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PATENT REQUEST : STANDARD PATENT

I/We being the person(s) identified below as the Applicant(s), request the grant of a patent to the person(s) identified below as the Nominated Person(s), for an invention described in the accompanying standard complete specification.

Full application details follow:

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[54] Invention Title:

AC motor control apparatus and control apparatus of electric rolling stock using the same

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AC MOTOR CONTROL APPARATUS AND CONTROL APPARATUS OF ELECTRIC ROLLING STOCK USING THE SAME

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(56) Prior Art Documents
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(57) Claim

1. A control apparatus of an AC motor driven by a power converter for delivering alternating current of variable voltage and variable frequency, characterized in that said apparatus comprises vector control means for controlling a vector of primary current of said motor and slip frequency control means for controlling a slip frequency of said motor, and that these two control means are used in combination.

10. A control apparatus for an electric vehicle having a power converter for delivering alternating current of variable voltage and variable frequency, and an AC motor energized by output of the power converter for driving the electric vehicle, characterized by the provision of means for determining values for exciting and torque components of primary current of said AC motor on the basis of a command from a master controller, means for determining a slip frequency for the AC motor from the exciting and torque current component values, means for determining values for exciting and torque components of output voltage of said power converter on the basis of the exciting and torque primary current component values, means for correcting the exciting component value of output voltage in accordance with a difference between said determined exciting component value of primary current and a measured exciting current component, means for correcting the torque component value of output voltage

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in accordance with a difference between said determined torque component value of primary current and a measured torque current component, and means for correcting said slip frequency on the basis of a difference between a measured magnitude of primary current of said AC motor and a predetermined value thereof based on the exciting and torque primary current components.

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COMPLETE SPECIFICATION

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INVENTION TITLE:

AC motor control apparatus and control apparatus of electric rolling stock using the same

The following statement is a full description of this invention, including the best method of performing it known to me/us:-

The present invention relates to an AC motor control apparatus.

In recent years, employment of an induction motor, which is controllable in variable voltage and variable frequency fashion by an inverter, as a motor for driving an electric vehicle has been gaining a widespread use in the field of railway vehicles.

Incidentally, when controlling rotation speed of the induction motor standing for an AC motor, a slip frequency control type PWM inverter unit is used with the aim of improving the voltage utilization rate of power supply, as described in, for example, JP-A-62-163589.

A technique for vector control of an induction motor for the sake of improving responsibility of torque control of an electric rolling stock driving motor, though not involved in the field of railway vehicle, is described in JP-A-2-266884.

As described in the former prior art, in the railway electric rolling stock, parameters for performing torque control of the motor are output voltage value and slip frequency with respect to inverter frequency and fluctuation of magnetic flux of the motor is not taken into consideration. As a result, a fluctuation of magnetic flux leads to a torque fluctuation, giving rise to occurrence of slip and the like. Especially, in the case of railway electric rolling stock, running at the adhesive limit is the most efficient and therefore torque fluctuation must be suppressed to as small a value as possible.

Thus, vector control capable of controlling magnetic flux inside the motor and current independently of each other is introduced into an AC motor control apparatus of railway electric rolling stock.

When the AC motor is vector controlled, responsibility of torque control is raised as described in the latter prior art.

However, mere application of the vector control as it is to the electric rolling stock control apparatus faces problems as below.

In typical vector control, the motor frequency is used upon decomposition of current into vector components. However, the number of pulses generated by a speed detector (encoder) mounted to a body such as an electric rolling stock having large exciting force upon acceleration is small.

Especially in the low speed region, therefore, a delay in speed detection occurs and it links directly to torque fluctuation.

In the electric rolling stock which requires the resolution of a single pulse of the encoder to improve voltage utilization rate it is difficult to control the motor with the vector control method in such a single pulse region.

In addition to the electric rolling stock exemplified herein, there are similar problems in the other systems driving an AC motor by using vector control method.

In accordance with the present invention, there is provided a control apparatus of an AC motor driven by a power converter for delivering alternating current of variable voltage and variable frequency, characterised in that said apparatus comprises vector control means for controlling a vector of motor primary current and slip frequency control means for controlling a slip frequency of the motor, and that both the control means are used in combination.

In one form of the invention the slip frequency control system controls the slip frequency of the motor in accordance with a difference in the motor primary current. Since very small delay occurs in detection of the motor current, an error due to vector control can be absorbed.

The invention also provides a control apparatus for an electric vehicle having a power converter for delivering alternating current of variable voltage and variable frequency, and an AC motor energized by output of the power converter for driving the electric vehicle, characterized by the provision of means for determining values for exciting and torque components of primary current of said AC motor on the basis of a command from a master controller, means for determining a slip frequency for the AC motor from the exciting and torque current component values, means for determining values for exciting and torque components of output voltage of said power converter on the basis of the exciting and torque primary current component values, means for correcting the exciting component value of output voltage in accordance with a difference between said determined exciting component value of primary current and a measured exciting current component, means for correcting the torque component value of output voltage in accordance with a difference between said torque component value of primary current and a measured torque current component, and means for correcting said slip frequency on the basis of a difference between a measured magnitude of primary current of said AC motor and a predetermined value thereof based on the exciting and torque primary current components.

Since the slip frequency control system based on the difference in the primary current comes into power in the region in which voltage control is invalidated, the motor can be controlled even in the vector control disabled region. Prior to describing embodiments of the present invention, a brief description will be given of the embodiments ~~invention.~~ .

Principally considered as factors hindering smooth torque control of the motor are:

firstly, change of primary resistance of the AC motor and change of inductance due to magnetic saturation of the core,

secondly, change of inverter DC power supply voltage,

thirdly, change of output voltage due to PWM pulse mode change of the inverter,

fourthly, setting error of slip angular frequency command due to change of secondary resistance of the AC motor, and

fifthly, speed detection error of the AC motor. Embodiments of the present invention are able to take ~~The present invention takes~~ advantage of the fact that the magnitude, exciting current component and torque current component of the motor primary current are inherently different in sensitivity to the factors of disturbance of torque control, and so provide ~~has~~ current control systems capable of controlling the respective current components independently of each other to thereby suppress torque fluctuation.

More particularly, in view of the aforementioned first to third disturbance factors, current control systems, independent of each other, are provided for the exciting current component of the motor primary current and the torque current component thereof, respectively, and they are used to control individual voltage components of the rotating magnetic field coordinate system of the motor, thereby ensuring that torque fluctuation can be suppressed with respect to the disturbance.

Further, in view of the aforementioned fourth and fifth disturbance factors, a closed loop control system for the magnitude of primary current is set up which has the highest sensitivity to a region starting from the low speed region and ending in the high speed region (starting from the multipulse mode and ending in one pulse mode) and it is used to control the primary angular frequency (inverter frequency).

In other words, according to embodiments of the invention, slip frequency control means for controlling the slip frequency of the motor and means for controlling respectively magnetic flux of the motor and current orthogonal thereto are controlled independently of each other so as not to interfere with each other. For example, the primary current is decomposed into an exciting component and a torque component which are controlled independently of each other or vector control means for controlling the magnitude and phase of the primary current is used in combination, thereby permitting excellent torque control over the entire operation region.

Especially, since in an AC motor control PWM inverter for driving a railway electric rolling stock, the use of PWM pulse is extended to even the mode of one pulse, voltage is fixed to power supply voltage and voltage control is invalidated. Even in such an event, the first to third disturbance factors compensated for by voltage control can be detected as a disturbance in the motor primary current and the primary angular frequency can be controlled to prevent over current to thereby keep stable running continuing.

Even in an application other than the electric rolling stock, ~~the present invention has the function~~ it is possible to prevent torque fluctuation due to at least the aforementioned fourth disturbance.

More particularly, for example, the control region of a control apparatus of an AC motor used for a rolling mill or the like is a variable voltage and variable frequency region and even in a use in such a region alone, a setting error of slip angular frequency command due to a change of the secondary resistance of the AC motor takes place. By correcting the slip angular frequency error on the basis of magnitude of the primary current, response of the vector current control system can be improved and as a result the setting error can be eliminated.

Incidentally, the previously-described JP-A-2-266884 gives a description that a vector control system switches to a slip frequency control system only when current detectors become faulty but does not teach a use of both the control systems in combination.

The invention is described in greater detail hereinbelow, by way of example only, with reference to the accompanying drawings, wherein:

Fig. 1 is a block diagram showing an embodiment of an AC motor control apparatus according to the invention;

Fig. 2 is a graph showing examples of patterns stored in a control command generator 6;

Fig. 3 is a block diagram showing details of construction of a voltage command arithmetic unit 5;

Fig. 4 is a block diagram showing details of construction of a current controller 8;

Fig. 5 is a vector diagram for explaining voltage and current components of the rotating magnetic field coordinate system;

Fig. 6 is a vector diagram for explaining magnetic flux (ϕ_{2q}) which develops when magnetic flux axis (m) of the motor does not coincide with coordinate axis (d) of control;

Fig. 7 is a graph showing the change of d-axis current with respect to the change of motor constants;

Fig. 8 is a graph showing the change of q-axis current with respect to the change of motor constants;

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Fig. 9 is a graph showing primary current ΔI_1 and magnetic flux ϕ_{2q} with respect to the change of motor constant r_1 ;

Fig. 10 is a graph showing primary current ΔI_1 and magnetic flux ϕ_{2q} with respect to the change of slip;

Fig. 11 is a graph showing torque τ with respect to the change of motor constants;

Fig. 12 is a graph showing the slip angular frequency with respect to the primary angular frequency;

Fig. 13 is a graph showing the relation of carrier frequency with respect to the primary angular frequency;

Fig. 14 is a fragmentary schematic block diagram showing a second embodiment of the AC motor control apparatus according to the invention;

Fig. 15 is a fragmentary schematic block diagram showing a third embodiment of the AC motor control apparatus according to the invention;

Fig. 16 is a fragmentary schematic block diagram showing a fourth embodiment of the AC motor control apparatus according to the invention;

Fig. 17 is a block diagram showing a fifth embodiment of the AC motor control apparatus according to the invention;

Fig. 18 is a block diagram showing details of construction of a coordinate transformation in the Fig. 17 embodiment;

Fig. 19 is a block diagram showing a sixth embodiment of the AC motor control apparatus according to the invention;

Fig. 20 is a block diagram showing details of a switching unit;

Fig. 21 is a block diagram showing another embodiment of the current controller 8;

Fig. 22 is a graph showing primary current ΔI_1 and magnetic flux ϕ_{2q} with respect to the change of motor constant $L_s\sigma^*$; and

Fig. 23 is a block diagram showing another embodiment of the current controller 8.

~~DESCRIPTION OF THE PREFERRED EMBODIMENTS~~

An embodiment of the present invention will now be described with reference to Figs. 1 to 11.

In Fig. 1, direct current fed from a stringing via a pantograph 11 is smoothed by a filter circuit comprised of a filter reactor 12 and a filter capacitor 13 and supplied to a pulse width modulation (hereinafter referred to as PWM) inverter 1 serving as a power converter. The inverter 1 converts DC voltage into three-phase AC voltage which is fed to an induction motor 2 standing for an AC motor to drive the same. An electric rolling stock runs with the induction motor 2 used as a drive source.

A forward/backward command signal D* and a power running/brake command signal N* which are delivered out of a master controll 7 are inputted to a control command generator 6.

On the basis of a filter capacitor voltage signal V_{FC}, a primary angular frequency command signal ω_1*, the power running/brake command signal N* and the forward/backward command signal D*, the control command generator 6 generates an exciting current command signal I_d* and a torque current command signal I_q* for the induction motor 2 which in turn are delivered to a voltage command arithmetic unit 5, a current controller 8 and a slip angular frequency arithmetic unit 19.

On the basis of the exciting current command signal I_d*, torque current command signal I_q* and primary angular frequency command signal ω_1*, the voltage command arithmetic unit 5 calculates V_d* and V_q* which are commands of two voltage components of a rotating magnetic field coordinate system and which are fed to the induction motor 2 and delivers them to adders 17 and 18.

In the adders 17 and 18, the two voltage component commands V_d* and V_q* are added with two voltage component correction commands ΔV_d* and ΔV_q*, respectively, to obtain V_d** and V_q** which in turn are delivered to a coordinate transformation 4. On the basis of coordinate transformation reference command signals, the coordinate transformation 4 converts the V_d** and V_q** into output voltage command signals v_u*,

v_v* and v_w* of a stator coordinate system which in turn are delivered to a PWM signal arithmetic unit 3.

In the PWM signal arithmetic unit 3, the output voltage commands v_u*, v_v* and v_w* are compared with a carrier which is determined in accordance with the output of a pulse mode generator 10 to obtain on/off pulses which in turn are supplied to the PWM inverter 1.

A coordinate transformation 9 receives inverter output currents i_u, i_v and i_w detected by current detectors 15u, 15v and 15w adapted to detect output current of the PWM inverter 1 and on the basis of the coordinate transformation reference signals, converts them into current components I_d and I_q of the rotating magnetic field coordinate system which in turn are delivered to the current controller 8.

On the other hand, an induction motor speed ω_r detected by a speed detector 16 is added to a slip angular frequency command signal ω_s* standing for the output of the slip angular frequency arithmetic unit 19 by means of an adder 22 to provide a primary angular frequency command ω_0* which is delivered to an adder 23.

In the adder 23, this primary angular frequency command signal ω_0* is added with a correction angular frequency command signal $\Delta\omega_1$* standing for the output of the current controller 8 to produce the primary angular frequency command signal ω_1*.

The primary angular frequency command signal ω_1* is supplied to an integrator 20, the control command generator 6, voltage command arithmetic unit 5, current controller 8 and pulse mode generator 10.

