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Optimal Unity Power Factor Control of Permanent Magnet Synchronous Motor with q-axis Field by Inverse LQ Method

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ABSTRACT

A synchronous motor (SM) with q-axis special field winding of which the q-axis field-current compensates and cancels armature reaction can be driven at unity power factor under the conditions of transient state as well as steady state. The motor operates in high efficiency in all conditions. However, in order to obtain maximum performance of the motor, it is required that the time constant of armature circuit corresponds to that of q-axis field circuit. Inverse LQ (ILQ) design method on a basis of the pole assignment is suitable for this problem: (1) The time constants of the output responses can be designed for desired specifications, (2) Relations between feedback gains and response of closed loop system are very clear and (3) Optimal solutions can be given by simple procedure of ILQ method without solving the Riccati's equation, compared to the usual LQ design method. Accordingly, the ILQ method can make the responses of armature current and q-axis field-current correspond. In this paper, it is proved by numerical simulations and experiments that the ILQ method is very effective for optimal regulator design of this plant and realizes a high-performance motor with unity power factor and high efficiency.

Key words: Permanent magnet synchronous motor, unity power factor, PWM inverter, optimal regulator, ILQ design method

1. Introduction

The motor is used as a clean mechanical-power source to a wide field such as electric vehicles, industrial machines, electrification products, information devices, equipments for welfare and environment etc. At present, the total electric energy due to the motor occupies more than 50 percent of all the electric energy in Japan, and improvement of motor efficiency is the most important subject from the view points of both energy saving and environment preservation of the earth [1].

DC motor is excellent in points of easy control and high efficiency, and has been used for long as the variable speed motor. But it is necessary to change periodically the brushes of commutator deteriorated by wear and spark. The fear of sparks is especially high in some peculiar environment, in which DC motors cannot be used, such as the presence of inflammable gas.

Many induction motors (IM) have been applied because of robust body, simple structure and low maintenance. However the IM is essentially excited from its primary to produce the main magnetic flux, and therefore has loss due to the excitation. In comparison with the IM, the permanent magnet (PM) synchronous motor (SM) is of higher efficiency, because the PM SM has no special exciter and produces the main magnetic flux with the PM itself. PM SM applications have been expanding rapidly in recent years [2].

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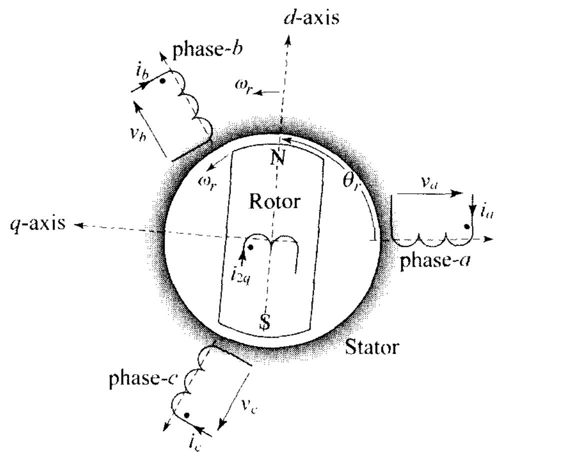


Fig. 1. PM synchronous motor with q-axis field winding

It is desirable that the SM is driven with unity power-factor in order to achieve higher efficiency and a smaller power-supply capacity. It is, however, impossible for ordinary PM SM to realize the unity power-factor in all operating conditions, because the magnetic field produced by the PM is constant and armature EMFs can not be reduced.

The PM SM with q-axis special field-winding (see Fig.1) of which the q-axis field-current cancels armature reaction, can be driven at unity power-factor under all the conditions including even transient state by making the voltage space-vector and the current space-vector on the same direction. Although this motor needs the small brushes which are simple structures compared with the commutator of the DC motor and have no flashover, it can be easily driven at unity power-factor and high efficiency in any operating condition. Moreover armature magnetomotive force (MMF) in the d-axis direction can be always null, accordingly there is no demagnetization of the PM.

Generally, the time constant of the armature circuit is not equal to that of the q-axis field circuit. But it is required that the responses of armature and q-axis field currents correspond to each other completely to achieve unity power-factor in the transient state.

Simple vector control strategy, by which time constants of the armature and q-axis field circuits in closed-loop correspond had been presented by the simulations and experiments [3][4]. But it remained a problem to solve: The power factor cannot satisfy unity by the deviation of motor parameters and the error of inverter output-voltages [4].

The optimal control has many excellent advantages such as low sensitivity to variation of parameters, robust stability etc. The solutions of optimal control can be obtained by solving the Riccati's equation as the LQ problem. Solving the Riccati's equation is complex and the relations between the weights of performance index and the responses of state variables in closed-loop are not clear. Therefore it is difficult to find the optimal gains which cause the time constants of the armature and q-axis field circuits in closed-loop to correspond.

