

DESCRIPTION

Title of Invention: POWER CONVERTER CONTROL DEVICE

Technical Field

5 [0001] The present invention relates to a power converter control device.

Background Art

[0002] Patent Documents 1 to 3 each disclose a direct power converter. This direct power converter includes a diode rectifier, an inverter, and a charge and discharge circuit.

[0003] The diode rectifier full-wave rectifies a single-phase AC voltage, and outputs a
10 rectified voltage after rectification across a pair of DC power supply lines (a DC link).

[0004] The charge and discharge circuit is provided to the DC link, and includes a buffer circuit and a boost circuit. The buffer circuit has a switch and a capacitor connected in series to each other between the pair of DC power supply lines. When the switch is conducting, the capacitor discharges to provide power to the DC link.

15 [0005] The boost circuit boosts the rectified voltage from the diode rectifier to charge the capacitor. This allows the charge and discharge circuit to receive power from the DC link. An input voltage (a DC voltage from the DC link) of the inverter matches the boosted voltage for the capacitor when the above-mentioned switch is conducting, and matches the rectified voltage when the switch is not conducting. The inverter receives
20 the DC voltage from the DC link as an input, converts it into an AC voltage, and outputs the AC voltage.

Prior Art Documents

Patent Documents

[0006] Patent Document 1: Japanese Patent Application Laid-Open No. 2015-084637

25 Patent Document 2: Japanese Patent Application Laid-Open No. 2015-065731

Patent Document 3: Japanese Patent Application Laid-Open No. 2015-076921

Summary of Invention

Problem to be Solved by Invention

[0007] It is desirable to drive a motor with high efficiency.

- 5 [0008] An object of the present application is thus to provide a direct power converter control device that enables driving of a motor with high efficiency.

Means to Solve the Problem

- [0009] A first aspect of a power converter control device according to the present invention is a device for controlling, in a power converter including first and second DC
- 10 power supply lines (LH, LL), a rectifying circuit (3), a discharge switch (Sc) and a capacitor (C4), a boost circuit (4a), and an inverter (5), the discharge switch, the rectifying circuit full-wave rectifying a single-phase AC voltage to a rectified voltage, and applying the rectified voltage across the first and second DC power supply lines, the discharge switch and the capacitor being connected in series to each other between the
- 15 first and second DC power supply lines, the boost circuit boosting the rectified voltage to charge the capacitor, the inverter receiving, as an input, the rectified voltage as a DC voltage (Vdc1) across the first and second DC power supply lines when the discharge switch is not conducting, receiving, as an input, a voltage across the capacitor as the DC voltage when the discharge switch is conducting, converting the DC voltage as input into
- 20 an AC voltage, and outputting the AC voltage to a motor (6), the power converter control device including: a switch control unit (11) that maintains the discharge switch not conducting over a first time period (T1), and switches the discharge switch between conducting and not conducting in a second time period (T2) other than the first time period; and a charge and discharge time period setting unit (12) that sets the first time
- 25 period when a rotational speed of the motor is higher than a speed threshold (ω_{ref}) shorter

than the first time period when the rotational speed is lower than the speed threshold.

[0010] A second aspect of the power converter control device according to the present invention is the power converter control device according to the first aspect in which, when the rotational speed (ω_m) is higher than the speed threshold (ω_{ref}), the first time period (T1) has a length of zero.

[0011] A third aspect of the power converter control device according to the present invention is the power converter control device according to the first or second aspect in which, when the rotational speed (ω_m) is lower than the speed threshold (ω_{ref}), the first time period (T1) is a time period in which the rectified voltage is greater than a predetermined value.

[0012] A fourth aspect of the power converter control device according to the present invention is the power converter control device according to any one of the first to third aspects in which the speed threshold is the rotational speed when amplitude of the AC voltage applied to the motor (6) reaches an upper limit.

Effects of Invention

[0013] According to the first and fourth aspects of the power converter control device according to the present invention, the voltage across the capacitor is higher than the rectified voltage, and thus the DC voltage can be higher than the rectified voltage when the discharge switch is conducting.

[0014] When the rotational speed is low, the DC voltage can be higher than the rectified voltage as the discharge switch is conducting in the second time period, but the DC voltage matches the rectified voltage as the discharge switch is maintained not conducting in the first time period. That is to say, the DC voltage higher than the rectified voltage cannot be input into the inverter in the first time period.

[0015] On the other hand, when the rotational speed is high, the discharge switch is

switched between conducting and not conducting in the second time period that is longer. The DC voltage when the rotational speed is high can thus be higher than the DC voltage when the rotational speed is low.

[0016] The maximum value of a voltage (maximum output applied voltage) that can be applied to the motor can be increased with increasing second time period, and thus high-speed rotation and high-load operation of the motor can be achieved while suppressing an increase in current of the motor. On the other hand, the DC voltage can become small when the rotational speed is low. A time period in which the discharge switch is conducting (i.e., a period in which the capacitor discharges) is required to be short so that the DC voltage becomes small, and the amount of operation of the boost circuit can become small accordingly. Loss caused in the boost circuit is reduced as the amount of operation of the boost circuit decreases. Efficiency when the rotational speed is low can thus be improved.

[0017] As described above, high-load operation can be achieved while suppressing the increase in current of the motor when the rotational speed is high, while efficiency is improved when the rotational speed is low.

[0018] According to the second aspect of the power converter control device according to the present invention, the maximum output applied voltage to the motor can be maximized.

[0019] According to the third aspect of the power converter control device according to the present invention, the DC voltage can be increased compared with a case where a time period in which the rectified voltage is smaller than a voltage reference value is used as the first time period.

[0020] The objects, features, aspects, and advantages of the present invention will become more apparent from the following detailed description and the accompanying

drawings.

Brief Description of Drawings

[0021] FIG. 1 is a block diagram showing the configuration of a direct power converter to which a control method shown in an embodiment is applied.

5 FIGS. 2A and 2B are diagrams schematically showing examples of various parameters in a first scheme.

 FIGS. 3A and 3B are diagrams schematically showing examples of various parameters in a second scheme.

 FIG. 4 is a flowchart showing an example of operation of a control unit.

10 FIG. 5 is a diagram showing an example of operation.

 FIG. 6 is a diagram showing an example of an equivalent circuit.

Description of Embodiments

[0022] <1. Configuration of Direct Power Converter>

 FIG. 1 is a block diagram showing the configuration of a direct power converter
15 to which a control method shown in the present embodiment is applied. The direct power converter includes a converter 3, a power buffer circuit 4, and an inverter 5.

[0023] The converter 3 is a single-phase rectifying circuit, and is connected to a single-phase AC power supply 1, for example, via a filter 2. The filter 2 includes a reactor L2 and a capacitor C2. The reactor L2 is connected between one of two output
20 terminals of the single-phase AC power supply 1 and the converter 3. The capacitor C2 is connected in series to the reactor L2 between the two output terminals of the single-phase AC power supply 1. The filter 2 removes a high-frequency component of a current. The filter 2 may be omitted. Description will be made below while ignoring the function of the filter 2 for simplicity.

25 [0024] The converter 3 uses, for example, a diode bridge, and includes diodes D31 to

D34. The diodes D31 to D34 constitute a bridge circuit, single-phase full-wave rectify a single-phase AC voltage V_{in} , which is an input voltage input from the single-phase AC power supply 1, to convert it into a rectified voltage V_{rec} , and output the rectified voltage V_{rec} across DC power supply lines LH and LL (they form a so-called DC link). A potential higher than that applied to the DC power supply line LL is applied to the DC power supply line LH. An input current I_{in} flows from the single-phase AC power supply 1 into the converter 3.

