1 9 8 8 年度 電氣·電子工學 學術大會 論文集 88/7

제어소자로서의 유도 서보 전동기

이 견 용*, 디•파울리
서독하겐 종합대학 전기공학과

INDUCTION SERVOMOTOR AS CONTROL COMPONENTS

Kunyong Lee, D.Pauly
Fachbereich Elektrotechnik, Gesamthochschule, Hagen, FRG

Abstract: This paper deals with some aspects for the design of induction servomotors with respect to permanent magnet servomotors and standard induction motors as control components. Induction servomotors have much lower weight and moment of inertia than the DC servomotor but cannot reach the excellent values of a brushless DC servomotor with rare earth magnets.

INTRODUCTION

In general DC servomotors are used for electric servomotor drives,but cmmutator and brush-holders are sensitive and expensive parts of a DC motor,requiring maintanance and increasing the total weight. The need for a better dynamic performance and for a maintenance free operation of servo drives sustains the growing demand for brush lees servomotors in the recent years.

In the meantime the brushless DC servomotor is widely used. In spite of its name, this motor type is a multiphase machine. The rectangular shape of the motor current allows a very simple signal processing and control,sometimes using components of the conventional DC servomotor,which saves costs and time for development. This simple control schems may lead to torque harmonics caused by the nonideal distribution of the air gap flux density and the nonideal commutation of the phase currents. Improvements then lead to a control scheme,similar to the synchronoous servomotor, ,which has a position sensor with a high resolution. The simple and rugged consideration of an induction motor is interesting for servomotors,too. Investigations are mainly concerned with the control of the induction servomotor.

MATHEMATICAL MODEL

The dynamic behaviour of an induction machine is described by a set of nonlinear differential equations. Using the theory of space vectors,these equations get simple and easy to survey. Considering the well-known field-oriented control,a transformation of these equations in a reference frame which rotates synchronously with the rotor flux is advantageous .Induction servomotors are always fed from a transistor invertec with a constant DC-link voltage. Speed,torque and frequency and amplitude of the fundamentals of the phase currents and the phase voltage vary in a wide range. With the DC-link voltage U_z and the peak current I_{max} it ,see appendix.

Eq.(2a) has the same form as the equation of the armture voltage and armature current of a DC-servomotor. One gets for the torque constant K_t,the voltage constant K_e,the electrical and mechanical time constant T_{e1} and T_{mech}. In contrast to permanent magnet servomotors,the torque and voltage constant are adjustable by the variable flux level. Therefore,rated values can be defined for normal saturation of the induction servomotor,which should be kept constant in the basic speed range. Operating points inthe field weakening range lead to lower values. The ohmic losses can be divided into the excitation losses p_{vx} and the load losses

p_{vy}, resulting in eq.(10) and (11).
Eq.(1) to (11) together with the conventional design procedure of induction motors are used for a special design procedure for induction servomotors. An operating point is then defined by the current components, the speed, the dimensions of the motor, the steel sheet laminations and the armature windings. To get a low phase current a high torque constant is favourable but this leads to a high voltage constant, too. This results in a low speed range for a given DC-link voltage. Therefore, the design procedure starts from given torque and voltage constants following from a given DC-link voltage of the inverter and a desired speed range without field weakening. A low electrical time constant is achieved by a low subtransient inductance σL_1. Therefore, a high number of stator and rotor slots, a fractional pitched armature winding and low slot depths are needed. Limiting by this the winding space, one gets high resistance R_1 and R_2 thus resulting in high ohmic losses p_{vx} and p_{vy}. A low subtransient inductance σL_1 requires a high switching frequency and a suitable modulation technique to get low current ripple and additional losses from inverter feeding. A low mechanical time constant leads to a good dynamic response of the speed control and can be achieved by a small bore diameter and a high stack length which lowers the moment of inertia. Such a design differs from standard induction motors significatly, which have a ratio stack length/bore diameter from 0.8 to 1.5 approximately. Induction servomotors need to have a value of 2 and more to get a moment of inertia which is comparable to other servomotors.
The output power of an electrical machine is determined by the cooling system. Standard induction motors operating at constant speed near the synchronous speed of the motor are effectively cooled by a fan wheel together with cooling fans and the fan shell. In contrast servomotors operate without a fan wheel, which is ineffective in any case at low speeds. Seperate fan drives are expensive and need for additional volume. Moreover, fans are disadvantageous at all because servomotors are often mounted to the machine tool in a way that oil, dust cutting compound and chips get onto the surface of the motor. Omitting the fan, the effectiveness of axial cooling fans on a horizontal mounted motor is severly diminished. Regarding a given flange size the outside diameter of the stator core laminations is restricted to a lower value thus resulting in higher losses of the servomotor. Cosequently most servomotors are designed without cooling fans or only small ones, thus excluding the use of housing of standard induction motors.

