Proposal and Preliminary Experimental Verification of Electrically Reversal Magnetic Pole Type Variable Magnetic Flux PM Motor

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This paper presents a variable magnetic flux PM motor in which space harmonic power is utilized for the magnetic flux weakening, automatically. The stator has a toroidally-concentrated winding structure, and the torque generation surfaces are composed of three air-gaps, which are single radial-gaps and double axial-gaps. The radial-gap rotor is a consist-magnetized PM rotor, and the axial-gap rotor is a self-excited wound-field rotor. These rotors are coaxially fastened with opposite magnetic pole position. The magnetomotive force of axial-gap rotor can automatically retrieve space harmonic power, which is inevitably generated by a concentrated winding structure. A mechanical design of the prototype is revealed, and the operation principle of the automated-magnetic flux weakening is clarified through the mathematical approach. Then, the operation principle of the electrified reversal magnetic pole is experimentally demonstrated, and the effects of the proposed variable magnetic flux technique are verified using the prototype machine in terms of the adjustable speed drive characteristics and the variable magnetic flux range.

Keywords: variable magnetic flux PM motor, self-excitation, space harmonics, toroidally-concentrated winding stator, three-dimensional magnetic flux path

1. Introduction

Recently, there has been active development of technology for 48-V mild hybrid (Mild HEV) systems, mainly in Europe. 48-V systems, compared to high voltage full HEV systems, are inferior in fuel efficiency improvement, but in terms of vehicle weight they represent small, lightweight, low-cost HEV systems for lightweight B-segment or smaller compact cars, and are likely to yield cost benefits and improved fuel efficiency. ISG systems combining claw-pole motors and inverters capable of variable magnetic flux are the most popular for efficiency. ISG systems combining claw-pole motors and inverters capable of variable magnetic flux are the most popular for high. Typical contemporary studies offer a memory motor type that varies the magnetic force of a PMSM magnet, an adjustable rotor skew angle type a variable magnetic field type with adjusting the amount of magnetization on an iron pole of consequent pole, and a variable magnetic flux type utilizing the leakage flux. In the case of , driving over a wide variable magnetic flux range is possible, making possible an optimal rotor magnetic field flux at each driving point, while magnetization and demagnetization are performed during driving by superimposing pulsed current on armature current, so it is necessary to have an inverter with a larger capacity than necessary to control instantaneous torque ripple and motor output. In the case of , a lock mechanism is required to prevent relative rotation of an external actuator or rotor in order to mechanically adjust the rotor skew angle. In the case of , a field coil and DC/DC converter are required in order to generate a static magnetic field, and in principle it is difficult to use the reluctance torque. In the case of , passive variable magnetic flux is possible with a simple construction. However, the problem is the narrow variable magnetic flux range. Furthermore, in Ref. (11), the torque density is low, and in Ref. (12), an actuator is required to drive the end plate for the short-circuited magnetic path.

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Proposal and Preliminary Experimental Verification (Masahiro Aoyama et al.)

The authors have already taken these problems into account and proposed a PM-motor for passively realizing variable magnetic flux that does not require a pulse current and any actuator, and forms an electromagnetic flux $\Psi_{\text{coil}}(\omega)$ with automatic polarity inversion (pole relative electrical angle 180 degrees) of a permanent magnet's magnetic flux as rotational speed increases. The proposed motor utilizes an asynchronous rotating magnetic field, and by utilizing a field current obtained from the electromagnetic induction principle of an inductor and diode rectification, achieves self-excited variable magnetic flux.

In this paper, the variable magnetic flux principle of the proposed motor are explained from a mathematical viewpoint. Next, the prototype for verification of the principle is revealed, and identifies actual operating principles and the effects of variable magnetic flux.

2. Structure of Proposed Motor and Variable Magnetic Flux Principle

2.1 Structure of Proposed Motor

A PM-rotor is arranged on the single radial gap of a toroidally-concentrated winding stator, and a self-excited wound field rotor (SE-WF rotor) is arranged on a double axial gap. The magnetic pole of the PM-rotor has a constant-magnetized permanent magnet, but the magnetic pole of the SE-WF rotor is a passive variable magnetic flux magnetic pole self-excited by a diode rectifier circuit utilizing the second-order space harmonics inevitably generated by the concentrated winding structure as a field energy source. Since this technique is based upon Faraday's law, the amount of field changes depending on the rotational speed and the armature magnetomotive force. Here, as described in Ref. (14), the second-order space harmonic generated by the concentrated winding structure are in reverse phase to the fundamental wave, and thus become a third harmonic wave when observed from above the fundamental wave rotating reference frame. Because the toroidally-concentrated winding stator core has a three-dimensional magnetic path, so a soft magnetic composite (SMC) is used for its core material. As shown in Fig. 3, it is possible to generate three-dimensional magnetic flux by applying reverse-direction current to opposing in-phase coils with an interposed salient pole. Figure 4 shows the rotor structure. Magnetic steel sheet is used for the PM-rotor core, and the SMC is used for the SE-WF rotor for a three-dimensional magnetic path. Each rotor is mechanically fastened to an output shaft and rotates with synchronous speed. As shown in Fig. 4(a), induction coils (I-coils) and field coils (F-coils) wound flatwise on the salient pole of SE-WF rotor are stacked axially, and wired to diodes arranged on the rotor back yoke in Fig. 4(d) in the rectifier circuit shown in Figs. 4(b) and (c). As for the second-order space harmonic, there are many interlinks near the gap, so the I-coils are arranged on the gap surface side, and F-coils are arranged on the salient pole base side of the SE-WF rotor.