[0025] The power buffer circuit 4 is a circuit that provides and receives power to and from the DC link, and includes a boost circuit 4a, a buffer circuit 4b, and a diode D43. The buffer circuit 4b includes a capacitor C4. The boost circuit 4a boosts the rectified voltage V_{rec} to charge the capacitor C4.

[0026] The buffer circuit 4b further includes a transistor (herein an insulated-gate bipolar transistor: hereinafter abbreviated as an “IGBT”) S_c connected in antiparallel to a diode D42. The transistor S_c is connected in series to the capacitor C4 between the DC power supply lines LH and LL at a side closer to the DC power supply line LH. Herein, connecting A and B in antiparallel refers to connecting A and B in parallel so that a forward direction of A and a forward direction of B are opposite to each other. Specifically, a forward direction of the transistor S_c is a direction from the DC power supply line LL towards the DC power supply line LH, and a forward direction of the diode D42 is a direction from the DC power supply line LH towards the DC power supply line LL. The transistor S_c and the diode D42 can collectively be understood as a single switch element (first switch).

[0027] The boost circuit 4a includes a diode D40, a reactor L4, and a transistor (herein an IGBT) S_L , for example. The diode D40 includes a cathode and an anode, and the cathode is connected between the first switch and the capacitor C4.

[0028] The reactor L4 is connected between the DC power supply line LH and the anode of the diode D40. The transistor SL is connected between the DC power supply line LL and the anode of the diode D40. A diode D41 is connected in antiparallel to the transistor SL, and they can collectively be understood as a single switch element (second switch). The boost circuit 4a is known as a so-called boost chopper.

[0029] The capacitor C4 is charged by the boost circuit 4a, and a voltage V_c higher than the rectified voltage V_{rec} is generated across the capacitor C4. Specifically, a current is allowed to flow from the DC power supply line LH to the DC power supply line LL via the second switch to store energy in the reactor L4, and then the second switch is turned off to store the energy in the capacitor C4 via the diode D40. This enables the power buffer circuit 4 to receive power from the DC link.

[0030] Since the voltage V_c is higher than the rectified voltage V_{rec} , a current basically does not flow through the diode D42. Whether the first switch is conducting or not thus depends solely on whether the transistor Sc is conducting or not. Not only the transistor Sc but also the first switch (a discharge switch) including the transistor Sc and the diode D42 is thus hereinafter also referred to as a switch Sc. When the switch Sc is conducting, the capacitor C4 discharges so that the power buffer circuit 4 can provide power to the DC link.

[0031] Since the DC power supply line LH is higher in potential than the DC power supply line LL, a current basically does not flow through the diode D41. Whether the second switch is conducting or not thus depends solely on whether the transistor SL is conducting or not. Not only the transistor SL but also the second switch including the transistor SL and the diode D41 is thus hereinafter also referred to as a switch SL.

[0032] The diode D43 is connected on the DC power supply line LH between the reactor L4 and the switch Sc. A forward direction of the diode D43 is a direction from

the converter 3 towards the inverter 5. The diode D43 prevents a short circuit of the reactor L4 when the switch Sc is turned on, for example.

[0033] The inverter 5 converts the DC voltage across the DC power supply lines LH and LL into an AC voltage, and outputs the AC voltage to output terminals Pu, Pv, and Pw. When this is described in terms of power, the inverter 5 receives DC power from the DC link as an input, and outputs AC power to a motor 6. The inverter 5 includes six switching elements Sup, Svp, Swp, Sun, Svn, and Swn. The switching elements Sup, Svp, and Swp are respectively connected between the output terminals Pu, Pv, and Pw and the DC power supply line LH, and the switching elements Sun, Svn, and Swn are respectively connected between the output terminals Pu, Pv, and Pw and the DC power supply line LL. The inverter 5 forms a so-called voltage source inverter, and includes six diodes Dup, Dvp, Dwp, Dun, Dvn, and Dwn.

[0034] The diodes Dup, Dvp, Dwp, Dun, Dvn, and Dwn are arranged with their cathodes being directed towards the DC power supply line LH and their anodes being directed towards the DC power supply line LL. The diode Dup is connected in parallel to the switching element Sup between the output terminal Pu and the DC power supply line LH. Similarly, the diodes Dvp, Dwp, Dun, Dvn, and Dwn are respectively connected in parallel to the switching elements Svp, Swp, Sun, Svn, and Swn.

[0035] IGBTs are used as the switching elements Sup, Svp, Swp, Sun, Svn, and Swn, for example.

[0036] The AC voltage is applied from the inverter 5 to the motor 6. The motor 6 rotates in accordance with the AC voltage. The motor 6 drives a load (e.g., a compressor), which is not illustrated. In a case where the compressor is used as the load, the compressor may be mounted on an air conditioner, for example. FIG. 1 shows an equivalent circuit of the motor 6. In a specific example, an inductance component Lu

and a resistance component R_u of a winding of a U phase are connected in series, and one end of the series-connected body is connected to the output terminal P_u . The same applies to windings of a V phase and a W phase. The other ends of these series-connected bodies are connected together.

5 [0037] A control device 10 outputs switch signals to the switches SL and Sc and the inverter 5 to control the power converter. A method of controlling the power converter will be described below.

[0038] <2. Control Method>

In such a power converter, the rectified voltage V_{rec} can be input into the
10 inverter 5 when the switch Sc is not conducting. On the other hand, the voltage V_c across the capacitor $C4$ can be input into the inverter 5 when the switch Sc is conducting. That is to say, a DC voltage V_{dc1} input into the inverter 5 matches the rectified voltage V_{rec} when the switch Sc is not conducting, and matches the voltage V_c when the switch Sc is conducting.

15 [0039] In the present embodiment, a time period $T1$ in which the switch Sc is maintained not conducting and a time period $T2$ in which the switch Sc is switched between conducting and not conducting are introduced. The time period $T2$ is a time period other than the time period $T1$. The time period $T2$ thus decreases with increasing time period $T1$.

20 [0040] FIGs. 2A and 2B schematically show examples of the DC voltage V_{dc1} input into the inverter 5 and a switch state of the switch Sc . In the example of FIG. 2A, the single-phase AC voltage V_{in} , the input current I_{in} , and a current i_L flowing through the reactor $L4$ are also shown for reference. As illustrated in FIG. 2A, the switch Sc is maintained not conducting (off in FIG. 2A) in the time period $T1$. Thus, in the time
25 period $T1$, the DC voltage V_{dc1} matches the rectified voltage V_{rec} ($= |V_{in}|$).

[0041] In the example of FIG. 2A, the switch state is blackened in the time period T2. As illustrated in FIG. 2B that is an enlarged view of the time Π , the switch S_c is repeatedly switched between conducting and not conducting in the time period T2. Thus, in the time period T2, the DC voltage V_{dc1} matches the voltage V_c when the switch S_c is conducting, and matches the rectified voltage V_{rec} when the switch S_c is not conducting. In the time period T2, an average of the DC voltage V_{dc1} can be controlled to be between the rectified voltage V_{rec} and the voltage V_c in accordance with a duty at which the switch S_c is conducting. This means that, in the time period T2, the average of the DC voltage V_{dc1} can be increased above the rectified voltage V_{rec} . The average of the DC voltage V_{dc1} herein refers to an average (hereinafter, also referred to as a first average) of the DC voltage V_{dc1} in switching periods of the switch S_c , for example.

