

Description

Title of Invention: POWER CONVERTER

Technical Field

[0001] The present invention relates to a power converter including a power buffer
5 circuit.

Background Art

[0002] A full-wave rectifying circuit is typically used to obtain a DC voltage from a single-phase AC voltage input from a single-phase AC power supply. Output of the full-wave rectifying circuit, however, has power pulsations having a frequency twice the
10 frequency of the single-phase AC voltage. To reduce the power pulsations, a power buffer circuit for buffering a power is required between an output of the full-wave rectifying circuit and a load. A capacitive element, for example, referred to as a smoothing capacitor is required for buffering of a power.

[0003] Non-patent Document 1 discloses a technique for connecting a buffer capacitor
15 to a smoothing capacitor via a current-reversible chopper to absorb a pulsating power. This technique reduces the electrostatic capacitance of the smoothing capacitor and further allows voltage ripples on a buffer side to significantly reduce the total electrostatic capacitance required for smoothing.

[0004] Non-patent Document 2 discloses a technique for connecting a buffer capacitor
20 to a DC link via a switching element while removing the smoothing capacitor of Non-patent Document 1. A direct conversion circuit that generates a voltage source by using this technique, and generates a high-frequency link along with a power supply voltage is shown.

[0005] Non-patent Document 3 further discloses a technique for converting an input
25 waveform into a sinusoidal waveform, and increasing efficiency.

[0006] Patent Documents 1 and 2 applied by the applicant of the present application are listed as documents that disclose techniques relating to the present application.

Prior Art Documents

Patent Documents

5 [0007] Patent Document 1: Japanese Patent Application Laid-Open No. 2011-193678

Patent Document 2: Japanese Patent Application Laid-Open No. 2012-135184

Non-patent Documents

[0008] Non-patent Document 1: Irie, Yamashita, and Takemoto, "Ripple Compensation for a Single-Phase Rectifier by 2-Quadrant Chopper and Auxiliary Capacitor", the
10 Institute of Electrical Engineers of Japan Transactions D, Vol. 112, No. 7, pp. 623-629 (1992)

Non-patent Document 2: Ohnuma, Itoh, "Circuit Configuration and Control Strategy of single-to-three Phase Power Converter with Active Buffer and Charge Circuit", the 2010 Annual Meeting of the Institute of Electrical Engineers of Japan, 4-057
15 (2010)

Non-patent Document 3: Ohnuma, Itoh, "Comparison between a Boost Chopper and an Active Buffer as a Single to Three Phase Converter", the 2011 Annual Meeting of the Institute of Electrical Engineers of Japan, 4-042 (2011)

Non-patent Document 4: Haga, Takahashi, Ohishi, "Unity Power Factor
20 Operation Control Method For Single-phase to Three-phase Matrix Converter", the Institute of Electrical Engineers of Japan Transactions D, Vol. 124, No. 5, pp. 510-516 (2004)

Summary of Invention

Problems to be Solved by Invention

[0009] In power conversion using the technique shown in Non-patent Document 1, a voltage of the buffer capacitor is not required to be controlled to be constant, and thus can be set to be smaller than a normal smoothing capacitor in required electrostatic capacitance, regardless of a direct conversion type or an indirect conversion type.

5 [0010] However, the amplitude of the pulsating power is as great as that of a constant component of a power. A value of an allowable ripple current required by the buffer capacitor is thus high. On the other hand, it is desirable to use an electrolytic capacitor as the buffer capacitor in view of device miniaturization and cost restraints. Due to the use of the electrolytic capacitor as the buffer capacitor, even the buffer capacitor has to
10 have an electrostatic capacitance that is larger than an actually-required electrostatic capacitance and is approximately equal to the electrostatic capacitance of the normal smoothing capacitor to meet the value of the allowable ripple current.

[0011] Besides, Non-patent Document 4 shows a single-phase to three-phase direct conversion circuit not including an active power buffer circuit. The relation between an
15 input current and output power ripples is shown herein. The need to provide a pulsating power having a frequency twice a power supply frequency to an output power is shown to convert an input waveform into a sinusoidal waveform.

[0012] In a household air conditioner of a 1.5 kW class, for example, speed ripples involved in pulsations of a load torque are expected to be suppressed to be approximately
20 10% by the moment of inertia of a motor. However, the stress of a compressor support system increases as a pulsating torque increases with increasing power capacity. It is therefore clear that power capacity that can be used has an upper limit.

[0013] Embodiments of the present application seek to make, in a power converter including an active power buffer circuit, a power buffered by the power buffer circuit smaller than a
25 pulsating power, and to thereby reduce the electrostatic capacitance and the power

capacity of the power buffer circuit. This may lead to expansion of the range of the power capacity in a case where the load pulsates.

Means for Solving the Problems

[0014] According to an aspect of the present invention, there is provided a power converter according to the present invention includes: a DC link (7) that includes a first power supply line (LH) and a second power supply line (LL); a converter (3) that receives a single-phase AC voltage (V_{in}), and outputs a pulsating power (P_{in}) to the DC link; an inverter (5) that receives a power from the DC link, and outputs an AC current (I_u , I_v and I_w); and a power buffer circuit (4) that receives a charge power (P_l) from the DC link, and outputs a discharge power (P_c) to the DC link.

[0015] When at least the pulsating power exceeds a predetermined value, a variation ($P_c - P_l$) of the discharge power to the charge power is smaller than an AC component ($\hat{P}_{in}$) of the pulsating power.

[0016] According to an embodiment of the present invention, an input power (P_{dc}) input from the DC link (7) into the inverter (5) takes a value ($P_{in} + P_c - P_l$) obtained by subtracting the charge power from a sum of the pulsating power and the discharge power. When at least the pulsating power exceeds the predetermined value, the charge power (P_l) takes a value ($k \cdot P_{in}$) which is a constant (k) (where the constant is a positive number less than $1/2$) times the pulsating power (P_{in}), and the discharge power (P_c) takes a value ($2k \cdot \hat{P}_{in} + P_l$) obtained by adding the charge power to (-2) times the AC component ($\hat{P}_{in}$) of the pulsating power multiplied by the constant.

[0017] According to an embodiment of the present invention, the converter (3) applies a rectified voltage (V_{rec}) obtained by full-wave rectifying the single-phase AC voltage ($V_m \cdot \sin(\omega t)$) to the DC link (7) so that a higher electric potential is applied to the first power supply line (LH) than the second

power supply line (LL), and allows, when at least the pulsating power exceeds the predetermined value, a current $((1 - k) \cdot I_m \cdot |\sin(\omega t)|)$ obtained by multiplying a first current by a value $(1 - k)$ obtained by subtracting the constant from 1 to flow through the DC link, the first current being a current $(I_m \cdot |\sin(\omega t)|)$ obtained by dividing the pulsating power (P_{in}) by the rectified voltage (V_{rec}). The power buffer circuit (4) includes: a discharge circuit (4a) that includes a capacitor (C4) and a first switch (Sc and D42) that is connected in series to the capacitor between the first power supply line and the second power supply line so as to be closer to the first power supply line than the capacitor is; and a charge circuit (4b) that charges the capacitor. When at least the pulsating power exceeds the predetermined value, the power buffer circuit (4) receives a charge current ($i_l = k \cdot I_m \cdot |\sin(\omega t)|$) which is the constant times the first current, and outputs a discharge current (i_c) that is a current (P_c/V_c) obtained by dividing the discharge power (P_c) by a both-end voltage (V_c) across the capacitor.

[0018] According to an embodiment of the present invention, the converter (3) applies a rectified voltage (V_{rec}) obtained by full-wave rectifying the single-phase AC voltage (V_{in}) to the DC link (7) so that a higher electric potential is applied to the first power supply line (LH) than the second power supply line (LL). The power buffer circuit (4) includes: a discharge circuit (4a) that includes a capacitor (C4) and a first switch (Sc and D42) that is connected in series to the capacitor between the first power supply line and the second power supply line so as to be closer to the first power supply line than the capacitor is; and a charge circuit (4b) that charges the capacitor. A rectifying duty (d_{rec}) that is a duty at which the converter is in conduction with the DC link (7) takes a value $(V_{dc}/V_m \cdot |\sin(\omega t)|)$ obtained by dividing a product of a predetermined voltage (V_{dc}) and a square ($\sin^2(\omega t)$) of a sine value of a phase (ωt) of the single-phase AC voltage by the rectified voltage (V_{rec}). A discharge

duty (dc) that is a duty at which the capacitor discharges takes a value ($V_{dc}/V_c \cdot \cos^2(\omega t)$) obtained by dividing a product of the predetermined voltage and a square ($\cos^2(\omega t)$) of a cosine value of the phase by a both-end voltage (V_c) across the capacitor.

