

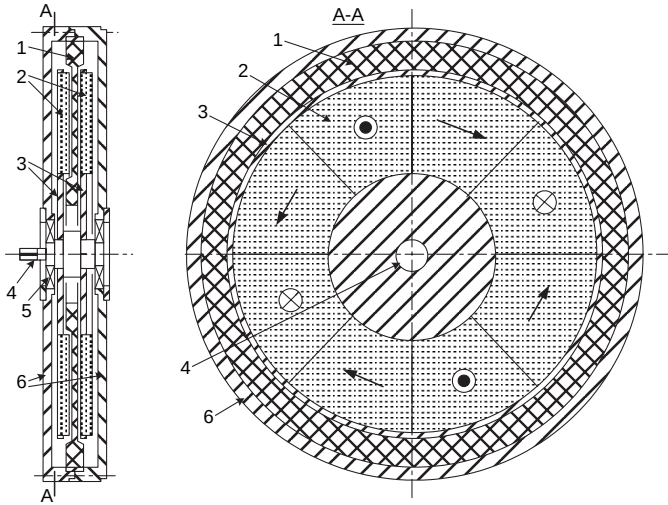


Fig. 6.1. Disc type coreless PM brushless machine: 1 — coreless stator (armature) winding, 2 — PMs, 3 — twin rotor, 4 — shaft, 5 — bearing, 6 — frame.

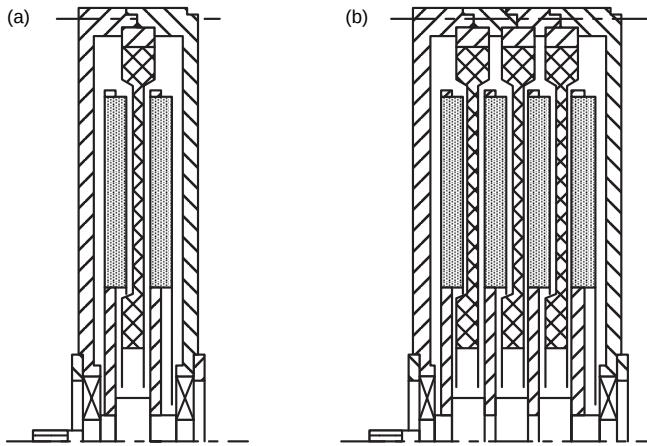


Fig. 6.2. Segmental construction of a disc type coreless PM brushless machine: (a) single module (segment), (b) three-module assembly.

and core losses. The only eddy current losses are losses in the stator winding conductors and metallic parts (if they exist) that reinforce the stator coreless winding.

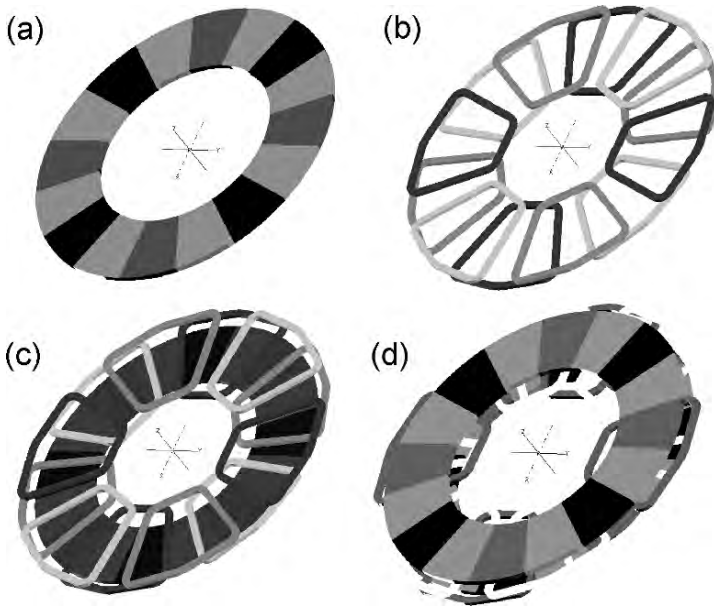


Fig. 6.3. Construction of a three-phase, 8-pole AFPM brushless machines with PMs arranged in Halbach array: (a) PM ring; (b) stator winding; (c) one half of the twin rotor; (d) stator winding and complete twin rotor.