[0042] As described above, the DC voltage V_{dc1} matches the rectified voltage V_{rec} in the time period T1, and the average of the DC voltage V_{dc1} can be controlled to be a desired value higher than the rectified voltage V_{rec} in the time period T2.

[0043] The relationship between the maximum value of an average (hereinafter, referred to as a second average) of the DC voltage V_{dc1} in a single period of the single-phase AC voltage V_{in} and the length of the time period T1 will be considered next. As described above, the DC voltage V_{dc1} matches the rectified voltage V_{rec} in the time period T1, and the DC voltage V_{dc1} can be controlled to be the value higher than the rectified voltage V_{rec} in the time period T2. The maximum value of the second average of the DC voltage V_{dc1} can thus be increased with decreasing time period T1. Conversely, the maximum value of the second average of the DC voltage V_{dc1} decreases with increasing time period T1.

[0044] In the present embodiment, the time period T1 is set to be shorter when a rotational speed ω_m of the motor 6 is higher than a speed threshold than when the

rotational speed ω_m is lower than the speed threshold. When the rotational speed ω_m is higher than the speed threshold, the time period T1 is set to be zero, for example. When zero is used as the time period T1, the DC voltage Vdc1 can be controlled to be maximized from a viewpoint of the length of the time period T1. Conversely, the time period T1 is set to be long when the rotational speed ω_m is low.

[0045] FIGs. 3A and 3B schematically show examples of the DC voltage Vdc1 and the switch state of the switch Sc when zero is used as the time period T1. Also in the example of FIG. 3A, the single-phase AC voltage Vin, the input current Iin, and the current iL are shown for reference. As illustrated in FIG. 3B that is an enlarged view of the time III, the switch Sc is repeatedly switched between conducting and not conducting in the entire time period (time period T2). The first average of the DC voltage Vdc1 can thus be higher than the rectified voltage Vrec in the entire time period.

[0046] Incidentally, the amplitude of the AC voltage applied to the motor 6 increases with increasing rotational speed ω_m of the motor 6. The maximum value of the amplitude of the AC voltage that the inverter 5 can output to the motor 6 depends on the DC voltage Vdc1. Specifically, the maximum value increases with increasing DC voltage Vdc1. Thus, the DC voltage Vdc1 is not required to be increased when the rotational speed ω_m is low.

[0047] In the present embodiment, the time period T1 is thus set to be long when the rotational speed ω_m is low (e.g., FIG. 2A). This can reduce the maximum value of the second average of the DC voltage Vdc1 when the rotational speed ω_m is low. If it is assumed that the duty in the time period T2 is not changed by the change of the length of the time period T1 for simplicity, for example, the second average of the DC voltage Vdc1 is reduced by extension of the time period T1.

[0048] When the time period T2 is reduced by increasing the time period T1 as

described above, a time period in which the switch S_c can be conducting, that is, a time period in which the capacitor C_4 can discharge decreases, and the amount of discharge can be reduced. The amount of charge and the amount of discharge of the capacitor C_4 are controlled to be balanced, and thus the amount of charge of the capacitor C_4 by the boost circuit 4a is reduced by reduction in amount of discharge. That is to say, a time period in which the switch S_L is on is shorter when the rotational speed ω_m is lower than the speed threshold than when the rotational speed ω_m is higher than the speed threshold. This reduces a current flowing through the boost circuit 4a. Loss caused in the boost circuit 4a is thus also reduced. Thus, when the time period T_1 is set to be long to reduce the DC voltage V_{dc1} , the motor 6 can be driven with higher efficiency compared with a case where the time period T_1 is set to be short.

[0049] In contrast, the time period T_1 is set to be short when the rotational speed ω_m is high (e.g., FIG. 3A). This can increase the maximum value of the second average of the DC voltage V_{dc1} when the rotational speed ω_m is high. That is to say, the DC voltage V_{dc1} can be controlled to be higher compared with a case where the time period T_1 is set to be long. An AC voltage having larger amplitude can thus be output to the motor 6. This enables the motor 6 to be driven properly at a high rotational speed ω_m .

[0050] When the amplitude of the AC voltage increases to the maximum value of the amplitude of the AC voltage that the inverter 5 can output, the rotational speed ω_m can be increased, for example, by performing field weakening control (or flux weakening control, the same applies to the following). In the field weakening control, the amplitude of the AC current flowing through the motor 6 increases rapidly with increasing rotational speed ω_m . The increase in current is undesirable, for example, because the current capacity of the circuit (the converter 3, the power buffer circuit 4, the inverter 5, and the motor 6) is required to be raised.

[0051] In the present embodiment, when the rotational speed ω_m is higher than the speed threshold, the DC voltage V_{dc1} can be controlled to be higher, and the inverter 5 can output the AC voltage having larger amplitude to the motor 6. Thus, a value of the rotational speed ω_m that requires the field weakening control can be increased. This can
5 reduce the amplitude of the AC current flowing through the motor 6. The current capacity that the circuit is required to have can be reduced, leading to reduction in manufacturing cost.

[0052] The configuration to perform the above-mentioned control will be described next. In the example of FIG. 1, a speed detection unit 9 and the control device 10 are
10 provided. The speed detection unit 9 detects the rotational speed ω_m of the motor 6. For example, AC currents i_u , i_v , and i_w flowing through the motor 6 are input into the speed detection unit 9. The AC currents i_u , i_v , and i_w are detected by a known current detection unit. The speed detection unit 9 calculates the rotational speed ω_m of the motor 6 on the basis of the AC currents i_u , i_v , and i_w , and outputs it to the control device
15 10.

[0053] The control device 10 includes a switch control unit 11 and a charge and discharge time period setting unit 12. The control device 10 herein includes a microcomputer and a storage device. The microcomputer performs processing steps (i.e., procedures) written in a program. The above-mentioned storage device can be formed
20 by one or more various storage devices, such as read only memory (ROM), random access memory (RAM), rewritable non-volatile memory (e.g., erasable programmable ROM (EPROM)), and a hard disk drive. The storage device stores various pieces of information, data, and the like, stores the program executed by the microcomputer, and provides a work area for executing the program. The microcomputer can be understood
25 to function as various units corresponding to the respective processing steps written in the

program, or to achieve various functions corresponding to the respective processing steps. The control device 10 is not limited to this, and some or all of various procedures performed by the control device 10 or various units or various functions achieved by the control device 10 may be achieved by hardware.

5 [0054] The rotational speed ω_m detected by the speed detection unit 9 is input into the charge and discharge time period setting unit 12. The charge and discharge time period setting unit 12 sets the length of the time period T1 (and, further, the time period T2) on the basis of the rotational speed ω_m as described above, and outputs information M showing the length of the time period T1 to the switch control unit 11.

10 [0055] FIG. 4 is a flowchart showing an example of operation of the charge and discharge time period setting unit 12, and FIG. 5 is a diagram for explaining the operation. As shown in FIG. 4, the charge and discharge time period setting unit 12 first determines whether the rotational speed ω_m is higher than a speed threshold ω_{ref} in step ST1. The speed threshold ω_{ref} is stored in a storage, for example, and the charge and discharge
15 time period setting unit 12 reads it, and determines whether the rotational speed ω_m is higher than the speed threshold ω_{ref} . The determination can be made, for example, using a comparator.

