Sensorless Control of Interior Permanent Magnet Synchronous Motor: An Overview and Design Study

Y. Kano, N. Matsui

Abstract — This paper provides an overview of the progress in sensorless permanent magnet synchronous machine (PMSM) control techniques in Japan. In the first stage, the techniques extracted rotor position information by back-EMF zero-crossing detection techniques. Since around 2000, model-based position-sensorless control techniques have been evolved and are widely used in home appliances as well as general industrial applications. Simultaneously, sensorless control based on the magnetic saliency has been developed and commercialized. However, owing to the considerable flux saturation in the rotors, the accuracy of saliency-based position detection depends on the design and structure of the motor. In future motor drives, motor design optimization for saliency-based sensorless control is expected to be the foremost issue for machines. In this paper, an example of design optimizations of application-specific IPMSMs for saliency-based sensorless control is introduced.

Index Terms—Sensorless control, permanent magnet machines

I. INTRODUCTION

In general, a high-performance ac variable-speed drive requires speed/position sensors. A vector-controlled induction motor (IM) basically uses a speed sensor and a permanent magnet (PM) brushless dc motor requires a position sensor for commutation and current control. However, the electrical connectors and signal wires from the sensor to the controller reduce the mechanical robustness of the overall system. In addition, the rotor position sensor has several disadvantages in terms of drive cost, machine size, and noise immunity. To overcome these problems, position-sensorless permanent magnet synchronous machine (PMSM) drive techniques have been studied since the early 1980s [1-4], and some of them have been commercialized and applied to industrial fields [5,6]. Two approaches are being utilized in industrial and automotive products in the marketplace: one is based on the back electromotive force (back-EMF) estimation and another is based on the position-dependency of the winding inductances in the low speed ranges including zero speed.

Interior PM synchronous machines (IPMSMs) are intentionally salient from the standpoint of their torque production, which makes them suitable for sensorless control. However, owing to the considerable flux saturation in the rotors, the accuracy of saliency-based position detection depends on the design and structure of the motor. The design of IPMSMs for sensorless control is receiving increasing attention and is becoming a field of great activity. Published work in this field has mainly focused on the design of the rotor including choice of rotor topology [7-9]. Other papers have shown that the type of winding configuration (i.e. concentrated or distributed) [10] and the slot shape [11] can have a large effect on the sensorless capability of a machine. However, there is little published work on how to design a motor for sensorless control. It is therefore very important to continue the research and development of guidelines for sensorless-oriented designs.

This paper provides an overview of the progress in sensorless PMSM control techniques in Japan. Subsequently, an example of design optimizations of application-specific IPMSM for saliency-based sensorless control is introduced.

II. PROGRESS IN SENSORLESS DRIVE TECHNIQUES IN JAPAN

A. Household Appliances.

In household appliances, especially air-conditioners and refrigerators, PMSMs have emerged as the standard ac motors for variable-speed drives owing to their superior power density, high efficiency, and reliability [12]. The first air conditioner employing an inverter-driven PMSM was developed in 1982 in Japan [13]. Since 1998, inverter-driven room air conditioners have accounted for more than 90% of the air conditioners sold in Japan. A similar trend become spread over the world for considering recent global energy and environmental problems [14].

In these applications, the motor is built in a completely sealed compressor, for which mechanical shaft position sensing is difficult to apply because of the low sensor reliability in high-temperature and refrigerant environments. Therefore, position-sensorless control of the motor is indispensable. Hence, many researchers have proposed various sensorless control approaches for PMSM drives.
techniques in air conditioners. The mass production of PMSM for compressor motors began in 1982. The position-sensorless trapezoidal current drive, i.e., 120° commutation drive, was the first to be applied to a compressor motor [3]. The basic concept underlying this method is to obtain the rotor position information indirectly from the instantaneous magnitude of the back-EMF, which is a function of the rotor position. In a brushless dc motor (BLDCM), the unexcited winding can be used as a sensor because two out of three windings are excited at a time. The rotor position is obtained from the detected zero-crossing points of the back-EMF in the un-excited winding.