[0019] According to an embodiment of the present invention, the constant (k) is independent
5 of the AC current (I_u , I_v and I_w).

[0020] According to an embodiment of the present invention, the constant (k) is inversely proportional to a magnitude of the AC current (I_u , I_v and I_w).

[0021] According to an embodiment of the present invention, the predetermined value is set so as to be smaller than a maximum rated power of the pulsating power.

10 [0022] According to an embodiment of the present invention, when the pulsating power is equal to or smaller than a predetermined value, the variation ($P_c - P_l$) of the discharge power to the charge power is equal to the AC component ($P_{in}^{\wedge}$) of the pulsating power.

[0023] According to an embodiment of the present invention, when the pulsating power exceeds a predetermined value, the variation ($P_c - P_l$) of the discharge power to the charge power
15 is independent of the pulsating power.

Effects of Invention

[0024] According to the aspect of the power converter according to the present invention, the variation of the discharge power to the charge power, which is a power buffered by the power buffer circuit, is smaller than the AC component of the pulsating power. The power capacity of
20 the power buffer circuit can thus be reduced compared with a conventional technique.

[0025] The power buffered by the power buffer circuit may be twice a value obtained by multiplying the AC component of the pulsating power by the constant. Since the constant is less than 1/2, the power buffered may be smaller than that buffered by a conventional power buffer circuit, and leads to achievement of the aspect.

- 5 [0026] A period in which a zero-phase current flows through the inverter can be reduced to thereby increase a period in which a voltage applied to a DC link is used.

[0027] The sum of the product of the rectifying duty and the rectified voltage and the product of the discharge duty and the both-end voltage corresponds to an average value of a DC voltage used by the inverter. The sum matches the predetermined voltage, and can be
10 set to a constant value independently of a phase of the single-phase AC voltage.

[0028] A ripple current of the capacitor can be reduced in a light-load range, and the life of the power converter can be increased.

[0029] When a torque of a motor serving as the load of the power converter is small, vibrations of the motor in a low-rpm region can be reduced.

- 15 [0030] Objects, features and advantages of the present invention will become more apparent from the following detailed description and the accompanying drawings.

Brief Description of Drawings

[0031] FIG. 1 is a block diagram showing configuration of a direct power converter according to an embodiment;

FIG. 2 is a block diagram showing a power balance in the direct power converter according to the embodiment;

FIG. 3 is a graph showing the relation among an input power, the electrostatic capacitance of a capacitor, and a ripple current in the direct power converter;

FIG. 4 is a circuit diagram showing an equivalent circuit of a circuit shown in FIG. 1;

FIGS. 5 and 6 are graphs showing operation of the direct power converter according to the embodiment;

FIG. 7 is a graph showing a virtual DC voltage,

FIG. 8 is a graph showing virtual operation of the direct power converter according to the embodiment;

FIGS. 9 and 10 are graphs showing operation of the direct power converter according to the embodiment;

FIG. 11 is a block diagram showing an example of configuration for performing control to achieve a DC current; and

FIG. 12 is a circuit diagram showing a modification of the direct power converter shown in FIG. 1.

Description of Embodiment

[0032] A. Configuration of Direct Power Converter

FIG. 1 is a block diagram showing configuration of a direct power converter 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, an inverter 5, and a DC link 7.

[0033] The converter 3 is connected to a single-phase AC power supply 1, for example, through a filter 2. The filter 2 includes a reactor L2 and a capacitor C2. The reactor L2 is provided between one of two output terminals of the single-phase AC power supply 1 and the converter 3. The capacitor C2 is provided 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. For the sake of simplicity, the function of the filter 2 is ignored in the following description.

[0034] The DC link 7 includes DC power supply lines LH and LL.

[0035] The converter 3 uses a diode bridge, for example, and includes diodes D31 to D34. The diodes D31 to D34 constitute a bridge circuit, and convert a single-phase AC voltage V_{in} , which is an input voltage input from the single-phase AC power supply 1, by single-phase full-wave rectifying into a rectified voltage V_{rec} , and output it across the DC power supply lines LH and LL. A higher electric potential is applied to the DC power supply line LH than to the DC power supply line LL. An input current I_{in} flows from the single-phase AC power supply 1 into the converter 3.

[0036] The power buffer circuit 4 includes a discharge circuit 4a and a charge circuit 4b, and provides and receives powers to and from the DC link 7. The discharge circuit 4a includes a capacitor C4, and the charge circuit 4b boosts the rectified voltage V_{rec} , and charges the capacitor C4.

[0037] The discharge circuit 4a 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 the side close to the DC power supply line LH. Herein, being connected in antiparallel refers to being connected in parallel with forward directions being opposite to each other. Specifically, a forward direction of the transistor Sc 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 Sc and the diode D42 can collectively be understood as a switch element (first switch). When the first switch is conductive, the capacitor C4 discharges to provide a power to the DC link 7.

[0038] The charge circuit 4b includes a diode D40, a reactor L4, and a transistor (herein an IGBT) Sl, for example. The diode D40 has a cathode and an anode, and the cathode is connected between the first switch and the capacitor C4. This configuration is known as a so-called boost chopper.

[0039] 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. The transistor Sl is connected in antiparallel to a diode D41, and they can collectively be understood as a switch element (second switch). Specifically, a forward direction of the transistor Sl is a direction from the DC power supply line LH towards the DC power supply line LL, and a forward direction of the diode D41 is a direction from the DC power supply line LL towards the DC power supply line LH.

[0040] The capacitor C4 is charged by the charge circuit 4b, and a both-end voltage Vc higher than the rectified voltage Vrec is generated. 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.

[0041] Since the both-end voltage V_c is higher than the rectified voltage V_{rec} , a current basically does not flow through the diode D42. Conduction/non-conduction of the first switch is thus solely dependent on that of the transistor Sc. Not only the transistor Sc but also the first switch including the transistor Sc and the diode D42 may be thus hereinafter also referred to as a switch Sc.

[0042] Since the DC power supply line LH is higher than the DC power supply line LL in an electric potential, a current basically does not flow through the diode D41. Conduction/non-conduction of the second switch is thus solely dependent on that of the transistor Sl. Not only the transistor Sl but also the second switch including the transistor Sl and the diode D41 may be thus hereinafter also referred to as a switch Sl.

[0043] The inverter 5 converts a DC voltage across the DC power supply lines LH and LL into an AC voltage, and outputs it to output terminals Pu, Pv and Pw. 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 comprises a so-called voltage source inverter, and includes six diodes Dup, Dvp, Dwp, Dun, Dvn and Dwn.

[0044] 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. AC currents Iu, Iv, and Iw are respectively output from the output terminals Pu, Pv, and Pw, and constitute a three-phase AC current. IGBTs are used as the switching elements Sup, Svp, Swp, Sun, Svn, and Swn, for example.

- 5 [0045] An inductive load 6 is a rotary machine, for example, and is shown by an equivalent circuit representing an inductive load. Specifically, a reactor Lu and a resistor Ru are connected in series to each other, and one end of the series body is connected to the output terminal Pu. The same applies to a reactor Lv and a resistor Rv and to a reactor Lw and a resistor Rw. Respective opposite ends of these series bodies
10 are connected to one another.

[0046] B. Control Method

(b-1) Basic Concept of Power Reduction

An instantaneous input power Pin input into the converter 3 is expressed by the following formula (1) with an input power factor being 1. Note that a crest value Vm
15 and a power supply angular velocity ω of the single-phase AC voltage Vin, a crest value Im of the input current Iin, and time t are introduced. The product ωt of the power supply angular velocity ω and the time t represents a phase of the single-phase AC voltage Vin. An AC waveform is understood as the product of a sine value of the phase ωt of the AC waveform and a crest value.