[0056] When determining that the rotational speed ω_m is lower than the speed threshold ω_{ref} , the charge and discharge time period setting unit 12 determines the time
20 period T1 as a first value in step ST2. Control based on the first value for the time period T1 is hereinafter also referred to as a first scheme. On the other hand, when determining that the rotational speed ω_m is higher than the speed threshold ω_{ref} in step ST1, the charge and discharge time period setting unit 12 sets the time period T1 to a second value (e.g., zero) that is smaller than the first value in step ST3. Control based
25 on the second value for the time period T1 is hereinafter also referred to as a second

scheme. As a result, the first scheme is used when the rotational speed ω_m is lower than the speed threshold ω_{ref} , and the second scheme is used when the rotational speed ω_m is higher than the speed threshold ω_{ref} as shown in FIG. 5.

[0057] The switch control unit 11 generates control signals SSc, SSL, SSUp, SSvp, SSwp, SSUn, SSvn, and SSwn for the switches SL and Sc of the power buffer circuit 4 and the switching elements Sup, Sun, Svp, Svn, Swp, and Swn of the inverter 5, and outputs them to the power buffer circuit 4 and the inverter 5. Not only the information M but also various parameters (e.g., crest values V_m and I_m , a power supply angular speed ω , and a d-axis current i_d and a q-axis current i_q) and various command values (e.g., a rotational speed command value ω_m^*) required for control are input into the switch control unit 11, and the above-mentioned control signals are generated on the basis of them. Any of techniques described in Patent Documents 1 to 3 can be used as a method of generating the control signals, for example. A specific example of the control will be described below.

[0058] The time period T1 (e.g., FIG. 2A) when the rotational speed ω_m is lower than the speed threshold ω_{ref} may be a time period in which the rectified voltage V_{rec} ($=|V_{in}|$) is greater than a voltage reference value V_{ref} (e.g., $1/\sqrt{2}$ of the crest value V_m). That is to say, the switch Sc is maintained not conducting when the rectified voltage V_{rec} is high. According to this, as illustrated in FIG. 2A, the rectified voltage V_{rec} near a peak value (crest value) appears as the DC voltage V_{dc1} in the time period T1. On the other hand, when a time period in which the rectified voltage V_{rec} is smaller than the voltage reference value V_{ref} is used as the time period T1, the rectified voltage V_{rec} near a bottom value (zero) appears as the DC voltage V_{dc1} in the time period T1. That is to say, the DC voltage V_{dc1} can be increased by using the time period in which the rectified voltage V_{rec} is greater than the voltage reference value V_{ref} as the time period T1. This

is desirable because the AC voltage from the single-phase AC power supply 1 is used effectively.

[0059] A plurality of speed thresholds may be provided. For example, the time period T1 may be set to the first value when the rotational speed ω_m is lower than a first speed threshold, may be set to the second value ($<$ the first value) when the rotational speed ω_m is higher than the first speed threshold and is lower than a second speed threshold, and may be set to a third value ($<$ the second value) when the rotational speed ω_m is higher than the second speed threshold.

[0060] <2-1. Specific Control Method>

An example of specific control of the switches Sc and SL and the inverter 5 will be described below for reference. As an example, in the first scheme used when the rotational speed ω_m is lower than the speed threshold ω_{ref} , the time period T1 is set to a time period in which the rectified voltage Vrec ($= |V_{in}|$) is greater than $1/\sqrt{2}$ of the crest value Vm (see FIG. 2A). In this case, the time periods T1 and T2 are switched each one-quarter period of the single-phase AC voltage Vin. On the other hand, the time period T1 is set to zero in the second scheme used when the rotational speed ω_m is higher than the speed threshold ω_{ref} (see FIG. 3A).

[0061] <2-1-1. Basic Concept>

The concept common to the first scheme and the second scheme will be described first. Input power Pin input into the converter 3 is expressed by an equation shown below with an input power factor being 1. Note that the crest value Vm and the power supply angular speed ω of the single-phase AC voltage Vin, the crest value Im of the input current Iin, and time t are introduced. The product ωt of the power supply angular speed ω and the time t represents a phase angle of the single-phase AC voltage Vin. An AC waveform is understood as the product of a sine value of the phase angle ωt

and the crest value of the AC waveform.

[0062]

$$\begin{aligned} P_{in} &= V_m \cdot I_m \cdot \sin^2(\omega t) \\ &= \frac{1}{2} V_m \cdot I_m - \frac{1}{2} V_m \cdot I_m \cdot \cos(2\omega t) \quad \dots (1) \end{aligned}$$

[0063] The second term of the rightmost side of the equation (1) indicates power ripple.

- 5 A variation range of the input power P_{in} is the amplitude of the second term of the rightmost side of the equation (1), and is indicated by $V_m \cdot I_m / 2$. Power that the converter 3 outputs to the DC link is ideally equal to the input power P_{in} input into the converter 3. The input power P_{in} can thus be understood as power input into the DC link. The power ripples as described above, and is thus hereinafter also referred to as
- 10 ripple power P_{in} . To reduce ripple of DC power P_{dc} output from the DC link to the inverter 5, the power buffer circuit 4 should provide and receive power to and from the DC link so that the amplitude of the second term of the rightmost side of the equation (1) is reduced. The power provided and received by the power buffer circuit 4 is hereinafter referred to as buffer power P_{buf} . The buffer power P_{buf} is determined by the following
- 15 equation, for example, by introducing a positive variable k equal to or smaller than 1.

[0064]

$$P_{buf} = \frac{k}{2} V_m \cdot I_m \cdot \cos(2\omega t) \quad \dots (2)$$

- [0065] That is to say, the buffer power P_{buf} is expressed by the product of a DC part ($V_m \cdot I_m / 2$) of the input power P_{in} input from the single-phase AC power supply 1 (or
- 20 further via the filter 2: the same applies to the following), a cosine value $\cos(2\omega t)$ of a value ($2\omega t$) that is twice the phase angle ωt , and the variable k .

[0066] The buffer power P_{buf} takes a positive value in a second time period T_{12} in

which the phase angle ωt of the single-phase AC voltage is 0 to $\pi/4$ inclusive, $3\pi/4$ to $5\pi/4$ inclusive, or $7\pi/4$ to 2π inclusive, and takes a negative value in a first time period T11 other than the second time period T12. That is to say, the power buffer circuit 4 outputs the absolute value of the buffer power P_{buf} to the DC link in the second time period T12, and receives the absolute value of the buffer power P_{buf} from the DC link in the first time period T11.

[0067] The single-phase AC voltage V_{in} is expressed by $V_m \cdot \sin(\omega t)$, and thus, by describing the above-mentioned range in another way, it can also be said that the power buffer circuit 4 outputs positive power when the absolute value of the single-phase AC voltage V_{in} is smaller than a value that is $1/\sqrt{2}$ times the crest value V_m thereof, and outputs negative power when the absolute value of the single-phase AC voltage V_{in} is greater than the value that is $1/\sqrt{2}$ times the crest value V_m thereof.

[0068] The output power P_{out} output from the inverter 5 is expressed by an equation shown below using the input power P_{in} and the buffer power P_{buf} . The DC power P_{dc} input into the inverter 5 and the output power P_{out} output from the inverter 5 are ideally equal to each other.