The integrator 20 calculates a coordinate reference signal ω_1* from the primary angular frequency command signal ω_1* and delivers it to a sine and cosine generator 21.

The sine and cosine generator 21 generates coordinate transformation reference signals sin ω*t and cos ω*t, which are delivered to the previously described coordinate transformation 4 and 9.

Further, on the basis of the exciting current command signal I_d*, torque current command signal I_q*, exciting current I_d, torque current I_q and primary angular frequency command signal ω_1*, the current controller 8 calculates two correction voltage component commands ΔV_d* and ΔV_q* and primary angular frequency correction signal $\Delta\omega_1$* which are respectively delivered to the adders 17, 18 and 23 as described previously.

Responsive to the primary angular frequency command signal ω_1*, the pulse mode generator 10 generates a PWM pulse number command N_p* which is delivered to the PWM pulse arithmetic unit 3.

With the control construction of Fig. 1 described as above, DC voltage of the PWM inverter 1 can be utilized most efficiently.

More particularly, in the low speed running region in which voltage control of the PWM inverter 1 can be effected highly accurately, it is possible to make the best of a two-current feedback system of vector control system (system for correcting voltage commands V_d* and V_q* determined from current commands, in accordance with a difference between exciting component of motor primary current and its command value and a difference between torque component of motor primary current and its command) and a feedback control system (slip frequency control) for controlling the inverter output frequency (primary angular frequency) in accordance with a difference between the magnitude of primary current (scalar quantity) and its command value.

Further, in the region in which primary current ripple increases owing to, for example, a decrease in pulse number complying with an increase in primary angular frequency, the gain of the two-current feedback control system of vector control system is reduced.

In the region in which the pulse number measures one pulse, the two-current feedback control system of vector control system is invalidated completely. However, since the voltage commands V_d* and V_q* take values calculated from the current command values I_d* and I_q* (these I_d* and I_q* change with the primary angular frequency as will be described later), vector control continues to proceed. As a result, only the phase relation is commanded but in contrast to the simple slip frequency control system, the command can fulfill itself even in the absence of the primary current control system, thus relieving load on the current control system.

Accordingly, a torque control system which can afford to provide high accuracies preserved even in the extreme one pulse mode can be constructed.

Details of individual components will now be described.

Fig. 2 shows examples of patterns stored in the control command generator 6.

The exciting current command signal I_d* and torque current command signal I_q* are changed with respect to the filter capacitor voltage V_{FC} and primary angular frequency command ω_1* so as to obtain accelerating performance of the rolling stock.

More specifically, the control command generator 6 receives a power running/braking force command N* calculated by the master controller 7 on the basis of a preset accelerating/decelerating acceleration command α* and a weight value which depends on the body of vehicle and the number of riders.

Then, in a range within which voltage fed to the induction motor 2 is smaller than a maximum value of output voltage of inverter 1 which is determined by a filter capacitor voltage V_{FC}, the exciting current command signal is made to take a constant value so as to keep magnitude ϕ* of magnetic flux of the induction motor 2 at a predetermined value.

As voltage fed to the induction motor 2 reaches the maximum value of inverter output voltage determined by the filter capacitor voltage V_{FC} (one pulse mode), the exciting current command signal I_d^* is made to be in inverse proportion to the primary angular frequency command signal ω_1^*. Through this, the voltage command can be maintained at a maximum value.

A torque current command signal I_q^* of the induction motor 2 is obtained by multiplying a ratio between power running/braking force command N* and magnitude ϕ^* of magnetic flux by a forward/backward command signal D* (+1 for forward running and -1 for backward running).

In addition, with respect to the primary angular frequency command ω_1^*, the exciting current command signal I_d^* and torque current command signal I_q^* are so limited that the maximum rated output of the inverter 1 or induction motor 2 is not exceeded.

The thus determined exciting current command signal I_d^*, torque current command I_q^* and primary angular frequency command ω_1^* are inputted to the voltage command arithmetic unit 5 for calculating a voltage command fed to the induction motor 2. Detailed construction of the voltage command arithmetic unit 5 is shown in Fig. 3.

A coefficient multiplier 500 multiplies the exciting current command signal I_d^* by a primary resistance r_1 to provide a product which is delivered to an adder 501.

A coefficient multiplier 502 multiplies the torque current command signal I_d* by a leakage inductance $L_{s\sigma}$ to provide a product delivered to a multiplier 505.

A coefficient multiplier 504 multiplies the exciting current command signal I_d* by a primary inductance L_1 to provide a product delivered to a multiplier 505.

A coefficient multiplier 506 multiplies the torque current command signal I_q* by the primary resistance r_1 to provide a product delivered to an adder 507.

The multiplier 503 multiplies an output signal of the coefficient multiplier 502 by the primary angular frequency ω_1* to provide a product delivered to the adder 501 and from this value and the output of the coefficient multiplier 500, the adder 501 calculates a voltage component V_d* of the rotating magnetic field coordinate system.

The multiplier 505 multiplies the output of the coefficient multiplier 504 by the primary angular frequency ω_1* to provide a product delivered to the adder 507 and from this value and the output of the coefficient multiplier 506, the adder 507 calculates a voltage component V_q* of the rotating magnetic field coordinate system.

A voltage equation of the rotating magnetic field coordinate system under the stationary state of AC motor 2 is expressed by the following equations:

$$I_d = \frac{1}{r_1 + L_{s\sigma} \cdot S} \{V_d + \omega_1 \cdot L_{s\sigma} \cdot I_q\} \quad \ldots (1)$$

$$I_q = \frac{1}{r_1 + L_{s\sigma} \cdot S} \{V_q - \omega_1 \cdot L_1 \cdot I_d\} \quad \ldots (2)$$

where S represents Laplace operator.

On the other hand, since the voltage command arithmetic unit 5 is constructed as shown in Fig. 3, there result the following equations by substituting $V_d = V_d^*$ and $V_q = V_q^*$ into equations (1) and (2):

$$I_d = \frac{1}{r_1 + L_{s\sigma} \cdot S} \{r_1 \cdot I_d^*\} \quad \ldots (3)$$

$$I_q = \frac{1}{r_1 + L_{s\sigma} \cdot S} \{r_1 \cdot I_q^*\} \quad \ldots (4)$$

Equations (3) and (4) indicate that $I_d = I_d^*$ and $I_q = I_q^*$ stand under the stationary state. To this end, it is necessary that constants of the coefficient multipliers 500, 502, 504 and 506 coincide with constants of the induction motor 2. However, winding temperature change and core magnetic saturation in the motor make it difficult to achieve the coincidence. Thus, in the present embodiment, the current controller

8 is provided to achieve the coincidence with motor constants.

Detailed construction of the current controller 8 is shown in Fig. 4.

Motor primary currents i_u to i_w detected by the current detectors 15_u to 15_w are converted by the coordinate transformation 9 into two phases representative of exciting current I_d and torque current I_q. An operation equation is given by equation (5),

$$\begin{bmatrix} I_d \\ I_q \end{bmatrix} = \begin{bmatrix} \cos\omega_1{*}t & \sin\omega_1{*}t \\ -\sin\omega_1{*}t & \cos\omega_1{*}t \end{bmatrix} \begin{bmatrix} 2/3 & -1/3 & -1/3 \\ 0 & 1/\sqrt{3} & -1/\sqrt{3} \end{bmatrix} \begin{bmatrix} i_u \\ i_v \\ i_w \end{bmatrix} \quad \cdots (5)$$

where the primary angular frequency command $\omega_1{*}$ defines values of sine and cosine.

Exciting current command signal $I_d{*}$ is inputted to an adder 801 and an arithmetic unit 807. Torque current command signal $I_q{*}$ is inputted to an adder 804 and the arithmetic unit 807. Exciting current I_d is inputted to the adder 801 and an arithmetic unit 810. Torque current I_q is inputted the adder 804 and arithmetic unit 810.

The adder 801 calculates a difference between the exciting current command signal $I_q{*}$ and a detected exciting current I_d and delivers the difference to a controller 803 via a multiplier 802.

The adder 804 calculates a difference between the torque current command signal $I_q{}^*$ and a detected torque current I_q and delivers the difference to an controller 806 via a multiplier 805.

The arithmetic units 807 and 810 respectively carry out operations expressed by the following equations (6) and (7), and an adder 808 calculates a difference between command signal $I_1{}^*$ for the magnitude of output current of the inverter and a detection signal I_1 and delivers a result to a controller 809.

$$I_1{}^* = \sqrt{(I_d{}^*)^2 + (I_q{}^*)^2} \quad \cdots (6)$$

$$I_1 = \sqrt{(I_d)^2 + (I_q)^2} \quad \cdots (7)$$

The controllers 803, 806 and 809 are each constructed of a proportional integration (P-I) arithmetic unit and the controllers 803 and 806 respectively deliver voltage command correction signals $\Delta V_d{}^*$ and $\Delta V_q{}^*$ for correcting the output voltage commands $V_d{}^*$ and $V_q{}^*$.

$$\Delta V_d{}^* = \left(P_d + \frac{K_d}{S}\right)(I_d{}^* - I_d) \quad \cdots (8)$$

$$\Delta V_q{}^* = \left(P_q + \frac{K_q}{S}\right)(I_q{}^* - I_q) \quad \cdots (9).$$

The controller 809 delivers a frequency command correction signal $\Delta\omega^*$ for correcting the primary angular frequency command ω_1^*:

$$\Delta\omega_1^* = \left(P_1 + \frac{K_1}{S} \right) (I_1^* - I_1) \qquad \ldots (10).$$

Incidentally, in the region in which the pulse number of PWM pulses of the PWM inverter 1 measures one, voltage control based on the exciting current and torque current is not permitted. As a result, the current difference is accumulated in integrators included in the controllers 803 and 806 and the correction values ΔV_d^* and ΔV_q^* are saturated.

Then, by taking advantage of the fact that the pulse number of PWM pulses can be controlled in accordance with the primary angular frequency command ω_1^* (a factor of power supply voltage must be added when the pulse mode switching frequency changes with power supply voltage), the output of a function generator 800 is used to make zero the input to each of the controllers 803 and 806, thus stopping them from operating. When nullification of the current difference by the action of the controllers 803 and 806 is prevented, the current difference develops in the adder 808, so that the primary angular frequency command ω_1^* is controlled by the controller 809 such that magnitude I_1 of the output current of inverter 1 is so controlled as to coincide with a command value I_1^*.

Under this condition, the vector control system controls input voltage vector to the motor, without resort to the correction loop.

In the current controller of the present embodiment, the magnitude of primary current is determined from the square root of the sum of the square of exciting component of primary current and the square of torque component of primary current but the primary current may also be determined directly from values of the current detectors 15u to 15w.

The operation of the controllers 803, 806 and 809 will now be described.

Fig. 5 is a vector diagram showing voltage components V_d and V_q and current components I_d and I_q of the rotating magnetic field coordinate system in the case where suitable primary angular frequency ω_1 and terminal voltage are fed to the AC motor 2.

When the primary angular frequency ω_1 and voltage components V_d and V_q are controlled such that d-axis standing for the coordinate reference of the control system coincides with magnetic flux vector ϕ of the induction motor 2, torque of the induction motor 2 is generated in proportion to q-axis current I_q and magnetic flux ϕ_{2d} to permit high-response torque control.

This state corresponds to ideal vector control.

However, as running of the induction motor 2 proceeds, not only temperature rises to cause primary resistance r_1 and secondary resistance r_2 to change but also magnetic saturation in the core causes leakage inductance to change. Therefore proper commanding of the primary angular frequency ω_1 and voltage components V_d and V_q is difficult to achieve.

More specifically, when a large value of command r_1^* in the control system is given to the primary resistance r_1 of the AC motor 2, output voltage vector v_1 lags in phase relative to vector v of Fig. 5 as shown in Fig. 6 and increases in magnitude. Accordingly, because of v_1, magnetic flux vector ϕ of the AC motor 2 assumes d-axis component ϕ_{2d} and q-axis component ϕ_{2q}, resulting in non-coincidence of d-axis of the control system with magnetic flux vector ϕ (m-axis). Under this condition, torque is generated in proportion to the product of I_q and ϕ_{2d} and the product of I_d and ϕ_{2q}, with the result that d-axis interferes with q-axis to prevent high-response torque control.

Therefore, in the present embodiment, a current control system is provided which corrects the respective primary angular frequency ω_1 and voltage components V_d and V_q of the rotating magnetic field system on the basis of inverter output currents.

Fig. 7 shows change ΔI_d of d-axis current with respect to changes of motor constants r_1^* and $L_s\sigma^*$.

In the figure, solid line represents ΔI_d with respect to change of primary resistance r_1 and dotted line represents ΔI_d with respect to change of leakage inductance $L_s\sigma^*$, the ΔI_d having the magnitude which is normalized by the rated current. When ΔI_d increases, the controller 803 calculates a voltage command ΔV_d^* ($<$ 0) for decreasing d-axis voltage so that d-axis voltage command may be corrected to control ΔI_d such that it becomes zero.

Fig. 8 shows change ΔI_q of d-axis current with respect to changes of motor constants r_1^* and $L_s\sigma^*$.

In the figure, solid line represents ΔI_q with respect to change of primary resistance r_1^* and dotted line represents ΔI_q with respect to change of leakage inductance $L_s\sigma^*$, the ΔI_q having the magnitude which is normalized by the rated current. When ΔI_q increases, the controller 806 calculates a voltage command ΔV_q^* ($<$ 0) for decreasing q-axis voltage so that q-axis voltage command may be corrected to control ΔI_q such that it becomes zero.

