Stator-Flux-Oriented Control of Induction Motor Considering Iron Loss

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Abstract—Recently, research to consider the influences of iron loss has been made in the vector control of an induction motor. However, little work has been done in the area of a stator-flux-oriented control system of an induction motor. This paper investigates the effects of iron loss in the direct stator-flux-oriented control system of an induction motor, and proposes a control algorithm considering iron loss. The iron loss is modeled by equivalent iron loss resistance in parallel to the magnetizing inductance. Torque control capability is much improved and the speed estimation error for a speed-sensorless drive is reduced by the proposed control algorithm. The effectiveness of the proposed method is verified by simulation and experimental results.

Index Terms—Induction motor, iron loss, stator-flux-oriented control.

I. INTRODUCTION

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ENERALLY, in the field-oriented control of an induction machine, it is assumed that iron loss may be neglected. Unfortunately, in practice, iron loss makes the output torque different from the reference torque, and the performance is deteriorated. To solve the problems caused by iron loss, the effects of iron loss in the vector-controlled induction motor have been investigated and compensated for in the last several years and considerable effectiveness has been achieved in performance and efficiency [1]–[9].

On the other hand, the consideration of iron loss has been applied only to the rotor-flux-oriented system, and little applied to the stator-flux-oriented system, since in the rotor-flux-oriented system, no decoupling control is required, but in the stator-flux-oriented system, there exists a coupling between the torque-producing component of the stator current and the stator flux [10]. However, stator-flux-oriented control drive has recently been receiving wide attention because it is insensitive to the variations in the parameters of the machine. In the stator-flux-oriented system, the speed estimation, decoupling compensation, and estimated torque are influenced by iron loss. As a result, the accurate torque control characteristics cannot be achieved and the performance is deteriorated.

This paper investigates the effects of iron loss in the direct stator-flux-oriented control system of an induction motor, and proposes a control algorithm considering iron loss. The iron loss is modeled by equivalent iron loss resistance in parallel with the magnetizing inductance. The estimated torque, slip angular frequency, and decoupling compensation current considering iron loss are derived. It is shown that torque control capability is much improved and the speed estimation error for a speed-sensorless drive system is reduced by the proposed control algorithm. In addition, the results in the case of detuned rotor resistance and rotor leakage inductance are shown. The effectiveness of the proposed method is verified by simulation and experimental results.

II. MODEL OF INDUCTION MOTOR CONSIDERING IRON LOSS

Fig. 1 shows the equivalent circuit of an induction motor considering iron loss in the rotating $d$–$q$ reference frame. Compared with the conventional model, it is seen that the iron loss resistance is connected in parallel with the magnetizing inductance. The voltage equations are written as

\[
\begin{align*}
\dot{v}_{ds} &= R_s \dot{i}_{ds} + p \lambda_{ds} - \omega_e \lambda_{qs} \\
\dot{v}_{qs} &= R_s \dot{i}_{qs} + p \lambda_{qs} + \omega_e \lambda_{ds} \\
0 &= \omega_r \dot{i}_{dr} + p \lambda_{dr} - (\omega_e - \omega_r) \lambda_{qr} \\
0 &= \omega_r \dot{i}_{qr} + p \lambda_{qr} + (\omega_e - \omega_r) \lambda_{dr} \\
R_s \dot{\lambda}_{dm} &= p \lambda_{qm} - \omega_e \lambda_{qm} \\
R_s \dot{\lambda}_{qm} &= p \lambda_{qm} + \omega_e \lambda_{dm}
\end{align*}
\]