[0069]

$$\begin{aligned} P_{out} &= P_{dc} = P_{in} + P_{buf} \\ &= \frac{1}{2} V_m \cdot I_m - \frac{1-k}{2} V_m \cdot I_m \cdot \cos(2\omega t) \quad \dots (3) \end{aligned}$$

[0070] As shown in the second term of the rightmost side of the equation (3), a variation range of ripple of the output power P_{out} is indicated by $(1 - k) \cdot V_m \cdot I_m / 2$. Since the variable k is greater than 0 and equal to or smaller than 1, the variation range is smaller than a variation range of the ripple power P_{in} (amplitude of the second term of the rightmost side of the equation (1)). That is to say, the power buffer circuit 4

provides and receives the buffer power P_{buf} shown in the equation (2) to and from the DC link, so that the DC link outputs the DC power P_{dc} having a variation range smaller than the variation range of the ripple power P_{in} to the inverter 5. The variable k corresponds to a magnitude of a difference between the variation range of the ripple power P_{in} and the variation range of the DC power P_{dc} , and indicates a degree of reduction in power ripple. The power ripple is reduced as the variable k increases. It can be seen that, when an equation $k = 1$ holds, for example, the ripple can be canceled out as the second term of the rightmost side of the equation (3) becomes zero. The variable k is hereinafter also referred to as a compensation rate k .

[0071] When the above-mentioned reduction in ripple is described qualitatively, the power buffer circuit 4 receives power from the DC link as an input to store the power when the ripple power P_{in} is large (in the first time period T_{11}), and outputs power to the DC link when the ripple power P_{in} is small (in the second time period T_{12}) to reduce the ripple of the DC power P_{dc} (= output power P_{out}) input into the inverter 5.

[0072] <2-1-2. Equivalent Circuit>

FIG. 6 shows an equivalent circuit of the circuit shown in FIG. 1. In the equivalent circuit shown in FIG. 6, a current i_{rec1} flowing from the converter 3 to the inverter 5 is equivalently expressed as a current i_{rec1} passing through a switch S_{rec} when the switch S_{rec} is conducting. Similarly, a discharge current of the capacitor C_4 is equivalently expressed as a current i_c passing through the switch S_c when the switch S_c is conducting. A current flowing to the motor 6 via the inverter 5 when the output terminals P_u , P_v , and P_w are connected in common to one of the DC power supply lines LH and LL in the inverter 5 is also equivalently expressed as a zero-phase current i_z flowing through a switch S_z when the switch S_z is conducting. FIG. 6 shows the reactor L_4 , the diode D_{40} , and the switch SL constituting the boost circuit 4a, and additionally

shows the current i_L flowing through the reactor L_4 .

[0073] In the equivalent circuit thus obtained, the following equation holds by introducing duties d_{rec} , d_c , and d_z at which the switches S_{rec} , S_c , and S_z are respectively conducting and a DC current I_{dc} input into the inverter 5.

5 [0074]

$$\begin{pmatrix} i_{rec} \\ i_c \\ i_z \end{pmatrix} = \begin{pmatrix} d_{rec} & 1 \\ d_c & 0 \\ d_z & 0 \end{pmatrix} \cdot \begin{pmatrix} I_{dc} \\ i_L \end{pmatrix} \quad \dots (4)$$

[0075] The currents i_{rec1} , i_c , and i_z are respectively obtained by multiplying the DC current I_{dc} by the duties d_{rec} , d_c , and d_z , and thus are average values in switching periods of the respective switches S_{rec} , S_c , and S_z . The current i_L is similarly an average value
10 in switching periods of the switch S_L .

[0076] The DC current I_{dc} is the sum of the currents i_{rec1} , i_c , and i_z respectively passing through the switches S_{rec} , S_c , and S_z , and thus an equation shown below holds. Note that inequations $0 \leq d_{rec} \leq 1$, $0 \leq d_c \leq 1$, and $0 \leq d_z \leq 1$ hold.

[0077]

$$d_{rec} + d_c + d_z = 1 \quad \dots (5)$$

15

[0078] The duties d_{rec} , d_c , and d_z can thus be seen as current distribution ratios of the DC current I_{dc} to the currents i_{rec1} , i_c , and i_z . The duty d_{rec} is a duty at which a time period in which the converter 3 is connected to the DC power supply lines LH and LL to allow a current to flow to the inverter 5 is set, and is thus hereinafter also referred to as a
20 rectification duty d_{rec} . The duty d_c is a duty at which the capacitor C_4 discharges, and is thus hereinafter also referred to as a discharge duty d_c . The duty d_z is a duty at which

the zero-phase current i_z always flows in the inverter 5 regardless of a voltage output from the inverter 5, and is thus hereinafter also referred to as a zero duty d_z .

[0079] The inverter 5 cannot use the DC voltage across the DC power supply lines LH and LL in a time period in which the zero-phase current i_z flows. An imaginary DC voltage (hereinafter, referred to as an “imaginary DC voltage”) V_{dc} of the DC voltage across the DC power supply lines LH and LL in a time period in which the inverter 5 can perform power conversion can thus be considered as shown below. The imaginary DC voltage V_{dc} can be understood as a voltage applied across the DC power supply lines LH and LL as an average of the maximum value of a voltage that the inverter 5 can output in a period in which switching of the switches S_c and S_L and the inverter 5 is controlled.

[0080]

$$V_{dc} = V_{rec} \cdot d_{rec} + V_c \cdot d_c + 0 \cdot d_z \quad \dots (6)$$

[0081] <2-1-3. First Scheme>

The buffer power P_{buf} takes the negative value in the first time period T_{11} as described above. That is to say, the power buffer circuit 4 receives the buffer power P_{buf} from the DC link in the first time period T_{11} . In this case, the power buffer circuit 4 is not required to output power to the DC link. This means that the capacitor C_4 is not required to discharge to the DC link. Thus, the switch S_c may be maintained not conducting in the first time period T_{11} . That is to say, the first time period T_{11} may be used as the above-mentioned time period T_1 . In this case, the discharge duty d_c is maintained zero in the time period T_1 .

[0082] The buffer power P_{buf} takes the positive value in the second time period T_{12} as described above. That is to say, in the second time period T_{12} , the switch S_c is switched between conducting and not conducting, so that the capacitor C_4 discharges to the DC

link. In other words, the second time period T12 is used as the above-mentioned time period T2.

[0083] In the second time period T12, the power buffer circuit 4 is not required to receive power from the DC link. Thus, the switch SL may be maintained not conducting in the second time period T12. In other words, a boost duty dL at which the switch SL is conducting may be maintained zero in the second time period T12.

[0084] In such control, the power buffer circuit 4 only receives power, and does not provide power to the DC link in the first time period T11. On the other hand, the power buffer circuit 4 only provides power, and does not receive power from the DC link in the second time period T12.

[0085] <2-1-4. Second Scheme>

In contrast to the above-mentioned example, the power buffer circuit 4 may allow power provision and reception over the entire time period. This means that the power buffer circuit 4 provides and receives power to and from the DC link also in the first time period T11, but the power buffer circuit 4 receives the buffer power Pbuf shown in the equation (2) from the DC link as the balance of power in the first time period T11. That is to say, power PL that the power buffer circuit 4 receives from the DC link is larger than power Pc that the power buffer circuit 4 provides to the DC link in the first time period T11. Similarly, the power buffer circuit 4 provides the buffer power Pbuf shown in the equation (2) to the DC link as the balance of power in the second time period T12. That is to say, the power Pc is larger than the power PL in the second time period T12. In this scheme, the switch Sc is switched between conducting and not conducting over the entire time period. This scheme thus corresponds to a case where the time period T1 in which the switch Sc is maintained not conducting is set to zero in the above-mentioned second scheme. This means that the entire time period corresponds to the time period

T2 in which the switch Sc is repeatedly switched between conducting and not conducting.