In the manner described as above, the controllers 803 and 806 so operate as to make zero the change of current due to the changes of motor constants but as shown in the Fig. 6 vector diagram, supply of proper voltage to the motor is prevented by the changes of motor constants to cause non-coincidence of d-axis of the control system with magnetic flux (m-axis) of the motor, so that the controller 803 for d-axis and the controller 806 for q-axis are not allowed to operate independently of each other but they sometimes interfere with each other to cause current to oscillate.

Accordingly, in the present embodiment, the controller 809 is provided which corrects the primary angular frequency ω_1 such that d-axis of the control system coincides with magnetic flux (m-axis) of the motor.

Fig. 9 shows change ΔI_1 in magnitude of the primary current with respect to change of motor constant r_1* and q-axis magnetic flux ϕ_{2q} of the motor. In the figure, solid line represents change ΔI_1 of the primary current which is normalized in magnitude by the magnitude of the rated current and chained line represents q-axis magnetic flux ϕ_{2q} which is normalized in magnitude by the magnitude of the rated magnetic flux.

As value r_1* of the primary resistance set by the control system increases, ΔI_1 increases in positive sense and q-axis magnetic flux ϕ_{2q} decreases in negative sense.

Accordingly, when ΔI_1 increases in positive sense, the controller 809 calculates angular frequency command $\Delta\omega_1$* for decreasing the primary angular frequency so that the primary angular frequency command may be corrected to control ϕ_{2q} such that it becomes zero. With ϕ_{2q} rendered to be zero, ΔI_1 is also rendered to be zero.

Fig. 10 shows magnitude ΔI_1 of the primary current with respect to change of motor constant r_2*, i.e., change of slip angular frequency command ω_s* and q-axis magnetic flux ϕ_{2q} of the motor. In the figure, solid line represents change ΔI_1 of the primary current which is normalized in magnitude by the magnitude of the rated current and chained line represents q-axis magnetic flux ϕ_{2q} which is normalized in magnitude by the magnitude of the rated magnetic flux. As slip angular frequency command ω_s* in the control system increases, ΔI_1 increases in positive sense and q-axis magnetic flux ϕ_{2q} decreases in negative sense.

Accordingly, when ΔI_1 increases in positive sense, the controller 809 calculates angular frequency command $\Delta\omega_1$* for decreasing the primary angular frequency as in the case of r_1* shown in Fig. 9 so that the primary angular frequency command may be corrected to control ϕ_{2q} such that it becomes zero.

In this manner, the change of current with respect to the changes of motor constants and the non-coincidence of d-axis with magnetic flux (m-axis) of the motor can be prevented to permit d-axis current control and q-axis current control to be effected independently of each other.

In the running range in which the PWM pulse number measures one pulse, control of the magnitude of output voltage of the inverter is not permitted but even in this range, phase angle θ_v* of output voltage determined by voltage component commands V_d** and V_q** can act on the inverter output voltage effectively. As a result, the non-coincidence of d-axis of the control system with magnetic flux (m-axis)of the AC motor can be prevented.

Fig. 11 shows torque τ with respect to changes of motor constants. In the figure, solid line represents torque with respect to change of slip angular frequency ω_s*, chained line represents torque with respect to change of primary resistance r_1*, and dotted line represents torque with respect to change of leakage inductance $L_s\sigma$, the torque being normalized in magnitude by the magnitude of the rated torque. As motor constants ω_s*, r_1* and $L_s\sigma$* increase, torque τ increases in proportion thereto. At that time, q-axis magnetic flux ϕ_{2q} of the motor deviates from zero as shown in Figs. 9, 10 and 22 and the non-coincidence of d-axis of the control system with magnetic flux (m-axis) of the AC motor, thus preventing torque and magnetic flux from being controlled independently of each other. In the present invention,control is effected such that the q-axis magnetic flux ϕ_{2q} is rendered to be zero to prevent torque from becoming excessively large or excessively small.

As a result, the interference of magnetic flux of the motor with torque current I_q can be suppressed to permit high-response torque control.

In the present embodiment, the controller 809 is operated in accordance with change ΔI_1 in magnitude of the primary current but instead q-axis current ΔI_q may be used.

When the q-axis current ΔI_q is used, r_1* is multiplied by 1.5 and $L_s\sigma$* is multiplied by 0.5. In such a case, variation in q-axis current can be cancelled out by the setting error between the two but the deviation of d-axis is not cancelled. Consequently, q-axis magnetic field is not rendered to be zero and the problem of the occurrence of interference between d- and q-axes still remains.

The slip angular frequency arithmetic unit 19 will now be described.

The slip angular frequency arithmetic unit 19 determines a slip angular frequency ω_s* by receiving exciting current command I_d* and torque current command I_q* and executing the following operation:

$$\omega_s^* = \frac{r_2 \cdot I_q^*}{M \cdot I_d^*} \quad \cdots (11)$$

where r_2 is secondary resistance design value of the induction motor 2 and M is exciting inductance.

This accounts for the slip angular frequency ω_s* having a characteristic with respect to primary angular frequency ω_1* as shown in Fig. 12.

The coordinate transformation 4 will now be described.

Correction values ΔV_d* and ΔV_q* determined by the current controller 8 are respectively added to output signals V_d* and V_q* of the voltage command arithmetic unit 5 by means of the adders 17 and 18 to provide voltage component commands V_d** and V_q** of the rotating magnetic field coordinate system which are delivered to the coordinate transformation 4.

The coordinate transformation 4 performs an operation expressed by equation (12) to deliver three-phase AC output voltage commands v_u*, v_v* and v_w* of the stator coordinate system.

$$\begin{bmatrix} v_u^* \\ v_v^* \\ v_w^* \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ -1/2 & \sqrt{3}/2 \\ -1/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} \cos\omega_1^*t & -\sin\omega_1^*t \\ \sin\omega_1^*t & \cos\omega_1^*t \end{bmatrix} \begin{bmatrix} V_d^{**} \\ V_q^{**} \end{bmatrix} \quad \ldots (12)$$

Then, the PWM signal arithmetic unit 3 compares the output voltage commands V_u*, v_v* and v_w* with a carrier proportional to a pulse number command signal N_p* standing for the output of the pulse mode generator 10 to produce on/off signals, which are delivered to the PWM inverter 1.

Responsive to the primary angular frequency command signal ω_1*, the pulse mode generator 10 generates the pulse number command signal N_p*. When the power supply voltage fluctuates, the switching frequency changes in accordance with this fluctuation (not shown).

An example of the carrier signal frequency with respect to the primary angular frequency command signal ω_1* is shown in Fig. 13.

Asynchronous PWM proceeds to make the carrier frequency constant for the primary angular frequency being in a range of from 0 to ω_0*, the pulse number command signal N_p* is sequentially decreased to 15, 9,3 and 1 for the primary angular frequency being in excess of ω_0*, and eventually running is carried out under PWM control of one pulse.

In this example, the asynchronous region turns into the synchronous region at the pulse number which is 15 but this may be done at the pulse number which is, for example, 9.

The characteristic of the function generator 800 shown in Fig. 4 is congruently depicted in Fig. 13.

As described previously, this function generator 800 converges the control variables ΔV_d* and ΔV_q* in order that the offset of the integrator in the region incapable of performing voltage control based on the current difference will not adversely affect the system.

In Fig. 13, gradual converging starts at a frequency at which $N_p* = 15$ changes to $N_p* = 9$ and zero is obtained concurrently with arrival at one pulse.

This converging region is a region in which voltage control is possible but current ripple is large because of a reduced pulse number and effective operation cannot be expected, and therefore control ratio (feedback control in the vector control system and in the slip frequency control system) is reduced. Thus, the timing for switching between the two systems is not particularly constrained by the pulse number and can be determined from a view point of whether voltage control is possible or not and whether current ripple is large or small.

In the region of one pulse, the vector control system does not operate completely and consequently only the slip frequency control system for controlling the primary angular frequency on the basis of the difference in primary current operates.

In the present embodiment, the gain of the vector control system starts dropping at a frequency where $N_p^* = 9$ takes place but this can be changed depending on the control scheme or the like factor.

To take care of the problem of the integrator offset and the like, the gain of the vector control system is dropped but by providing the function to reset the offset such as for example notch-off or notch return operation, the function generator 800 in the current controller 8 may be omitted.

In the case where the PWM pulse is decreased to the ultimate one pulse as described above, utilization rate of DC power supply voltage of the PWM inverter can be improved and DC current on the DC power supply side can be decreased.

Incidentally, regenerative running can be accomplished by making torque current command negative and inverting the polarity of output signal of the adder 808 (not shown).

As described above, since in the present embodiment terminal voltage necessary for the motor is fed and correction of the slip frequency (differences insensitivity and controllable object between the aforementioned control systems) is effected, excellent torque control can be carried out over the entire running region.

Fig. 14 shows a second embodiment of the invention. Only differential points from the first embodiment are shown. A torque command signal T* is calculated through a speed commander 23, an adder 24 for calculating the speed difference and a speed controller 25 and delivered to control command arithmetic unit 6. The speed controller 25 calculates proportional integration (P-I) of output signal $\Delta\omega_r$ of the adder 24 and delivers the torque command signal T*:

$$T^* = \left(P_A + \frac{K_A}{S} \right) (\omega_r^* - \omega_r) \qquad \ldots (13).$$

A control command arithmetic unit 6 calculates torque current command I_q^* for the induction motor 2 from the torque command signal T^* pursuant to the following equation:

$$I_q^* = K_T \cdot \frac{T^*}{I_d^*} \qquad \ldots (14)$$

where K_T is constant.

According to the present embodiment, forward running and backward running are determined in accordance with positive and negative polarities of the speed commander 23 and even when load on the induction motor 2 changes abruptly, speed can be so controlled as to coincide with a command signal.

The present embodiment can also be combined with Automatic Train Operation (ATO) easily, thus making it easy to practice constant speed running control.

Fig. 15 shows a third embodiment of the present invention. Like the second embodiment, only differential points are depicted. The third embodiment differs from the first embodiment of Fig. 1 in that a switching signal generator 26 for switching the control types is provided in order to forcibly make output signals ΔV_d^* and ΔV_q^* of the current controller 8 zero and multipliers 270 and 271 are provided which multiply an output signal of the switching signal generator 26 by the output signals ΔV_d^* and ΔV_q^* of the current controller 8. It is to be noted that when ΔV_d^* and ΔV_q^* are forcibly rendered to be zero, inputs to the controllers 803 and 806 shown in Fig. 4 are obviously rendered to be zero simultaneously forcibly, though not shown.

An electric rail car applied with the present embodiment is improved in torque characteristic in the low speed region in comparison with a conventional electric rail car of slip frequency control type so as to relieve load imposed on the conventional type electric rail car but on the other hand large requisite torque is needed and slip is liable to occur.

Accordingly, with a view of permitting combination of the electric rail cars of different control schemes, the control type of the present embodiment is forcibly applied with the slip frequency control to permit driving covering the entire running region.

The operation will now be described. Output signals of the controllers 803 and 806 shown in Fig. 4 are forcibly rendered to be zero and hence current components I_d and I_q do not coincide with command signals I_d* and I_q* so that the primary angular frequency may be controlled on the basis of the current difference between magnitude I_1 of output current and I_1*. As a result, the slip frequency control can proceed in the entire running region.

In the conventional electric rolling stock, the inverter output voltage is adjusted in accordance with the primary current difference only when starting and in order to permit coordination with that control system, the vector control system may be operated only in that region.

According to the present embodiment, when an electric car applied with the present invention and a conventional electric rail car of slip frequency control type are organized in the same train and caused to run, localization of load sharing between the electric rail cars can be prevented. Especially, this advantage is remarkable in electric locomotives.

Fig. 16 shows a fourth embodiment of the invention. As in the precedence, only differential points are depicted. The present embodiment differs from the second embodiment shown in Fig. 14 in that a speed detector 28 for detecting the actual speed of an electric vehicle, for example, the speed of a non-driven shaft is provided and torque command signal T* is calculated on the basis of the difference between speed command signal ω_r* and detection value ω_r of the speed detector 28. The torque command signal T* is expressed by the following equation:

$$T^* = \left(P_A + \frac{K_A}{S} \right) (\omega_r^* - \omega_T) \qquad \ldots (15)$$

According to the present embodiment, even when driving wheels slip, the torque command can be prepared accurately.

Fig. 17 shows a fifth embodiment of the invention.

The present embodiment differs from the Fig. 1 first embodiment in that in place of the coordinate transformation 4, an output voltage command arithmetic unit 30 and a coordinate transformation 29 are provided.

The output voltage command arithmetic unit 30 has a pattern based on characteristic curves of induction motor 2 which is representative of magnitude V^* of output voltage fed in accordance with primary angular frequency command signal ω_1^* and the magnitude V^* delivered to the coordinate transformation 29 is variable with filter capacitor voltage V_{FC}.

Detailed construction of the coordinate transformation 29 is shown in Fig. 18. The contents of calculation of the coordinate transformation 29 is expressed by the following equation:

$$\left.\begin{aligned}
v_u^* &= V_1^{**} \cdot \cos(\omega_1^* t + \theta_v^*) \\
v_v^* &= V_1^{**} \cdot \cos(\omega_1^* t - 2\pi/3 + \theta_v^*) \\
v_w^* &= V_1^{**} \cdot \cos(\omega_1^* t - 4\pi/3 + \theta_v^*) \\
\theta_v^* &= \pi/2 - \tan^{-1}\left(\frac{V_d^{**}}{V_q^{**}}\right) \\
V_1^{**} &= V^* + K_v(v_1^* - V^*)
\end{aligned}\right\} \quad \cdots (16)$$

where K_v represents output signal of a function generator 2900.

According to the present embodiment, with K_v rendered to be zero forcibly, control type equivalent to the slip frequency control type used in the conventional electric rail car is available and when the electric rail car applied with the present invention and the conventional electric rail car are organized in the same train the load is shared in each car in uniform.