20 [0047]

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

[0048] The instantaneous input power Pin has an AC component $(-1/2) \cdot V_m \cdot I_m \cdot \cos(2\omega t)$ (hereinafter, also referred to as an “AC component Pin[^]”) shown

by the second term of the right-hand side of the formula (1). The instantaneous input power P_{in} is thus hereinafter also referred to as a pulsating power P_{in} .

[0049] The power converter shown in FIG. 1 can be understood as follows:

[0050] The DC link 7 includes the DC power supply lines LH and LL;

5 The converter 3 receives the single-phase AC voltage V_{in} , and outputs the pulsating power P_{in} ;

 The power buffer circuit 4 receives a charge power P_l from the DC link 7, and outputs a discharge power P_c to the DC link 7; and

 The inverter 5 receives, from the DC link 7, an input power P_{dc} ($= P_{in} + P_c - P_l$) obtained by subtracting the charge power P_l from the sum of the pulsating power P_{in} and the discharge power P_c , and outputs the AC currents I_u , I_v , and I_w .

[0051] FIG. 2 is a block diagram schematically showing a power balance in the direct power converter shown in FIG. 1. A power subjected to buffering (hereinafter, referred to as a “buffering power P_{buf} ”) is equal to a power difference ($P_c - P_l$) obtained by
15 subtracting the charge power P_l from the discharge power P_c . A power P_{rec} directed from the converter 3 to the inverter 5 is equal to $P_{in} - P_l$. Thus, $P_{dc} = P_{rec} + P_c$ holds.

[0052] In conventional techniques shown in Non-patent Document 2 and Patent Document 1, the charge power P_l and the discharge power P_c each take the same value as the AC component $\hat{P}_{in}$, and are provided and received between the power buffer circuit 4
20 and the DC power supply lines LH and LL in mutually exclusive periods of a power supply phase to counteract the above-mentioned AC component $\hat{P}_{in}$. A power subjected to buffering by the power buffer circuit 4 is thus equal to the absolute value of the AC component $\hat{P}_{in}$, and the power buffer circuit 4 is required to have a power capacity equal to or larger than the absolute value of the AC component $\hat{P}_{in}$.

[0053] Therefore, described is which technique can reduce the buffering power P_{buf} to reduce the power capacity of the power buffer circuit 4.

[0054] It is of course undesirable to set the buffering power P_{buf} to completely zero as a power input into the inverter 5 pulsates with the second term of the right-hand side of the formula (1), and there is no need to reduce the power capacity of the power buffer circuit 4 in the first place. The buffering power P_{buf} is determined by the following formula (2) by introducing a positive constant k less than $1/2$. In other words, achievement of such buffering power P_{buf} leads to reduction in power capacity of the power buffer circuit.

10 [0055]

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

[0056] This means that a difference between powers provided and received between the power buffer circuit 4 and the DC power supply lines LH and LL satisfies the formula (2). In other words, if the absolute value of a power ($P_c - P_l$) that is a variation of the discharge power P_c to the charge power P_l is smaller than the absolute value of the AC component $\hat{P}_{in}$, the power capacity of the power buffer circuit 4 can be reduced compared with a conventional technique.

[0057] By introducing the constant k less than $1/2$ as described above, the buffering power P_{buf} can be selected to be small compared with the techniques disclosed in Non-patent Documents 2 and 3. This results in the following advantages of the power buffer circuit 4.

[0058] (b-2) Description of Advantages in Discharge Circuit 4a

In this section, reduction of the buffering power P_{buf} allows the use of an electrolytic capacitor as the capacitor C_4 and achieving the discharge circuit 4a inexpensively is described.

[0059] FIG. 3 is a graph showing the relation between the electrostatic capacitance (also simply referred to as a “capacitor capacity”) (a left vertical axis) of the capacitor used or required in the direct power converter receiving and processing an input power (an average value of the instantaneous input power P_{in} : a horizontal axis) and a ripple current and an allowable value (a right vertical axis) of the ripple current in the capacitor.

[0060] Symbols H1 and H2 each represent the capacitor capacity used in a case where a single-phase power factor correction circuit is used. The symbols H1 and H2 respectively represent data in cases of using air conditioners having air conditioning abilities of 6 kW and 11.2 kW. The power factor correction circuit can herein be understood to have configuration in which the switch S_c has been shorted and removed from the power buffer circuit 4, and a connection point between the reactor L4 and the converter 3 is not directly connected to the inverter 5. In this configuration, the diode D40 and the reactor L4 connected in series to each other are provided between the converter 3 and the inverter 5 in the DC power supply line LH, and the capacitor C4 is connected in parallel to the inverter 5 between the DC power supply lines LH and LL.

[0061] In view of the fact that $P_{dc} = P_{in} - P_{in}^{\wedge}$ holds when $k = 1/2$, the electrostatic capacitance C required for the capacitor C4 can be obtained from Non-patent Document 3 by the following formula (3). Note that a maximum value V_{cmax} and a minimum value V_{cmin} of the both-end voltage V_c are introduced.

[0062]

$$C = \frac{V_m \cdot I_m}{\omega(V_{cmax}^2 - V_{cmin}^2)} \quad \dots(3)$$

[0063] On a graph G1, calculated values required for capacitor capacities when $k = 1/2$ (i.e., the buffering power P_{buf} is equal to the absolute value of the AC component $P_{in}^{\wedge}$) are plotted. Note that the maximum value $V_{cmax} = 400 + 50 = 450$ (V) and the

minimum value $V_{cmin} = 400 - 50 = 350$ (V) hold. Solid circles on the graph G1 indicate the same cases indicated by the symbols H1 and H2 in the input power. However, capacitor capacities that are actually used are plotted as the symbols H1 and H2, and, when they are back-calculated from the formula (3), the both-end voltage V_c varies
 5 by approximately $\pm 5\%$.

[0064] As can be understood from comparison between the symbols H1 and H2 and the solid circles on the graph G1, the capacitor capacities are reduced to approximately $1/3$ to $1/4$ times compared with a normal power factor correction circuit when $k = 1/2$.

[0065] When $k = 1/2$, however, a ripple current shown by a graph G2 flows through the
 10 capacitor C4. On the other hand, a ripple current (e.g., values obtained at 45°C using an electrolytic capacitor of a GW series (105°C specifications) from Nichicon Corporation) allowed when the capacitor capacity obtained by the graph G1 is obtained using the electrolytic capacitor is shown by a graph G3. Comparison between the graphs G2 and G3 suggests that using the electrolytic capacitor cannot achieve the capacitor C4 when k
 15 $= 1/2$ from the perspective of allowing the ripple current.

[0066] However, the ripple current flowing through the capacitor C4 can be reduced by setting the constant k so that the constant k is less than $1/2$ compared to a case where $k = 1/2$. The capacitor C4 can be achieved by the electrolytic capacitor by setting the constant k in accordance with a desired capacitor capacity, leading to achieving the power
 20 buffer circuit 4 inexpensively.

[0067] (b-3) Description of Advantages in Charge Circuit 4b

In this section, reduction of the buffering power P_{buf} to achieve the charge circuit 4b inexpensively is described.

[0068] All the instantaneous input powers P_{in} pass through a normal power factor
 25 correction circuit. On the other hand, when the power buffer circuit 4 is used, a power

passing through the charge circuit 4b is the charge power P_l , which corresponds to the buffering power P_{buf} expressed by the formula (2) in a case where the buffering power P_{buf} is positive (see FIG. 2).

[0069] The power capacity required for the charge circuit 4b decreases as the buffering power P_{buf} decreases. By reducing the buffering power P_{buf} , a current i_l (this can be understood as a charge current input into the charge circuit 4b) flowing through the reactor L4 is reduced, and thus the power capacity required for the reactor L4 is also reduced.

[0070] (b-4) Specific Examples of Charge Power P_l and Discharge Power P_c

10 In and after this section, the above-mentioned charge power P_l and discharge power P_c are respectively determined by the formulas (4) and (5) as examples.