[0086] <2-1-4-1. Specific Control Method in First Scheme>

The method of generating the control signals described in Patent Document 2 will be described herein. A ratio of the imaginary DC voltage Vdc to the crest value Vm is first introduced as a voltage utilization ratio R. An increase in voltage utilization ratio R means an increase in use of the single-phase AC voltage Vin. The rectification duty drec and the discharge duty dc to maximize the voltage utilization ratio R are set by the equation (6) and an equation (7) in the first time period T11 (time period T1). This, however, is setting to maximize the voltage utilization ratio R in a case where the imaginary DC voltage Vdc is constant as suggested in Patent Document 2. The voltage utilization ratio R is (1/√2) in this case. In the first time period T11 (time period T1), an equation dc = 0 holds, and thus the switch Sc is not conducting. In the second time period T12 (time period T2), the capacitor C4 is not charged, and thus the current iL does not flow (see also FIG. 2A).

15 [0087]

$$\left. \begin{aligned} d_{rec} &= \sqrt{2} |\sin(\omega t)| \\ d_c &= \frac{V_m}{V_c} \cdot \frac{\cos(2\omega t)}{\sqrt{2}} \end{aligned} \right) \cdots (7)$$

$$\left. \begin{aligned} d_{rec} &= \frac{1}{\sqrt{2} |\sin(\omega t)|} \\ d_c &= 0 \end{aligned} \right) \cdots (8)$$

[0088] In this “first setting”, the DC current Idc input into the inverter 5 is allowed to

ripple to obtain the buffer power P_{buf} satisfying the equation (2) while maintaining the above-mentioned voltage utilization ratio R , that is, while maintaining the duties in the equations (7) and (8).

[0089] Specifically, the DC current I_{dc} satisfying an equation shown below is input
 5 into the inverter 5. Control of the inverter 5 for satisfying the equation shown below can be achieved by controlling a current command value. Control of the current command value will be described below.

[0090]

$$I_{dc} = \frac{V_m \cdot I_m}{2 \cdot V_{dc}} (1 - (1-k) \cos(2\omega t)) \cdots (9)$$

10 [0091] In the second time period T_{12} (time period T_2), the current i_L is zero as described above, and thus a current i_{rec} is equal to the current i_{rec1} , and is equal to the product $d_{rec} \cdot I_{dc}$. Thus, an equation shown below holds. Note that an equation $V_m/V_{dc} = \sqrt{2}$ is used in view of the voltage utilization ratio R .

[0092]

$$\begin{aligned} i_{rec} &= i_{rec1} = d_{rec} \cdot I_{dc} \\ &= \sqrt{2} |\sin(\omega t)| \cdot \frac{V_m \cdot I_m}{2 \cdot V_{dc}} \cdot (1 - (1-k) \cos(2\omega t)) \\ &= I_m |\sin(\omega t)| \cdot (1 - (1-k) \cos(2\omega t)) \cdots (10) \end{aligned}$$

15

[0093] The current i_{rec} appears as the absolute value of the input current i_{in} , and thus the input current i_{in} differs from a sinusoidal current. In the present embodiment, the crest value I_m is thus a crest value of a current having a sinusoidal waveform with an rms value of the input current i_{in} (i.e., a time average of the square of the input current i_{in}) as

its rms value.

[0094] In the equation (10), however, when an equation $\cos(2\omega t) = 0$ holds independently of the value k , an equation $|\sin(\omega t)| = 1/\sqrt{2}$ holds, and the current i_{rec} takes a value $I_m/\sqrt{2}$. In the present embodiment, the crest value I_m of the current i_{rec} can be understood as a value that is $\sqrt{2}$ times a value that the current i_{rec} (i.e., the absolute value of the current I_{in}) takes when the phase angle ωt is $\pi/4$ or $3\pi/4$ (i.e., at the boundary between the second time period T_{12} and the first time period T_{11}).

[0095] From the equation (10), power input from the single-phase AC power supply 1 in the second time period T_{12} (time period T_2) is obtained by an equation (11):

10 [0096]

$$P_{in} = V_m \cdot I_m \cdot \sin^2(\omega t) \cdot (1 - (1-k) \cos(2\omega t)) \cdots (11)$$

[0097] On the other hand, power input into the inverter 5, that is, the output power P_{out} consumed by the inverter 5 is obtained as the product of the DC current I_{dc} and the imaginary DC voltage V_{dc} , and is obtained by an equation (12):

15 [0098]

$$\begin{aligned} P_{out} &= V_{dc} \cdot I_{dc} \\ &= \frac{V_m \cdot I_m}{2} (1 - (1-k) \cos(2\omega t)) \cdots (12) \end{aligned}$$

[0099] When a difference between the equations (11) and (12) is obtained, an equation (13) holds, and matches the equation (2).

[0100]

$$\begin{aligned}
P_{out} - P_{in} &= \frac{V_m \cdot I_m}{2} [(1 - (1-k) \cos(2\omega t)) - \sin^2(\omega t)] \\
&= \frac{k \cdot V_m \cdot I_m}{2} \cdot \cos(2\omega t) \quad \dots (13)
\end{aligned}$$

[0101] The validity of setting the DC current I_{dc} by the equation (9) is thus explained.

[0102] In the first time period T11 (time period T1), the current i_L is set as shown in an equation (14). Accordingly, power stored in the power buffer circuit 4 in the first time

5 period T11 (time period T1) is calculated as shown in an equation (15), and is equal in absolute value and has polarity opposite to that in the equation (2), which shows the buffer power P_{buf} . It can thus be understood that setting the current i_L by the equation (14) is valid.

[0103]

$$i_L = k \cdot I_m \cdot \left[|\sin(\omega t)| - \frac{1}{2|\sin(\omega t)|} \right] \quad \dots (14)$$

10

$$\begin{aligned}
|V_{in}| \cdot i_L &= V_m |\sin(\omega t)| \cdot k \cdot I_m \cdot \left[|\sin(\omega t)| - \frac{1}{2|\sin(\omega t)|} \right] \\
&= V_m \cdot I_m \cdot k \cdot \left[\sin^2(\omega t) - \frac{1}{2} \right] \\
&= \frac{1}{2} V_m \cdot I_m \cdot k \cdot [-\cos(2\omega t)] \quad \dots (15)
\end{aligned}$$

[0104] The current i_L is controlled on the basis of whether the switch SL is conducting or not. That is to say, the current i_L can have a value shown in the equation (14) through

15 appropriate control of the switch SL. Control of the switch SL for causing the current i_L

to have a value closer to a desired value is known as control of a boost chopper circuit, and thus detailed description thereof is omitted.