The present invention may also be applied to synchronous motors in addition to the induction motor in a similar way.

Further, the present invention may also be applicable to rolling mills, elevators and other motors for general use in addition to the electric rail car.

Fig. 19 shows a sixth embodiment of the invention.

Differential points from the fifth embodiment shown in Fig. 17 will be described.

There are provided a first control command arithmetic unit 6 for delivering an exciting current command signal Id * and a torque current command signal I_q* and a second control command arithmetic unit 31 for delivering a primary current command signal I_1* and a slip angular frequency command signal ω_s*, and vector control is switched to slip frequency control or vice versa in accordance with a primary angular frequency signal ω_1* or a corresponding value.

The second control command arithmetic unit 31 delivers the primary current command signal I_1* based on the forward/backward command signal D*, power running/braking force command signal N*, ω_1* and filter capacitor voltage signal and the slip angular frequency command signal ω_s* to adders 32 and 33, respectively.

The adder 32 delivers a difference between the primary current command signal I_1* and a primary current detection signal I_1 to a controller 34.

The controller 34 calculates a signal $\Delta\omega_s$* for correcting the slip angular frequency and delivers it to the adder 33.

The adder 33 calculates a slip angular frequency command signal ω_{s2}* for controlling the slip frequency and delivers it to a switching unit 34.

As will be described later, the switching unit 34 responds to the primary angular frequency command signal ω_1* to sequentially switch a slip angular frequency command signal ω_{s1}* during vector control and the slip angular frequency command signal ω_{s2}* during slip frequency control so as to deliver a slip angular frequency command signal ω_s** to an adder 22. 803 is I_d controller, 806 is I_q controller, and 809 is I_1 controller. 810 calculates $I_1 = \sqrt{I_d^2 + I_q^2}$.

Fig. 20 shows detailed construction of the slip angular frequency switching unit 34.

A function generator 3400 responds to the primary angular frequency command signal ω_1* to produce a switching gain $K\omega$ which is delivered to a multiplier 3402.

The contents of operation by the switching unit 34 is given by the following equation:

$$\omega_s^{**} = \omega_{s2}^{*} + K\omega\ (\omega_{s1}^{*} - \omega_{s2}^{*}) \qquad \dots (17)$$

The operation will now be described.

On the basis of output signal I_d^* and I_q^* of the first control command arithmetic unit 6, output voltage command V_1^* and slip angular frequency command ω_s^* during vector control are calculated. On the basis of output signals I_1^* and ω_s^* of the second control command arithmetic unit 31, output voltage command V^* and slip angular frequency command ω_{s2}^* during slip frequency control are calculated.

For switching the vector control system and the slip frequency control system, switching of output voltage commands is effected by a coordinate transformation 29 and switching of slip angular frequency is effected by the switching unit 34. Further, the slip frequency control system can be maintained by a command from an operation console.

The present embodiment differs from the other first to fifth embodiments in that while in the other embodiments the inverter 1 is applied with the voltage commands having phase relation, i.e., vectors even in the region where closed loop control of exciting current and torque current is difficult to achieve (extending from the region in which ripple of output current increases to the one pulse region), the control is completely switched to the slip frequency control system, if necessary in the present embodiment (this switching can be accomplished by changing switching gains K_v and K_w from 1 to 0).

By completely switched to the slip frequency control system, the present embodiment can be combined with the conventional electric rail car applied with the control system having slip frequency control based on primary current feedback control.

Fig. 21 shows a seventh embodiment of the invention. The present embodiment differs from the first embodiment shown in Figs. 1 and 4 in that output signal of an adder 811 is used as input signal to the controller 809.

The adder 811 adds output signal ΔI_d of the adder 801 and output signal ΔI_q of the adder 804 to produce output signal $\Delta I_1{}^*$ which is delivered to the controller 809.

The operation of the present embodiment will now be described. Fig. 22 shows change ΔI_1 in magnitude of the primary current with respect to change of motor constant $L_s\sigma^*$ and q-axis magnetic flux ϕ_{2q}. In the figure, solid line represents change ΔI_1 which is normalized in magnitude by the the magnitude of the rated current and chained line represents q-axis magnetic flux ϕ_{2q} which is normalized in magnitude by the magnitude of the rated magnetic flux. As set value

$L_s\sigma^*$ in the control system of leakage inductance increases, ΔI_1 and ϕ_{2q} both increase in positive sense. This tendency differs from that with respect to r_1^* and ω_s^* explained in connection with Figs. 9 and 10 in that when ω_1^* is corrected on the basis of ΔI_1, ϕ_{2q} increases conversely. Thus, in the present embodiment, ΔI_d is added with ΔI_q to provide the sum which is ΔI_1^* and the primary angular frequency is corrected with respect to a set error of $L_s\sigma^*$ on the basis of the ΔI_1^*. This takes advantage of the fact that ΔI_d and ΔI_q are opposite in polarity and substantially equal in magnitude with respect to $L_s\sigma^*$ as shown at dotted line in Figs. 7 and 8, thereby ensuring that correction sensitivity of the primary angular frequency ω_1^* to the set error of $L_s\sigma^*$ can be rendered to be substantially zero so as to secure correction sensitivity to r_1^* and ω_s^*.

In this manner, according to the present embodiment, even when the $L_s\sigma^*$ has a large setting error, stable correction of ω_1^* can be effected without increasing ϕ_{2q}.

Fig. 23 shows an eighth embodiment of the invention. The present embodiment differs from the first embodiment of Figs. 1 and 4 in that a multiplier 812 for multiplying output signal of the adder 808 by output signal of the adder 801 is provided and output signal of the multiplier 812 is inputted to the controller 809.

The operation of the present embodiment will now be described. As shown in Fig. 22, both of primary current ΔI_1 and q-axis magnetic flux ϕ_{2q} with respect to change of set value $L_s\sigma^*$ in the control system of leakage inductance increase or decrease. Therefore, when the primary angular frequency is corrected on the basis of ΔI_1, ϕ_{2q} increases conversely.

Thus, the present embodiment takes advantage of the fact that ΔI_d and Δq with respect to motor constants r_1^* and ω_s^* both increase or decrease but ΔI_d and ΔI_q with respect to change of $L_s\sigma^*$ increase or decrease in opposite senses so as to suppress ϕ_{2q} with respect to change of $L_s\sigma^*$ by utilizing the polarity of ΔI_d. More particularly, as $L_s\sigma^*$ increases, ΔI_1 also increases but ΔI_d decreases in negative sense as shown in Fig. 7. When the ΔI_1 is multiplied by the polarity of ΔI_d to calculate ΔI_1^*, the relation can be obtained in which ΔI_1^* and q-axis magnetic flux ϕ_{2q} increase or decrease in opposite senses. Consequently, as ΔI_1^* decreases in negative sense, the controller 809 increases the primary angular frequency ω_1^* to control ϕ_{2q} such that it becomes zero.

According to the present embodiment, since d-axis of the control system can be controlled with respect to the motor constants such that it coincides with magnetic flux (m-axis) of the motor, the d-axis controller 87 and q-axis controller 88 do not interfere with each other and can operate independently of each other.

As described above, according to the first to eighth embodiments, there are provided the controller for controlling the primary angular frequency such that d-axis of the control system coincides with magnetic flux (m-axis) of the motor and the current controllers for correcting d-axis voltage and q-axis voltage, respectively, whereby the individual controllers correct the voltage errors with respect to changes of motor constants so that as shown in Fig. 11 r_1^*/r_1, $L_s\sigma^*/L_s\sigma$ and ω_s^*/ω_s may be rendered to be 1.0 to produce torque exactly as commanded and to permit high-response torque control.

The foregoing first to eighth embodiments have been described by way of example of electric rail car driving AC motor control apparatus but the essence of the present invention may be applied also to apparatus for other purposes in addition to the electric rail car, for example, an AC motor control apparatus of rolling mill.

According to the above described embodiments of the invention, the effect of controlling torque more accurately can be attained.

Further, even in the region in which inverter output voltage control is not permitted, the control can advantageously continue to proceed.

Further, torque fluctuation in the low speed region can advantageously be prevented and in the high speed region, the power supply voltage utilization rate can advantageously be improved by 115% or more.

THE CLAIMS DEFINING THE INVENTION ARE AS FOLLOWS:

1. A control apparatus of an AC motor driven by a power converter for delivering alternating current of variable voltage and variable frequency, characterized in that said apparatus comprises vector control means for controlling a vector of primary current of said motor and slip frequency control means for controlling a slip frequency of said motor, and that these two control means are used in combination.

2. A control apparatus according to claim 1, wherein said slip frequency control means controls the slip frequency on the basis of a difference between a selected magnitude for the primary current of said motor and a detected magnitude of the primary current.

3. A control apparatus according to claim 1, wherein said vector control means comprises closed loop control systems for respective vector components including an exciting component and a torque component constituting the primary current.

4. A control apparatus according to claim 3, characterized in that the slip frequency of said motor is controlled on the basis of a closed loop control system for the magnitude of primary current of said motor, an exciting component command for output voltage of said power converter is controlled on the basis of the closed loop control system for the exciting component of said primary current, and a torque component command for the output voltage of said power converter is controlled on the basis of the closed loop control system for the torque component of said primary current.

5. A control apparatus according to claim 1, wherein

said slip frequency control means controls the slip frequency on the basis of a difference between a selected magnitude for the primary current of said motor and a detected magnitude of the primary current,

said vector control means has means for outputting an exciting component command and a torque component command for an output voltage of said power converter on the basis of vector components of exciting current and torque current constituting the primary current, means for correcting the exciting component command for the output voltage in accordance with a difference between the detected exciting component of said primary current and a selected value thereof, and means for correcting the torque component command for the output voltage in accordance with a difference between the detected torque component of said primary current and a selected value thereof, and

said apparatus further comprises means for preventing outputs of said exciting component correcting means and torque component correcting means in a region wherein the output voltage of said power converter gets out of control.

6. A control apparatus according to claim 5, wherein said means for outputting an exciting component command and a torque component command for the output voltage of said power converter outputs the commands over the entire operation frequency region of said motor including the region wherein the output voltage of said power converter gets out of control.

7. A control apparatus according to claim 1, further including means for switching between said vector control means and said slip frequency control means on the basis of an output frequency of said power converter.

8. A control apparatus according to claim 1, wherein

said vector control means has means for producing a first output voltage command for said power converter in accordance with predetermined values for the exciting component and torque component of the primary current of the motor,

said slip frequency control means has means for controlling the slip frequency of the motor in accordance with a difference between the predetermined magnitude for the motor primary current and a detected magnitude of the primary current,

said apparatus further comprises means for calculating an operating frequency of said power converter from said slip frequency and a rotation frequency of the motor, means for producing a second output voltage command for said power converter in accordance with the operating frequency, and means for producing a new output voltage command for said power converter from said first and second output voltage commands.

9. A control apparatus according to claim 1, wherein

said vector control means outputs an exciting component command and a torque component command for the voltage outputted from said power converter in accordance with predetermined values for the exciting component and torque component of the primary current of the motor and a feedback value thereof,

said slip frequency control means outputs a slip frequency command on the basis of the primary current of the motor and the feedback value thereof,

said apparatus further comprises means for commanding the output frequency of the power converter obtained by calculation with a predetermined slip frequency and a detected rotation frequency of the motor or commanding an equivalent value of said output frequency, and means for increasing or decreasing the exciting component command and the torque component command of said voltage in accordance with the output frequency of the power converter or an equivalent value of said output frequency.

10. A control apparatus for an electric vehicle having a power converter for delivering alternating current of variable voltage and variable frequency, and an AC motor energized by output of the power converter for driving the electric vehicle, characterized by the provision of means for determining values for exciting and torque components of primary current of said AC motor on the basis of a command from a master controller, means for determining a slip frequency for the AC motor from the exciting and torque current component values, means for determining values for exciting and torque components of output voltage of said power converter on the basis of the exciting and torque primary current component values, means for correcting the exciting component value of output voltage in accordance with a difference between said determined exciting component value of primary current and a measured exciting current component, means for correcting the torque component value of output voltage in accordance with a difference between said determined torque component value of primary current and a measured torque current component, and means for correcting

said slip frequency on the basis of a difference between a measured magnitude of primary current of said AC motor and a predetermined value thereof based on the exciting and torque primary current components.

11. A control apparatus according to claim 10, further comprising means for decreasing outputs of said means for correcting said exciting component value and means for correcting said torque component value of the output voltage in a predetermined inverter frequency region.

12. A control apparatus for an AC motor, substantially as hereinbefore described with reference to the accompanying drawings.