where the superscript “$e$” is the rotating $d$–$q$ reference frame; $p$ is the differential operator; $\dot{v}_{ds}$ and $\dot{v}_{qs}$ are the stator $d$- and $q$-axes input voltages; $R_s$ is the stator resistance; $R_r$ is the rotor
resistance; \( R_i \) is the equivalent iron loss resistance; \( \omega_e \) is the excitation angular frequency; \( \omega_r \) is the rotor angular frequency; \( \dot{\xi}_{ds} \) and \( \dot{\xi}_{qs} \) are the \( d \)- and \( q \)-axes stator currents; \( \dot{\xi}_{dr} \) and \( \dot{\xi}_{qr} \) are the \( d \)- and \( q \)-axes rotor currents; \( \dot{\xi}_{ds}^c \) and \( \dot{\xi}_{qs}^c \) are the \( d \)- and \( q \)-axes currents flowing through \( R_i \); \( \lambda_{ds} \) and \( \lambda_{qs} \) are the \( d \)- and \( q \)-axes stator fluxes; \( \lambda_{dr} \) and \( \lambda_{qr} \) are the \( d \)- and \( q \)-axes rotor fluxes; and \( \lambda_m \) and \( \lambda_{gm} \) are the \( d \)- and \( q \)-axes air-gap fluxes.

The current equations are given as

\[
\begin{align*}
\dot{\xi}_{ds} + \dot{\xi}_{dr} &= \dot{\xi}_{dm} + \dot{\xi}_{ds} \\
\dot{\xi}_{qs} + \dot{\xi}_{qr} &= \dot{\xi}_{qm} + \dot{\xi}_{qs}
\end{align*}
\] (7) (8)

where \( \dot{\xi}_{dm} \) and \( \dot{\xi}_{qm} \) are the \( d \)- and \( q \)-axes magnetizing currents.

The flux equations are given as

\[
\begin{align*}
\lambda_{ds} &= \lambda_{dm} + L_is\dot{\xi}_{ds} \\
\lambda_{qs} &= \lambda_{qm} + L_is\dot{\xi}_{qs} \\
\lambda_{dr} &= \lambda_{dm} + L_\tau\dot{\xi}_{dr} \\
\lambda_{qr} &= \lambda_{qm} + L_\tau\dot{\xi}_{qr} \\
\lambda_{dm} &= L_m\dot{\xi}_{dm} \\
\lambda_{qm} &= L_m\dot{\xi}_{qm}
\end{align*}
\] (9) (10) (11) (12) (13) (14)

where \( L_m \) is the magnetizing inductance, \( L_{ds} \) is the stator leakage inductance, and \( L_\tau \) is the rotor leakage inductance.

The torque equation is given as

\[
T_e = \frac{3}{4} P \left( \dot{\lambda}_{qr} \dot{\xi}_{dr} - \dot{\lambda}_{dr} \dot{\xi}_{qr} \right)
\] (15)

where \( P \) is the number of poles.

The consideration of iron loss in the control scheme requires the knowledge of the equivalent iron loss resistance. The equivalent iron loss resistance can be determined experimentally from a no-load test with sinusoidal supply or pulsewidth modulated (PWM) supply. The PWM voltage source is known to cause a significant increase in total iron loss, compared with a purely sinusoidal supply. However, the iron loss produced by higher order harmonics of the PWM voltage supply has no impact on the accuracy of the vector control. Accordingly, the component of iron loss that is relevant for detuning studies is only the first harmonic component, which represents the equivalent iron loss resistance in the modeling. In addition, the iron loss of the first-order harmonic is almost equal to that of sinusoidal voltage supply [1], [2].

At no-load condition, the motor is rotated lower than the synchronous speed, owing to the mechanical losses. In this case, the loss with PWM supply is the sum of iron loss, mechanical loss, and copper loss of the stator resistance. In this paper, to remove the mechanical loss from the input power, the induction motor is rotated at synchronous speed by a dc motor connected to the shaft of the induction motor. The input power at this time becomes the sum of the iron loss and copper loss of the stator resistance. The copper loss of the stator resistance is calculated by measuring the stator current and stator resistance. The measurement of the first-order harmonic of iron loss is achieved by using a power analyzer. The test results for the induction motor shown in Table I are shown in Fig. 2. It can be seen that the equivalent iron loss resistance \( R_i \) is a function of operating frequency.