[0105] The duties d_{rec} , d_c , and d_z to control the switches S_{rec} , S_c , and S_z are provided by the equations (5), (7), and (8). The switch S_{rec} , however, is not actually provided,
 5 and only appears in the equivalent circuit. The switch S_{rec} is thus equivalently controlled by controlling the switch S_c and the inverter 5 respectively on the basis of the duty d_c and the duty d_z . That is to say, whether the switch S_c is conducting or not is controlled on the basis of the duty d_c shown in the equation (7) in the second time period T_{12} , and the switch S_c is controlled to be not conducting on the basis of the duty d_c
 10 shown in the equation (8) in the first time period T_{11} .

[0106] In the second time period T_{12} , the inverter 5 is caused to use a zero voltage vector on the basis of the zero duty d_z obtained by the equations (5) and (7). The zero voltage vector is a voltage vector when all the switching elements S_{up} , S_{vp} , and S_{wp} or all the switching elements S_{un} , S_{vn} , and S_{wn} are not conducting. In the first time
 15 period T_{11} , the inverter 5 is caused to use the zero voltage vector on the basis of the zero duty d_z obtained by the equations (5) and (8).

[0107] As described above, the switches S_{rec} , S_c , and S_z can equivalently be caused to be conducting respectively on the basis of the duties d_{rec} , d_c , and d_z through control of the switches S_L and S_c and the inverter 5. This leads to reduction in ripple of the DC
 20 power P_{dc} .

[0108] In this first scheme, the voltage utilization ratio R is maximized as described above, but a value thereof is $1/\sqrt{2}$. That is to say, the maximum value of the imaginary DC voltage V_{dc} is $1/\sqrt{2}$ of the crest value V_m .

[0109] <2-1-5. Second Scheme>

25 A case where the switch S_c is switched between conducting and not conducting

over the entire time period in the second scheme will be described next. The method of generating the control signals described in Patent Document 1 will be described herein. The buffer power P_{buf} is expressed by a difference ($P_c - P_L$) between the power P_L input from the DC link into the power buffer circuit 4 and the power P_c output from the power buffer circuit 4 to the DC link. The capacitor C_4 is charged when the power P_L is input from the DC link into the power buffer circuit 4, and thus the power P_L is hereinafter also referred to as charge power P_L . On the other hand, the capacitor C_4 discharges when the power P_c is output from the power buffer circuit 4 to the DC link, and thus the power P_c is hereinafter also referred to as discharge power P_c .

[0110] As examples of the above-mentioned charge power P_L and discharge power P_c , they are determined by the following equations (16) and (17):

[0111]

$$P_L = k_1 \cdot V_m \cdot I_m \cdot \sin^2(\omega t) \quad \dots (16)$$

$$\begin{aligned} P_c &= k_1 \cdot V_m \cdot I_m \cdot \cos^2(\omega t) \\ &= k_1 \cdot V_m \cdot I_m \cdot \cos(2\omega t) + P_L \quad \dots (17) \end{aligned}$$

15

[0112] One half of the compensation rate k is k_1 . The charge power P_L is power that is k_1 times the input power P_{in} as shown in the equation (16), and the discharge power P_c is power that is obtained by adding the charge power P_L to power that is $(-2 \cdot k_1)$ times an AC component of the input power P_{in} as shown in the equation (17).

20 [0113] It is obvious, from the equations (16) and (17) and an equation $P_{buf} = P_c - P_L$, that the equation (2) is satisfied.

[0114] In contrast to Patent Documents 2 and 3 in which the charge power P_L and the

discharge power P_c as described above are provided and received between the power buffer circuit 4 and the DC link in exclusive time periods (the time period T1 and the time period T2) in a power supply phase, the exclusive time periods in the power supply phase are not set.

5 [0115] However, discharge is dominant over charge in the second time period (hereinafter, also referred to as a “discharge dominant time period”) T12 in which an inequation $P_c > P_L$ (i.e., $P_{buf} > 0$) holds, and charge is dominant over discharge in the first time period T11 (hereinafter, also referred to as a “charge dominant time period”) in which an inequation $P_c < P_L$ (i.e., $P_{buf} < 0$) holds. As understood from the equation
 10 (17), a time period in which an inequation $(n + 1/4)\pi \leq \omega t \leq (n + 3/4)\pi$ holds is the charge dominant time period, and a time period in which an inequation $(n + 3/4)\pi \leq \omega t \leq (n + 5/4)\pi$ holds is the discharge dominant time period (n is an integer: the same applies to the following).

[0116] In the second scheme, the time period T1 is zero, and both the first time period
 15 T11 and the second time period T12 are included in the time period T2.

[0117] An art for setting the current i_{rec1} flowing from the converter 3 to the inverter 5 of the current i_{rec} output from the converter 3 dependently of the value k_1 (= one half of the compensation rate k) will be described next.

[0118] The rectified voltage V_{rec} shown in an equation (18) is applied to an output of
 20 the converter 3.

[0119]

$$V_{rec} = |V_{in}| = V_m \cdot |\sin(\omega t)| \quad \dots (18)$$

[0120] Power P_{rec} directly heading from the converter 3 to the inverter 5 is obtained by subtracting the charge power P_L from the input power P_{in} . That is to say, an equation

$P_{rec} = P_{in} - P_L$ holds, and thus the following equation (19) holds.

[0121]

$$P_{rec} = (1-k_1) \cdot V_m \cdot I_m \cdot \sin^2(\omega t) = |V_{in}| \cdot i_{rec1} \dots (19)$$

[0122] The current i_{rec1} is thus expressed by the following equation (20):

5 [0123]

$$i_{rec1} = (1-k_1) \cdot I_m \cdot |\sin(\omega t)| \dots (20)$$

[0124] The equation (1) is based on the assumption that the input current i_{in} is expressed by $I_m \cdot \sin(\omega t)$, that is, the input current i_{in} has a sinusoidal waveform, and thus the current i_L satisfies an equation (21) shown below. This is because, as can be seen from FIG. 1, the current i_{rec} output from the converter 3 is equal to the sum of the current i_{rec1} and the current i_L .

[0125]

$$i_L = k_1 \cdot I_m \cdot |\sin(\omega t)| \dots (21)$$

[0126] The discharge power P_c output from the power buffer circuit 4 is expressed by the product $V_c \cdot i_c$ by introducing the discharge current i_c flowing from the capacitor C4 to the inverter 5. The discharge current i_c should satisfy the following equation (22) so that the discharge power P_c satisfies the equation (17).

[0127]

$$i_c = \frac{P_c}{V_c} = \frac{k_1 \cdot V_m \cdot I_m}{2V_c} (1 + \cos(2\omega t)) \dots (22)$$

20 [0128] The rectification duty d_{rec} and the discharge duty d_c are respectively set by the following equations (23) and (24) from the equations (20), (22), and (4).

[0129]

$$d_{rec} = (1-k_1) \cdot \frac{I_m}{I_{dc}} \cdot |\sin(\omega t)| \quad \dots (23)$$

$$d_c = k_1 \cdot \frac{V_m}{2V_c} \cdot \frac{I_m}{I_{dc}} \cdot (1 + \cos(2\omega t)) \quad \dots (24)$$

5 [0130] That is to say, the equations (20), (22), (23), and (24) are used upon request for the balance of power, and the equation (21) is used upon further request to make the input current I_{in} sinusoidal so that the equation (1) on which the above-mentioned various equations are based is satisfied.

[0131] In a case where the converter 3 uses the diode bridge, the converter 3 cannot
10 actively perform switching at the rectification duty d_{rec} shown by the equation (23) as described above. The current i_{rec1} shown by the equation (20) can thus be obtained by the inverter 5 and the switch S_c performing switching respectively in accordance with the zero duty d_z and the discharge duty d_c determined by the equations (23), (24), and (5).