[0071]

$$P_l = k \cdot V_m \cdot I_m \cdot \sin^2(\omega t) \quad \dots(4)$$

[0072]

15
$$\begin{aligned} P_c &= k \cdot V_m \cdot I_m \cdot \cos^2(\omega t) \\ &= k \cdot V_m \cdot I_m \cdot \cos(2\omega t) + P_l \quad \dots(5) \end{aligned}$$

[0073] This means that the charge power P_l is a power which is the constant k times the pulsating power P_{in} , and the discharge power P_c is a power obtained by adding the charge power P_l to (-2) times the AC component $\hat{P}_{in}$ multiplied by the constant k .

[0074] The charge power P_l can be understood as a power $k \cdot P_{in}$ distributed from the pulsating power P_{in} to the power buffer circuit 4 via the DC link 7 at a distribution ratio of the constant k . The constant k is thus hereinafter also referred to as a buffer distribution ratio k .

20

[0075] It is obvious from the formulas (4) and (5) and $P_{buf} = P_c - P_l$ that the formula (2) is satisfied.

[0076] In contrast to Patent Document 1 and Non-patent Document 2 in which the charge power P_l and the discharge power P_c are provided and received between the power buffer circuit 4 and the DC link 7 in the exclusive periods of the power supply phase, the exclusive periods of the power supply phase for the charge power P_l and the discharge power P_c are not set.

5 [0077] However, discharge is dominant over charge in a period (hereinafter, also referred to as a “discharge dominant period”) in which $P_c > P_l$ (i.e., $P_{buf} > 0$), and charge is dominant over discharge in a period (hereinafter, also referred to as a “charge dominant period”) in which $P_c < P_l$ (i.e., $P_{buf} < 0$). As can be understood from the formulas (4) and (5), a period in which $(n+1/4)\pi \leq \omega t \leq (n+3/4)\pi$ is the charge dominant period, and a period in which $(n+3/4)\pi \leq \omega t \leq$
 10 $(n+5/4)\pi$ is the discharge dominant period (n is an integer: the same applies hereinafter).

[0078] By determining the charge power P_l and the discharge power P_c as described above, a description of the power capacity required for the charge circuit 4b can be more quantitatively than the previous section.

[0079] An average value of a power, input into a conventional power factor correction circuit,
 15 i.e., input into the reactor, per period of the power supply frequency can be obtained from the formula (6) in view of the formula (1).

[0080]

$$\frac{V_m \cdot I_m}{2\pi} \int_0^{2\pi} \frac{1}{2} (1 - \cos 2\theta) d\theta = \frac{V_m \cdot I_m}{2} \quad \dots (6)$$

[0081] On the other hand, when the power buffer circuit 4 is used, the charge power P_l is input
 20 into the power buffer circuit. In view of the formula (4), an average value of a power, input into the power buffer circuit 4, i.e., input into the charge circuit 4b, per period of the power supply frequency can be obtained from the formula (7).

[0082]

$$\frac{V_m \cdot I_m}{2\pi} \int_0^{2\pi} |k \cdot \sin^2 2\theta| d\theta = \frac{V_m \cdot I_m}{2} \cdot k \quad \dots(7)$$

[0083] As can be understood from comparison between the formulas (6) and (7), the power capacity required for the charge circuit 4b is k times compared with the power factor correction circuit. By setting the buffer distribution ratio k to less than 1/2, the power capacity required for the charge circuit 4b is reduced to be smaller than one half of the power capacity in the power factor correction circuit.

[0084] (b-5) Current Distribution Dependent on Constant k

In this section, a technique for setting a current irec1 flowing from the converter 3 to the inverter 5 out of a current irec output from the converter 3 dependently on the buffer distribution ratio k is described.

[0085] The rectified voltage Vrec expressed by the formula (8) is applied to an output side of the converter 3.

[0086]

$$V_{rec} = |V_{in}| = V_m \cdot |\sin(\omega t)| \quad \dots(8)$$

[0087] The following formula (9) holds as $P_{rec} = P_{in} - P_l$.

[0088]

$$P_{rec} = (1 - k) \cdot V_m \cdot I_m \cdot \sin^2(\omega t) = |V_{in}| \cdot irec1 \quad \dots(9)$$

[0089] The current irec1 is thus expressed by the following formula (10).

[0090]

$$irec1 = (1 - k) \cdot I_m \cdot |\sin(\omega t)| \quad \dots(10)$$

[0091] The formula (1) is based on the premise that the input current Iin is expressed by $I_m \cdot \sin(\omega t)$, i.e., has a sinusoidal waveform, and thus the current il satisfies the following formula (11). As can be seen from FIG. 2, this is because the current irec output from the converter 3 is equal to the sum of the current irec1 and the current il.

[0092]

$$i_l = k \cdot I_m \cdot |\sin(\omega t)| \quad \dots(11)$$

[0093] A peak value of the current i_l is thus $k \cdot I_m$. In view of the fact that the peak value of the current input into the conventional power factor correction circuit is I_m , it is understood that the power capacity required for the reactor L4 is reduced compared with a conventional technique.

[0094] When a discharge current i_c flowing from the capacitor C4 to the inverter 5 is introduced, the discharge power P_c output from the power buffer circuit 4 is expressed by the product $V_c \cdot i_c$. In order for the discharge power P_c to satisfy the formula (5), the discharge current i_c thus satisfies the following formula (12).

[0095]

$$i_c = \frac{P_c}{V_c} = \frac{k \cdot V_m \cdot I_m}{2V_c} (1 + \cos(2\omega t)) \quad \dots(12)$$

[0096] FIG. 4 shows an equivalent circuit of the circuit shown in FIG. 1. The equivalent circuit is introduced in Patent Document 1, for example. In the equivalent circuit, the current i_{rec1} is equivalently expressed as a current i_{rec1} passing through a switch S_{rec} when the switch S_{rec} is conductive. Similarly, the discharge current i_c is equivalently expressed as a current i_c passing through the switch S_c when the switch S_c is conductive. A current flowing through the inverter 5 to the inductive load 6 when the output terminals P_u , P_v and P_w are connected in common to any 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 conductive. FIG. 4 also shows the reactor L4, the diode D40 and the switch S1 constituting the charge circuit 4b, and additionally shows the current i_l flowing through the reactor L4.

[0097] In the equivalent circuit thus obtained, the following formula holds by introducing duties d_{rec} , d_c , and d_z at which the switches S_{rec} , S_c , and S_z are respectively conductive and a DC current I_{dc} input into the inverter 5.

[0098]

$$5 \quad \begin{bmatrix} i_{rec} \\ i_c \\ i_z \end{bmatrix} = \begin{bmatrix} d_{rec} & 1 \\ d_c & 0 \\ d_z & 0 \end{bmatrix} \cdot \begin{bmatrix} I_{dc} \\ i_l \end{bmatrix} \quad \dots(13)$$

[0099] The currents i_{rec1} , i_c and i_z are obtained by multiplying the DC current I_{dc} by the duties d_{rec} , d_c and d_z , respectively, and thus are average values in switching cycles of the respective switches S_{rec} , S_c and S_z .

[0100] The DC current I_{dc} is the sum of the currents i_{rec1} , i_c , and i_z flowing through
10 by the respective switches S_{rec} , S_c and S_z , and thus the following formula holds. Note that $0 \leq d_{rec} \leq 1$, $0 \leq d_c \leq 1$, and $0 \leq d_z \leq 1$.

[0101]

$$d_{rec} + d_c + d_z = 1 \quad \dots(14)$$

[0102] The duties d_{rec} , d_c and d_z can thus be considered as current distribution ratios
15 of the DC current I_{dc} to the respective currents i_{rec1} , i_c , and i_z . The duty d_{rec} is hereinafter also referred to as a rectifying duty d_{rec} , as the duty d_{rec} is a duty at which a period in which the converter 3 is connected to the DC link 7, and allows a current to flow through the inverter 5 is set. The duty d_c may be hereinafter also referred to as a discharge duty d_c as the duty d_c is a duty at which the capacitor C4 discharges. The
20 duty d_z may be hereinafter also referred to as a zero duty d_z as the duty d_z is a duty at which the zero-phase current i_z always flows through the inverter 5 regardless of a voltage output from the inverter 5.

