussion of the design focused on the number of cells and the MMC output voltage.
References