On the other hand, the inverse LQ (ILQ) design method [5-9] is the strategy to find the optimal gains on a basis of pole assignment and has the advantages of

1) time constants of the output responses can be designed for desired specifications,
2) relations between feedback gains and responses of closed loop system are very clear, and optimal solutions can be given by simple procedure of the ILQ design method without solving the Riccati's equation.

For the reasons given above, the ILQ design method is suitable for the optimal control design in the system [10].

This paper presents a simple linearized-model for the PM SM with q-axis field winding and the optimal controller for unity instantaneous power-factor by the ILQ design method. It is proved by numerical simulations and experiments that the ILQ design method is very effective and excellent speed control with unity power-factor can be achieved.

2. Basic equations of PM SM with *q*-axis field winding

2.1 Voltage equation and linear state equation

Figure 2 shows an analysis model for PM SM with q-axis field winding. The PM can generally be analyzed by replacing it with a coil located on the d-axis region of the rotor. Using d-q transformation, the voltage equation in the rotor reference frame is given as follows:

$$\begin{bmatrix} v_{1d} \\ v_{2d} \\ v_{1q} \\ v_{2q} \end{bmatrix} = \begin{bmatrix} r_1 + L_{1d}p & M_{12d}p & -\omega_r L_{1q} & -\omega_r M_{12q} \\ M_{12d}p & r_{2d} + L_{2d}p & 0 & 0 \\ \omega_r L_{1d} & \omega_r M_{12d} & r_1 + L_{1q}p & M_{12q}p \\ 0 & 0 & M_{12q}p & r_{2q} + L_{2q}p \end{bmatrix} \begin{bmatrix} i_{1d} \\ i_{2d} \\ i_{1q} \\ i_{2q} \end{bmatrix} \tag{1}$$

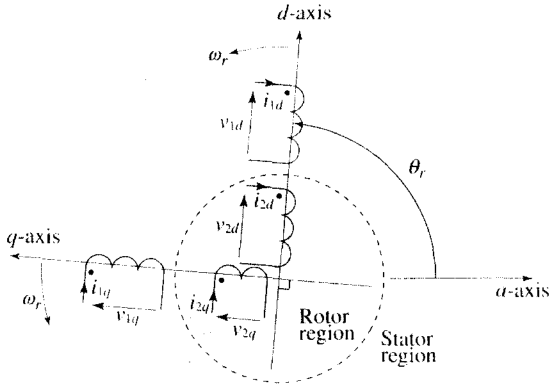


Fig. 2. Analysis model of proposed SM

where

v_{1d}, v_{1q}	d- and q-axis components of armature voltages
v_{2d}, v_{2q}	input voltages of d- and q-axis field windings
i_{1d}, i_{1q}	d- and q-axis components of armature currents
i_{2d}, i_{2q}	currents of d- and q-axis field windings
r_1	armature resistance
r_{2d}, r_{2q}	resistances of d- and q-axis field windings
L_{1d}, L_{1q}	d- and q-axis self-inductances of the armature windings
L_{2d}, L_{2q}	self-inductances of d- and q-axis field windings
M_{12d}, M_{12q}	mutual inductances between armature and d- and q-axis field- windings
ω_r	rotor speed represented in electrical angular velocity
p	differential operator ($=d/dt$).

Equation (1) is non-linear, with coupling elements for d and q axes. Equation (1) is linearized and decoupled by transforming it into a new reference frame with independent input-voltages as follows [11]-[13]:

$$\begin{bmatrix} v_{1d} \\ v_{2d} \\ v_{1q} \\ v_{2q} \end{bmatrix} = \begin{bmatrix} v_{1d}^c \\ v_{2d}^c \\ v_{1q}^c \\ v_{2q}^c \end{bmatrix} + \begin{bmatrix} 0 & 0 & -\omega_r L_{1q} & -\omega_r M_{12q} \\ 0 & 0 & 0 & 0 \\ \omega_r L_{1d} & \omega_r M_{12d} & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} i_{1d} \\ i_{2d} \\ i_{1q} \\ i_{2q} \end{bmatrix} \tag{2}$$

where v_{1d}^c, v_{1q}^c, v_{2d}^c, v_{2q}^c are the compensated input-voltages of v_{1d}, v_{1q}, v_{2d}, v_{2q} for decoupling in the new reference frame, respectively.