[0103] The rectifying duty d_{rec} and the discharge duty d_c are respectively set by the following formulas (15) and (16) from the formulas (10), (12) and (13).

[0104]

$$d_{rec} = (1-k) \cdot \frac{I_m}{I_{dc}} \cdot |\sin(\omega t)| \quad \dots(15)$$

5 [0105]

$$d_c = k \cdot \frac{V_m}{2V_c} \cdot \frac{I_m}{I_{dc}} \cdot (1 + \cos(2\omega t)) \quad \dots(16)$$

[0106] That is to say, the formulas (10), (12), (15) and (16) are used to meet the demand for the power balance, and the formula (11) is used to further meet the demand for making the input current I_{in} have a sinusoidal waveform and satisfaction of the formula (1) on which the above-mentioned formulas are based.

[0107] In a case where the diode bridge is used as the converter 3, the converter 3 cannot actively be switched at the rectifying duty d_{rec} expressed by the formula (15). By switching the inverter 5 and the switch S_c respectively in accordance with the zero duty d_z and the discharge duty d_c determined by the formulas (14), (15) and (16), the current i_{rec1} expressed by the formula (10) can be obtained.

[0108] The inverter 5 cannot utilize the DC voltage at the DC link 7 in a period in which the zero-phase current i_z flows. A virtual DC voltage (hereinafter, referred to as a “virtual DC voltage”) V_{dc} in a period in which the inverter 5 can perform power conversion out of the DC voltage across the DC power supply lines LH and LL can be considered as follows.

[0109]

$$V_{dc} = V_{rec} \cdot d_{rec} + V_c \cdot d_c + 0 \cdot d_z \quad \dots(17)$$

[0110] This can be understood as a voltage applied to the DC link 7 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 Sc and Sl and the inverter 5 is controlled. In FIG. 4, the virtual DC voltage Vdc is understood as a voltage generated across a current source Idc (which flows the DC current Idc) representing the inverter 5 and a load thereof.

[0111] A ratio R (= Vdc/Vm) of the virtual DC voltage Vdc to the crest value Vm is hereinafter referred to as a voltage utilization ratio. It is understood from the formulas (13), (14) and (16) that, by reducing the zero duty dz, a period in which the inverter 5 flows the zero-phase current can be reduced, and a period in which a voltage applied to a DC link 7 is utilized can be thereby increased. This leads to an increase in voltage utilization ratio R.

[0112] Incidentally, the input power Pdc input from the DC link 7 into the inverter 5 is the product of the virtual DC voltage Vdc and the DC current Idc. Since the inverter 5 obtains, from the DC link 7, the input power Pdc (= Pin + Pc - Pl) obtained by subtracting the charge power Pl from the sum of the pulsating power Pin and the discharge power Pc, the following formula (18) holds.

[0113]

$$\begin{aligned} I_{dc} &= \frac{P_{in} + P_c - P_l}{V_{dc}} \\ &= \frac{V_m \cdot I_m}{2} (1 - (1 - 2k) \cos(2\omega t)) \cdot \frac{1}{V_{dc}} \quad \dots(18) \end{aligned}$$

[0114] Note that neither the virtual DC voltage Vdc nor the DC current Idc is limited in the settings having described so far. This means that, once the buffer distribution ratio k is determined, the duties drec, dc, and dz can be set regardless of settings of the virtual DC voltage Vdc and the DC current Idc as long as the virtual DC voltage Vdc and the DC current Idc satisfy the formula (18).

[0115] Requirements for an increase in voltage utilization ratio R as a new effect is now described. As described above, the voltage utilization ratio R can be increased by reducing the zero duty d_z , and a lower limit of the zero duty d_z is zero. The duties d_{rec} and d_c when the virtual DC voltage V_{dc} is the highest per buffer distribution ratio k can be obtained when the zero duty d_z is zero.

[0116] If the zero duty d_z is zero, the formula (19) is obtained from the formulas (14), (15) and (16).

[0117]

$$\frac{I_m}{I_{dc}} \left\{ (1-k) \cdot |\sin(\omega t)| + \frac{k \cdot V_m}{2V_c} (1 + \cos(2\omega t)) \right\} = 1 \quad \dots (19)$$

[0118] The both-end voltage V_c can herein be considered to be approximately constant (vary by approximately 5%, for example, as described above) even if charge and discharge are performed by the power buffer circuit 4. The DC current I_{dc} when the virtual DC voltage V_{dc} is the highest per buffer distribution ratio k is determined from the formula (19). As a result, the duties d_{rec} and d_c when the virtual DC voltage V_{dc} is the highest per buffer distribution ratio k are determined from the formulas (15) and (16). The virtual DC voltage V_{dc} in this case is determined by the formulas (18) and (19).

[0119] The DC current I_{dc} when the virtual DC voltage V_{dc} is the highest as described above takes a minimum value in view of the formula (18). This is desirable to reduce power ratings of the switching elements S_{up} , S_{vp} , S_{wp} , S_{un} , S_{vn} and S_{wn} and the diodes D_{up} , D_{vp} , D_{wp} , D_{un} , D_{vn} and D_{wn} used in the inverter 5.

[0120] FIGs. 5 and 6 are graphs showing operation of the direct power converter shown in FIG. 1, and show the operation in a case where the duties d_{rec} , d_c , and $d_z (= 0)$ are set to satisfy the formulas (14) to (19). Note that $k = 1/6$ in FIG. 5, and $k = 1/2$ is set in FIG. 6.

[0121] In each of FIGs. 5 and 6, the duties d_{rec} , d_c and d_z are shown in the first tier, the virtual DC voltage V_{dc} , a first component $d_{rec} \cdot V_{rec}$ and a second component $d_c \cdot V_c$ of the virtual DC voltage V_{dc} , and the DC current I_{dc} are shown in the second tier, the current i_{rec} (being equal to the absolute value of the input current I_{in}) output from the
 5 converter 3 and the currents i_{rec1} , i_l and i_c are shown in the third tier, and the powers P_{in} , P_c , P_{buf} and P_{rec} are shown in the fourth tier. In each of the graphs, the horizontal axis represents the phase ωt with “degree” as the unit.

[0122] Note that the first component $d_{rec} \cdot V_{rec}$ of the virtual DC voltage V_{dc} is a voltage appearing in the first term of the formula (17), and indicates contribution of the
 10 converter 3 to the virtual DC voltage V_{dc} . The second component $d_c \cdot V_c$ of the virtual DC voltage V_{dc} is a voltage appearing in the second term of the formula (17), and indicates contribution of the capacitor C_4 to the virtual DC voltage V_{dc} . Calculation is performed on the assumption that $V_m = 1$, $I_m = \sqrt{2}$, and $V_c = 1.5V_m = 1.5$ for convenience sake.

15 [0123] Since $k = 1/2$ in FIG. 6, pulsations (in the second term of the rightmost-hand side of the formula (1)) of the pulsating power are all cancelled by the power buffer circuit 4. The input power P_{dc} input into the inverter 5 is thus maintained constant. In this case, $i_{rec1} = i_l$ holds from the formulas (10) and (11).

[0124] FIG. 7 is a graph showing the virtual DC voltage V_{dc} in cases where $k = 1/2$,
 20 $1/4$ or $1/6$, and $d_z = 0$. This means that the virtual DC voltage V_{dc} determined by the formulas (18) and (19) is shown.

[0125] As can be understood from the formula (15), $d_{rec} = 0$ holds when $\omega t = n\pi$, and thus $d_c = 1$ holds by setting $d_z = 0$. Thus, $V_{dc} = V_c (= 1.5)$ holds when the phase ωt is 0 degrees, 180 degrees, and 360 degrees in FIG. 7. As can be understood from the

formula (16), $dc = 0$ holds when $\omega t = (2n-1)\pi/2$, and thus $drec = 1$ holds by setting $dz = 0$.

Thus, $V_{dc} = V_m (= 1)$ holds when the phase ωt is 90 degrees and 270 degrees in FIG. 7.