### III. ESTIMATION OF IRON LOSS CURRENT

The iron loss currents \( \dot{\xi}_{ds} \) and \( \dot{\xi}_{qs} \) can be estimated by (5) and (6). In the rotating \( d-q \) reference frame, the flux is kept constant in the constant torque region, and it varies in the field-weakening region. However, the time derivative \( d\lambda_{dm}/dt \) is small enough compared with the other terms in (5) and (6), even in the field-weakening region. Therefore, the iron loss current can be written as

\[
\begin{align*}
\dot{\xi}_{ds}^c &\approx -\frac{\omega_e \dot{\lambda}_{dm}}{R_i} = -\frac{\omega_e}{R_i} \left( \dot{\lambda}_{ds} - L_is\dot{\xi}_{qs} \right) \\
\dot{\xi}_{qs}^c &\approx -\frac{\omega_e \dot{\lambda}_{qm}}{R_i} = -\frac{\omega_e}{R_i} \left( \dot{\lambda}_{qs} - L_is\dot{\xi}_{ds} \right)
\end{align*}
\] (16) (17)

where the superscript “\( \lambda \)” is the estimation value.

### IV. SPEED-SENSORLESS STATOR-FLUX-ORIENTED CONTROL SYSTEM

The control scheme of the proposed drive system is a stator-flux-oriented control system [10]. The stator-flux-oriented control of an induction motor is used more in industrial variable-speed drive systems because stator-flux estimation accuracy is dependent only on the stator resistance variation. In addition, it is insensitive to the variation in the leakage inductance of the machine. Fig. 3 shows the block diagram of a stator-flux-oriented control system considering iron loss.
The stator flux in the stationary \(d-q\) reference frame can be estimated by the integration of back EMF as

\[
\dot{\lambda}_{ds}^s = \int \left( v_{ds}^s - R_s i_{ds}^s \right) dt
\]  

(18)

where the superscript “\(s\)” is the stationary \(d-q\) reference frame.

The integration by pure integrator has drift and saturation problems. To solve these problems, a low-pass filter is commonly used instead of a pure integrator. In this paper, a cascaded low-pass filter such as equation (19) is used for exact estimation of stator flux and solving the drift and saturation problems [11]

\[
\dot{\lambda}_{ds}^s = \frac{2}{s + |\omega_e|} \left( v_{ds}^s - R_s i_{ds}^s \right).
\]  

(19)

The stator voltage is reconstructed from the inverter switching states [12]. A switching function \(S_A\) for phase \(A\) is 1 when the upper switch of phase \(A\) is ON, \(S_A = 0\) when the lower switch of phase \(A\) is ON. A similar definition is adopted for phases \(B\) and \(C\). The stator voltages can be given in terms of switching states and the dc-link voltage as follows:

\[
v_{ds}^s = \frac{V_{dc}}{3} \left( 2S_A - SB - SC \right),
\]  

(20)

\[
v_{qs}^s = \frac{V_{dc}}{\sqrt{3}} (SB - SC).
\]  

(21)

The transformation angle \(\hat{\theta}_e\) is given as (22), and the synchronous angular frequency is estimated by the derivative of \(\hat{\theta}_e\) as (23)

\[
\dot{\lambda}_{ds}^s = v_{ds}^s - R_s i_{ds}^s
\]  

(24)

\[
\dot{i}_{ds}^s = \frac{L_n \dot{\theta}_e}{L_m} \left( \lambda_{ds}^s - \sigma L_s i_{ds}^s \right) + L_n \dot{\lambda}_{ds}^s
\]  

(25)

where \(\sigma = 1 - L_2/L_1\) is the total leakage factor, \(L_s\) is the stator self-inductance, and \(L_n\) is the rotor self-inductance.