Substituting Eq. (2) into Eq. (1) yields

$$\begin{bmatrix} v_{1d}^c \\ v_{2d}^c \\ v_{1q}^c \\ v_{2q}^c \end{bmatrix} = \begin{bmatrix} r_1 + L_{1d}p & M_{12d}p & 0 & 0 \\ M_{12d}p & r_{2d} + L_{2d}p & 0 & 0 \\ 0 & 0 & r_1 + L_{1q}p & M_{12q}p \\ 0 & 0 & M_{12q}p & r_{2q} + L_{2q}p \end{bmatrix} \begin{bmatrix} i_{1d} \\ i_{2d} \\ i_{1q} \\ i_{2q} \end{bmatrix} \tag{3}$$

In the PM motor, the current i_{2d} can be considered constant, thus the simple state equation in the d-axis coordinate system can be obtained from the first line in Eq (3) as follows:

$$\frac{d}{dt} i_{1d} = -\frac{r_1}{L_{1d}} i_{1d} + \frac{1}{L_{1d}} v_{1d}^c \tag{4}$$

On the other hand, from the third and fourth lines in Eq. (3) the state equation in the q-axis coordinate system can be obtained as follows:

$$\frac{d}{dt}\boldsymbol{x}_q = \boldsymbol{A}_q \boldsymbol{x}_q + \boldsymbol{B}_q \boldsymbol{u}_q, \quad \boldsymbol{y}_q = \boldsymbol{C}_q \boldsymbol{x}_q \tag{5}$$

where

$\boldsymbol{x}_q = [i_{1q} \quad i_{2q}]^T$, $\boldsymbol{u}_q = [v_{1q}^c \quad v_{2q}^c]^T$, $\boldsymbol{y}_q$: Output and

$$\boldsymbol{A}_q = \begin{bmatrix} -\dfrac{r_1}{\sigma_q L_{1q}} & \dfrac{r_{2q}}{\bar{\sigma}_q M_{12q}} \\ \dfrac{r_1}{\bar{\sigma}_q M_{12q}} & -\dfrac{r_{2q}}{\sigma_q L_{2q}} \end{bmatrix}, \quad \boldsymbol{B}_q = \begin{bmatrix} \dfrac{1}{\sigma_q L_{1q}} & -\dfrac{1}{\bar{\sigma}_q M_{12q}} \\ -\dfrac{1}{\bar{\sigma}_q M_{12q}} & \dfrac{1}{\sigma_q L_{2q}} \end{bmatrix},$$

$\boldsymbol{C}_q = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$ and $[\quad]^T$: Transpose of matrix.

σ_q is leakage coefficient of q-axis coordinate system as follows:

$$\sigma_q = 1 - \frac{M_{12q}^{\ 2}}{L_{1q} L_{2q}}, \quad \bar{\sigma}_q = \frac{\sigma_q}{1 - \sigma_q} \tag{6}$$

2.2 Instantaneous power-factor and torque

Space vectors of armature voltage and current are defined as follows:

$$\vec{V}_1 = \begin{bmatrix} v_{1d} & v_{1q} \end{bmatrix}^T, \ \vec{I}_1 = \begin{bmatrix} i_{1d} & i_{1q} \end{bmatrix}^T \tag{7}$$

Figure 3 shows the space vectors of voltage and current in case of lag power- factor. We define the instantaneous power-factor as cosϕ, where ϕ is the angle between voltage and current vectors. In the steady state condition, since the space vectors correspond to the time vectors, the instantaneous power-factor is just an extrapolation of the steady state's power factor.

From the inner product by space vectors

$$\vec{V}_1 \cdot \vec{I}_1 = |\vec{V}_1||\vec{I}_1|\cos\phi = v_{1d}i_{1d} + v_{1q}i_{1q} \tag{8}$$

Hence, instantaneous power-factor can be given as

$$\cos\phi = \frac{v_{1d}i_{1d} + v_{1q}i_{1q}}{\sqrt{v_{1d}^2 + v_{1q}^2}\sqrt{i_{1d}^2 + i_{1q}^2}} \tag{9}$$

When the instantaneous power-factor is unity i.e. cosϕ becomes ± 1 (motoring= +1, regenerating= −1), the voltage space-vector and the current space-vector are on the same direction. Moreover, an indefinite set of solutions for unity power-factor derived from Eq. (9), can be obtained according to the operating conditions.

Torque τ is given by

$$\tau = \frac{P_0}{2}\left(M_{12d}i_{2d}i_{1q} - M_{12q}i_{2q}i_{1d}\right) + \frac{P_0}{2}\left(L_{1d} - L_{1q}\right)i_{1d}i_{1q} \tag{10}$$

where P_0 : number of poles.

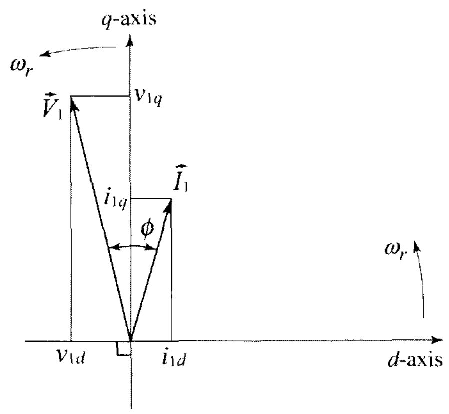


Fig. 3. Space vectors of voltage and current

And motion equation is given as follows:

$$\tau = J\frac{d}{dt}\omega_m + D\omega_m + \tau_L, \quad \omega_m = \frac{2}{P_0}\omega_r \tag{11}$$

where

J, D	inertia and friction coefficients
τ_L	load torque
ω_m	rotor speed represented in mechanical angular velocity.