2.2 Variable Magnetic Flux Principle

Figure 5 shows the magnetic flux vectors of second-order space harmonic superimposing the armature magnetomotive force of the toroidal stator. By interlinking the second-order space harmonic on the static coordinates to the I-coils, induced electromotive force is generated, and through the diode rectification of this induced electromotive force to obtain a field current, a magnetic pole is organized via the self-excitation. The induced electromotive force of an I-coil is based upon Faraday’s law, so the permanent magnet of the SE-WF rotor is intensified as rotational speed increases. As a result, since the magnetic flux vector and magnetic poles of the PM-rotor are arranged in reverse, the magnetic flux vectors of the SE-WF rotor interlink with the armature winding in opposition,
so it is possible to decrease the armature interlink magnetic flux as rotational speed increases, and as a result it is possible to automatically suppress line voltage. On the other hand, in extremely low speed areas, the induced electromotive force due to the space harmonics is wasted as Joule loss in SE-WF rotor winding resistance, and no reversed magnetic poles due to the space harmonics exist as Joule loss in SE-WF rotor winding resistance. Thus, it is possible to decrease the armature interlink magnetic flux in high-speed areas and achieve high torque with a multi-gap structure with playing as reluctance axial-gap rotor due to its saliency.

For the above understanding, mathematically described below. It is possible to express the equation for dq-axis voltage between the SE-WF rotor and stator disregarding the PM-rotor as in Eq. (1) below.

\[
\begin{align*}
\begin{bmatrix}
    v_{sd} \\
    v_{sq} \\
    v_{rd} \\
    v_{rq}
\end{bmatrix} &=
\begin{bmatrix}
    R_s & 0 & 0 & 0 \\
    0 & R_s & 0 & 0 \\
    0 & 0 & R_r & 0 \\
    0 & 0 & 0 & R_r
\end{bmatrix}
\begin{bmatrix}
    i_{sd} \\
    i_{sq} \\
    i_{rd} \\
    i_{rq}
\end{bmatrix} \\
&+ \left[ \begin{array}{cccc}
    L_{sd}(t) & 0 & 0 & 0 \\
    0 & L_{sq}(t) & 0 & 0 \\
    0 & 0 & L_{rd}(t) & 0 \\
    0 & 0 & 0 & L_{rq}(t)
\end{array} \right]
\begin{bmatrix}
    \omega L_{sd}(t) & -\omega L_{sq}(t) & pM_{sdzrd} & -\omega M_{rdzrd} \\
\omega L_{sq}(t) & pL_{sq}(t) & \omega M_{sqzrd} & pM_{qrzrd} \\
0 & pM_{rdzrd} & -\omega L_{rd}(t) & pL_{rd}(t) \\
0 & 0 & pM_{qrzrd} & -\omega L_{rq}(t)
\end{array} \right]
\begin{bmatrix}
    i_{sd} \\
    i_{sq} \\
    i_{rd} \\
    i_{rq}
\end{bmatrix} \\
&+ \begin{bmatrix}
    pL_{sd}(t) \\
    pL_{sq}(t) \\
0 \\
0
\end{bmatrix}
\begin{bmatrix}
    \omega L_{sd}(t) & -\omega L_{sq}(t) & pM_{sdzrd} & -\omega M_{rdzrd} \\
\omega L_{sq}(t) & pL_{sq}(t) & \omega M_{sqzrd} & pM_{qrzrd} \\
0 & pM_{rdzrd} & -\omega L_{rd}(t) & pL_{rd}(t) \\
0 & 0 & pM_{qrzrd} & -\omega L_{rq}(t)
\end{array} \right]
\end{align*}
\]

(1)

Here, \(v_{sd}, v_{sq}, v_{rd}, \) and \(v_{rq}\) are the respective voltages of a stator and rotor on a dq-axis. \(i_{sd}, i_{sq}, i_{rd}, \) and \(i_{rq}\) are the respective currents of a stator and rotor on the dq-axis; \(R_s, R_r, L_{sd}(t), L_{sq}(t), L_{rd}(t), \) and \(L_{rq}(t)\) are the respective DC components of stator and rotor inductance on the dq-axis; \(M\) is the mutual inductance between the rotor and stator; \(p\) is a differential operator; and \(\omega\) is angular velocity. \(L_{sd}(t), L_{sq}(t), L_{rd}(t), L_{rq}(t)\) are the respective sums of DC components and high-frequency components of stator and rotor inductance on a dq-axis. In the case of the proposed motor, when an approximate expression of \(L_{sd}(t)\) and \(L_{sq}(t)\) is consist of only the static inductance term and the dynamic third harmonic inductance term, it may be expressed as in Eqs. (2) and (3).

\[
\begin{align*}
L_{sd}(t) & \approx L_{sd0} + L_{sda} \cos(3\omega t - \gamma) \quad \text{(2)} \\
L_{sq}(t) & \approx L_{sq0} + L_{sqa} \cos(3\omega t - \gamma - \pi/2) \quad \text{(3)}
\end{align*}
\]

Here, \(L_{sda}\) and \(L_{sqa}\) are ripple amplitudes for high-frequency inductance, and \(\gamma\) is a high-frequency inductance phase. Consequently, the term of the product with the differential operator is as in Eqs. (4) and (5).

\[
\begin{align*}
pL_{sd}(t) &= \frac{d}{dt} \{ L_{sd0} + L_{sda} \cos(3\omega t - \gamma) \} \\
&= -3\omega L_{sda} \sin(3\omega t - \gamma) \quad \text{(4)} \\
pL_{sq}(t) &= -3\omega L_{sqa} \sin(3\omega t - \gamma - \pi/2) \quad \text{(5)}
\end{align*}
\]