[0132] Incidentally, the DC power P_{dc} input from the DC link into the inverter 5 is the
15 product of the imaginary DC voltage V_{dc} and the DC current I_{dc} . The inverter 5 obtains, from the DC link, the DC power P_{dc} ($= P_{in} + P_c - P_L$) obtained by subtracting the charge power P_L from the sum of the ripple power P_{in} and the discharge power P_c , and thus the following equation (25) holds.

[0133]

$$I_{dc} = \frac{P_{in} + P_c - P_L}{V_{dc}}$$

$$= \frac{V_m \cdot I_m}{2} (1 - (1-2 \cdot k_1) \cos(2\omega t)) \cdot \frac{1}{V_{dc}} \dots (25)$$

[0134] Note that neither the imaginary DC voltage V_{dc} nor the DC current I_{dc} is limited in setting so far. That is to say, once the value k_1 (= one half of the compensation rate k) is determined, the duties d_{rec} , d_c , and d_z can be set whatever the
 5 imaginary DC voltage V_{dc} and the DC current I_{dc} are set as long as they satisfy the equation (25).

[0135] In Patent Document 1, the duties d_{rec} , d_c , and d_z are calculated by further adding conditions to maximize the voltage utilization ratio (= V_{dc}/V_m). For example, the duty d_z is set to zero. This is because, as can be understood from the equations (5)
 10 and (6), the imaginary DC voltage V_{dc} is maximized when the zero duty d_z is zero.

[0136] When the zero duty d_z is zero, an equation (26) can be obtained from the equations (5), (23), and (24).

[0137]

$$\frac{I_m}{I_{dc}} \{ (1-k_1) \cdot |\sin(\omega t)| + \frac{k_1 \cdot V_m}{2V_c} (1+\cos(2\omega t)) \} = 1$$

$$\dots (26)$$

15 [0138] The zero duty d_z is a duty at which all the switching elements S_{up} , S_{vp} , and S_{wp} or all the switching elements S_{un} , S_{vn} , and S_{wn} are turned on in the inverter 5. The discharge duty d_c is a duty at which the switch S_c is turned on, and the boost duty d_L is a duty at which the switch S_L is turned on. For example, the switch control unit 11 generates the control signals for the inverter 5 and the power buffer circuit 4 on the basis
 20 of each of the duties d_z , d_c , and d_L . For example, each control signal is generated on

the basis of comparison between each duty and a carrier.

[0139] In Patent Document 1, the control signals for the inverter 5 are generated as described below. First, a deviation of the rotational speed ω_m from the rotational speed command value ω_m^* is calculated, and proportional integral control is performed on the deviation $\Delta\omega$. In parallel with this, the output power P_{out} is calculated on the basis of the AC currents i_u , i_v , and i_w and the AC voltage (command value) of the motor 6, and a ripple component thereof is extracted. A deviation ΔP of the ripple component from a target ripple component (i.e., the second term of the rightmost side of the equation (3)) is calculated, and proportional integral control is performed on the deviation ΔP . Next, a result of the proportional integral control performed on the deviation $\Delta\omega$ and a result of the proportional integral control performed on the deviation ΔP are added to generate a current command value. The control signals for controlling the inverter 5 are generated on the basis of the current command value. Generation of the control signals based on the current command value is known, and thus detailed description thereof is omitted.

[0140] In the second scheme, the imaginary DC voltage V_{dc} can be maintained at a constant of the crest value V_m , for example, by adjusting each of the duties d_{rec} , d_c , d_z , and d_L . On the other hand, in the first scheme, the equation $d_c = 0$ holds in the first time period T_{11} (time period T_1), and thus the imaginary DC voltage V_{dc} cannot be higher than the rectified voltage V_{rec} (see the equation (6)). In the case of FIG. 2A, for example, the maximum value of the imaginary DC voltage V_{dc} is $1/\sqrt{2}$ of the crest value V_m as described above. The imaginary DC voltage V_{dc} can thus be increased more in the second scheme.

[0141] In the above-mentioned example, the time period T_1 is set to be shorter when the rotational speed is higher than the speed threshold than when the rotational speed is lower than the speed threshold. As described above, there is a substantially positive

correlation between the rotational speed and the amplitude of the AC voltage applied to the motor 6. The charge and discharge time period setting unit 12 may set the time period T1 so that the time period T1 is reduced when the amplitude of the AC voltage applied to the motor 6 reaches an upper limit. That is to say, a value of the rotational speed ω_m when the amplitude of the AC voltage applied to the motor 6 reaches the upper limit may be used as the speed threshold ω_{ref} . The upper limit is determined by the DC voltage Vdc1. The fact that the amplitude has reached the upper limit can be detected through comparison between the amplitude of the AC voltage and the DC voltage Vdc1.

[0142] By setting the time period T1 so that the time period T1 is reduced when the amplitude of the AC voltage applied to the motor 6 reaches the upper limit, the DC voltage Vdc1 can be increased to increase the upper limit of the amplitude of the AC voltage that can be output to the motor 6 to thereby further increase the rotational speed ω_m . Thus, the rotational speed can be increased without using the field weakening (magnetic flux weakening) control. As the amplitude, a command value on the AC voltage output from the inverter 5 may be used. That is to say, the time period T1 may be set to be reduced when an amplitude command value on the AC voltage reaches an upper limit (e.g., the DC voltage Vdc1).

[0143] Although control is performed on the basis of the equation (3) in the above-mentioned example, control is not limited to this.

[0144] While the power converter control device has been described in detail, the foregoing description is in all aspects illustrative, and does not limit the power converter control device. It is therefore understood that numerous modifications not having been described can be devised without departing from the scope of the power converter control device.

CLAIMS

1. A power converter control device for controlling a discharge switch in a power converter including first and second DC power supply lines, a rectifying circuit, said discharge switch and a capacitor, a boost circuit, and an inverter, said rectifying
5 circuit full-wave rectifying a single-phase AC voltage to a rectified voltage, and applying said rectified voltage across said first and second DC power supply lines, said discharge switch and said capacitor being connected in series to each other between said first and second DC power supply lines, said boost circuit boosting said rectified voltage to charge said capacitor, said inverter receiving, as an input, said rectified voltage as a DC voltage
10 across said first and second DC power supply lines when said discharge switch is not conducting, receiving, as an input, a voltage across said capacitor as said DC voltage when said discharge switch is conducting, converting said DC voltage as input into an AC voltage, and outputting said AC voltage to a motor, said power converter control device comprising:

15 a switch control unit that maintains said discharge switch not conducting over a first time period, and switches said discharge switch between conducting and not conducting in a second time period other than said first time period; and

a charge and discharge time period setting unit that sets said first time period when a rotational speed of said motor is higher than a speed threshold shorter than said
20 first time period when said rotational speed is lower than said speed threshold.

2. The power converter control device according to claim 1, wherein

when said rotational speed is higher than said speed threshold, said first time period has a length of zero.

3. The power converter control device according to claim 1 or 2, wherein
when said rotational speed is lower than said speed threshold, said first time
period is a time period in which said rectified voltage is greater than a predetermined
value.

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4. The power converter control device according to any one of claims 1 to 3,
wherein
said speed threshold is said rotational speed when amplitude of said AC voltage
applied to said motor reaches an upper limit.

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