Tables 1 and 2 show the rated parameters and constants of the tested motor, respectively.

Table 1. Rated parameters of tested motor

Output power	2 (kW)
No. of poles	4
Terminal voltage	220 (V)
Phase current	6.5 (A)
Frequency	60 (Hz)
Tourque	10.6 (N-m)

Table 2. Constants of tested motor

r_1 (Ω)	0.859
r_{2d} (Ω)	13.3
r_{2q} (Ω)	1.17
L_{1d}=L_{1q} (H)	0.104
L_{2d} (H)	2.73
L_{2q} (H)	0.138
M_{12d} (H)	0.49725
M_{12q} (H)	0.1035
σ_d	0.1291
σ_q	0.2537
J (kg-m^2)	0.153
D (N-m/rad/s)	0.0012

3. Unity instantaneous power-factor and torque control

As mentioned above, the condition of unity power-factor yields an indefinite set of solutions for different operating conditions. To seek a unique solution, the following conditions are assumed:

$$v_{1d} = 0 \ \text{ and } \ i_{1d} = 0 \tag{12}$$

With the condition i_{1d}=0, only the q-axis component of

MMF remains. This component can also be cancelled by the injection of the appropriate current i_{2q}. As a result, the total MMF in the q-axis and therefore the armature reaction become null. Therefore, the input voltage is decreased to its strict minimum.

Controlling the motor with $i_{1d} = 0$ has many advantages, among which, the demagnetizing effect of i_{1d} on the PM will be suppressed. Accordingly, the d-axis control system is constructed so that i_{1d} becomes zero (refer to section 4.1).

When i_{1d} satisfies the zero condition, voltage v_{1d}^c is equal to zero too. Thus, we can derive the condition for unity instantaneous power-factor from the first line in Eq. (2) as follows:

$$L_{1q} i_{1q} + M_{12q} i_{2q} = 0 \tag{13}$$

If Eq. (13) is satisfied, the voltage v_{1d} is equal to v_{1d}^c thus, becoming null, and the instantaneous power-factor is kept unity. The first and second terms in Eq. (13) represent the MMFs of currents i_{1q} and i_{2q}, respectively. It means that armature MMF is cancelled by MMF due to i_{2q}.

As a result, the control law for unity instantaneous power factor is given as:

$$i_{2q}^* = -\frac{L_{1q}}{M_{12q}} i_{1q}^* \tag{14}$$

where, i_{1q}^* and i_{2q}^* are commands of i_{1q} and i_{2q}, respectively.

When the control of unity instantaneous power-factor is carried out, actual currents i_{1q} and i_{2q} may not satisfy the condition of Eq. (13), because i_{1q}^* and i_{2q}^* in Eq. (14) are ideal target values. To achieve completely unity instantaneous power-factor even at transient, the transfer function from i_{1q}^* to i_{1q} must correspond to the one from i_{2q}^* to i_{2q}.

Since current i_{2d} is kept constant in the PM motor, by controlling $i_{1d} = 0$ the motor's torque-equation can be simplified, and expressed only as a function of i_{1q}.

$$\tau = \frac{P_0}{2} M_{12d} i_{2d} i_{1q} = P_0 \lambda i_{1q} \propto i_{1q} \tag{15}$$

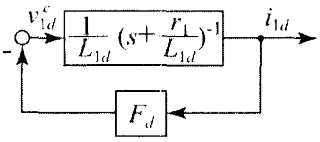


Fig. 4. Control block diagram for d-axis coordinate system

where

λ : number of flux interlinkage of armature windings.

In the section 4.2, the optimal servo-system for q-axis system is designed by the ILQ design method.

4. Design of an optimal control system

4.1 Design for *d* axis system

Figure 4 shows the control block diagram for d-axis coordinate system. The feedback gain F_d is determined by simple pole-assignment method. The state equation of d-axis coordinate system is rewritten as a unforced system:

$$\frac{d}{dt} i_d = -\frac{r_1 + F_d}{L_{1d}} i_d = -\frac{1}{T_d} i_d \tag{16}$$

The solution is an exponential response that decreases by a time constant T_d. As the gain F_d becomes larger, the current i_{1d} converges to zero faster and faster.

One-phase PWM-inverter controls the q-axis field current, operating at a sampling time $T_s = 0.5$ms. The time constant T_d is set at 4ms, which is eight times larger than T_s. This yields the gain $F_d = 25.141$. Since the armature resistance $r_1 = 0.859\ \Omega$ increases at a rate of about 10 percent per 25 degrees of temperature, this system is very robust to variation of r_1.