LOSSES OF MOTORS

Losses from the fundamentals of current and magnetic field. Disregarding additional losses from inverter feeding, ohmic losses and iron losses can be calculated in the same way as for a normal induction motor. Obviously one has to consider variable frequencies and flux levels. Although a high switching frequency is commonly used, these additional losses cannot be neglected especially when regarding the restrictions of the total admissible losses by omitting the fan. For calculation of these losses one has to consider the signicant influence of the skin effect in the stator windings and the rotor bars. The influence of eddy currents on the distribution of the flux density in the stator laminations has to be considered when calculating the addtional iron losses from inverter feeding. The effect of the skin effect, especially inthe rotor bars can be summed up by the following aspects:

Steady stator: No influence on the current distribution of the fundamental of the bar current because of the low slip frequency.

Current ripple from inverter feeding: The effective slot scattering conductance is reduced to approximately 50 % of its value at low frequency, the effective resistance of the rotor bar increase with a factor of approximately 30 to 100.

Transient state: the rise time of current and torque is significantly diminished, which is advantageous for induction servo motors. Mains operated standard induction motors show this effect at starting from standstill,too. Maximum addtional losses from inverter feeding result from modulation techniques which exclude the voltage space vector null. Calculations and measurements for such a modulation technique are shown in Fig. 1 for the test motor,see appendix. Fig. 1 shows on the one hand a distinct rise of the additonal losses for a low switching frequency of the transistor inverter and on the other hand a low limit for these losses at a high switching frequency . This is caused by the skin effect,which compensates a lower current ripple at a higher switching frequency by increasing the effective resistance of the armature winding and the rotor bars. For that reason it is not necessaryto exceed the switching frequency too much if there are no other restrictions to meet noise. Especially high speed servomotors have a low number of turns and subtransient inductance. Measured results for such motors show significant additional losses reaching up to 20% and more of the total admissable losses.

Overload capacity of an induction servo drive is determined on the one hand by the motor and on the other hand by the inverter. Usually an inverter is charaterized by the rated current limit which may not be exceeded to prevent damage of the power transistors. Furthermore an actual peak current can be defined,which can be delivered by the inverter for a few milliseconds up to a few minutes followed by a pause. Usually permanent magnet motors have a current limit which may not be exceeded to prevent demagnetization of the magnet system. In contrast to magnet motors,induction servomotors have no strict limit neither for the whose exceeding damages the motor. For demonstration of the behaviour of an induction motor at high overload,Fig.2 shows two torque-cuerent curves for a fractional horse power inductance motor,Fig.2a and for the test motor,Fig 2b. The curves marked with (1) show the linear dependence of the torque M from the current component i_{1y} for field-oriented control. It was found,that induction motors normally can deliver approximately ten times of the rated torque defined without fan cooling, without significant deviation from this linear dependence,correct parameter tuning assumed. Disregarding any restriction from field-oriented control and tuning the slip frequency for a given current amplitude for maximum torque output,one gets both curves marked with (2). The iron is unsaturated in the lower,parabolic part of the curves. Summing up the results from Fig.2,induction servomotors can deliver a peak torque much higher than ten times of the rated torque provided that the current limit of the inveter is high enough. This is advantageous compared to permanent magnet motors. The limit temperature of the winding insulation determines the thermal overload capacity of an induction servomotor. The time to reach this limit temperature for a given load depends on the thermal time constant of the iron stack,the coils in the slots and the coil ends. At heavy overload the temperature of the coil ends differs significantly from that of the coils in the slots,Fig.3 shows this for an overload of 3.6 times of the rated stall torque for the test motor. Obiously the temperature of the coil ends determine the thermal overload capacity of the motor. The temperatures in the motor can be calculated by the well-known methods of thermal modelling.

TORQUE OF MOTOR

Magnetizing current eq.(12) and neglecting the diffence between total and differntial permeability the rotor eq.(1b) results in eq.(13) and (14). Fig.4 shows K_t for the test motor as a function of magnetiring current where the dashed line holds for infinite permeability of the iron.

Most applicationsof servomotors reqiure smooth running even at very low speed. Therefore,low frequency torque harmonics have to be avoided to prevent the corresponding low frequency speed harmonics. These can be only be damped with the gain of the closed-loop speed control. Ripple torques can be distinguished in those caused by non-symmetrical and or non-sinusoidal phase currents and those caused from slotting and winding distribution. Offset,hysteresis,variable gain and other inaccuraies of current measuring devices, analogue-digital converters together with inaccuracies of the modulation technique itself may result in current harmonics. The stator current space vector is then discribed by eq.(15). Positive values for f_ν give positive rotating, negative values give negative rotating space vectors. DC components of the phase currents, which may arise despite of a floating neutral are described by a space vector with $f_\nu = 0$. For constant speed n the torque can be written as eq.(16). Neglecting slot opening, the torque harmonicscaused by the space harmonics can be calculated in a simple way. Fig. 5 shows some experimental results for the test motor. Experiment were carried out for locked rotor and a digital current control. Usually the value of rippöe torque should be below one percent of the rated torque. Experiments on the test motor and several standard induction motor showed, that ripple torques caused by space harmonics of the air gap flux density are below two percent of the rated torque even for motors without fractional pitched armature winding and only two slots per pole and phase. The design rules for standard induction motors to prevent noise, synchronous and asynchronous torques are with only slight modifications valid for induction servomotors, too. Then the substantial ripple torques result from current harmonics.