[0126] As can be understood from these graphs, the voltage utilization ratio R higher than the voltage utilization ratio $(1/\sqrt{2})$ introduced in Non-patent Document 2 and Patent Document 1 can be obtained.

[0127] In a case where dz is not set to zero, the virtual DC voltage V_{dc} can be set to a constant value. A specific example thereof and an upper limit of the virtual DC voltage V_{dc} as constant are described below.

[0128] As can easily be understood from FIG. 7, the virtual DC voltage V_{dc} increases with increasing buffer distribution ratio k , and thus a case where $k = 1/2$ is considered. In this case, the virtual DC voltage V_{dc} approximately takes the crest value V_m or more in a whole section of the phase ωt , but takes the crest value V_m or less when the phase ωt is in a range of 30 degrees to 150 degrees and in a range of 210 degrees to 330 degrees. Even when $k = 1/2$ and $dz = 0$, the virtual DC voltage V_{dc} cannot be set to a constant value equal to or more than the crest value V_m .

[0129] FIG. 8 is a graph showing operation of the direct power converter in a case where $dz < 0$ is virtually allowed, and $V_{dc} = V_m$ holds when $k = 1/2$ while satisfying the formulas (15) and (16). As in FIGs. 5 to 7, calculation is herein performed on the assumption that $V_m = 1$, $I_m = \sqrt{2}$, and $V_c = 1.5V_m = 1.5$. With reference to FIG. 7, it is understood that $dz < 0$ holds when the phase ωt is in the range of 30 degrees to 150 degrees and in the range of 210 degrees to 330 degrees.

[0130] A maximum value of the virtual DC voltage V_{dc} that is constant is thus obtained by setting $dz = 0$ as follows. Since the virtual DC voltage V_{dc} is the highest when $k = 1/2$ as described above, a condition to satisfy $drec + dc = 1$ in this case is

considered. The formula (20) holds by substituting $k = 1/2$ in the formula (19). Note that $x = |\sin(\omega t)|$ is introduced.

[0131]

$$d_{rec} + d_c = \frac{1}{2} \cdot \left[x + \frac{V_m}{V_c} (1 - x^2) \right] = 1 \quad \dots(20)$$

- 5 [0132] $P_{dc} = V_m \cdot I_m / 2$ holds as $P_{dc} = V_{dc} \cdot I_{dc}$ and $k = 1/2$, and thus the formula (21) is obtained.

[0133]

$$V_{dc} = \frac{V_m}{2} \cdot \frac{I_m}{I_{dc}} = \frac{V_m}{x + \frac{V_m}{V_c} (1 - x^2)} \quad \dots(21)$$

- [0134] The phase ωt providing the maximum value of the virtual DC voltage V_{dc} provides a minimum value of a denominator of the right-hand side of the formula (21). Therefore, the phase ωt when a value of the formula (22) indicating a derivative of the denominator is zero should be obtained.

[0135]

$$\begin{aligned} \frac{d}{d(\omega t)} \left[\left| \sin(\omega t) \right| + \frac{V_m}{V_c} (1 - \sin^2(\omega t)) \right] \\ = \cos(\omega t) \cdot \left[1 - \frac{2V_m}{V_c} \sin(\omega t) \right] \quad \dots(22) \end{aligned}$$

- 15 [0136] In this case, $x = V_c / (2 \cdot V_m)$. In the above-mentioned example, it is understood that the maximum value of the virtual DC voltage V_{dc} that is constant is approximately $0.96V_m$ as $V_c = 1.5V_m$.

- [0137] FIG. 9 is a graph showing operation of the direct power converter in a case where $V_c = 1.5V_m = 1.5$ and $V_{dc} = 0.96V_m = 0.96$ ($V_m = 1$). It is understood that $dz \geq 0$ holds in a whole section of the phase ωt .

[0138] When the buffer distribution ratio k is fixed to $1/2$ as described above, by fixing the duties in accordance with the formulas (14), (15) and (16), the virtual DC voltage V_{dc} can be set to be constant. It is understood that, with the technique, the voltage utilization ratio R can be set to $\alpha/(1+\alpha \cdot \alpha/4)$ (where $\alpha = V_c/V_m$) by setting x to $V_c/(2 \cdot V_m)$ in the
 5 formula (21) by reducing the zero duty d_z .

[0139] A condition that a ratio α should satisfy to let the voltage utilization ratio $R = \alpha/(1+\alpha \cdot \alpha/4)$ be larger than the voltage utilization ratio $(1/\sqrt{2})$ introduced in Non-patent Document 2 and Patent Document 1 is $2\sqrt{2} - 2 < \alpha < 2\sqrt{2} + 2$. The both-end voltage V_c is charged to be higher than the crest value V_m , and the ratio α is larger than 1. To set
 10 the virtual DC voltage V_{dc} to be constant, $\alpha \leq 2$ holds as $x = |\sin(\omega t)| = \alpha/2$. The condition is thus satisfied, and the voltage utilization ratio R larger than the voltage utilization ratio $(1/\sqrt{2})$ introduced in Non-patent Document 2 and Patent Document 1 can be obtained.

[0140] It is understood that the voltage utilization ratio R takes a maximum value of 1
 15 when a value obtained by differentiating the voltage utilization ratio R with respect to the ratio α is zero, i.e., $\alpha = 2$. That is, the maximum value of 1 of the voltage utilization ratio R can be obtained by setting the both-end voltage V_c to be twice the crest value V_m .

[0141] (b-6) Setting of Duties Independent of Constant k

In this section, a technique for setting the duties d_{rec} , d_c , and d_z independently
 20 of the buffer distribution ratio k is described. In this section, the duties d_{rec} and d_c are respectively set by the formulas (23) and (24). The zero duty d_z is determined by the formulas (23), (24) and (14).

[0142]

$$\begin{aligned} d_{rec} &= \frac{V_{dc}}{V_m} |\sin(\omega t)| \\ &= \frac{V_{dc}}{V_{rec}} \cdot \sin^2(\omega t) \quad \dots (23) \end{aligned}$$

[0143]

$$\begin{aligned} d_c &= \frac{V_{dc}}{V_c} \cdot \cos^2(\omega t) \\ &= \frac{V_{dc}}{2V_c} (1 + \cos(2\omega t)) \quad \dots (24) \end{aligned}$$

[0144] The rectifying duty d_{rec} can be expressed as a value obtained by dividing the product of the square of the sine value of the phase ωt and the virtual DC voltage V_{dc} by the rectified voltage V_{rec} , for example. The discharge duty d_c can be expressed as a value obtained by dividing the product of the square of a cosine value of the phase ωt and the virtual DC voltage V_{dc} by the both-end voltage V_c , for example.

[0145] It is clear from the formula (25) that the rectifying duty d_{rec} and the discharge duty d_c thus set satisfy the formula (17).

[0146]

$$\begin{aligned} d_{rec} \cdot V_{rec} + d_c \cdot V_c &= V_{dc} |\sin(\omega t)| \cdot |\sin(\omega t)| + V_{dc} \cdot \cos^2(\omega t) \\ &= V_{dc} \quad \dots (25) \end{aligned}$$

[0147] The virtual DC voltage V_{dc} can thus arbitrarily be set, e.g., to a constant value independent of the phase ωt , while the rectifying duty d_{rec} and the discharge duty d_c are set to values independent of the buffer distribution ratio k .

[0148] In this setting, the DC current I_{dc} is expressed by the formula (18) as the relation $P_{dc} = V_{dc} \cdot I_{dc} = P_{in} + P_c - P_l$ holds.

[0149] It is assumed that the input current I_{in} is expressed by $I_m \cdot \sin(\omega t)$, i.e., the input current I_{in} has a sinusoidal waveform, and thus the current i_l satisfies the following formula (26) dependently on the DC current I_{dc} .