In 1996, an IPMSM with a rare-earth magnet was employed to provide reluctance torque and flux-weakening control, yielding high efficiency and power factor improvement simultaneously [15,16]. IPMSM with sinusoidal flux distributions excites three windings at a time; therefore, back-EMF zero-crossing detection cannot be used. The position-sensorless method using the neutral point signal of IPMSM was employed to overcome this problem; thus, the sinusoidal current (180° commutation) drive mode became available [16].

Moreover, since around 2000, model-based position-sensorless sinusoidal current drives have emerged as the standard compressor drives in air conditioners [17]. The basis of the sensorless control algorithm is to exploit the difference between the detected actual state variables and the estimated state variables, which are calculated from an equivalent motor model of the IPMSM in the controller. The rotor position is estimated by measuring only the stator voltages and currents. As a result, a high-efficiency control strategy, namely maximum torque per ampere (MTPA) control, has been widely used in this application to achieve high torque output and minimization of copper losses.

Generally, the robustness against parameter variations always becomes a problem in the model-based position sensorless drives. To solve this disadvantage, several approaches have been proposed and found wide applications, for example, by using the extended Kalman filter [18] and the model reference adaptive system (MRAS) [19].

On the other hand, low-price household appliances require a low-cost control system without using a position sensor and high-performance microcontroller unit (MCU). The simplified vector-control method, which is suitable for implementation with using a typical low-cost MCU, has been adopted in these applications [20]. Fig. 2 shows the simplified vector-control method. The control structure is simplified by eliminating the automatic speed regulator (ASR) and automatic current regulator (ACR) from the conventional vector controller. The output voltage references $v_{dc}$ and $v_{qc}$ are determined by feed-forward-like calculation using the motor parameters, rotation speed command, and current references. Therefore, the drive performance of this method under the steady state condition is almost the same as that of the conventional vector control. A practical estimation equation for the rotor position has been proposed, and the phase-locked loop control approach has been adopted to drive IPMSM.

Further simplification has been achieved by removing the current sensors from the system, instead of detecting the motor current [21]. This system detects the motor current from the direct current using the A/D converter of a microcomputer. When the direct current cannot be detected, the motor current is reverse-converted from the current value of the dq-axis. At present, sensorless control without position and current sensors is widely adopted in home appliances such as refrigerators, cordless cleaners, and washing machines.

**B. Industrial Drive Applications**

In the 1980s, there was a clear difference in use between the induction motors and PM motors, the induction motors has been exclusively used as the variable speed ac motors for air conditioners, pumps, lifts, etc., while the surface-mounted PM synchronous motor (SPMSM) has been exclusively used in servo applications. Since the 1990s, the IPM synchronous motor (IPMSM) has been used in many industrial drives and servo applications. This is because it has the following advantages over IM and SPMSM.

1. The torque density of IPMSM is significantly higher than that of general-purpose IM (30% higher in the range of several tens of kW, near 1800-rpm operation). In addition, the efficiency of IPMSM is 5% higher than that of general-purpose IM.
2. The stainless-steel can, which is usually used in SPMSM to prevent the magnet from scattering, can be eliminated in IPMSM. Since the rotor is made of laminated core and can is eliminated, significant iron loss reduction has been realized.
3. Square-shaped PM can be used for IPMSM; it decreases the magnet cost compared to the arc-shaped magnet.
4. Owing to the inherent saliency of IPMSM, initial position estimation and stable sensorless control at very low speed can be realized.