Asynchronous armature magnetic flux attributable to high-frequency inductance per Eqs. (4) and (5) is an asynchronous rotating magnetic field (dq-qh axis coordinates) rotating in the third harmonic of a fundamental wave synchronously rotating axis (dq-q axis coordinates). Here, a rotor winding rectifier circuit is a full-bridge rectifier circuit, but for the purpose of simplifying the drawing, a half-wave rectifier circuit is depicted in Fig. 6. The angular velocity electromotive force term attributable to space harmonics generated in (4) and (5) is as shown in Fig. 6(a), and is present on a third harmonic asynchronous rotating reference frame relative to the fundamental wave synchronously rotating rotor windings. Thus, induced electromotive force is produced by magnetic coupling due not only to the pLsd(t)iqd term but the pLsq(t)iqd term. When this happens, as shown in Fig. 6(b), and the rotor salient pole is assumed to be a d-axis reference, in addition to rotor salient pole permeance distribution, when the rotor salient pole is self-excited, the d-q coordinate and dq-qh coordinate magnetic coupling (magnetic resistance) changes
according to the positional relationship of the rotor field magnetic pole vector and the armature magnetic flux vector. As shown in Fig. 7, the above may be confirmed through the electromagnetic field analysis from the results of determining changes in the magnetic flux of the rotor salient pole (d-axis) when the armature current phase is set to 0 degrees and 90 degrees. Since a forced field arises when the armature magnetic flux vector is facing in the same direction as the rotor field magnetic pole vector (current phase 0 degrees), as for the rotor permeance distribution $\rho_i(\theta)$, the permeance distribution attributable to the field magnetic pole becomes dominant, and turns into a cos $\theta$ distribution in a section of a single period of the electrical angle. $\theta$ is the spatial phase difference (electrical angle) of the d-axis and dh-axis. Accordingly, the dh-qh coordinates rotate in the third harmonic relative to the d-q coordinates, and the coefficient of magnetic coupling attributable to the magnetic pole changes by a cos $3\theta$ function. On the other hand, when the armature magnetic flux vector is facing orthogonal to the rotor field magnetic pole vector (current phase 90 degrees), as for the rotor permeance distribution, the percentage of permeance distribution attributable to the salient pole increases, and becomes a cos $2\theta$ distribution in a section of a single period of the electrical angle. Since the dh-qh coordinates rotate in the third harmonic relative to the d-q coordinates, the coefficient of magnetic coupling attributable to the salient pole changes by cos $6\theta$. That is, as for the rotor permeance distribution, in addition to the static permeance term, the permeance distribution attributable to the magnetic pole and salient pole overlap, and become the “total” for rotor permeance distribution in Fig. 7(b). These percentages change depending upon the armature current phase, and may be expressed by Eq. (6).

$$\rho_i(\theta) = \rho_{r0} + \rho_{ps} \cos 3\theta + \rho_{sat} \cos 6\theta$$

When the above is mathematically modeled, the mutual inductance $M_{rdad}$ between the rotor and stator in Eq. (1) contributes to the transfer of the self-excited field energy source, and is expressed by Eq. (7).

$$pM_{rdad} = \frac{N_{is}}{N_i} \left[ \kappa_d pL_{ad}(t) i_{ad} + \kappa_q pL_{aq}(t) i_{aq} \right] \rho_i(\theta)$$

Here, $N_{is}$ and $N_i$ are the number of turns in the I-coils and armature coils, and $\kappa_d$ and $\kappa_q$ are the respective coupling coefficients of the high-frequency magnetic flux between the d-dh axis and between the q-qh axis. $\rho_{r0}$, $\rho_{ps}$, and $\rho_{sat}$ are respectively the static content terms for rotor permeance distribution, and the dynamic ripple amplitudes of the permeance distribution attributable to the magnetic pole and the salient pole, wherein each permeance distribution coefficient changes depending upon the current phase $\delta$. Therefore, the rotor voltage equation in Eq. (8) holds.

$$v_{rd} = -pM_{rdad}i_{ad} = R_r i_{rd} + L_{rd0}\phi_{rd}$$

Here, a coil is not wound around the q-axis, so $M_{raq} = 0$, and induced electromotive force is generated only on the d-axis according to space harmonics. In addition, this closely resembles $L_{rd}(t) = L_{rd0}$. For simplification, when $\gamma = 0$ is assumed for Eqs. (4) and (5) to rearrange Eq. (8), the result is Eq. (9).

$$R_r i_{rd} + L_{rd0}\phi_{rd} = -3\omega \frac{N_{id}}{N_i} \left\{ \kappa_d L_{rad} i_{ad} \sin 3\omega t - \kappa_q L_{raq} i_{aq} \cos 3\omega t \right\} \rho_i(\theta)$$

The Laplace transform of Eq. (9) is Eq. (10), and comparing variables to solve it results in Eq. (11).

$$\left( R_r + sL_{rad} \right) i_{ad}(s) = -3\omega \frac{N_{id}}{N_i} \rho_i(\theta) \left\{ \kappa_d L_{rad} \left( \frac{3\omega}{s^2 + (3\omega)^2} \right) \right\} - \kappa_q L_{raq} i_{aq} \cos 3\omega t \rho_i(\theta)$$

$$i_{ad} = \frac{-3\omega \frac{N_{id}}{N_i} \rho_i(\theta)}{Z} \left\{ \kappa_d L_{rad} (e^{-\tau \phi} \sin + \sin(3\omega t - \phi)) \right\} - \left\{ - \kappa_q L_{raq} \left( e^{-\tau \phi} \cos \phi + \cos(3\omega t - \phi) \right) \right\}$$

$$Z = \sqrt{R_r^2 + (3\omega L_{rad})^2}$$

$$\cos \phi = \frac{R_r}{Z} \sin \phi = \frac{3\omega L_{rad}}{Z} \phi, \quad \tau = \frac{L_{rad}}{R_r}$$

Here, Eq. (11) expresses only the induction current generated by space harmonics interlinking with rotor winding when a rotor winding is short-circuit connected. Actually, field current is self-excited by the diode rectifier circuit shown in Fig. 4(c), that is, the field current $i_{rd}$ in Eq. (14) is obtained with a neutral point clamp type rectifier circuit, and is applied to the F-coil, obtaining the field magnetic flux $\Psi_{coil}$ in Eq. (15) so a magnetic pole is formed on the rotor. Equation (15) obtains the average value of the field current and disregards the field magnetic flux ripple.