Substituting (24) and (25) into (4), the steady-state form of the slip angular frequency can be given as

\[
\dot{\omega}_s = \frac{L_s i_{qs}^s - L_m i_{qs}^s}{T_r \left( \dot{\lambda}_{ds}^s - \sigma L_s \dot{i}_{ds}^s \right) + L_n \dot{\lambda}_{ds}^s}
\]  

(26)

where \(T_r = L_n/R_e\) is the rotor time constant.

The rotor angular frequency can be estimated as

\[
\omega_r = \dot{\omega}_e - \dot{\omega}_s
\]  

(27)

Rearranging (15) as to stator current and stator flux, the torque equation can be written as

\[
T_e = \frac{3}{4} P \left[ \lambda_{ds}^s \left( v_{qs}^s - \dot{\lambda}_{qs}^s \right) - \lambda_{qs}^s \left( v_{ds}^s - \dot{\lambda}_{ds}^s \right) - L_s \left( \dot{\lambda}_{ds}^s i_{qs}^s - \dot{\lambda}_{qs}^s i_{ds}^s \right) \right].
\]  

(28)

Equation (28) can be rewritten as (29) since \(\lambda_{qs}^s = 0\) in the rotating \(d-q\) reference frame

\[
T_e = \frac{3}{4} P \left[ \lambda_{ds}^s \left( v_{qs}^s - \dot{\lambda}_{qs}^s \right) - L_s \left( \dot{\lambda}_{ds}^s i_{qs}^s - \dot{\lambda}_{qs}^s i_{ds}^s \right) \right].
\]  

(29)

From (29), the \(q\)-axis current command can be given as

\[
\dot{\omega}_r = \frac{4}{3B T_e} \left( \dot{\lambda}_{ds}^s - L_s i_{qs}^s \right) \left( \lambda_{ds}^s - L_s i_{qs}^s \right)
\]  

(30)
where

\[
\alpha = \frac{\dot{\gamma}qs (\lambda_{ds} - L_{ds} \dot{\gamma}_{ds})}{\lambda_{qs} - L_{ds} \dot{\gamma}_{qs}}.
\]

“\( \dot{\gamma} \)” is the reference value, and \((K_p + K_f/s)\) is the transfer function of the speed regulator (proportional plus integral (PI) type).

Substituting (24) and (25) into (3), (31) can be derived with \(\lambda_{qs} = 0\) in the rotating \(d-q\) reference frame

\[
(1 + T_p \dot{\gamma}qs) \lambda_{qs} = (1 + \sigma T_r \dot{\gamma}qs) L_{ds} \dot{\gamma}_{qs} - \sigma L_{ds} T_r \dot{\gamma}qs + \lambda_{qs} \dot{\gamma}_{ds} - L_{ds} \dot{\gamma}_{ds} - L_{m} \dot{\gamma}_{ds} - \frac{L_{m}}{R_r} (p L_{rs} \dot{\gamma}_{ds} - \omega_{rs} L_{rs} \dot{\gamma}_{qs}).
\]

Equation (31) shows that the \(q\)-axis stator current \(\dot{\gamma}_{qs}\) and the stator flux \(\lambda_{qs}\) are coupled. It means that any change in \(\dot{\gamma}_{ds}\) without changing \(\dot{\gamma}_{qs}\) will cause a transient in the stator flux. From Fig. 3, \(\dot{\gamma}_{ds}\) can be written as

\[
\dot{\gamma}_{ds} = \left( K_p + \frac{K_f}{s} \right) (\lambda_{ds} - \hat{\lambda}_{qs}) + i_{dq} \tag{32}
\]

where \((K_p + K_f/s)\) is the transfer function of the flux regulator (PI type).

Substituting (32) into (31), the steady-state form of the decoupling compensation current can be derived as

\[
i_{dq} = \sigma T_r \dot{\gamma}qs \frac{\dot{\gamma}_{qs}}{L_{rs}} + \frac{L_{m}}{L_{rs}} \left( \dot{\gamma}_{ds} - \frac{\dot{\gamma}_{ds}}{R_r} \right). \tag{33}
\]

V. SIMULATION RESULTS

The proposed control scheme was studied by simulation with the drive system shown in Fig. 3. The induction motor parameters are shown in Table I.