Since the early 2000s, several Japanese motor drive companies have developed general industrial IPMSM drives in the power range of 0.4–500 kW [22-25]. Here, a typical example is introduced.
Fig. 3 shows the sectional view of Yaskawa’s IPMSM [22]. It can be considered as a typical example having the structural advantages stated in points 2 and 3 above. The back-EMF disturbance observer-based sensorless control was employed at medium and high speeds, and at low and zero speeds, the high-frequency signal injection method was employed. Fig. 4 shows the control block diagram of the signal injection technique. The high-frequency voltage \( v_{\text{yaj}} \) is superimposed onto the control voltage, and the high frequency component of the stator current vector \( i_q^* \) is obtained from the measured current \( i_q \) through the band-pass filter (BPF). The estimated rotor position is detected using the field-oriented controller, as shown in Fig. 4(b). This commercialized sensorless control system has the following features:

- Initial position detection error within ±5°.
- Stable sensorless drive in all speed and torque ranges.
- Smooth restart after free-run condition.

III. DESIGN STUDY OF SALIENCY-BASED SENSORLESS IPMSM DRIVE FOR TRACTION APPLICATION

At present, several automobile companies are commercializing hybrid electric vehicles (HEVs) and electric vehicles (EVs) in which an IPMSM is the main traction motor because of its high power density and high efficiency. To achieve the best performance of the machine, an IPMSM drive requires a relatively expensive position transducer for machine vector control. However, the rotor position sensor has several disadvantages in terms of drive cost, machine size, and noise immunity. Thus, various sensorless IPMSM drive schemes have been developed for automobiles [26], but there is not a decisive idea as to whether or not to use a sensorless drives of traction motor in its entirety because of the automobile special demand for extremely high safety. It is possible to use position-sensorless control as a backup for a position sensor when sensor failure occurs. Saliency-based position-sensorless control methods are used at zero and low speeds [27-29]. However, there are certain problems and limitations when saliency is used for position-sensorless control under loaded conditions, especially under the heavy loaded condition. Saturation within the machine causes the magnitude of the saliency to decrease up to the point where position-sensorless control is no longer possible. Although saliency-based position-sensorless control methods are only used at zero and low speeds, it is important to investigate their limitations. Standstill or pull-away torque is very important for HEVs and EVs; thus, the ability to produce high torque at zero speed is important in the case of a sensorless drive.

In this section, design methods that can achieve the maximum torque under the sensorless drive while retaining the required torque-speed capability are studied. Finite element analysis (FEA) is first used to evaluate both the self-sensing capability and the torque-speed capability. Finally, experiments on these two aspects are conducted to verify the results.

\[
\begin{bmatrix}
\Delta I_{\phi} \\
\Delta I_{\alpha}
\end{bmatrix} = T^{-1}(\Delta \theta) \begin{bmatrix}
\Delta I_{\alpha} \\
\Delta I_{\phi}
\end{bmatrix} = \pm \Delta T \cdot V_s 
\begin{bmatrix}
\cos(\Delta \theta) & \sin(\Delta \theta)
\end{bmatrix} \\
\begin{bmatrix}
L_{\alpha} \\
L_{\phi}
\end{bmatrix} \begin{bmatrix}
L_{\phi} \\
L_{\alpha}
\end{bmatrix} \\
\end{bmatrix} = \begin{bmatrix}
L_{\phi} \\
L_{\alpha}
\end{bmatrix} \sin(2 \Delta \theta)
\]

\[
L_{\phi} = \frac{L_{\phi} - L_{\alpha}}{2}
\]

where \( I_{\phi}, I_{\alpha} \) are the high-frequency components of the d-q

A. Position Estimation Method for Wide-Torque-Range Operation at Low Speed

The authors have already proposed the signal-injection-based IPM traction drive for wide-torque-range operation at low speed [30]. Fig. 5 shows the proposed sensorless drive system. In the target HEV application, the controller has only a current feedback loop. At low speed, the IPMSM operates along the MTPA trajectory. The proposed control scheme shown in Fig. 5 switches from the harmonic-reactive-power-based method [31] to the “ellipse area method [30]”. In the low-torque region with \( L_{\phi} > L_{\alpha} \), the switch SW in Fig. 5 is connected to terminal 0, and the rotor position is estimated using the conventional harmonic-reactive-power-based technique. The injected voltage is shown in Fig. 6. In this technique, sampling of the current and updating of the pulse-width modulation (PWM) is done twice in a PWM switching period. The sampling period is half of the PWM switching period. The difference between successively sampled currents (\( \Delta I_{\phi}, \Delta I_{\alpha} \) in the estimated rotor reference frame includes rotor position information given by
axes currents, respectively, and $L_{dh}, L_{qh}$ are the $d$- and $q$-axis incremental self-inductances, respectively. The harmonic reactive power can be calculated as