COMPARISON OF THREE SERVOMOTORS

Designed for the same DC-link voltage and the same rated speed the data of the motors are given in Table 1. A DC servomotor, a brushless DC servomotor(BLDC) and an induction servomotor have the same flange size. The rated current of the induction servomotor is much higher than of the BLDC motor, because the current component i_{1x} is significant. At high overload however, the phase current of the induction servomotor is only determined by M and K_t and are then nearly the same. Compared to the BLDC motor, the moment of inertia of the induction servomotor is 18% and of that the DC servomotor is 147% higher, although the BLDC motor was not optimized for minimum moment of inertia but for minimum magnet and copper volume. A lower value is possible, so that the difference to the induction servomotor and to the DC servomotor gets higher.

CONCLUSIONS

The difference between a DC servomotor, BLDC motor, and an induction servomotor were dicussed. Induction servomotors have much lower weight and moment of inertia than the DC servomotor but cannot reach the excellent values of a BLDC motor with rare earth magnets.

APPENDIX

Space vector equations for the symmetrical induction machine in a rotating(u_1) reference frame/1/:

$$\underline{u}_1 = R_1 \underline{i}_1 + \dot{\underline{\psi}}_1 + j\omega_1 \underline{\psi}_1 \qquad (1a)$$

$$0 = R_2 \underline{i}_2 + \dot{\underline{\psi}}_2 + j(\omega_1 - 2\pi pn) \qquad (1b)$$

$$u_{1x} = -\sigma L_1 (2\pi pn + R_2 i_{1y}/(L_2 i_{1x})) i_{1y} + R_1 i_{1x} \qquad (2a)$$

$$u_{1y} = \sigma L_1 \dot{i}_{1y} + (R_1 + R_2(L_1/L_2)) i_{1y} + (2\pi p L_1 i_{1x}) n \qquad (2b)$$

$$\psi_{2x} = L_h i_{1x} \qquad (2c)$$

$$M = (1.5\, p L_h^2 i_{1x} / L_2)\, i_{1y} \qquad (2d)$$

$$\sqrt{u_{1x}^2 + u_{1y}^2} < U_z/\sqrt{3} \qquad (3)$$

$$\sqrt{i_{1x}^2 + i_{1y}^2} < I_{max} \qquad (4)$$

$$K_t = 1.5\, p L_h \psi_{2x}/L_2 \qquad (5)$$

$$K_e = 2\pi p L_1 \psi_{2x}/L_h \qquad (6)$$

$$T_{el} = \sigma L_1/(R_1 + R_2 L_1/L_2) \qquad (7)$$

$$K_e/K_t = 4\pi/(3(1-\sigma)) \qquad (8)$$

$$T_{mech} = 2\pi J (R_1 + R_2 L_1/L_2)/(K_t K_e) \qquad (9)$$

$$P_{vx} = 1.5\, R_1 i_{1x}^2 \qquad (10)$$

$$P_{vy} = 1.5 (R_1 + R_2 (L_h/L_2)^2) i_{1y}^2 \qquad (11)$$

$$i_\mu = \psi_{2x}/L_h \qquad (12)$$

$$L_2/R_2\, \dot{i}_\mu + i_\mu = i_{1x} \qquad (13)$$

$$K_t = 1.5\, p L_h^2 i_\mu / L_2 \qquad (14)$$

$$\underline{i}_1 = \sum_\nu i_{1\nu} \exp[j(2\pi f_\nu t + \phi_{i\nu}] \qquad (15)$$

$$M = \frac{3}{2} p \,\mathrm{Im}\{ \sum_\nu \sum_\mu i_{1\nu} i_{1\mu} \exp[j(2\pi (f_\mu - f_\nu) t + \phi_{i\nu} - \phi_{i\mu}] \cdot L_1 \frac{R_2 - j\sigma L_2 2\pi (f_\nu - pn)}{R_2 - jL_2 2\pi (f_\nu - pn)} \} \qquad (16)$$

REFERENCE

/1/ Kovacs, K.P.: Transiente Vorgänge in Wechselstrommaschinen, Verlag Akademi, Budapest, 59.