4.2 Optimal control design of *q*-axis coordinate system by ILQ method

Figure 5 shows the optimal control block diagram based on ILQ design method for q-axis coordinate system. In the servo-system, $\boldsymbol{y}_q^*$ and $\boldsymbol{\Sigma}$ are objective and gain adjusting parameter, respectively. $\boldsymbol{F}_q^0$, $\boldsymbol{K}_q^0$ are the basic optimal gains. $\boldsymbol{F}_q$ and $\boldsymbol{K}_q$ are the optimal feed back gains represented as

$$F_q = \Sigma F_q^0 = \begin{bmatrix} f_{11} & f_{12} \\ f_{21} & f_{22} \end{bmatrix},\ K_q = \Sigma K_q^0 = \begin{bmatrix} k_{11} & k_{12} \\ k_{21} & k_{22} \end{bmatrix} \quad (17)$$

This system can be analyzed by the following extended state equation:

$$\frac{d}{dt}x_e = A_e x_e + B_e u_e \quad (18)$$

and

$$x_e = [\Delta x_q^T \quad \Delta z_q^T]^T,\quad u_e = u_q - u_{q0},$$

$$A_e = \begin{bmatrix} A_q & 0 \\ C_q & 0 \end{bmatrix},\ B_e = \begin{bmatrix} B_q \\ 0 \end{bmatrix},\quad \Delta x_q = x_q - x_{q0},$$

$$\Delta y_q = y_q - y_q^* = \frac{d}{dt}\Delta z_q$$

where, x_{q0} and u_{q0} represent the steady state values of x_q and u_q, respectively.

The problem is then summarized to the minimization of the quadratic performance index as follows:

$$J_e = \int_0^\infty \left(x_e^T Q x_e + u_e^T R u_e\right)dt,\quad Q = C_e^T C_e,\quad R > 0 \quad (19)$$

and the optimal gains are obtained by

$$u_e = -\left[F_q \quad K_q\right]x_e = -\Sigma\left[F_q^0 \quad K_q^0\right]x_e = -K_e x_e \quad (20)$$

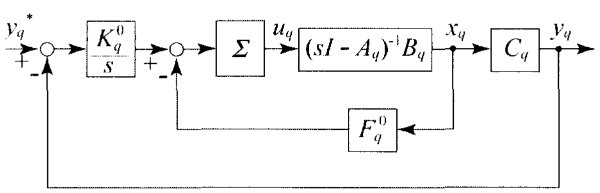


Fig. 5. Control block diagram for q-axis coordinate system

The feedback gains can be obtained by solving the Riccati's equation, but due to trial and error, it is difficult to find the solutions by which the transfer function from i_{1q}^* to i_{1q} corresponds to the one from i_{2q}^* to i_{2q}.

This system satisfies

1) controllable and observable system,
2) no zero at origin,
3) minimal-phase system,
4) the orders in difference between denominator and numerator of the plant, are unity,

so that it is possible to apply the ILQ design method, which can assign desired poles to optimal servo system (refer to an Appendix). Especially from 4), objective transfer functions are given by time-lag of first order as follows:

$$\frac{1}{1+T_i s} = \frac{1}{1 - s/s_i} \quad (i = 1, 2) \quad (21)$$

where, T_i are time constants and s_i assigned poles defined as $s_i = -1/T_i$.

Consequently, the condition for unity instantaneous power-factor is derived as $s_1 = s_2$.

In addition, necessary and sufficient conditions for the ILQ servo system are given regardless of any pole assignment as follows [9]:

1) $E = \Sigma - KB_q - (KB_q)^T > 0$,
2) $\mathrm{Re}\lambda(F) = \mathrm{Re}\lambda(A_K - GH) < 0$,
3) $\left\|H(sI - F)^{-1}G\right\|_\infty < 1$,

where Re $\lambda(X)$ denotes the real part of eigenvalue of matrix X, K is the decoupling gain-matrix, and constants are defined as $A_K := A_q - B_q K$, $G := A_q - B_q E^{-1/2}$ and $H := E^{-1/2} K A_K$.

Therefore the optimality of the ILQ servo system is always guaranteed by following condition:

$$\Sigma > \Sigma_{min} \quad (22)$$

where Σ_{min} is inferior value of gain adjusting parameter Σ and is derived from the above conditions 1) – 3).