$$Q_h = v_a \Delta i_{ah} - v_a \Delta i_{nh} = v_a \Delta i_{nh}$$

$$= -\Delta T \frac{V_s}{L_{dh} L_{nh}} \sin(2\Delta \theta). \quad (3)$$

Note that the harmonic reactive power is proportional to the sine function of twice the rotor position estimation error $\Delta \theta$. If the harmonic reactive power is controlled to be zero, the error in the estimated rotor position converges to zero.

To support stable sensorless control, the controller uses the estimated rotor position correction based on an error in the harmonic reactive power. When $L_{dh}$ is lower than $L_{qh}$, as is generally the case in conventional IPMSMs, the gradient of the magnitude of $Q_h$ versus $\Delta \theta$ characteristics will be negative for $-20^\circ < \Delta \theta < 20^\circ$. Hence, the estimated rotor position $\theta_e$ can be corrected by

$$\theta_e(n) = \theta_e(n-1) + T_c \omega_c(n) = \theta_e(n-1) + T_c K_\theta \Delta Q_h(n) \quad (4)$$

where $K_\theta$ is the correction factor and $T_c$ is the control interval. The position estimation error $\Delta \theta = \pm 20^\circ$ corresponds to the maximum permissible position error in the target application.

It is well known that the conventional harmonic-reactive-power-based method does not work well in the high-torque region with $L_{dh} > L_{qh}$, because the gradient of the magnitude of $Q_h$ versus $\Delta \theta$ characteristics become positive for $-20^\circ < \Delta \theta < 20^\circ$.

To overcome this problem, the proposed control scheme shown in Fig. 5 switches from the harmonic-reactive-power-based method to the “ellipse area method” proposed in [30]. The “ellipse area” means the area of high-frequency current trajectories as the response to the high-frequency rotating voltage. Fig.7 shows the measured high-frequency current trajectories of a 12-pole test motor as a consequence of injected rotating voltage vector at motor speed $N=100$ r/min.

The amplitude of the rotating voltage vector and its frequency were 10 V and 500Hz, respectively. In the measurements, the current lead angle $\beta$ was changed from 0° to 60° under 150% and 300% of the rated currents.

It is clear in Fig. 7 that the variation of high-frequency current trajectories with $\beta$ is small under 150% rated current. In contrast, under 300% rated current, the areas of elliptical current trajectories $S$ change dramatically with the current lead angle $\beta$. The reason for this is that the dynamic inductances vary during the current lead angle owing to magnetic saturation. The decrease of the current lead angle corresponds to the decrease of the negative $d$-axis current. Thus, the smaller current lead angle provides heavy magnetic saturation. As a consequence, the gradient of the ellipse area $S$ versus the current lead angle characteristics is negative.

Based on these phenomena, the sensorless position detection algorithm is developed. In the proposed algorithm, for simplicity, the value of the product of the two current vector amplitudes $|I_{h,q}|$ and $|I_{h,d}|$ is used as a substitute for the ellipse area $S$. Fig. 8 shows the relations between the applied voltage vector and the current responses $I_{h,q}$ and $I_{h,d}$.