$$i_{rd} = i_{rd} \left( 0 \leq t < \frac{T_r}{2} \right) + i'_{rd} \left( \frac{T_r}{2} \leq t < T_r \right)$$

$$\Psi_{pole} = N_r L_{rd0} \text{average}(i_{rd})$$

Here, the effect of the rotor permeance distribution in Fig. 7 is received, and the peak value is different on the positive and negative sides of the induced current. The rotor rectifier circuit of the proposed motor is not a full-bridge rectifier circuit using a diode bridge. Since a field current is obtained by the neutral point clamp type rectifier circuit shown in Fig. 4(c), when full-bridge rectification is mathematically modeled with absolute values, errors occur. Accordingly, as for $i'_{rd}$ in Eq. (14), the third harmonic wave interlinks with the rotor winding rectifier circuit in Fig. 4(c), so the first term on the right-hand side is considered forward direction induced current, the second term on the right-hand...
The waveform in Eq. (11) is defined as a waveform with a 60 degree shift in the electrical angle over time in order to model full-bridge rectification. In addition, $T_i$ is the period of the induced current. When the voltage equation in Eq. (1) is expanded with the addition of field magnetic flux $\Psi_{e-coil}$ that can be obtained by self-excitation phenomenon and the magnetic flux $\Psi_{mag}$ of the magnet of the radial gap PM-rotor, it becomes as in Eq. (16).

$$\begin{bmatrix}
  v_{sd} \\
  v_{sq} \\
  v_{rd} \\
  v_{rq}
\end{bmatrix} = \begin{bmatrix}
  R_s & 0 & 0 & 0 \\
  0 & R_s & 0 & 0 \\
  0 & 0 & R_p & 0 \\
  0 & 0 & 0 & R_p
\end{bmatrix} \begin{bmatrix}
  i_{sd} \\
  i_{sq} \\
  i_{rd} \\
  i_{rq}
\end{bmatrix} + \begin{bmatrix}
  L_{sd} & 0 & 0 & 0 \\
  0 & L_{sq} & 0 & 0 \\
  0 & 0 & L_{rd} & 0 \\
  0 & 0 & 0 & L_{rq}
\end{bmatrix} \begin{bmatrix}
  i_{sd} \\
  i_{sq} \\
  i_{rd} \\
  i_{rq}
\end{bmatrix} + p \begin{bmatrix}
  L_{sd}(t) & -\omega L_{sq}(t) & pM_{srd} & -\omega M_{sdr} \\
  \omega L_{sq}(t) & L_{sq}(t) & 0 & L_{sd}(t) \\
  0 & 0 & pL_{rd}(t) & 0 \\
  0 & 0 & 0 & pL_{rq}(t)
\end{bmatrix} \begin{bmatrix}
  i_{sd} \\
  i_{sq} \\
  i_{rd} \\
  i_{rq}
\end{bmatrix} + \omega(\psi_{mag} - \psi_{e-coil})...
$$

Here, as shown in Fig. 2(a), $\Psi_{mag}$ and $\Psi_{e-coil}$ have reversed signs due to the positional relationship of the reversed magnetic pole. In the third term on the right-hand side in Eq. (16), the time derivative terms (p$L_{sd}(t)$, p$L_{sq}(t)$) for d-axis and q-axis inductance go from transformer electromotive force terms to speed electromotive force terms. Accordingly, torque $T$ is determined by the third term on the right-hand side of Eq. (16), the speed electromotive force term in the fourth term, and the cross product of the armature current, and is expressed by Eq. (17).

$$T = \frac{P_p}{f_p} \begin{bmatrix}
  -3L_{sd} \sin(3\omega t - \gamma) \\
  L_{sd}(t) - 3L_{sq} \sin(3\omega t - \gamma - \pi/2) \\
  \theta_{mag} - \theta_{e-coil} \\
  0
\end{bmatrix} \begin{bmatrix}
  i_{sd} \\
  i_{sq} \\
  i_{rd} \\
  i_{rq}
\end{bmatrix} = (L_{sd0} - L_{sq})i_{sd}i_{sq} + (\theta_{mag} - \psi_{e-coil})i_{sq} + L_{sq}(\cos(3\omega t - \gamma) - L_{rd} \cos(3\omega t - \gamma - \pi/2)) i_{sd}i_{sq} - 3(L_{sd} \sin(3\omega t - \gamma) + L_{sq} \sin(3\omega t - \gamma - \pi/2)) i_{rd}i_{sq}$$

Here, $P_p$ is the number of pole pairs, and the third and fourth terms on the right-hand side in Eq. (17) represent torque ripple. Thus, it is found that the second-order space harmonic inevitably generated due to the concentrated winding structure from the fourth term on the right-hand side in Eq. (16) are utilized as an energy source for reversed magnetic pole formation, and it is possible to automatically reduce the armature interlink magnetic flux as rotational speed increases. Furthermore, it can be confirmed that the magnetic torque ($\psi_{mag}i_{sq}$) from the second term on the right-hand side of Eq. (17) is reduced by the electromagnetic torque ($\psi_{e-coil}i_{sq}$) due to the inverted magnetic pole position.