DATED this 13th day of October, 1994

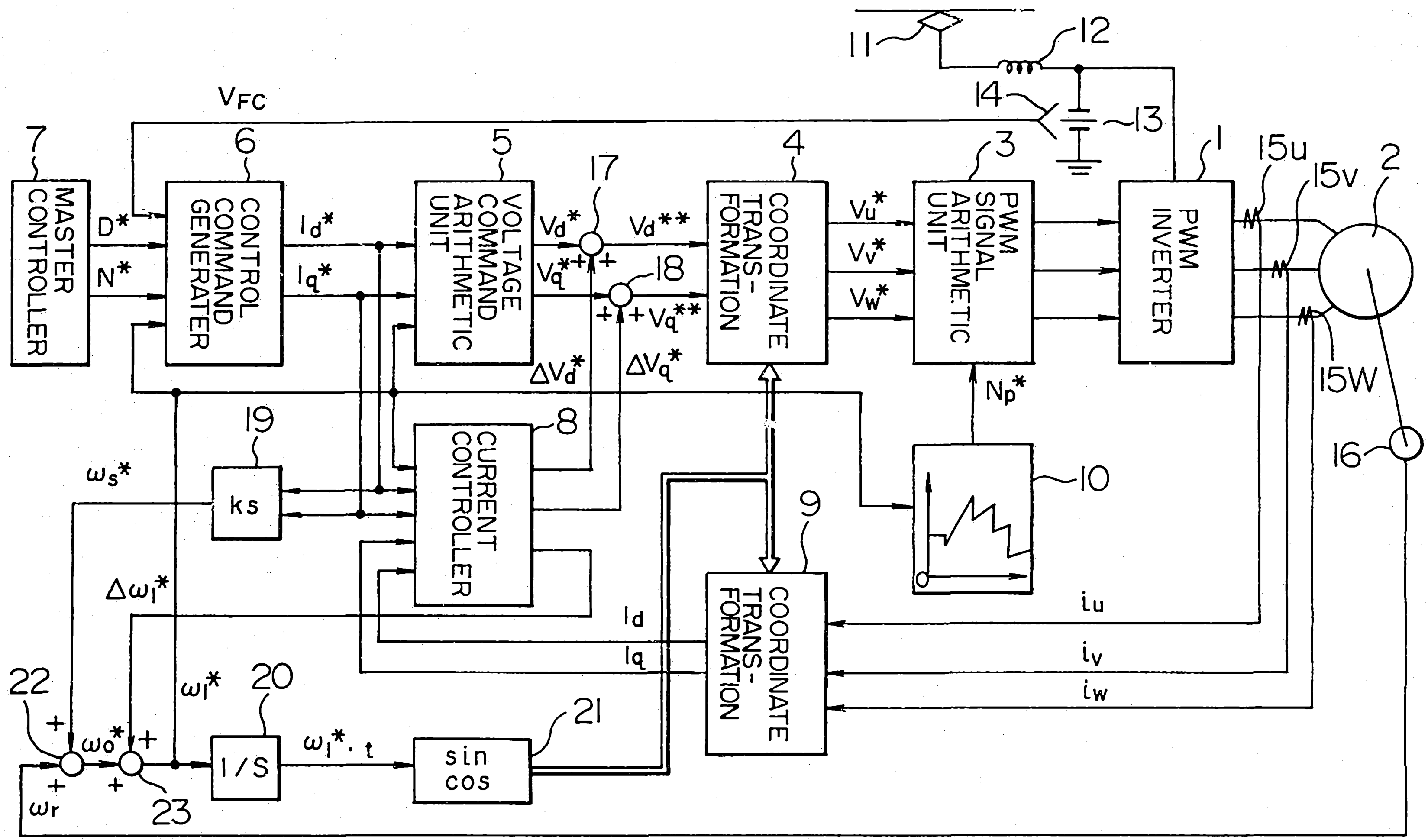
HITACHI, LTD. ~~AND HITACHI TECHNO ENGINEERING CO. LTD.~~ AND HITACHI MITO ENGINEERING CO. LTD.

by DAVIES COLLISON CAVE
Patent Attorneys for the Applicant

ABSTRACT OF THE DISCLOSURE

There is provided a control apparatus of an AC motor capable of performing stable torque control even over the entire running region of the AC motor. Structurally, a feedback control system for controlling the magnitude of motor primary current, a feedback control system for controlling the exciting component of motor primary current and a feedback control system for controlling the torque component of motor primary current are provided. Torque control is effected by a vector control system in the low speed running region and torque control is effected by a slip frequency control system in the high speed running region, so that torque can be controlled excellently over the entire running region.

FIG. 1



1/14

24532/92

FIG. 2

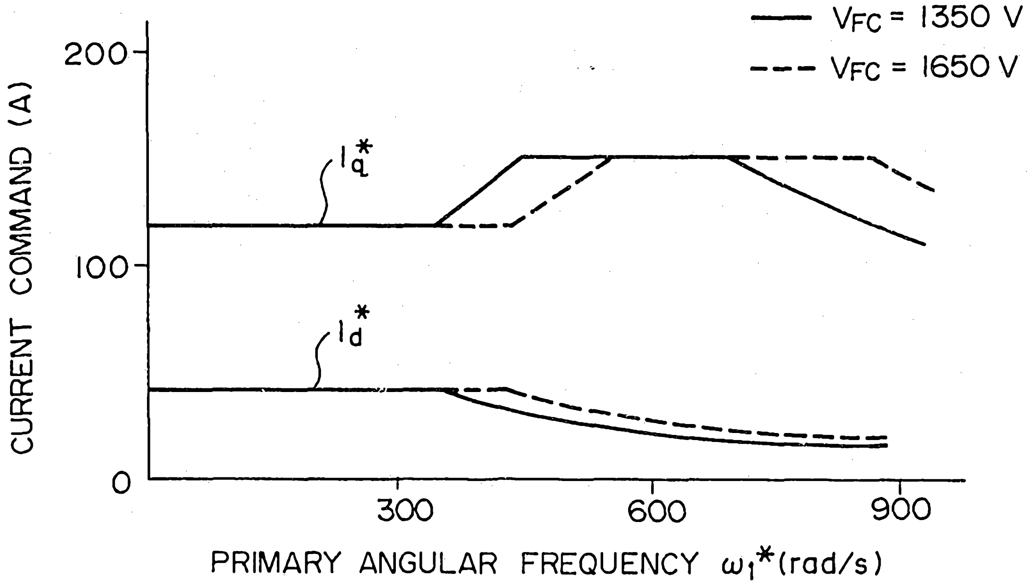

VFC = 1350 V
VFC = 1650 V
200
100
0
CURRENT COMMAND (A)
Iq*
Id*
300
600
900
PRIMARY ANGULAR FREQUENCY ω1*(rad/s)

FIG. 3

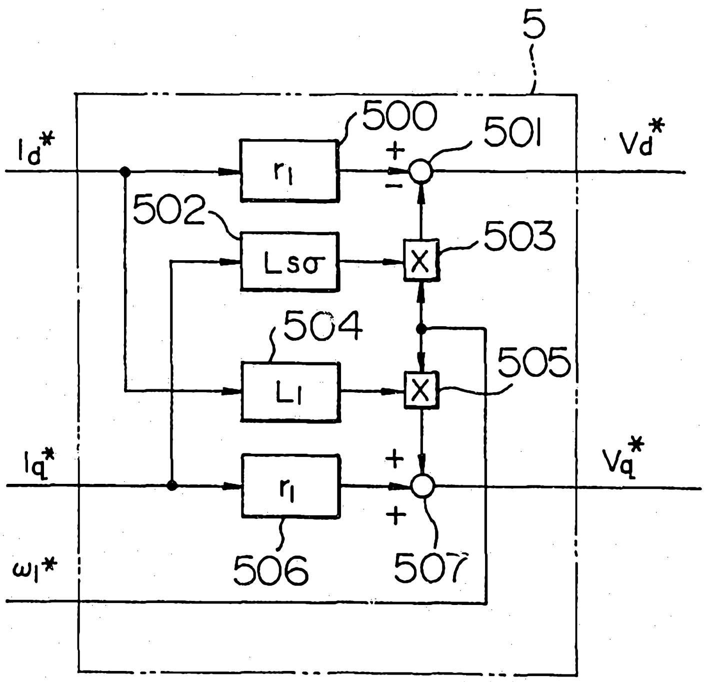
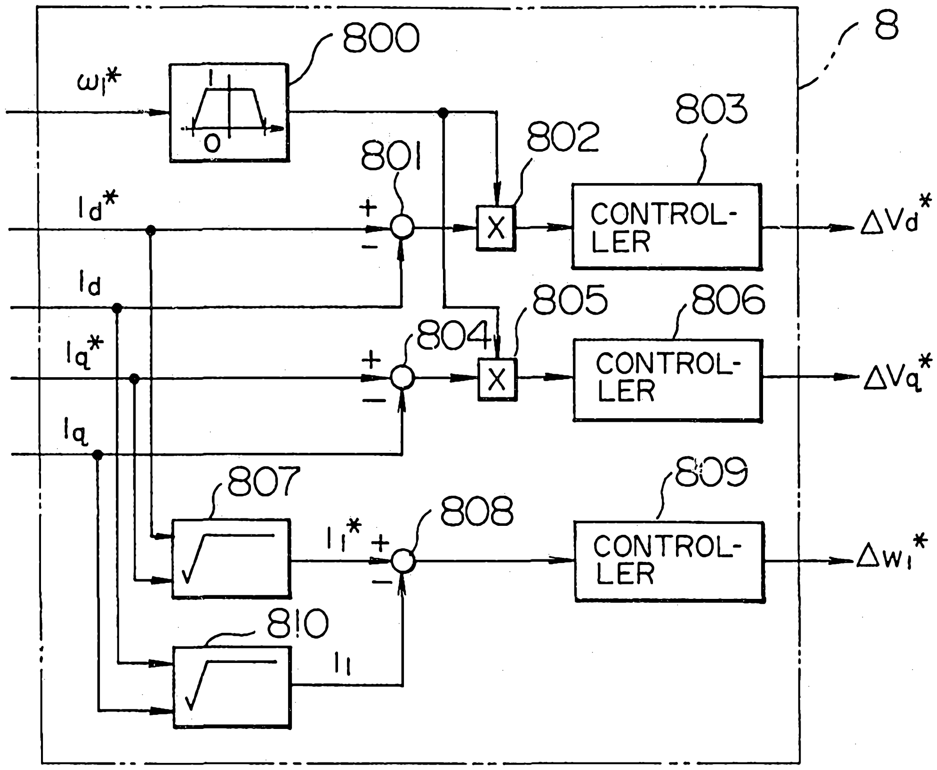
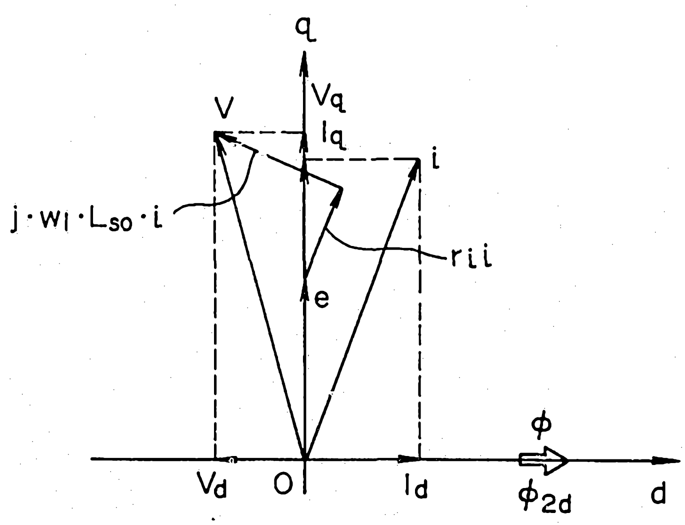

5
Id*
500
501
Vd*
502
Lsσ
503
504
L1
505
Iq*
r1
Vq*
506
507
ω1*
+
−
X

FIG. 4



8
800
ω_1*
1
0
801
802
803
CONTROL-LER
I_d*
+
−
ΔV_d*
I_d
805
806
804
I_q*
+
−
CONTROL-LER
ΔV_q*
I_q
807
809
808
I_1*
+
−
CONTROL-LER
Δw_1*
810
I_1

FIG. 5



q
V
Vq
Iq
i
j·w1·Lso·i
r1i
e
φ
Vd
O
Id
φ2d
d

FIG. 6

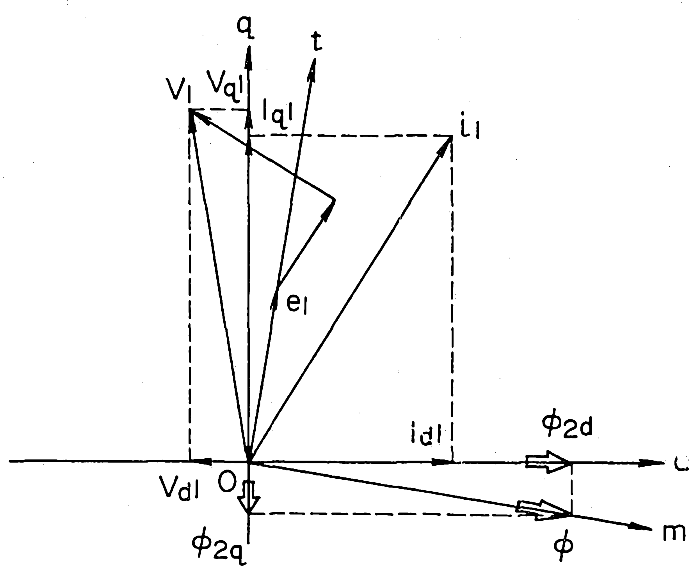

q
t
V_1
V_{q1}
I_{q1}
i_1
e_1
i_{d1}
ϕ_{2d}
V_{d1}
O
ϕ_{2q}
ϕ
m

FIG. 7

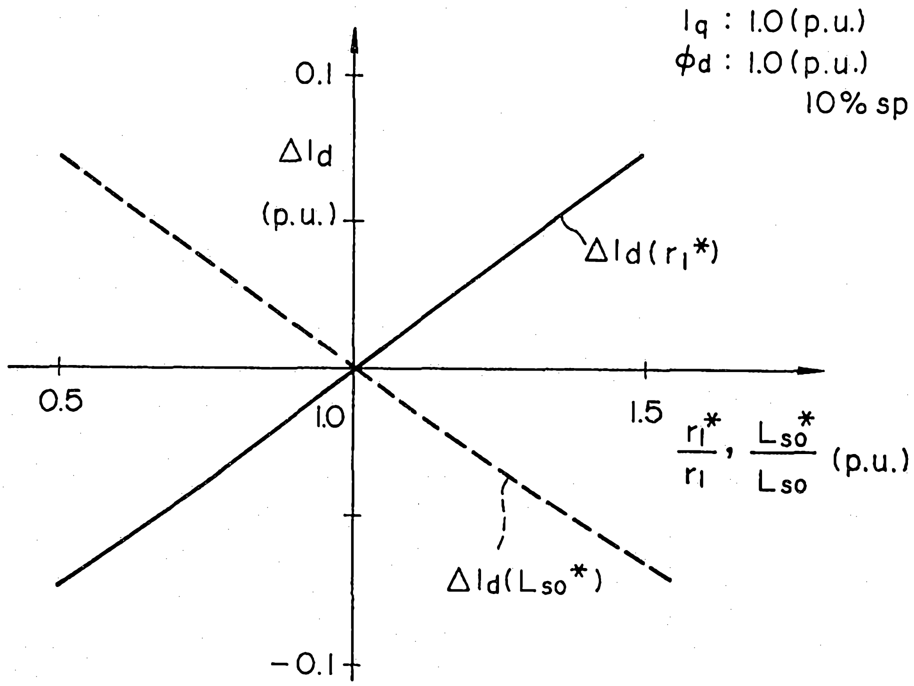

I_q : 1.0 (p.u.)
ϕ_d : 1.0 (p.u.)
10% sp
0.1
ΔI_d
(p.u.)
$\Delta I_d(r_1^*)$
0.5
1.0
1.5
$\frac{r_1^*}{r_1}$, $\frac{L_{so}^*}{L_{so}}$ (p.u.)
$\Delta I_d(L_{so}^*)$
−0.1