Fig. 8. Relations between applied voltage vector and current responses $I_{h,q}$ and $I_{h,d}$.
where $\theta_m$ is the cross-saturation angle induced by the cross-saturation effect. The value of the product of the two current vector amplitudes $|I_h|\alpha$ and $|I_h|\delta$ can be approximately calculated using Eq. (5) as follows:

$$
S \equiv \Delta I_x \cdot \Delta I_y = \left( \frac{\Delta T \cdot V_x}{L_{sh} L_{ph} - L_{qph}^2} \right)^2 \\
L_{sh} = \frac{L_{sh}^2 + 2I_{sh}^2 (L_{sh}^2 + L_{qsh}^2) \cos 2(2 \Delta \theta + \theta_m)}{(L_{sh}^2 + L_{qsh}^2) / 8 + I_{sh}^2 (L_{sh}^2 + L_{qsh}^2) \cos 2(2 \Delta \theta + \theta_m)}
$$

(7)

(8)

Fig. 10 shows the calculated ellipse area $S$ vs. $\Delta \theta$ characteristics of the test motor. In the simulation, the amplitude of the injected voltage and its frequency were 80 V and 5 kHz, respectively. It is clear that the ellipse area $S$ changes drastically with the position estimation error $\Delta \theta$ in the high-torque region with $L_s > 100$. As a result, if the detected ellipse area $S$ is controlled to be $S(\Delta \theta = 0)$, the rotor position error will be zero. In addition, the estimated rotor position $\theta_e$ can be corrected by

$$
\theta_e(n) = \theta_e(n-1) + T \cdot K_{th} (S(n) - S'(\Delta \theta = 0'))
$$

(9)

where $K_{th}$ is the correction factor and $S'(\Delta \theta = 0')$ is the reference value of $S$ under MTPA operation.

B. Design Restrictions and Requirements for Target HEV Drive Application

The design restrictions and target specifications for HEV applications are summarized in Table I. Most of values in the table are determined using an IPMSM installed in 2-ton-class lightweight tracked HEVs sold in Japan [32]. From the standpoint of compatibility with motor assembly, the stator dimension and winding configuration of the machines are set to be the same as those of the existing IPMSM for EV application. As for the motor size constraints, the stator outer diameter $D_s$ and stack length $L_s$ are 400 mm and 48 mm, respectively. Sintered neodymium-iron-boron ($\text{NdFeB}$) magnets were selected because of their high remnant flux density.

As for the target specifications, the maximum torque, power, and speed are set to 350 Nm, 36 kW, and 3,850 rpm, respectively. The motor must satisfy the maximum torque of 350 Nm under the saliency-based sensorless drive in the shaded area shown in Fig. 11. Moreover, the motor should operate with higher efficiency at the frequent operating points shown as No. 1, 2 and 3 in the figure.

C. Design Parameters

Fig. 12 shows a sectional view of the IPM motor for the design study. The motor has one interior magnet per pole. Further, it has 12-pole, 72-slot construction and distributed windings. The number of turns of armature winding per phase is 10. Because the stator dimension and winding configuration are set to be the same as those of the existing IPMSM, the rotor design is the main focus of this study. The important rotor design parameters are the rotor tooth opening $\theta_t$, magnet length, and depth of the embedded magnet, $d_e$.

The cost of NdFeB magnets has risen sharply in recent years; hence, the magnet length is fixed at 8.7 mm under the considerations of demagnetization and cost. Thus, the rotor tooth opening $\theta_t$ and the depth of the embedded magnet, $d_e$, are treated as the design parameters.

In the following design study, a commercial FEA package (JMAG-Studio ver.10.0, released by JSOL Corporation as a 2D-FEA electromagnetic solver) is used.
D. Design Procedures

Sensorless wide-torque-range operation at low speed requires a changeover between two techniques, i.e., the conventional harmonic-reactive-power-based method and the ellipse area method. Therefore, the transition region between the two techniques should be increased by as much as possible for a stable sensorless drive. In the controller, the switching scheme for transition between the two methods is shown in Fig. 13. Because the hysteresis width is set to ±10 Nm, at least 20 Nm of the transition region is required. In this design, allowing for a margin of 10 Nm, the required transition region will be more than 30 Nm. The transition region is calculated in the following steps:

1. Calculate both the harmonic reactive power $Q_h$ and the ellipse area $S$ versus $\Delta \Theta$ characteristics considering the cross-coupling and space harmonic effects;
2. Obtain both the gradients of magnitude of $Q_h$ and $S$ versus $\Delta \Theta$ characteristics in the range of -20° to 20° for increasing load torque;
3. Determine the zero-crossing point for the gradient versus load-torque characteristics as shown in Fig. 14 for both estimation techniques;
4. The transition region $\Delta T_{\text{sd}}$ for which both estimation techniques exhibit reasonable performance is given by

$$\Delta T_{\text{sd}} = T_0 - T_5$$

where $T_0$, $T_5$ are the torque values at the zero-crossing point shown in Fig. 14.

The design procedure is as follows:

1. The combination of the rotor tooth opening $\theta_r$ and the depth of embedded magnet, $d_r$, is determined to fulfill the condition $\Delta T_{\text{sd}} > 30$ Nm.
2. The optimum combination of $\theta_r$ and $d_r$ should be determined for increasing the motor efficiency at the frequent operating points.
3. The flux-barrier arrangement should be designed to fulfill the required withstand voltage of the switching device, i.e., 530 V_{dc-p}.

E. Design Study

Fig. 15(a) shows the calculated transition region $\Delta T_{\text{sd}}$ for changing the two design parameters, namely $\theta_r$ and $d_r$. Fig. 15(b) shows the maximum torque under the harmonic-reactive-power-based sensorless drive and (c) shows the lower torque limit under the ellipse-area-based sensorless drive.
It can be seen that the characteristics of the transition region are strongly dependent on the maximum torque capability under the harmonic-reactive-power-based sensorless drive. In general, as the rotor tooth opening decreases, the q-axis inductance also decreases, and the maximum torque under the harmonic-reactive-power-based sensorless drive will be smaller. However, the decrease in the rotor tooth opening from 78° to 36° gives satisfactory results, as shown in Fig. 15(b). The reason is that lower rotor iron saturation is occurred when the q-axis inductances are low.

In a previous study [28], the maximum torque was found to decrease with increasing \( d \)-regardless of the rotor tooth opening. However, the maximum torque increases with \( d \) (see Fig. 15(b)) when the rotor tooth opening increases from 24° to 54°. According to some numerical-analysis-based investigations using (3), a larger \( d \) produces a decrease in the rotor-position-dependent fluctuation of the cross-coupling inductance \( L_{dR} \). As a result, the increase in \( d \) yields satisfactory results at the maximum torque capability.

Fig. 15(a) shows four design candidates that fulfill the requirement of the transition region \( \Delta T_{el} > 30 \text{ Nm} \), namely “R30D11”, “R36D9”, “R42D7”, and “R42D11”; the number after R indicates the value of \( \theta_r \) and the number after D indicates the value of \( d \).

In Fig. 11, three frequent operating points of the motor in an urban-traffic driving situation are indicated as No. 1 to No. 3. The motor efficiency at these operating points should be as high as possible because it plays an important role in improving the fuel consumption of vehicles. Motor efficiency is calculated by 2D-FEA considering the copper losses in the armature winding, the iron losses in all the laminated cores, and the eddy current loss in the permanent magnet.

Fig. 16 shows the calculated motor efficiencies of the four selected design candidates at the frequent operating points at a motor temperature of 100°C. From Fig. 16, it can be seen that there are no significant differences among the four candidates in terms of efficiency. The lower magnet volume results in a low back-EMF constant and low PM cost. Accordingly, R42D7 can be selected as the optimal solution. Finally, the triangular flux barrier is applied and designed to decrease the harmonic content of the back-EMF waveforms.

IV. EXPERIMENTAL VERIFICATION

A. Prototype Motor Specifications and Experimental Setup

According to the optimum design results, a prototype of 12-pole, 72-slot IPMSM was constructed. Fig. 17 shows the measured dimensions and an exterior photograph of the prototype motor.