3. Construction and Main Specifications of Verification Model

The core material is used an SMC (Höganäs AB Somaloy) for the three-dimensional magnetic path. The stator core in Fig. 8(a) is constructed split into 12 salient poles and yoke segments in the circumferential direction and two segments in the axial direction, such that the weight density of the core piece when molded from powdered magnetic granules with a 100 ton press reaches 7.4 to 7.5 g/cm³. Similarly, the rotor core also assumes a split form as shown in Fig. 8(b). By segmentally constructing the stator core, it is possible to attach a toroidal coil pre-formed by edgewise winding to yoke segments via an insulator (made of PPS), and assemble them toroidally, thereby realizing an improved coil space factor. As shown in Fig. 5, the second-order space harmonic that serve as the field energy source for the wound-field magnetic pole of the SE-WF rotor mainly interlink in the vicinity of the rotor tooth tip gaps, so the I-coils for generating induced electromotive force are arranged on the gap surface side. For holding of the SE-WF rotor core, as shown in Fig. 8(b), the stainless steel (SUS303) holding component is used as a mechanical strength measure in the axial direction. The rotor core is fixed with holding component through a bolt hole provided on the core, wherein the holding component was mechanically coupled to the shaft. In addition, in order to hold the rotor winding, it is similarly covered with an SUS303 protective component which prevented dropout of the winding. Then, a wiring board is placed on the back surface of the rotor core holding component, and then covered it on top with a brass cover providing both wiring protection and a rotor balance correction function. For axial direction gap control, the spatial position of PM-rotor and axial rotor on one side is fixed on the convex portions provided on the shaft, and on the other side positioned on the flange sections of the holding members of the SE-WF rotor. In addition, for the convenience of workability of balance correction, the SE-WF rotor
positioned on the flange section is structurally designed with clearance fit with the shaft, tightening in the axial direction with a locknut. Therefore, it becomes possible after balance correction of the rotor assembly to remove the clearance fit side of the SE-rotor, assemble a toroidal stator, and thus complete the motor. Table 1 shows the main specifications of the verification model. For the core material, the SMC was H"ogan"as AB Somaloy 700-3P, the magnetic steel sheet was Nippon Steel & Sumitomo Metal Corporation 30DH, and the PM rotor magnet was Shin-Etsu Chemical Co., Ltd. N39UH ($B_r = 1.22$ T, $H_{cb} = 965.7$ kA/m@293 K).

4. Actual Prototype Model

Figure 10 shows an actual prototype machine. When molding and shaping the SMC cores, it is difficult to achieve dimensional accuracy with edge sections and core mating surfaces, so they are machined after molding. The magnet is built with a pre-magnetized element embedded within the PM-rotor core. In order to improve the mechanical strength of the SE-WF rotor, as well as its heat dissipation and insulation performance, it is mold treated with a PPS resin. After the stator coil connecting process has been similarly mold treated as in Fig. 10(i), it is covered with a PPS cover.

Table 1. Specifications of prototype

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of rotor poles</td>
<td>8</td>
</tr>
<tr>
<td>Number of stator slots</td>
<td>12</td>
</tr>
<tr>
<td>Motor core outer diameter</td>
<td>120 mm</td>
</tr>
<tr>
<td>Air-gap length</td>
<td>Radial 0.7 mm, Axial 0.9 mm</td>
</tr>
<tr>
<td>Axial length of core</td>
<td>51.8 mm (without axial-gap core), 107.6 mm (with axial-gap core)</td>
</tr>
<tr>
<td>Maximum magnetomotiveforce</td>
<td>900 A_{max} T (30 s)</td>
</tr>
<tr>
<td>Number of stator coil-turn</td>
<td>9</td>
</tr>
<tr>
<td>Armature coil winding connection</td>
<td>4 series</td>
</tr>
<tr>
<td>Number of rotor induction coil-turn</td>
<td>30</td>
</tr>
<tr>
<td>Number of rotor field coil-turn</td>
<td>30</td>
</tr>
<tr>
<td>Armature coil size (with insulation coating)</td>
<td>5.26 mm×0.56 mm</td>
</tr>
<tr>
<td>Rotor coil size (with insulation coating)</td>
<td>2.57 mm×0.47 mm</td>
</tr>
<tr>
<td>Core material</td>
<td>SMC (stator and axial rotor), Magnetic steel sheet (radial rotor)</td>
</tr>
</tbody>
</table>

5. Verification of Operating Characteristics of Actual Prototype Machine

5.1 Rotor Current Measurement

First, in order to confirm that the SE-WF rotor can be self-excited with second-order space harmonic inevitably generated by the toroidally-concentrated winding stator, the rotor current is measured with the prototype model with slip ring model shown in Fig. 11. It was configured connecting one of four diodes in Fig. 10(e) to a 3ch slip ring, and then connecting to a diode external to the motor via a brush. Figure 11(a) shows the configuration of measuring rotor current using a current sensor. Figure 12 shows the rotor current waveform measured under conditions of rotational speed 1000 r/min, armature magnetomotive force 405 A_{max} T, and current phase −20 degrees. In the same figure, the forward-direction current and reverse-direction current generated by the third harmonic of...
Proposal and Preliminary Experimental Verification (Masahiro Aoyama et al.)

5.2 Current Phase-vs.-Torque

The reluctance torque is measured under conditions in which a universal inverter (Myway Plus Cop.) was used with carrier frequency 10 kHz. In order to separate the reluctance torque and magnet torque, a non-magnetized magnet is used, and opened the SE-WF rotor winding. Next, we measured each torque under conditions using a magnetized magnet, and under conditions releasing and connecting the SE-WF rotor winding. For torque measurement, the rotation speed is controlled on the motor bench side, torque control with a test motor driving inverter, and used a HBM torque flange (T10FS) for torque detection. Torque measurement values were averaged over five times with one-second intervals via a 30 Hz low pass filter within a torque meter amplifier. Measurement temperature conditions were within 90° the normal temperature, which was the detection temperature of a thermocouple attached to the end of the stator coil. The magnetic pole position was adjusted by the position at which the N-pole of the PM-rotor was facing the U-phase armature winding direction. In order to confirm the automatic flux weakening function of the proposed motor, the DC-bus voltage is set to a rather high 200 V within the driving speed range and conducted measurements. Figure 13 shows the current phase-vs.-torque characteristics from opening the SE-WF rotor winding and measuring the current phase in 10 degree increments under conditions of rotational speed 500 r/min and armature magnetomotive force 630 Aₘₘₜ,T or less. The figure shows the current phase-vs.-torque characteristics of a SPM (surface permanent magnet) motor with the armature magnetomotive force in a low area, and the magnet torque dominant since the magnet of the PM-rotor is arranged in the salient pole section. On the other hand, it is possible to confirm that as the armature magnetomotive force increases, the reluctance torque of a SE-WF rotor having forward salient polarity is added, and an MTPA point is present in the flux intensifying area.