FIG. 8

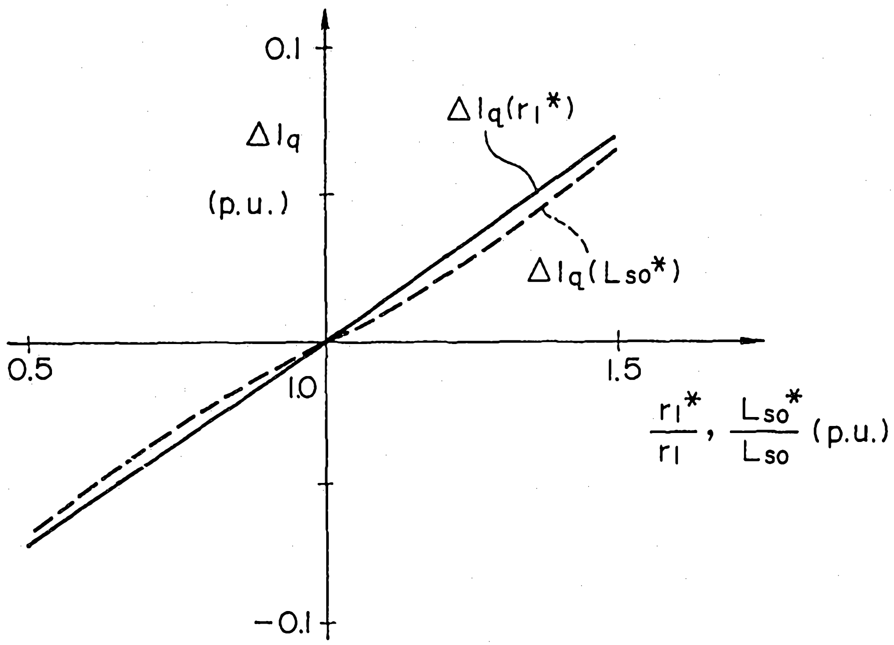

0.1
ΔIq
(p.u.)
ΔIq(r1*)
ΔIq(Lso*)
0.5
1.0
1.5
r1*/r1, Lso*/Lso (p.u.)
−0.1

FIG. 9

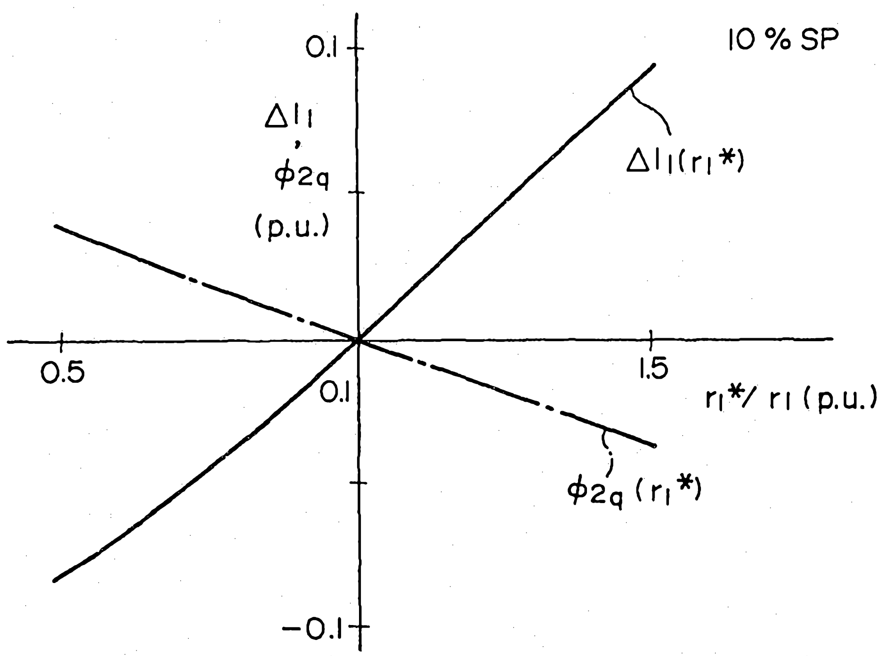

0.1
ΔI1, φ2q
(p.u.)
10 % SP
ΔI1(r1*)
0.5
0.1
1.5
r1*/ r1 (p.u.)
φ2q (r1*)
−0.1

FIG. 10

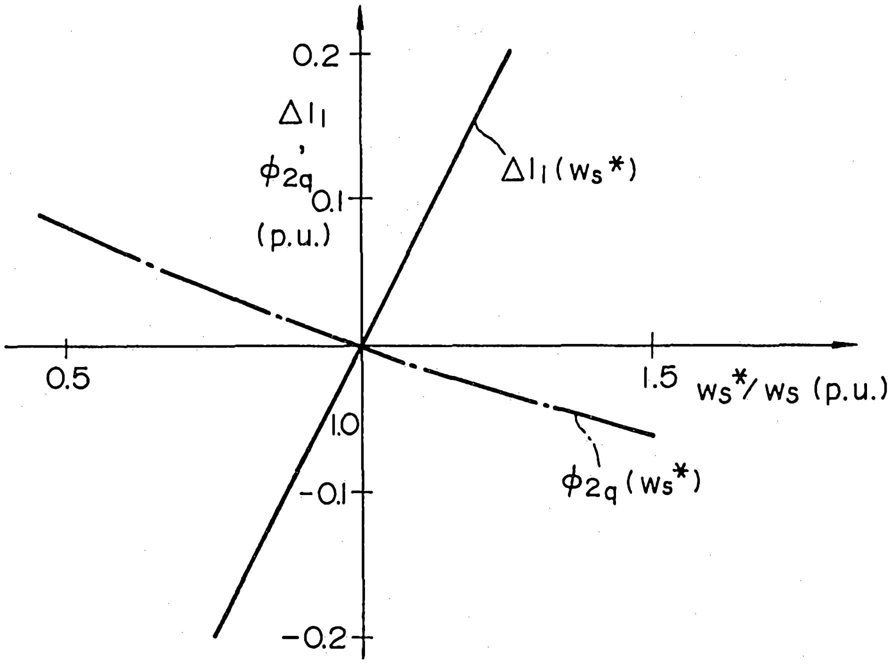


FIG. 11

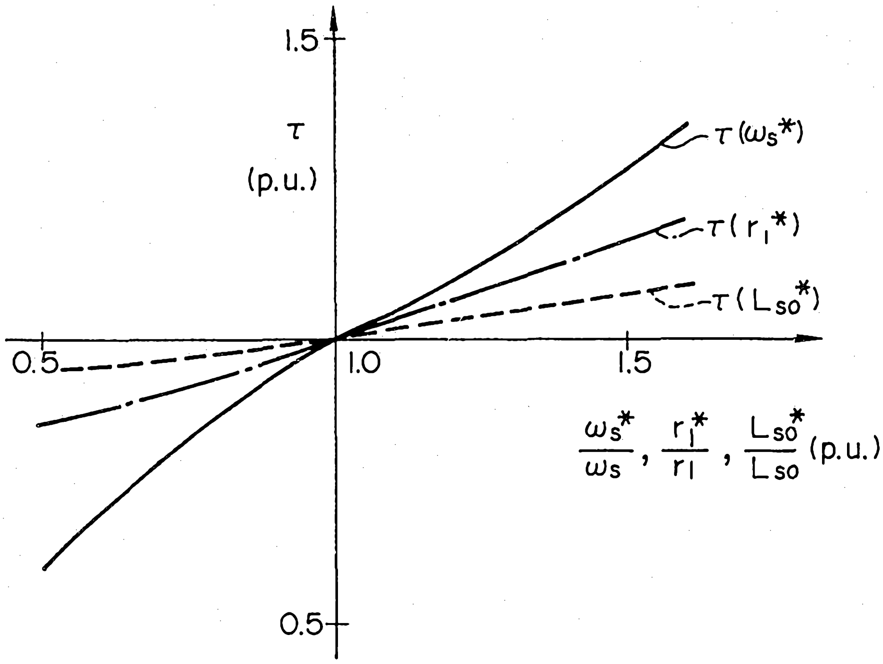

1.5
τ
(p.u.)
τ (ωs*)
τ (r1*)
τ (Lso*)
0.5
1.0
1.5
ωs*/ωs , r1*/r1 , Lso*/Lso (p.u.)
0.5

FIG. 12

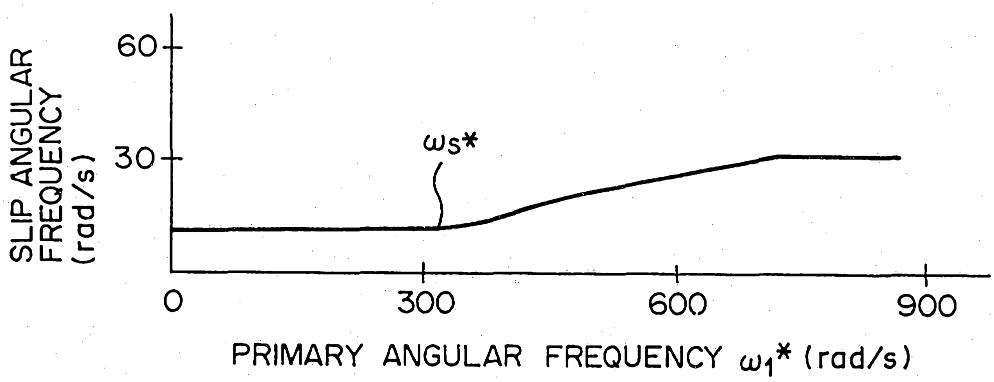

SLIP ANGULAR FREQUENCY (rad/s)
60
30
ωs*
0
300
600
900
PRIMARY ANGULAR FREQUENCY ω1* (rad/s)

FIG. 13

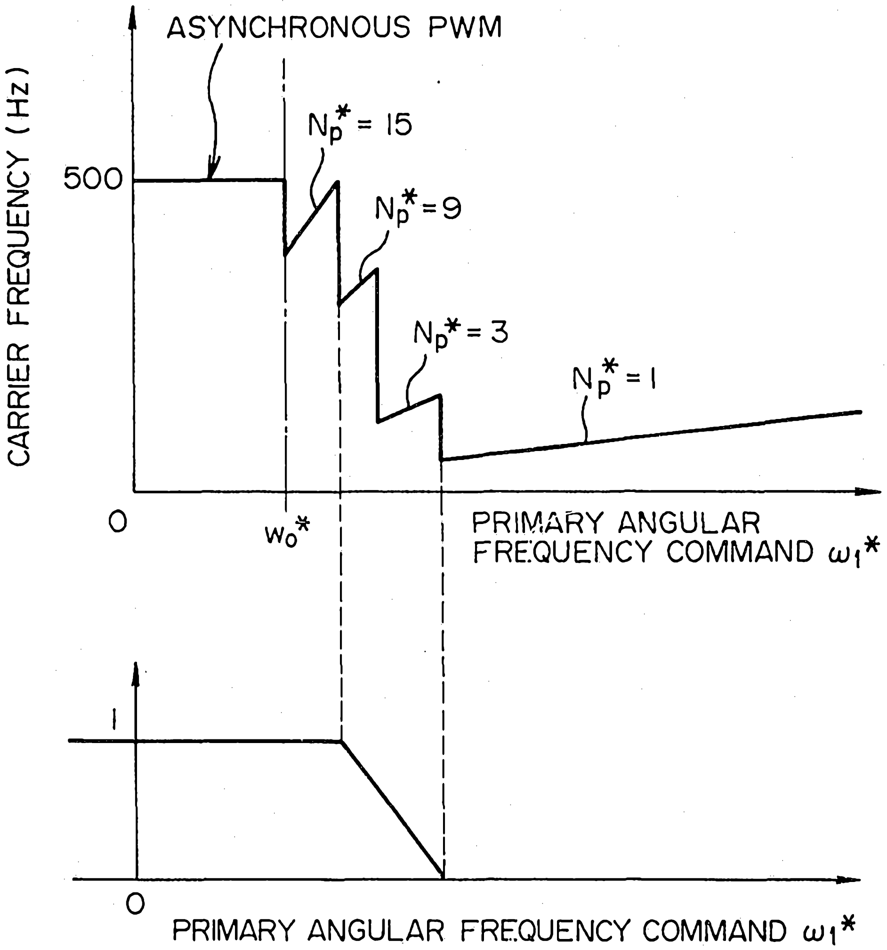

ASYNCHRONOUS PWM
CARRIER FREQUENCY (Hz)
500
$N_p^* = 15$
$N_p^* = 9$
$N_p^* = 3$
$N_p^* = 1$
0
w_0^*
PRIMARY ANGULAR FREQUENCY COMMAND ω_1^*
1
0
PRIMARY ANGULAR FREQUENCY COMMAND ω_1^*