Fig. 18 shows the experimental control system configuration. The induction motor (Meidensha, FC-95, 160 kW) was used for the load machine. The digital signal processor TMS320C6713 controller performed the sensorless algorithm. The PWM switching frequency was 5kHz, and the control loop cycle was 200 \( \mu \text{s} \). The amplitude of the injected voltage and its frequency were 80 V and 5 kHz, respectively.

The motor shaft was equipped with an encoder (2,000 ppr) for monitoring the actual rotor position. The practical load changes in the target HEV drive applications were realized by the vector-controlled load induction motor. The torque transducer and its amplifier (SS-100 and Ts-3100, Ono Sokki) achieved accurate torque measurements.

B. Induced Voltage Waveforms

The open-circuit-induced terminal voltage waveforms were measured under a constant speed of 3,850 rpm as shown in
The position error \( \Delta \theta \) increases temporarily, immediately after changing the position estimation method at torque command \( \tau^* = 200 \text{ Nm} \), but after that, it converges to zero. The maximum error is within 13.3°. The controller can fulfill the desired acceleration.

Fig. 22 shows the position-sensorless characteristics of the test motor when the load torque changes from 0 Nm to 350 Nm for 0.117 s at 100 rpm. These load changes are the standard requirements in the target application. The position error \( \Delta \theta \) increases temporarily, immediately after a load change, but it converges to a steady-state level without delay. The steady-state position error under a load of 350 Nm is approximately within ±9°.

V. CONCLUSION

In this paper, we presented an overview of the progress in sensorless PMSM control techniques in Japan. The application of sensorless PMSM drives has expanded to various fields in the power range of 0.4–500 kW. In future motor drives, sensorless-oriented design optimization will be the foremost issue for machines. As one of the solutions, this paper introduced a design method of sensorless-oriented IPMSM. This design method achieves not only stable sensorless drive capability but also torque-speed capability.

VI. ACKNOWLEDGMENT

Authors would like to express their sincere gratitude to their research colleagues for the continuous cooperation.

VII. REFERENCES


VIII. BIOGRAPHIES

Yoshiasaki Kano was born in Aichi, Japan, on Feb. 15, 1977. He received the B.S. degree in electrical engineering from Tokyo University of Agriculture and Technology, Tokyo, in 1999, and M.S. and Ph.D. degrees in electrical and computer engineering from Nagoya Institute of Technology, Nagoya, in 2001 and 2004, respectively. From 2004 to 2007, he worked with the Nagoya Industrial Science Research Institute, Aichi. In 2007, he joined the Nagoya Institute of Technology, Nagoya, as a Research Associate. In 2009, he joined Toyota National College of Technology, Aichi, as a Research Associate in the Department of Information and Computer Engineering. In 2013, he became an Assistant Professor. Since 2016, he has been with Daido University, Aichi, where he is an Associate Professor with the Department of Electrical and Electronic Engineering. His current research interests include designs of application-oriented electrical machines. Dr. Kano is a member of the IEEE and IEEJ.

Nobuyuki Matsui was born in Wakayama, Japan, on May 7, 1943. He received the B.S. degree in electrical engineering from Tokyo University of Agriculture and Technology, Tokyo, in 1966 and 1968, respectively, and the Ph.D. degree from Tokyo Institute of Technology, Tokyo, Japan, in 1976. Since 1968, he has been a Professor with the Department of Electrical and Computer Engineering, NITech. As a Professor, he has been engaged in research and education on power electronics and motion control. From April 2000 to October 2002, he was the Vice President of NITech. From January 2004 to March 2010, he was the President of NITech. From 2010, he has been granted the title of emeritus professor in NITech.

Prof. Matsui is a Life Fellow, IEEE and a Fellow of the Institute of Electrical Engineers of Japan. He received the Best Paper Award at IEEE IECION in 2002, the Best Transactions Paper Award from the IEEE Industrial Electronics Society in 2004, and the Outstanding Achievement Award from the IEEE Industry Applications Society in 2005.