The coreless machine designed as a segmental (modular) machine is shown in Fig. 6.2. The output shaft power can easily be adjusted to desired level by adding more modules.

To obtain a high power (or high torque) density motor, the magnetic flux density in the air gap should be as high as possible. This can be achieved by using PMs arranged in “Halbach array” (Figs 3.15, 6.3 and Fig 6.4). The magnetic flux density distribution excited by Halbach arrays is described by eqns (3.42) to (3.46). In practice, the angle between magnetization vectors of adjacent magnets is 90° , 60° or 45° (Fig. 6.5).

6.3 Air Gap Magnetic Flux Density

Fig. 6.6 shows the results of the 2D FEM modelling of the magnetic field in the air gap of a coreless AFPM brushless machine. NdFeB magnets with remanent flux density $B_r = 1.2$ T and coercivity $H_c = 950$ kA/m have been

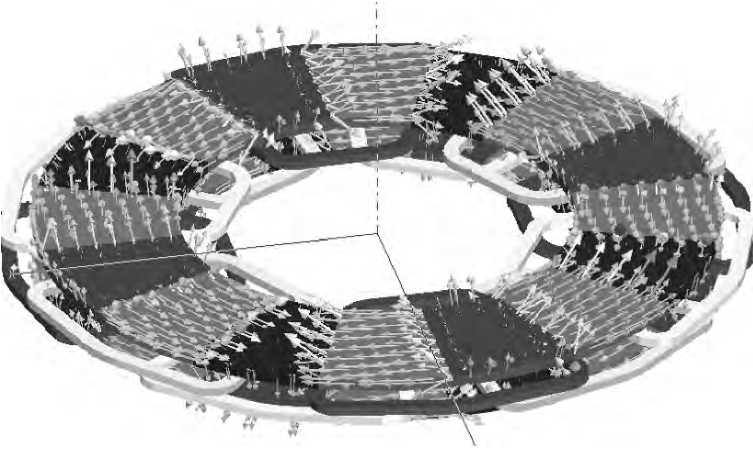


Fig. 6.4. 3D magnetic flux density distribution excited by an 8-pole twin rotor with PMs arranged in Halbach array.

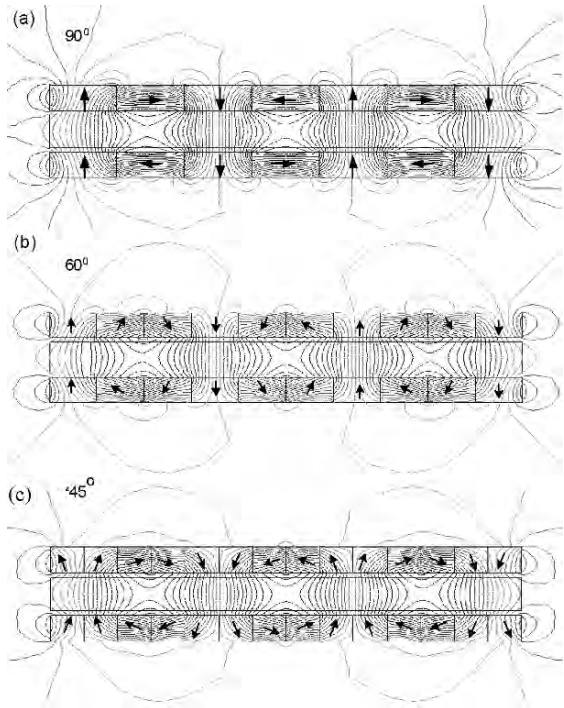


Fig. 6.5. Arrangement of the twin rotor PMs in 90°, 60°, and 45° Halbach array.