FIG. 14

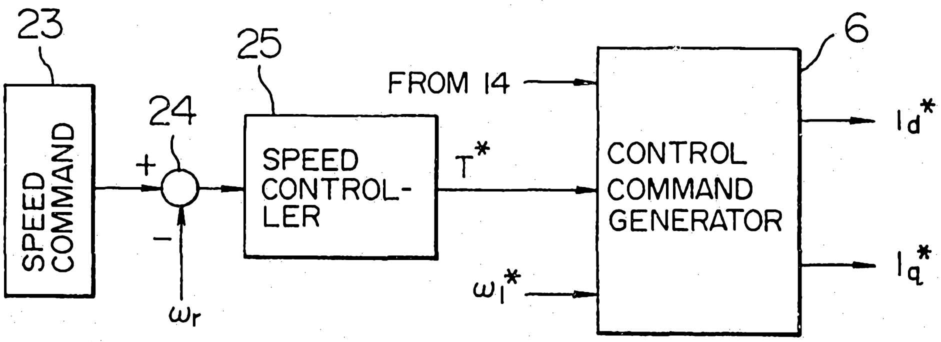

23
SPEED COMMAND
24
+
−
ω_r
25
SPEED CONTROLLER
T^*
FROM 14
ω_1^*
6
CONTROL COMMAND GENERATOR
I_d^*
I_q^*

FIG. 15

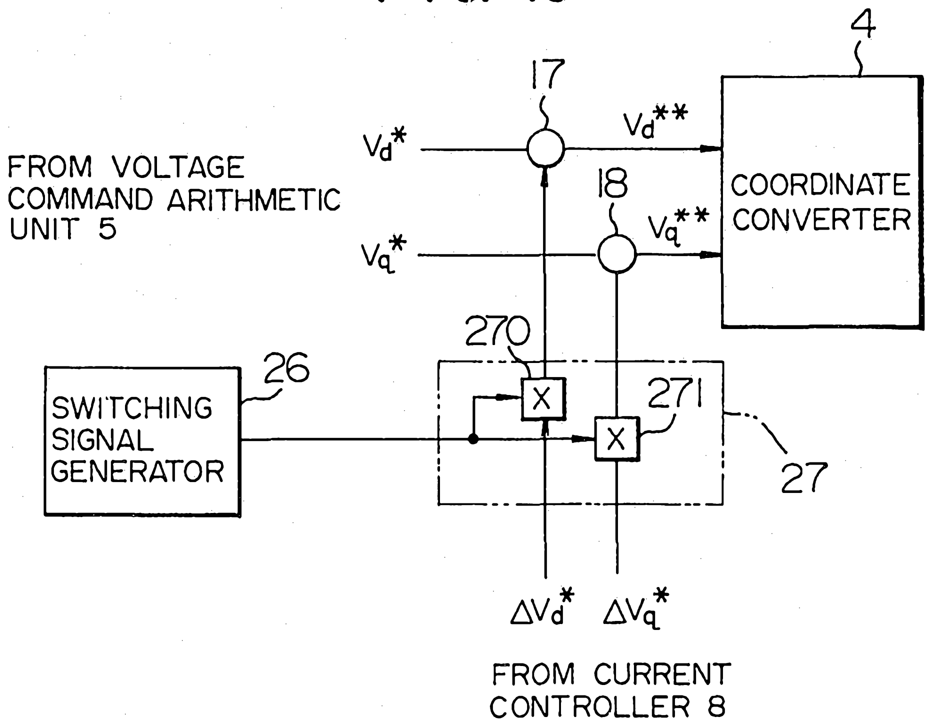

4
17
V_d^{**}
V_d^*
FROM VOLTAGE
COMMAND ARITHMETIC
UNIT 5
COORDINATE
CONVERTER
18
V_q^{**}
V_q^*
270
26
SWITCHING
SIGNAL
GENERATOR
X
271
X
27
ΔV_d^*
ΔV_q^*
FROM CURRENT
CONTROLLER 8

FIG. 16

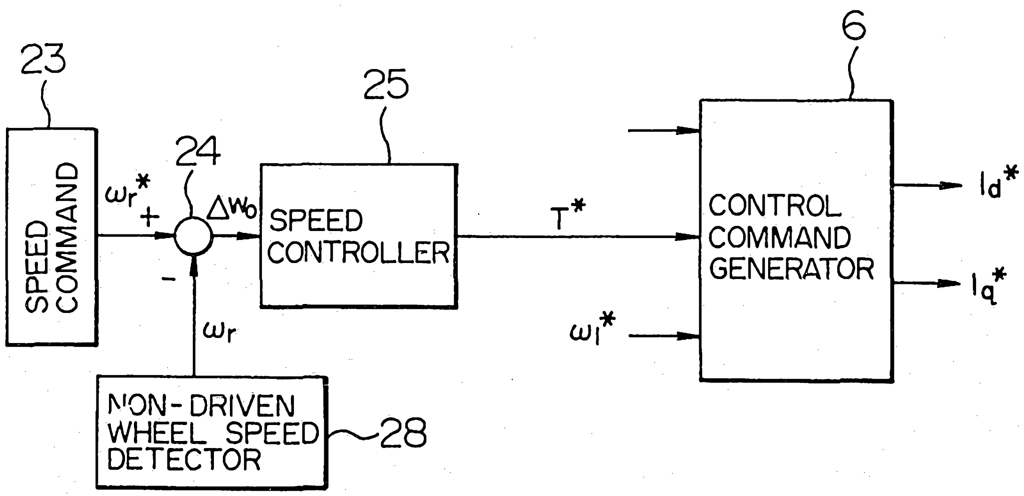

6
23
25
24
ω_r^*
+
$\Delta\omega_0$
SPEED
COMMAND
SPEED
CONTROLLER
T^*
CONTROL
COMMAND
GENERATOR
I_d^*
I_q^*
-
ω_r
ω_1^*
NON-DRIVEN
WHEEL SPEED
DETECTOR
28

FIG. 17

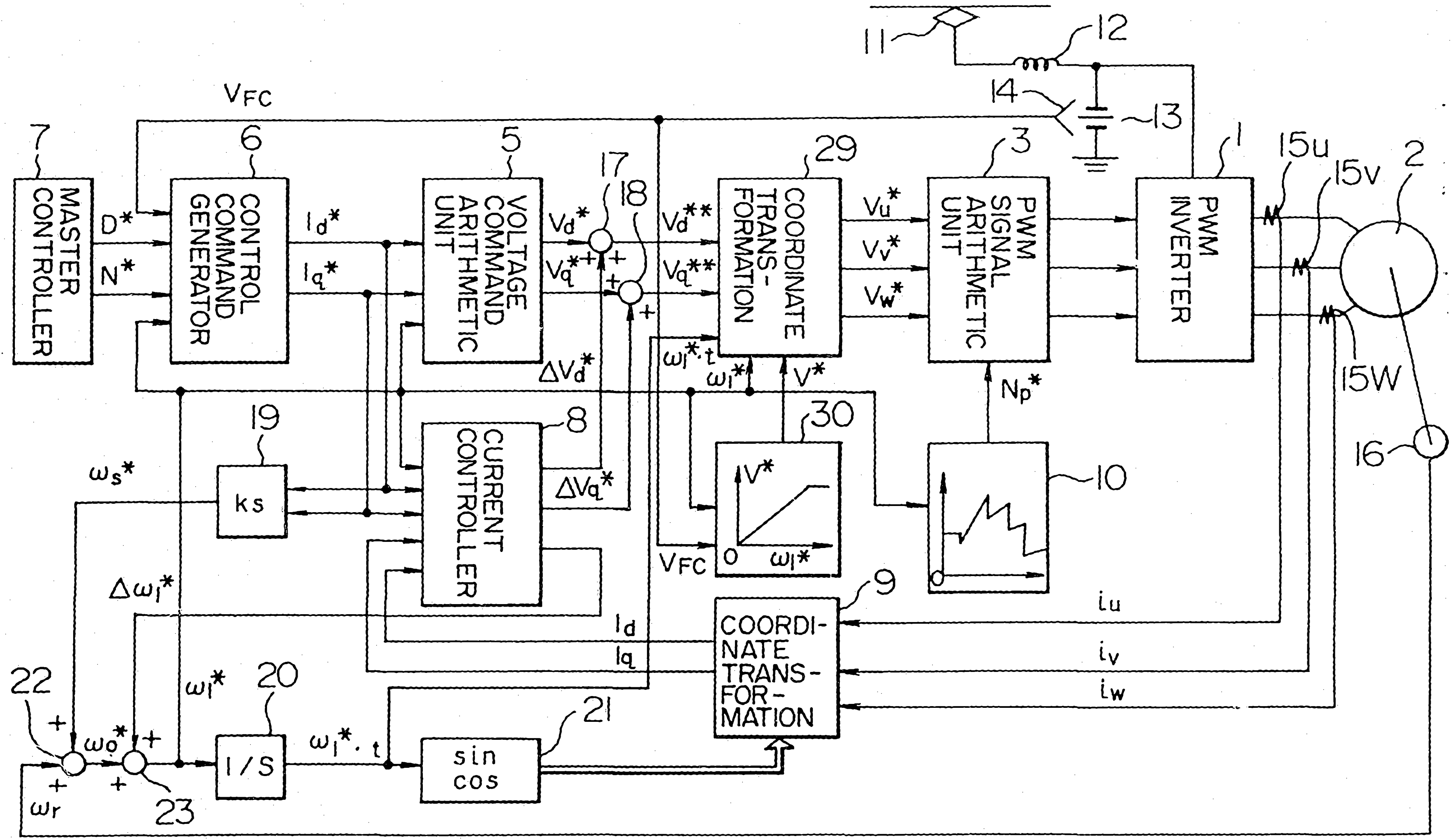

MASTER CONTROLLER
CONTROL COMMAND GENERATOR
VOLTAGE COMMAND ARITHMETIC UNIT
COORDINATE TRANS-FORMATION
PWM SIGNAL ARITHMETIC UNIT
PWM INVERTER
CURRENT CONTROLLER
COORDI-NATE TRANS-FOR-MATION
ks
I/S
sin cos
VFC
D*
N*
Id*
Iq*
Vd*
Vq*
Vd**
Vq**
ΔVd*
ΔVq*
Vu*
Vv*
Vw*
V*
Np*
ω1*
ω1*·t
ωs*
Δω1*
ωo*
ωr
Id
Iq
iu
iv
iw
0
1
2
3
5
6
7
8
9
10
11
12
13
14
15u
15v
15W
16
17
18
19
20
21
22
23
29
30