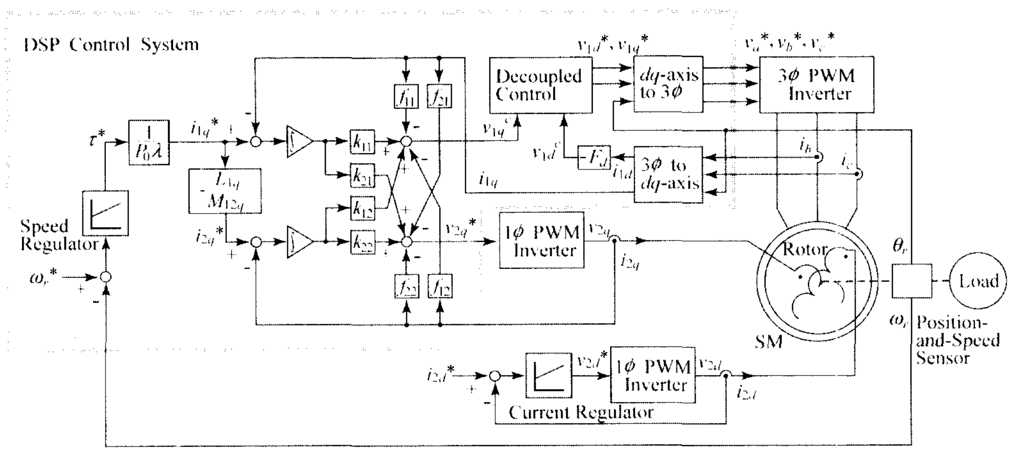


Fig. 6. Control block diagram for speed control under the instantaneous unity-power-factor control

Table 3. Gains determined by ILQ design method

gains	(s_1, s_2)= (-50, -50)	(s_1, s_2)= (-50, -75)
k_{11}	2602.9	3123.5
k_{12}	1832.3	3288.4
k_{21}	3664.7	4397.6
k_{22}	3453.8	6200.9
f_{11}	51.986	62.383
f_{12}	36.588	43.906
f_{21}	73.176	87.811
f_{22}	68.987	82.777

The ILQ optimal gains are calculated by using ILQ design tools, developed by Prof. Takao Fujii *et al.*, Graduate School of Engineering Science, Osaka University. The gains are determined by the following procedure:

1) Calculation of basic optimal gains $\boldsymbol{F}_q^0$, $\boldsymbol{K}_q^0$.
2) Determination of inferior values $\boldsymbol{\Sigma}_{min}$ of gain adjusting parameter $\boldsymbol{\Sigma}$.
3) Modification of optimal gains due to $\boldsymbol{\Sigma} > \boldsymbol{\Sigma}_{min}$ by numerical simulations.

Table 3 shows the determined optimal gains $\boldsymbol{F}_q$ and $\boldsymbol{K}_q$ in the two cases of (s_1, s_2)=(-50, -50) as compared with (s_1, s_2)=(-50, -75).

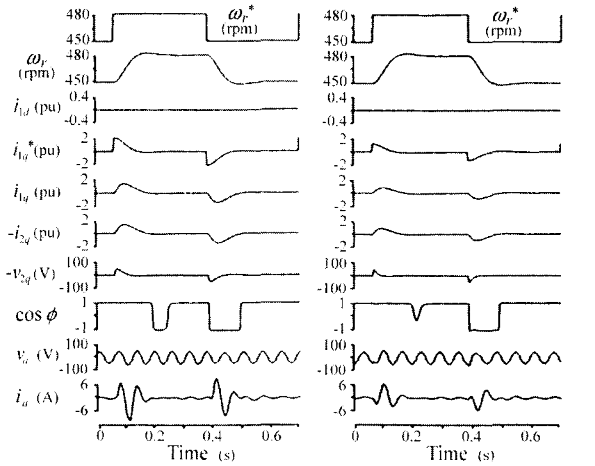
Figure 6 shows the block diagram for speed control of the PM SM with q-axis special field-winding in the unity instantaneous power-factor. armature windings and two single-phase PWM inverters control of the currents of d- and q-axis field windings, A three-phase PWM inverter supplies the power to the respectively. The current i_{2d} in the d-axis field, is kept constant by PI controller in order to imitate the PM. Control computation is carried out by using the digital signal processor (DSP: TI, TMS320C25, 40MHz).

5. Numerical simulations and experiments

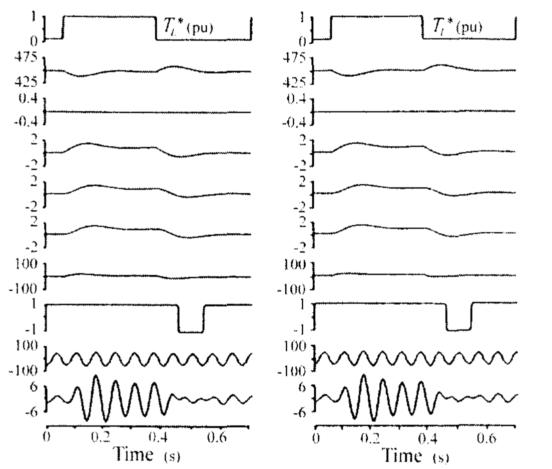
Figure 7 shows the numerical simulations. Figure 7 (a) is the step response of the speed in the case of the assigned field circuits have the same time-constants in closed loop, pole (s_1, s_2)=(-50, -50), in which the armature and q-axis and Fig. 7 (b) is in the case of (s_1, s_2)=(-50, -75), in which the different time-constants are applied on purpose to compare with Fig. 7 (a). The condition s_1=s_2 of Fig. 7 (a) leads to favorable response of unity power factor at transient state. Figures 7 (c) and (d) illustrate the step responses of load torque in the assigned poles (s_1, s_2)=(-50, -50) and (s_1, s_2)=(-50, -75), respectively. The responses on both figures are almost same even at the transient state.