5.3 Adjustable Speed Drive Characteristics

Figure 14 shows an example of the torque characteristics during adjustable speed drive operation of the proposed motor. Figure 14(a) shows the results of assuming the armature magnetomotive force to be 90 Aₘₘₜ,T, 180 Aₘₘₜ,T, 270 Aₘₘₜ,T, 360 Aₘₘₜ,T, 450 Aₘₘₜ,T, and 630 Aₘₘₜ,T, and comparing the current phase-vs.-torque characteristics when the rotational speed is 750 r/min and 2000 r/min. Figure 14(b) shows the results of comparing torque at 750 r/min and 2000 r/min at the MTPA point of each armature magnetomotive force. From the figure, it is possible confirm that torque is decreasing due to increased rotational speed and the armature magnetomotive force. As stated in section 2, the induced electromotive force of the SE-WF rotor of the proposed motor is based upon Faraday’s Law, so the amount of magnetization of the reversed magnetic pole is automatically intensified by increased rotational speed, and torque is decreased by counteracting magnet magnetic flux. Furthermore, the second-order space harmonic acting as the field energy source of the reversed magnetic pole passively increases with increases in...
the armature magnetomotive force, so the extent of decreased torque increases as the armature magnetomotive force increases.

5.4 Efficiency Maps Figure 15 shows the measured efficiency map. For the convenience of the test environment and the mechanical structural design of the principle verification prototype machine, the experimental test is set at the rotational speed upper limit to 3000 r/min during reluctance torque measurement, and setting 2000 r/min as the upper limit when releasing and connecting the SE-WF rotor winding. The efficiency map is drawn by plotting the results of measuring the MTPA (max torque per ampere) when the armature magnetomotive force was 90 AmT, 180 AmT, 270 AmT, 360 AmT, 450 AmT and 630 AmT, in 250 r/min rotational speed increments from 500 r/min to 2000 r/min. Here, the upper limit of the armature magnetomotive force is 630 AmT (70% load of design value). But, the stator was lacking in the strength of its electromagnetic vibration and cooling performance since the prototype was made with a split core for improving the magnetic characteristics and improving mechanical strength of the SMC core pieces as shown in Figs. 9(c) and (d), so it was not possible to conduct experiments with the largest loads during desk study. When comparing Figs. 15(b) and (c), it can be confirmed that the largest torque decreases as rotational speed increases in the reference model (SE-WF rotor winding released) of the proposed motor. The proposed motor utilized the space harmonics conventionally lost due to self-excitation as the field energy source for the SE-WF rotor. Thus, the form of the loss changes. As for the conventional 100% iron loss, there is \( x\% \) iron loss depending upon the amount of self-excitation, while the remaining (100-\( x \))% turns into rotor winding copper loss. As described in Ref. (15), if there is self-excitation due to diode rectification, the phase difference of the induced electromotive force on the rotor side is determined by the rotor electrical time constant, but the rotor induced electromotive force generated by the phase difference at this time acts so that it instantly counteracts space harmonics, so the loss form changes as above. The space harmonics content passively changes depending upon the armature magnetomotive force, so the form of the loss (percentage of iron loss and copper loss) does not affect torque according to Eq. (17). However, the torque changes depending upon the electromagnetic flux \( \psi_{e-coil} \) of the reversed magnetic pole self-excitedly formed by the SE-WF rotor. From this, it is found that the amount of magnetization of the reversed magnetic pole of the SE-WF rotor increases as rotational speed increases, and a flux weakening is automatically formed.

On the other hand, in terms of efficiency, as the prototype was mainly designed to verify the principle of variable magnetic flux from the reversed magnetic pole, the magnetic circuit has not been optimized, resulting in low efficiency compared to a common electric rotary machine.
Consideration has been given to the fact that magnetic resistance increases due to the split core form (see Fig. 10) given so that the weight density of the core piece reaches 7.4 to 7.5 g/cm³ when molded with a 100 ton press, and that eddy current loss in holding member SUS and metal bolts is a cause of increased loss, but detailed loss analysis remains a problem to address in future.

When comparing with the reference model and the proposed motor, it is possible to confirm that efficiency decreases slightly as rotational speed increases. Decreased efficiency is particularly pronounced in low load areas. Here, in order to verify the principle of the automatic flux weakening effect due to space harmonics, the efficiency with MTPA points without considering voltage limits was plotted. On the other hand, the original purpose of variable magnetic flux technology is to adjust the amount of field so that line voltage may be kept within voltage limits, and to achieve expanded adjustable speed drive characteristics and improved efficiency. In future, there needs to be a comparison of performance with flux weakening from conventional armature current phase advance (weak field with vector suppression according to −Ld) and an automatic weak field due to the electrical reversed magnetic pole proposed here (weak field due to motor hardware), and a clarification of the advantages and disadvantages of the proposed variable magnetic flux technique. On the other hand, as the results show that efficiency decreases with the reversed magnetic pole in low load areas without voltage limits, there needs to be an analysis of the correlation between the amount of variable magnetic flux and efficiency through design balancing the magnetomotive force of a magnet, armature, and electromagnet. Additionally, it is felt that this may be resolved during low rotations and in low load areas where reversed magnetic pole formation is not desired by using a Zener diode or similar means to provide a threshold for field current generation.

Furthermore, in the case of applying SMC material to a motor, while there are merits to the utilization of a three-dimensional magnetic path and eddy current loss reduction effects, the magnetic characteristics are inferior to magnetic steel sheets. In other words, the armature magnetomotive force needs to be greater than that with magnetic steel sheets to achieve the same magnetic flux density, so copper loss increases. In addition, magnetic characteristics greatly change depending upon the manufacturing method, so it is important that magnetic circuit and structural design includes considerations of aspect ratio in which there are fewer splits, while satisfying the core piece weight density to achieve the necessary magnetic characteristics and mechanical strength.