[0150]

$$\begin{aligned} i_l &= I_m \cdot |\sin(\omega t)| - d_{rec} \cdot I_{dc} \\ &= (I_m - \frac{V_{dc}}{V_m} \cdot I_{dc}) \cdot |\sin(\omega t)| \quad \dots (26) \end{aligned}$$

[0151] The above-mentioned description shows that the rectifying duty d_{rec} and the discharge duty d_c are set to the values independent of the buffer distribution ratio k , and the virtual DC voltage V_{dc} can be set to be constant. However, the virtual DC voltage V_{dc} that is constant has an upper limit. As described in the previous section, the formula (27) holds from the formulas (23) and (24) by considering $d_{rec} + d_c = 1$ and introducing $x = |\sin(\omega t)|$.

[0152]

$$d_{rec} + d_c = \frac{V_{dc}}{V_m} \cdot x + \frac{V_{dc}}{V_c} \cdot (1 - x^2) = 1 \quad \dots (27)$$

[0153] The formula (28) is thus obtained, and is the same as the formula (21). In the above-mentioned example, it is understood that the maximum value of the virtual DC voltage V_{dc} that is constant is approximately $0.96V_m$ as $V_c = 1.5V_m$.

[0154]

$$V_{dc} = \frac{1}{\frac{x}{V_m} + \frac{1-x^2}{V_c}} = \frac{V_m}{x + \frac{V_m}{V_c}(1-x^2)} \quad \dots (28)$$

[0155] A graph showing operation of the direct power converter obtained when $k = 1/2$ thus matches the graph of FIG. 9 described in the previous section. For example, the formula (26) matches the formula (11) with $k = 1/2$ in the formula (18).

[0156] In this section, however, the duties d_{rec} , d_c , and d_z are set independently of the buffer distribution ratio k , and thus the buffer distribution ratio k is reduced, and the virtual DC voltage V_{dc} can be controlled to be constant while the formula (28) is satisfied.

[0157] FIG. 10 is a graph showing operation of the direct power converter in a case where $V_c = 1.5V_m = 1.5$ and $V_{dc} = 0.96V_m = 0.96$ ($V_m = 1$). It is understood that $dz \geq 0$ holds in a whole section of the phase ωt . The formulas (10) and (11) do not hold in contrast to the technique described in the previous section as k is herein $1/6$, which is different from $1/2$. Specifically, the current i_{rec1} is expressed by the formula (29) with reference to the formulas (13), (18) and (23).

[0158]

$$\begin{aligned} i_{rec1} &= d_{rec} \cdot I_{dc} = \frac{V_{dc}}{V_m} |\sin(\omega t)| \cdot \frac{V_m \cdot I_m}{2V_{dc}} (1 - (1 - 2k) \cos(2\omega t)) \\ &= \frac{I_m}{2} |\sin(\omega t)| \cdot (1 - (1 - 2k) \cos(2\omega t)) \quad \dots(29) \end{aligned}$$

[0159] (b-7) Setting of DC Current I_{dc}

10 Not only the duties d_{rec} , d_c , and d_z but also the DC current I_{dc} is required to be set based on the formula (18) in a case where any of the techniques described in the above-mentioned sections (b-5) and (b-6) is used. An example to achieve the formula (18) used for the DC current I_{dc} is described in this section.

[0160] As for operation of a normal AC load, a case where well-known dq-axis control is performed is taken as an example. A power formula on the dq axes is typically expressed by the formula (30). Symbols V^* and I respectively represent a command value of a voltage applied to the AC load and a current flowing through the AC load. They are both alternating, and thus a dot indicating that they are each expressed as a complex number is attached to each of the symbols V^* and I . Note that a q-axis voltage ideally follows a command value V_q^* thereof, and a d-axis voltage ideally follows a command value V_d^* thereof.

[0161]

$$P + jQ = \dot{V}^* \cdot \dot{I} = V_d^* \cdot I_d + V_q^* \cdot I_q + j(V_q^* \cdot I_d - V_d^* \cdot I_q) \quad \dots(30)$$

[0162] A power supplied from the DC power supply lines LH and LL to the inverter 5 does not have a reactive power, and thus the power is expressed by the formula (31) by ignoring the third term of the formula (30).

[0163]

$$5 \quad P_{dc} = V_d \cdot I_d + V_q \cdot I_q \quad \dots(31)$$

[0164] On the other hand, the formula (32) holds from the formula (18) as $P_{dc} = V_{dc} \cdot I_{dc}$.

[0165]

$$\begin{aligned} P_{dc} &= \frac{V_m \cdot I_m}{2} (1 - (1 - 2k) \cos(2\omega t)) \\ &= \frac{V_m \cdot I_m}{2} - \frac{(1 - 2k) \cdot V_m \cdot I_m}{2} \cos(2\omega t) \quad \dots(32) \end{aligned}$$

10 [0166] Control to achieve the formula (18) can thus be performed by performing control so that an AC component of the formula (31) and the second term of the rightmost-hand side of the formula (32) match each other. FIG. 11 is a block diagram showing an example of configuration to perform the above-mentioned control. This configuration is provided, for example, in configuration shown as a controller 10 in FIG. 1.

[0167] A part of the configuration of FIG. 11 showing a well-known technique is briefly described. A current phase command value β^* is input to obtain trigonometric function values $\cos\beta^*$ and $-\sin\beta^*$, and, from these and a current command value I_a^* , a q-axis current command value I_q^* and a d-axis current command value I_d^* are generated.

20 On the assumption that the inductive load 6 is a rotary machine, the q-axis voltage command value V_q^* and the d-axis voltage command value V_d^* are obtained based on a rotation angular velocity ω_m of the rotary machine, a field magnetic flux Φ_a of the rotary machine, d-axis inductance L_d and q-axis inductance L_q of the rotary machine, the q-axis

current command value I_q^* and the d-axis current command value I_d^* , and a q-axis current I_q and a d-axis current I_d . Voltage command values V_u^* , V_v^* , and V_w^* for controlling the inverter 5 are generated from the q-axis voltage command value V_q^* and the d-axis voltage command value V_d^* .

5 [0168] In the configuration shown in FIG. 1, for example, a velocity detector 9 detects the AC currents I_u , I_v and I_w flowing through the inductive load 6, and, based on these, the rotation angular velocity ω_m , the q-axis current I_q and the d-axis current I_d are provided to the controller 10.

[0169] The controller 10 obtains, based on the voltage command values V_u^* , V_v^* , and
10 V_w^* , signals SS_{up} , SS_{vp} , SS_{wp} , SS_{un} , SS_{vn} and SS_{wn} (see FIG. 1) for respectively controlling operation of the switching elements Sup , Svp , Swp , Sun , Svn and Swn of the inverter 5 through arithmetic processing, which is not shown (see, for example, Patent Document 1).

[0170] The controller 10 also generates signals SS_c and SS_l for respectively controlling
15 operation of the switches Sc and Sl , which are generated based on the duties d_{rec} , d_c , d_z , and d_l (see, for example, Patent Document 1).

[0171] A processor 71 for performing control so that the AC component of the formula (31) and the second term of the rightmost-hand side of the formula (32) match each other is described below. The processor 71 includes a DC power calculator 711, a pulsating
20 component extractor 712, a pulsating component calculator 713, a subtracter 714, an adder 715 and a PI processor 716.

[0172] The DC power calculator 711 receives the q-axis voltage command value V_q^* and the d-axis voltage command value V_d^* and the q-axis current I_q and the d-axis current I_d , calculates the input power P_{dc} based on the above-mentioned formula (31),
25 and provides the input power P_{dc} to the pulsating component extractor 712.

[0173] The pulsating component extractor 712 extracts and outputs the AC component of the formula (31). The pulsating component extractor 712 is achieved by a high pass filter HPF, for example.

[0174] The pulsating component calculator 713 receives the crest values V_m and I_m ,
 5 the power supply angular velocity ω , and the buffer distribution ratio k , and obtains the second term of the rightmost-hand side of the formula (32). The crest values V_m and I_m and the power supply angular velocity ω can be input into the pulsating component calculator 713 as information obtained from the single-phase AC power supply 1 (see FIG. 1).

10 [0175] As described above, the desired processing is causing the AC component of the formula (31) and the second term of the rightmost-hand side of the formula (32) to match each other, and thus control should be performed to reduce a difference between output of the pulsating component extractor 712 and output of the pulsating component calculator 713. The subtracter 714 thus obtains the difference, and a value obtained by performing
 15 integral proportional control on the difference by the PI processor 716 is output to the adder 715.