Fig. 4 shows torque responses. The motor is accelerated and decelerated between 2500–3500 r/min by torque control without load. When the motor is accelerated, the torque reference is \(3\) N-m. When the motor is decelerated, the torque reference is \(-3\) N-m. When the motor is being accelerated, if the motor speed becomes 3500 r/min, the torque reference is changed from \(3\) N-m to \(-3\) N-m. The time when the torque reference is changed from \(3\) to \(-3\) N-m is determined by real motor speed because the estimated speed has an estimation error. If it is determined by estimated speed, motor speed is not varied between 2500–3500 r/min exactly. Fig. 4(a) is the case not considering iron loss. In this simulation, the estimated torque not considering iron loss is controlled. It is observed that the real torque is lower than the reference torque. As a result, the deceleration is faster than the acceleration. The iron loss in the motoring area can be regarded as unrecognized power that has to be supplied to the machine by the inverter, so that the actual output torque is lower than the reference torque. In contrast to this, iron loss is covered from the converted mechanical power in the braking area, so that the actual torque is higher than the reference torque. Fig. 4(b) is the case considering iron loss. In this simulation, the estimated torque considering iron loss is controlled. It is seen that the real torque is nearly equal to the reference torque by the proposed method.

Fig. 5 shows the results of detuned operation. The condition for torque regulation is identical to that of Fig. 4, and iron loss is considered. Fig. 5(a) is the case of detuned rotor leakage inductance. Fig. 5(b) is the case of detuned rotor resistance.
are only rotor parameter detunings between Figs. 5 and 4(b). From the comparison between Figs. 5 and 4(b), it can be seen that the system is not deteriorated by the rotor parameter detuning. Rotor parameter detuning is not compensated by the proposed method. The system is insensitive to the rotor parameter detuning because the stator-flux-oriented control drive is insensitive to the variations in the parameters of the machine [10], [13]. Stator-flux estimation accuracy is dependent only on the stator-resistance variation. The voltage drop in stator resistance should be considered in the low-speed region because back EMF is small. As a result of that, in the low-speed region, the stator-flux-oriented system is mainly affected by stator resistance. However, the voltage drop in stator resistance can be negligible as the speed increases because back EMF is high enough. In Fig. 5, speed is varied between 2500–3500 r/min. In this speed region, the system is not affected by stator resistance, even though stator resistance is detuned. Consequently, Figs. 5 and 4 mean that, in the stator-flux-oriented system, iron loss is mainly responsible for the detuning effects in torque response rather than rotor parameters.

Fig. 6 shows the speed estimation error between the real rotor speed and the estimated rotor speed in the speed-sensorless stator-flux-oriented control drive. It is seen that speed estimation error is reduced by the proposed method. It is also observed that the speed estimation error occurs under the proposed scheme with the consideration of the iron loss. Stator flux is precisely estimated by (19). However, the estimation is not perfect. Equation (19) is transformed into the sampled-data model using the first-forward difference approximation. The sampled-data model has a modeling error which, in turn, produces an error in the stator flux estimation. In addition, (23) is also transformed into the sampled-data model using the first-forward difference approximation. The sampled-data model has a modeling error which, in turn, produces an error in the rotor speed estimation. In Fig. 6, estimated stator flux is controlled and the current control period is 125 µs. If real stator flux is controlled and the current control period is changed from 125 to 50 µs, the speed estimation error is less than 1 r/min at 3500 r/min.