In addition, the variable magnetic flux principle of the proposed motor uses space harmonics as a field energy source for the reversed magnetic pole of the SE-WF rotor, so as described above, the loss conventionally represented by 100% iron loss changes in form depending upon the amount of self-excitation, wherein χ% is represents iron loss, and the remaining (100-χ)% represents the rotor winding copper loss. It could not clarify the effect of loss form changes upon efficiency, so further investigation is required in future.

5.5 Variable Magnetic Flux Range Width

It is experimentally verified the proposed variable magnetic flux effects by comparing the reference model (SE-WF rotor winding released) and the results of the proposed motor through measuring the line voltage. Figure 16 shows the results of converting and plotting voltage with a circular locus based on the current phase command value versus the root mean square value V1 of the fundamental wave component of line voltage measured to facilitate visual relative comparison. Here it should be noted that this is not a conversion to dq-axis voltage using the voltage phase. According to Fig. 16(a), when the reference model and proposed motor are compared, it is possible to confirm that the voltage ellipse of the reference model and proposed motor overlap at 750 r/min, and hardly any difference is seen. However, as the rotational speed increases, the amount of magnetization of the reversed magnetic pole increases, the armature interlink magnetic flux automatically weakens, and the voltage ellipse becomes smaller. When the voltage ellipse of the reference model and forced field model (flux intensifying: SE-WF rotor arranged so that it becomes an N-pole relative to the N-pole of the PM rotor) in Fig. 16(b) are compared, the amount of change in the voltage ellipse is found to be larger than that in Fig. 16(a). In the case of the forced field model, the second-order space harmonics acting as a field energy source for the SE-WF rotor tend to increase as a result of the forced field effect from self-excitation. On the other hand, in the case of the proposed motor, the second-order space harmonics tend to decrease as a result of the weak field effect from self-excitation. As a result, the energy source for reversed magnetic pole formation itself is weakened, and the amount of variable magnetic flux becomes smaller than in the forced field model. Based upon this, as described in the preceding section, it is found that design balancing of each magnetomotive force (magnet, armature, self-excited electromagnet) against the amount of variable magnetic flux required is important for self-excited variable magnetic flux technique.
utilizing space harmonics.

Next, Fig. 17 shows the results of comparing the voltage ellipse during variable speed driving per each armature magnetomotive force between the reference model and proposed motor. According to the drawing, there is no difference in the voltage ellipse even when rotational speed increases at driving points where the armature magnetomotive force is low. This is due to the fact that the induced electromotive force generated by the I-coils is consumed as Joule loss by the resistance of the SE-WF rotor winding and the resistance of the SiC diodes (ROHM SCS230AE2). It is possible to confirm that the induced electromotive force obtained as the armature magnetomotive force increases is larger than the amount of loss consumed by Joule loss, so a field current is produced, thus a reversed magnetic pole is formed, and the armature interlink magnetic flux is automatically adjusted.

6. Conclusion

This paper describes a PM-motor in which an actuator is not necessary, an electromagnet magnetic pole is formed automatically reversing the magnetic pole of a permanent magnet as rotational speed increases, such that variable magnetic flux is passively realized. A diode rectifier circuit-connected wound-field rotor is arranged on the double axial gaps of a toroidally-concentrated winding stator, second-order space harmonics inevitably generated by the concentrated winding structure are interlinked with the rotor coil for self-excited magnetization, and a radial gap face PM-rotor and reversed magnetic pole are arranged, in order to realize variable magnetic flux. Describing this variable magnetic flux principle from a mathematical viewpoint, it is experimentally verified the above variable magnetic flux principle using a actual prototype machine. As a result, it is demonstrated variable magnetic flux in which space harmonics were utilized as a variable magnetic flux energy source.

In future, Comparing performance with flux weakening from conventional armature current phase advance (weak field with vector suppression according to $-i_d$) and an automatic flux weakening due to the electrical reversed magnetic pole proposed here (weak field due to motor hardware) will be demonstrated, and clarify the advantages and disadvantages of the proposed variable magnetic flux technology. In addition, it is possible to change the magnetic pole relative angle of the axial gap rotor and radial gap rotor as a feature of the multi-gap structure, so it may be possible to improve torque density by changing the phase difference of the magnet torque and reluctance torque. In future, the verification of variable speed characteristics when changing the magnetic pole relative angle of the PM-rotor and SE-WF rotor should be demonstrated. In addition, it is also important to verify the relationship between torque characteristics and variable magnetic flux range. Furthermore, the optimization of magnetic circuit design with considering the manufacturing process of an SMC core is important. At the same time, it should be discuss the relationship between balancing armature magnetomotive force, magnet magnetomotive force, and self-excited electromagnet magnetomotive force and motor performance, and research design balancing optimal magnetomotive forces.
\[ v_{rd} = -pM_{rdad}i_{ad} = R_{r}i_{rd} + L_{rd0}i_{rd} + pL_{rd}(t)i_{rd} \]  \hspace{1cm} \text{(A2)}

Here, when it is performed direct current term approximation of the rotor \(d\)-axis self-inductance \(L_{rd0} = L_{rd0} + L_{rd}(t)\), then \(pL_{rd}(t) = 0\), so it is possible to organize (Eq. (A2)) with \(\text{(Eq. (A3))}\), yielding Eq. (8).

\[ v_{rd} = -pM_{rdad}i_{ad} = R_{r}i_{rd} + L_{rd0}i_{rd} \]  \hspace{1cm} \text{(A3)}

2. Derivation of Eqs. (10) and (11)

By performing a Laplace transform of Eq. (9) as follows, it is possible to derive Eq. (10).

\[ \mathcal{L}\left[ R_{r}i_{rd} + L_{rd0}i_{rd} \right] = -\frac{3N_{r}^{2}}{N_{s}}p_{r}(\theta) \left\{ \frac{\kappa_{d}L_{rd}i_{sd}}{s^2} \left[ \sin 3\omega t \right] \right\} \]  \hspace{1cm} \text{(A4)}