[0176] The adder 715 performs processing to correct the current command value I_a^* in normal processing with output of the PI processor 716. Specifically, first, a subtracter 701 obtains the deviation of the rotation angular velocity ω_m from a command value ω_m^*
 20 thereof in normal processing to obtain the current command value I_a^* . A PI processor 702 performs integral proportional control on the deviation to once obtain the current command value I_a^* . The adder 715 then performs processing to increase the current command value I_a^* with the output of the PI processor 716.

[0177] The above-mentioned well-known technique is applied to the current command
 25 value I_a^* corrected by the processor 71 as described above to generate the q-axis voltage

command value V_q^* and the d-axis voltage command value V_d^* . Such control is control to provide feedback regarding the q-axis voltage command value V_q^* and the d-axis voltage command value V_d^* , and the q-axis current I_q and the d-axis current I_d , and to cause the difference output from the subtracter 714 to be closer to zero. This means that the AC component of the formula (31) and the second term of the rightmost-hand side of the formula (32) can be caused to match each other by performing such control.

[0178] C. Modifications

(c-1) Selection of Buffer Distribution Ratio k

The buffer distribution ratio k can be set regardless of the magnitude of the load. Such setting is desirable to reduce a ripple current of the capacitor C_4 in a light-load range, and to increase the life of the power converter.

[0179] On the other hand, the buffer distribution ratio k may be set to decrease with increasing magnitude of the load. In the configuration shown in FIG. 11, for example, the buffer distribution ratio k may be set to be inversely proportional to the current command value I_a^* obtained by the adder 715, or may be set to be proportional to an inverse for the sum of the square of the q-axis current I_q and the square of the d-axis current I_d . Such setting of the buffer distribution ratio k can be achieved using a well-known technique.

[0180] In this case, the direct power converter performs operation to increase the buffer distribution ratio k in inverse proportion to a current flowing through the inductive load 6. Such operation is desired in the following situation, for example. That is to say, as the load of the power converter, there is a case where one having a large inertia, such as a motor for driving a compressor, is used. The large inertia suppresses vibrations of the motor, and eventually suppresses vibrations of the compressor caused by torque

variations. However, the effect of the inertia to deal with the torque variations decreases in a low-speed range. The buffer distribution ratio k can be increased to thereby suppress torque variations of the motor, which is a load, in the low-speed range of the motor, i.e., a range in which a small current flows through the inverter.

5 [0181] Alternatively, the buffer distribution ratio k may be set to $1/2$ when the load is equal to or smaller than a predetermined threshold. The buffer distribution ratio k of $1/2$ means that the buffering power P_{buf} (corresponding to the power $P_c - P_l$ as the variation: see (b-1)) is equal to the AC component $\hat{P}_{in}$ of the pulsating power. The predetermined value may be set to zero. In this case, the power $P_c - P_l$ is zero if the pulsating power
10 $\hat{P}_{in}$ is zero. In this case, the power $P_c - P_l$ is controlled to be smaller than the pulsating power $\hat{P}_{in}$ if the pulsating power $\hat{P}_{in}$ is positive.

[0182] By setting the threshold to a value corresponding to the predetermined value smaller than a maximum rated power $\hat{P}_{in(max)}$ of the pulsating power $\hat{P}_{in}$, buffering of a power can be performed when the pulsating power $\hat{P}_{in}$ takes the maximum rated power
15 $\hat{P}_{in(max)}$.

[0183] When the load exceeds the threshold (the pulsating power $\hat{P}_{in}$ exceeds the predetermined value corresponding to the threshold), for example, the buffering power P_{buf} may be set to a constant value independently of the pulsating power $\hat{P}_{in}$.

[0184] (c-2) Modification of Circuit Configuration

20 In a case where any of the above-mentioned techniques is used, the filter 2 can also be provided between the converter 3 and the power buffer circuit 4.

[0185] FIG. 12 is a circuit diagram in a case where the filter 2 is provided between the converter 3 and the power buffer circuit 4 as the modification, showing only the vicinity thereof.

[0186] When such configuration is used, it is desirable to provide a diode D0 in the DC power supply line LH between the filter 2 and the discharge circuit 4a. An anode of the diode D0 is located on the filter 2 side, and a cathode of the diode D0 is located on the discharge circuit 4a side.

- 5 [0187] The diode D0 can prevent a both-end voltage across the capacitor C2 from being affected by the both-end voltage Vc across the capacitor C4 by switching of the switch Sc.

[0188] While the present invention has been described in detail, the foregoing description is in all aspects illustrative and not restrictive. It is therefore understood that
10 numerous modifications that have not been described can be devised without departing from the scope of the present invention.

CLAIMS

1. A power converter comprising:

a DC link that includes a first power supply line and a second power supply line;

5 a converter that receives a single-phase AC voltage, and outputs a pulsating power to said DC link;

an inverter that receives a power from said DC link, and outputs an AC current;

and

a power buffer circuit that receives a charge power from said DC link, and

10 outputs a discharge power to said DC link, wherein

when at least said pulsating power exceeds a predetermined value, a variation of said discharge power to said charge power is smaller than an AC component of said pulsating power.

15 2. The power converter according to claim 1, wherein

an input power input from said DC link into said inverter takes a value obtained by subtracting said charge power from a sum of said pulsating power and said discharge power, and

20 when at least said pulsating power exceeds said predetermined value, said charge power takes a value which is a constant times said pulsating power, and said discharge power takes a value obtained by adding said charge power to (-2) times said AC component of said pulsating power multiplied by said constant,

wherein said constant is a positive number less than $1/2$.

25 3. The power converter according to claim 2, wherein

said converter

applies a rectified voltage obtained by full-wave rectifying said single-phase AC voltage to said DC link so that a higher electric potential is applied to said first power supply line than said second power supply line, and

5 allows, when at least said pulsating power exceeds said predetermined value, a current obtained by multiplying a first current by a value obtained by subtracting said constant from 1 to flow through said DC link, said first current being a current obtained by dividing said pulsating power by said rectified voltage,

said power buffer circuit includes:

10 a discharge circuit that includes a capacitor and a first switch that is connected in series to said capacitor between said first power supply line and said second power supply line so as to be closer to said first power supply line than said capacitor is; and

a charge circuit that charges said capacitor, and

15 when at least said pulsating power exceeds said predetermined value, said power buffer circuit receives a charge current which is said constant times said first current, and outputs a discharge current that is a current obtained by dividing said discharge power by a both-end voltage across said capacitor.

20 4. The power converter according to claim 2, wherein

said converter applies a rectified voltage obtained by full-wave rectifying said single-phase AC voltage to said DC link so that a higher electric

potential is applied to said first power supply line than said second power supply line,

said power buffer circuit includes:

a discharge circuit that includes a capacitor and a first switch that is connected in series to said capacitor between said first power supply line and said second

5 power supply line so as to be closer to said first power supply line than said capacitor is;
and

a charge circuit that charges said capacitor,

a rectifying duty that is a duty at which said converter is in conduction with said DC link takes a value obtained by dividing a product of a predetermined voltage and a
10 square of a sine value of a phase of said single-phase AC voltage by said rectified voltage,
and

a discharge duty that is a duty at which said capacitor discharges takes a value obtained by dividing a product of said predetermined voltage and a square of a cosine value of said phase by a both-end voltage across said capacitor.

15

5. The power converter according to any one of claims 2 to 4, wherein
said constant is independent of said AC current.

20

6. The power converter according to any one of claims 2 to 4, wherein
said constant is inversely proportional to a magnitude of said AC current.

7. The power converter according to any one of claims 1 to 6, wherein

said predetermined value is set so as to be smaller than a maximum rated power of said pulsating power.

5 8. The power converter according to any one of claims 1 to 7, wherein
when said pulsating power is equal to or smaller than the predetermined value,
said variation of said discharge power to said charge power is equal to said AC
component of said pulsating power.

10 9. The power converter according to any one of claims 1 to 8, wherein
when said pulsating power exceeds the predetermined value, said variation ($P_c - P_l$) of said discharge power to said charge power is independent of said pulsating power.