FIG. 18

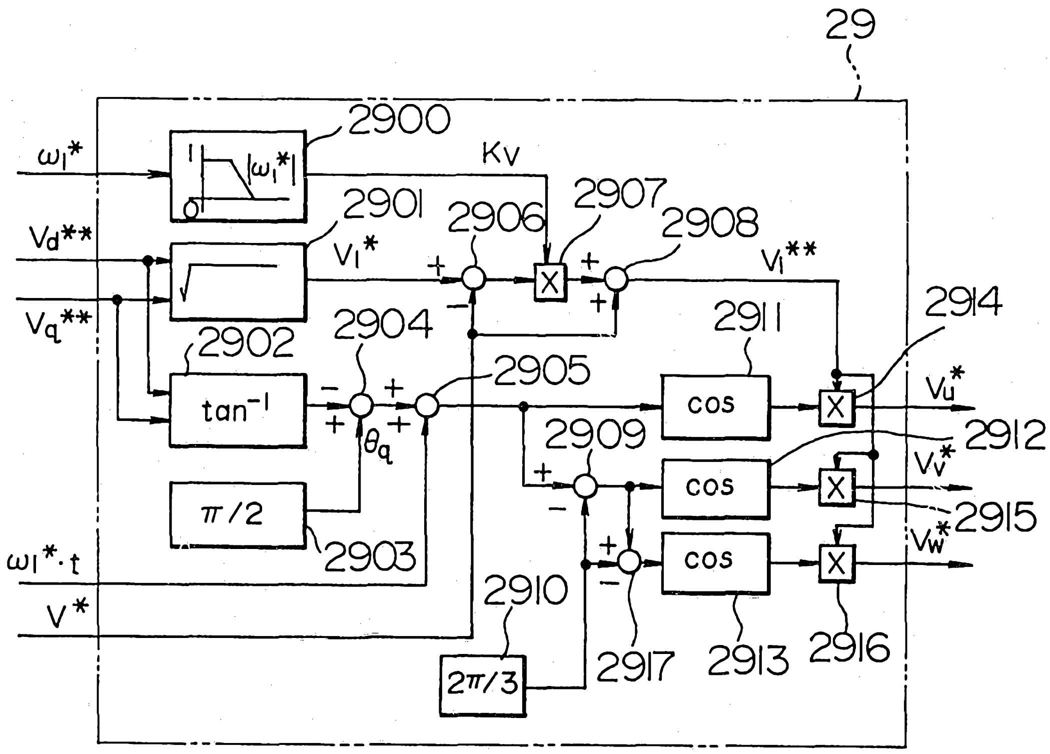
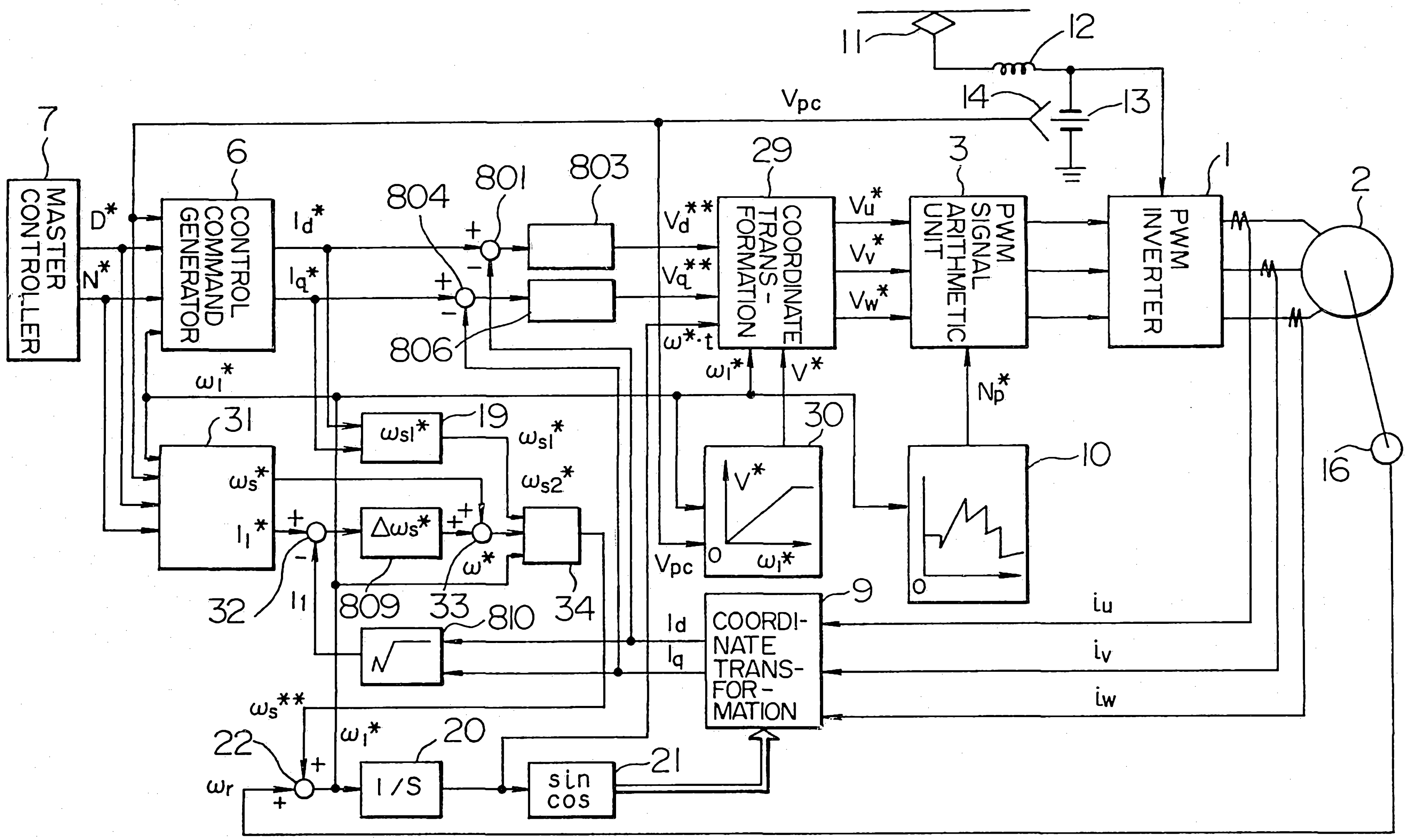

29
2900
ω1*
|ω1*|
Kv
2901
Vd**
Vq**
V1*
2906
2907
2908
V1**
2902
tan⁻¹
2904
θq
2905
2911
cos
2914
Vu*
2909
cos
2912
Vv*
π/2
2903
2915
cos
Vw*
ω1*·t
V*
2910
2π/3
2917
2913
2916

FIG. 19



MASTER CONTROLLER
CONTROL COMMAND GENERATOR
COORDINATE TRANS-FORMATION
PWM SIGNAL ARITHMETIC UNIT
PWM INVERTER
COORDI-NATE TRANS-FOR-MATION
V_{pc}
D^*
N^*
I_d^*
I_q^*
V_d^{**}
V_q^{**}
$\omega^* \cdot t$
ω_1^*
V^*
V_u^*
V_v^*
V_w^*
N_p^*
ω_{sl}^*
ω_s^*
ω_{s2}^*
I_1^*
$\Delta\omega_s^*$
ω^*
I_1
I_d
I_q
i_u
i_v
i_w
ω_s^{**}
ω_r
1/S
sin
cos
1
2
3
6
7
9
10
11
12
13
14
16
19
20
21
22
29
30
31
32
33
34
801
803
804
806
809
810

FIG. 20

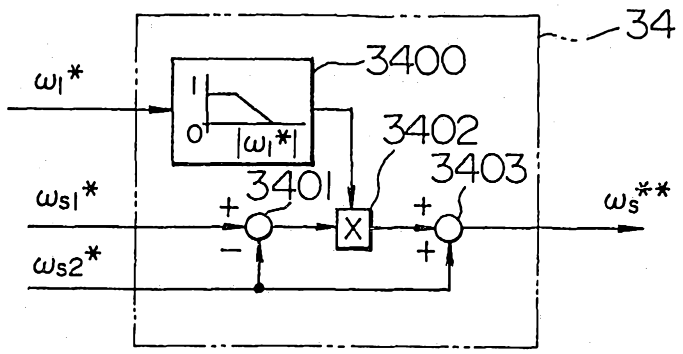

34
3400
1
0
|ω1*|
ω1*
3402
3403
3401
ωs1*
ωs2*
ωs**
X
+
−
+
+

FIG. 21

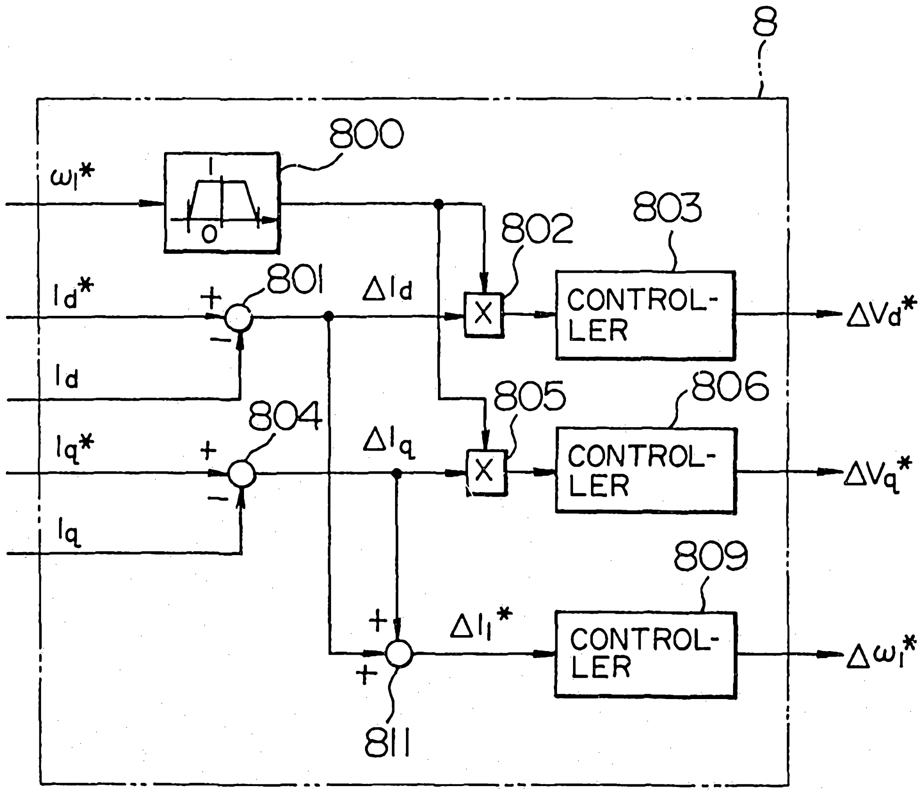
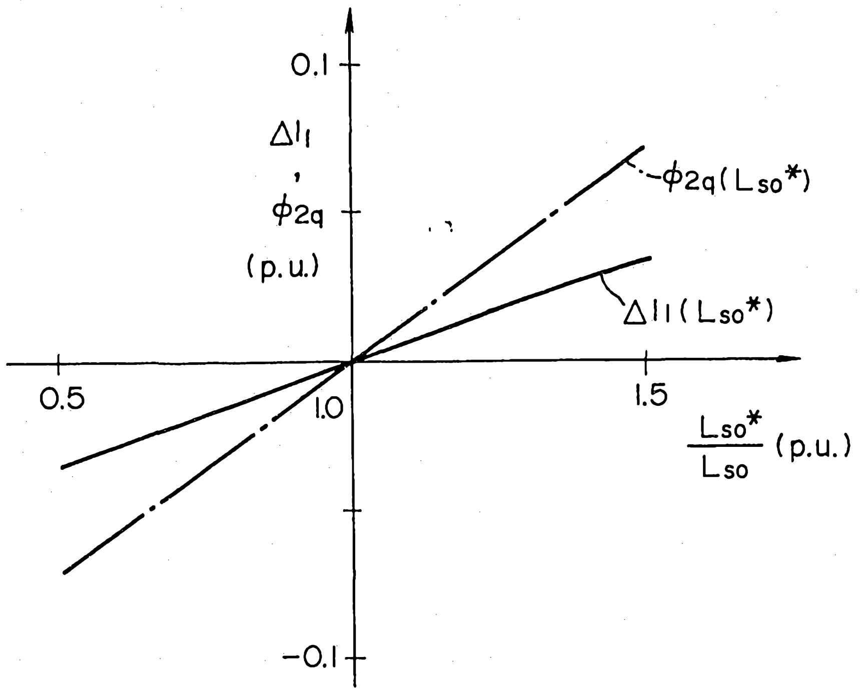
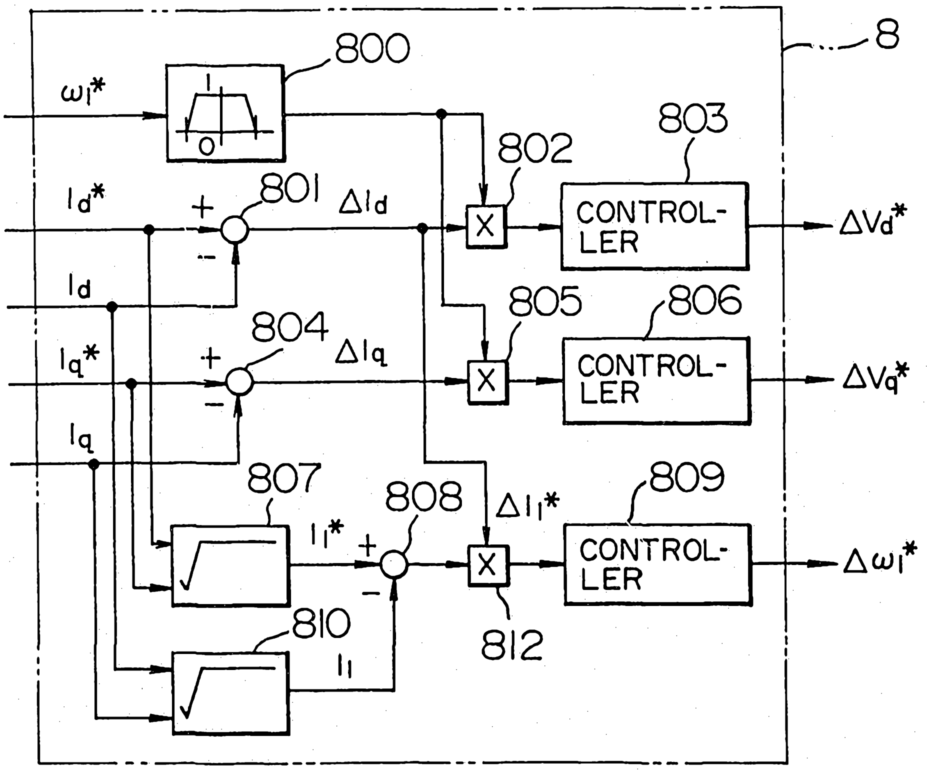

8
800
ω1*
0
1
803
802
801
ΔId
Id*
Id
X
CONTROL-
LER
ΔVd*
805
806
804
ΔIq
Iq*
Iq
X
CONTROL-
LER
ΔVq*
809
ΔI1*
CONTROL-
LER
Δω1*
811
+
−
+
−
+
+

FIG. 22



0.1
ΔI_1 , ϕ_{2q} (p.u.)
$\phi_{2q}(L_{so}^*)$
$\Delta I_1(L_{so}^*)$
0.5
1.0
1.5
$\frac{L_{so}^*}{L_{so}}$ (p.u.)
-0.1

FIG. 23



8
800
ω_1^*
1
0
I_d^*
801
ΔI_d
802
803
CONTROL-LER
ΔV_d^*
I_d
I_q^*
804
ΔI_q
805
806
CONTROL-LER
ΔV_q^*
I_q
807
I_1^*
808
ΔI_1^*
809
CONTROL-LER
$\Delta\omega_1^*$
810
I_1
812