Here, when each coefficient in (Eq. (A4)) is placed as in (Eq. (A5)) and (Eq. (A6)), Eq. (10) may be organized as in (Eq. (A7))

\[ V_{rd} = -3N_{r}^{2}L_{rd0}p_{r}(\theta) \kappa_{d}L_{rd}i_{rd} \]  \hspace{1cm} \text{.........(A5)}

\[ V_{rq} = 3\omega N_{r}^{2}p_{r}(\theta) \kappa_{q}L_{rd}i_{sq} \]  \hspace{1cm} \text{.........(A6)}

\[ \frac{R_{r} + sL_{r}(s)}{3\omega V_{rd} + sV_{rd}} = \frac{3\omega V_{rd} + sV_{rd}}{s^2 + (3\omega)^2} \]  \hspace{1cm} \text{.........(A7)}

When we rearrange (Eq. (A7)) for \(I_{rd}(s)\), it becomes (Eq. (A8)).

\[ I_{rd}(s) = \frac{3\omega V_{rd} + sV_{rd}}{s^2 + (3\omega)^2} \]  \hspace{1cm} \text{.........(A8)}

We then expand (Eq. (A8)) using variables \(a\) and \(b\) as in (Eq. (A9)).

\[ I_{rd}(s) = \frac{3\omega V_{rd}}{(R_{r} + sL_{rd0})} \left\{ \frac{a_{1}}{(R_{r} + sL_{rd0})^2} \right\} + \frac{b_{1}s + c_{1}}{s^2 + (3\omega)^2} \]

\[ + V_{rq} \left\{ \frac{a_{2}}{(R_{r} + sL_{rd0})} \right\} + \frac{b_{2}s + c_{2}}{s^2 + (3\omega)^2} \]

\[ = \frac{3\omega V_{rd}}{(R_{r} + sL_{rd0})} \left\{ \frac{a_{1} + b_{1}L_{rd0} + (b_{1}R_{r} + c_{1}L_{rd0}) + (9a_{1}\omega^2 + c_{1}R_{r})}{(R_{r} + sL_{rd0})^2} \right\} \]

\[ + V_{rq} \left\{ \frac{a_{2} + b_{2}L_{rd0} + (b_{2}R_{r} + c_{2}L_{rd0}) + (9a_{2}\omega^2 + c_{2}R_{r})}{(R_{r} + sL_{rd0})^2} \right\} \]  \hspace{1cm} \text{.........(A9)}

Here, when it is compared the variables in (Eq. (A8)) and (Eq. (A9)), the following holds true.

\[ a_{1} + b_{1}L_{rd0} = 0, \quad b_{1}R_{r} + c_{1}L_{rd0} = 0, \quad 9a_{1}\omega^2 + c_{1}R_{r} = 1 \]  \hspace{1cm} \text{.........(A10)}

\[ a_{2} + b_{2}L_{rd0} = 0, \quad b_{2}R_{r} + c_{2}L_{rd0} = 1, \quad 9a_{2}\omega^2 + c_{2}R_{r} = 0 \]  \hspace{1cm} \text{.........(A11)}

When solving the simultaneous equations (Eq. (A10)) and (Eq. (A11)), it is possible to organize them as in (Eq. (A12)) and (Eq. (A13)).
\[ a_1 = \frac{L_{rd0}^2}{R_r^2 + (3\omega L_{rd0})^2}, \quad b_1 = -\frac{L_{rd0}}{R_r^2 + (3\omega L_{rd0})^2}. \]
\[ c_1 = \frac{R_r}{R_r^2 + (3\omega L_{rd0})^2} \]
\[ a_2 = -\frac{R_r L_{rd0}}{R_r^2 + (3\omega L_{rd0})^2}, \quad b_2 = \frac{R_r}{R_r^2 + (3\omega L_{rd0})^2}. \]
\[ c_2 = \frac{(3\omega)^2 L_{rd0}^2}{R_r^2 + (3\omega L_{rd0})^2} \]  \hspace{1em} (A12)
\[ (A13) \]

Accordingly, when substituting (Eq. (A12)) and (Eq. (A13)) into (Eq. (A9)), it becomes as in (Eq. (A14)).

\[
I_\theta(s) = 3\omega V_{id} \left[ \frac{L_{rd0}^2}{R_r^2 + (3\omega L_{rd0})^2} + \frac{L_{rd0}}{R_r^2 + (3\omega L_{rd0})^2} \right] + V_{rq} \left[ -\frac{R_r L_{rd0}^2}{R_r^2 + (3\omega L_{rd0})^2} \frac{R_r}{s^2 + (3\omega)^2} \right].
\]
\[ \]
\[ (A14) \]

Here, when
\[
\cos \varphi = \frac{R_r}{Z}, \quad \sin \varphi = \frac{3\omega L_{rd0}}{Z}, \quad \tau = \frac{L_{rd0}}{R_r}, \quad \]  \hspace{1em} (A15)

it is possible to organize (Eq. (A14)) as in (Eq. (A17)).

\[
I_\varphi(s) = \frac{1}{Z} \left[ V_{id} \left\{ \frac{\sin \varphi}{s^2 + (3\omega)^2} \left\{ \frac{\sin \varphi}{s^2 + (3\omega)^2} \right\} \right. \right. \]
\[ + V_{rq} \left\{ \frac{-\cos \varphi}{s^2 + (3\omega)^2} \left\{ \cos \varphi + \sin \varphi \right\} \right\} \right] \]
\[ \]
\[ \]  \hspace{1em} (A17)

Accordingly, when an inverse Laplace transform is performed on (Eq. (A17)), it becomes (Eq. (A18)), making it possible to derive Eq. (11).

\[
i_{\varphi}(t) = \frac{1}{Z} \left[ V_{id} \left\{ e^{-\tau} \sin \varphi + \sin(3\omega t - \varphi) \right\} \right. \]
\[ + V_{rq} \left\{ -e^{-\tau} \cos \varphi + \cos(3\omega t - \varphi) \right\} \right] \]
\[ \]
\[ \]  \hspace{1em} (A18)