VI. EXPERIMENTAL RESULTS

In order to verify the proposed scheme, the control system is implemented in the software of digital signal processor (DSP) TMS320C31. The inverter input voltage is \( V_{dc} = 325 \) V. The switching frequency is 4 kHz. The current control period is \( T_C = 125 \mu s \). The speed control period and the flux control period are each \( T_s = 1.25 \) ms. An encoder of 1024 pulses per revolution is mounted on the rotor. Load torque is produced by a dynamometer. The stator currents are detected through Hall-type sensors. The stator currents are sampled and held at every sampling instant, then A/D converted with 3.5-µs conversion time. The motor is a 2.2-kW three-phase induction motor as shown in Table I.

Fig. 7 shows torque responses. The motor is accelerated and decelerated between 2500–3500 r/min by torque control without load. When the motor is accelerated, the torque reference is 3 N·m. When the motor is decelerated, the torque reference is −3 N·m. To investigate the torque control capability, real motor speed is used because the estimated speed has an estimation error. Fig. 7(a) shows the results of the case without considering iron loss. In this experiment, the estimated torque not considering iron loss is controlled. The estimated torque shown in Fig. 7(a) is the estimated torque considering iron loss. It is seen that the deceleration is faster than the acceleration. Fig. 7(b) shows the results of the case considering iron loss. In this experiment, the estimated torque not considering iron loss is controlled. The estimated torque shown in Fig. 7(b) is the estimated torque
considering iron loss. It can be seen that the acceleration time is nearly equal to the deceleration time. It implies that the exact torque estimation is achieved by the proposed method.

Fig. 8 shows the applied load torque and torque estimation error between the applied load torque and the estimated torque. The applied load torque is the actual measured torque by the use of torque transducer. In this experiment, the estimated speed is controlled. The speed reference is 2500 r/min and the load torque of 5 N·m is applied to the induction motor by a dynamometer, then, after about 4 s, the load is removed from the induction motor. In Fig. 8(a), it can be seen that notable torque estimation error is produced because iron loss is not considered. In Fig. 8(b), torque estimation error is much reduced by the proposed method.

Fig. 9 shows the speed and speed estimation error in the case not considering iron loss. In this experiment, estimated speed is controlled and measured speed is monitored. The speed reference is changed from 2000—2750—3500 r/min under no-load condition. It is seen that the speed estimation error is about 23 r/min at 3500 r/min. It is also observed that the speed estimation error is increased in transient. The estimated speed has an error because the sampled-data model of the estimated speed has a modeling error when it is transformed into the sampled-data model using the first-forward difference approximation. The error included in the estimated speed is removed by the use of a low-pass filter. The speed estimation error in transient is increased because of the delay of the estimated speed by the use of the low-pass filter.

Fig. 10 shows the speed and speed estimation error in the case considering iron loss. The conditions for the experiment are identical to those of Fig. 9. It is seen that the speed estimation error is about 13 r/min at 3500 r/min. About 10 r/min of speed estimation error is reduced at 3500 r/min by the proposed method compared with Fig. 9.

Fig. 11 shows the speed and torque characteristics in the speed-sensorless drive system by the proposed method. The speed reference is 100 r/min. The load torque of 4 N·m is applied, then, after about 8 s, the load is removed from the system. Fig. 11(a) shows that the speed is reduced by about 20 r/min when the load torque is applied to the system, then
the speed is converged to the speed reference. Fig. 11(d) shows that the flux estimation is stable when the load torque is applied and removed. In the low-speed region, speed estimation error due to iron loss appears to be very small because iron loss at low operating frequency is very small. The speed estimation error is mainly affected by the detuning of stator resistance in the low-speed region because back EMF is very low.

VII. Conclusion

The stator flux-oriented control system of an induction motor considering iron loss has been proposed in this paper. The torque equation, decoupling compensation current, and slip angular frequency considering iron loss were derived. From the simulation and experimental results, it was shown that the torque control capability was much improved and the speed estimation error for the speed-sensorless stator-flux-oriented control system was reduced. It was also shown that the iron loss is mainly responsible for the detuning in the torque response. Further, it was seen that the torque was exactly estimated by the proposed method.

REFERENCES


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