Variations of 10 percent for each motor parameter did not almost cause the difference of responses. Accordingly, these simulated results have been abbreviated in this paper.

Figure 8 shows the experimental results under the same conditions in the numerical simulations. The experiments in Fig. 8 favorably correspond with the simulations in Fig. 7.

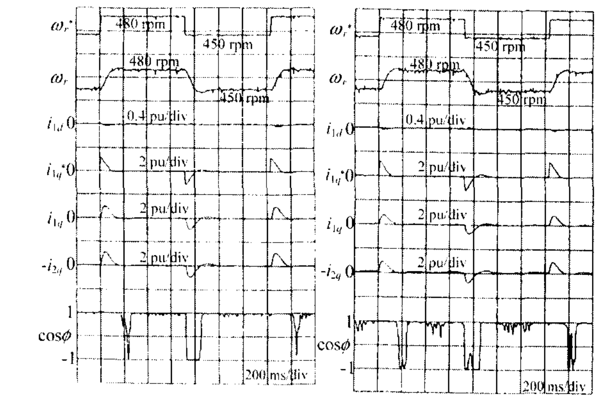


(a) step response of speed $(s_1, s_2) = (-50, -50)$ (b) step response of speed $(s_1, s_2) = (-50, -75)$

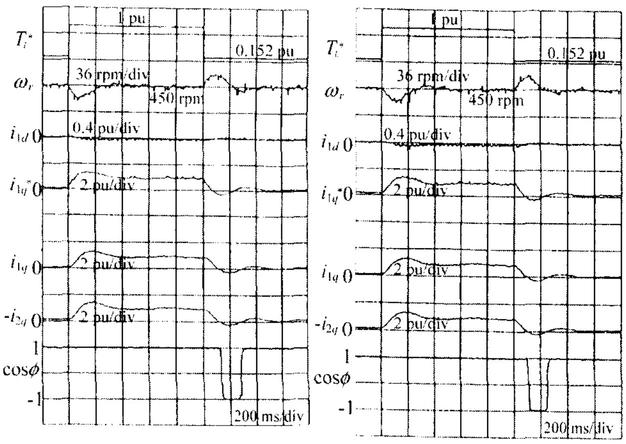


(c) step response of torque $(s_1, s_2) = (-50, -50)$ (d) step response of torque $(s_1, s_2) = (-50, -75)$

Fig. 7. Simulated results

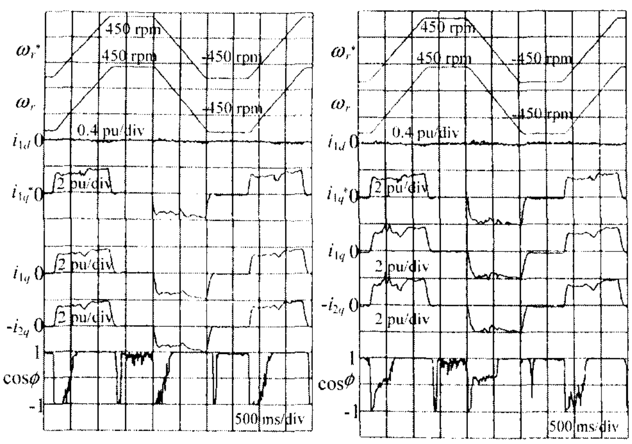


(a) step response of speed $(s_1, s_2) = (-50, -50)$ (b) step response of speed $(s_1, s_2) = (-50, -75)$



(c) step response of torque $(s_1, s_2) = (-50, -50)$ (d) step response of torque $(s_1, s_2) = (-50, -75)$

Fig. 8. Experimental results



(a) $(s_1, s_2) = (-50, -50)$ (b) $(s_1, s_2) = (-50, -75)$

Fig. 9. Four-quadrant operation

Figure 9 shows the four-quadrant operation which is composed of positive-rotation in motoring, positive-rotation in regenerating, negative-rotation in motoring and negative-rotation in regenerating. In the condition (s_1, s_2)=(-50, -50), all the operations are excellent and the favorable optimal control for the unity instantaneous power-factor is accomplished.

As a result, to achieve unity power factor even at transient state, the conditions s_1=s_2 is very important. It is proved that the ILQ design method is very effective for optimal regulator design for this plant.

6. Conclusions

Unity instantaneous power-factor control for PM SM with q-axis field winding by the ILQ design method has been presented in this paper and it has been shown that the ILQ design method is very effective. The proposed strategy enables to drive at high efficiency, and the evaluation of its effectiveness has been carried out by numerical simulations and experiments. To accomplish unity instantaneous power-factor, it is very important to make the time constants of the armature and q-axis field circuits in the closed loop correspond by assigning the system poles in the same position. The designed system is very robust to variations of motor parameters and high and precise speed-control has been achieved. Although this motor needs small brushes through which the q-axis field winding is driven by one-phase PWM inverter, it has the advantage of operating at unity power-factor, which corresponds to an optimum active power supply and therefore high efficiency. The high efficiency of this motor makes it suitable for numerous applications such as electric vehicles for city use or high-efficiency servo-motors.

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Appendix

The conditions for the ILQ design method are given as follows:

1) Controllable and observable system

Following controllable and observable conditions can be derived:

$$\mathrm{rank}\begin{bmatrix} \boldsymbol{B}_q & A_q\boldsymbol{B}_q \end{bmatrix} = 2, \quad \mathrm{rank}\begin{bmatrix} \boldsymbol{C}_q \\ \boldsymbol{C}_q A_q \end{bmatrix} = 2 \tag{A1}$$

Because, from Eq. (5) both matrices $\boldsymbol{B}_q$ and $\boldsymbol{C}_q$ are nonsingular.

2) No zero at origin

In Eq. (5), the matrix A_q is also nonsingular, that is

$$\mathrm{rank}\begin{bmatrix} A_q & \boldsymbol{B}_q \\ \boldsymbol{C}_q & \boldsymbol{0} \end{bmatrix} = 4 \tag{A2}$$

Therefore, this system has no zeros at origin.

3) Minimal-phase system

In the transfer function $\boldsymbol{G}(s) = \boldsymbol{C}_q(s\boldsymbol{I}-A_q)^{-1}\boldsymbol{B}_q$ of Eq. (5), a solution z satisfying rank$\boldsymbol{G}(z) < 2$ represents zeros of the function $\boldsymbol{G}(s)$, that is, it means that the lines of $\boldsymbol{G}(s)$ become linear dependence [9]. Where, matrix $\boldsymbol{I}$ represents 2-by-2 identity matrix. In the equation,

$$\boldsymbol{\Gamma}(z) = \begin{bmatrix} z\boldsymbol{I} - A_q & -\boldsymbol{B}_q \\ \boldsymbol{C}_q & \boldsymbol{0} \end{bmatrix} \tag{A3}$$

following condition can be formed:

$$\mathrm{rank}\,\boldsymbol{\Gamma}(z) = 4 \quad \left(\because \det \boldsymbol{\Gamma}(z) = \frac{1}{\sigma_q L_{1q} L_{2q}} \neq 0 \right) \tag{A4}$$

Therefore this system does not have unstable zeros and is minimal-phase system.

4) The orders in difference between denominator and numerator of the plant are unity

In a decoupling matrix,

$$\Lambda = \begin{bmatrix} \boldsymbol{c}_{q1} A_q^{d_1-1} \boldsymbol{B}_q \\ \boldsymbol{c}_{q2} A_q^{d_2-1} \boldsymbol{B}_q \end{bmatrix}, \quad d_i = \min_{k=1,2}\left\{ k \,\middle|\, \boldsymbol{c}_{qi} A_q^{k-1} \boldsymbol{B}_q \neq 0 \right\} \text{ and}$$

$\boldsymbol{c}_{qi}$: the i-th line vector ($i = 1, 2$) of $\boldsymbol{c}_q$, (A5)

when, $k = 1$, the following conditions are obtained as:

$$\left.\begin{aligned} \boldsymbol{c}_{q1} A_q^0 \boldsymbol{B}_q &= \begin{bmatrix} \dfrac{1}{\sigma_q L_{1q}} & -\dfrac{1}{\bar{\sigma}_q M_{12q}} \end{bmatrix} \neq 0 \\ \boldsymbol{c}_{q2} A_q^0 \boldsymbol{B}_q &= \begin{bmatrix} -\dfrac{1}{\bar{\sigma}_q M_{12q}} & \dfrac{1}{\sigma_q L_{2q}} \end{bmatrix} \neq 0 \end{aligned}\right\} \tag{A6}$$

and

$$\det \Lambda = \frac{1}{\sigma_q L_{1q} L_{2q}} \neq 0 \tag{A7}$$

Accordingly, the matrix Λ is nonsingular, thus the orders in difference between denominator and numerator of the plant become unity ($d_1 = d_2 = 1$).

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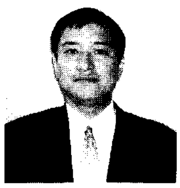

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