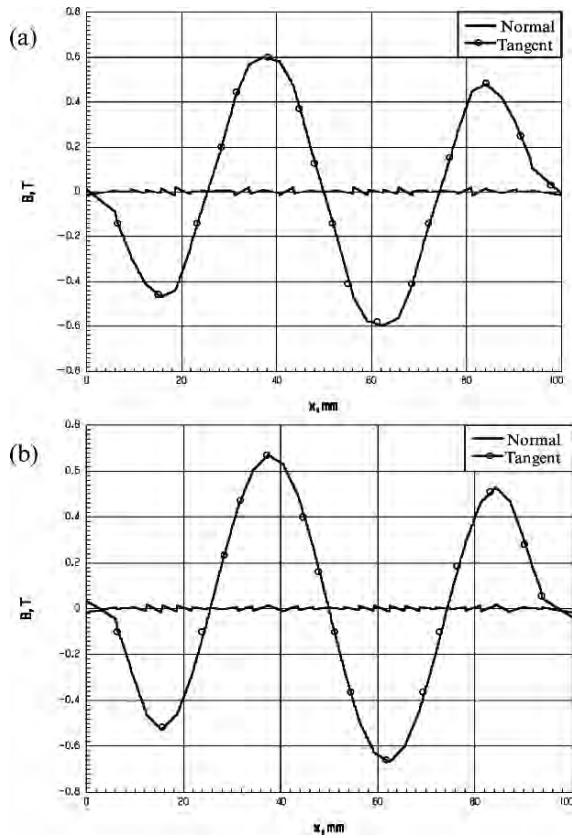


Fig. 6.6. Normal and tangential components of the magnetic flux density in the centre of the air gap of a double-sided coreless AFPM brushless machine excited by Halbach arrays of PMs: (a) 90°; (b) 45°.

considered. The thickness of each PM has been assumed 6 mm, the coreless stator winding thickness is 10 mm and one-sided air gap thickness equals to 1 mm.

With the aid of Halbach array a high peak value (over 0.6 T) of the normal component of the magnetic flux density has been excited. This value is sufficient to obtain a high electromagnetic torque. The peak value of the flux density can be even higher for the optimized magnetic circuit of an AFPM machine. In practice, 60° and 45° Halbach arrays produce similar peak magnetic flux density (Fig. 3.22). The peak value of the normal component of the magnetic flux density is higher than that in a standard arrangement of surface PMs. The backing ferromagnetic discs added to the double-sided PM structure with a large nonmagnetic gap do not increase the flux density as much as Halbach array excitation does.

6.4 Electromagnetic Torque and EMF

The electromagnetic torque of a coreless AFPM brushless machines with Halbach array PM excitation can be calculated on the basis of eqns (2.32), (2.33), (2.34), (2.127) and (2.128). Similarly, the EMF can be calculated according to eqns (2.36), (2.37), (2.125) and (2.126).

6.5 Commercial Coreless AFPM Motors

Coreless AFPM machine architecture known as *segmented electro-magnetic array* (SEMA) has been proposed by *Lynx Motion Technology*, Greenville, IN, U.S.A. for electric motor, actuator and generator applications requiring high power density and high efficiency [151]. Applications that require an electromechanical drive with high power density, high efficiency and low torque ripple include precision motion control, naval propulsion systems and acoustically sensitive applications. SEMA technology may also find applications in distributed generation systems and energy storage systems, such as *flywheel motor-generator systems*. Motors with high torque and efficiency are demanded for gearless electromechanical drives. The use of direct drive motors without speed reducers eliminates gear noise, oil leaks, positioning errors from backlash and lack of torsional stiffness.



Fig. 6.7. Exploded view of the SEMA AFPM brushless machine. Courtesy of *Lynx Motion Technology*, Greenville, IN, U.S.A.

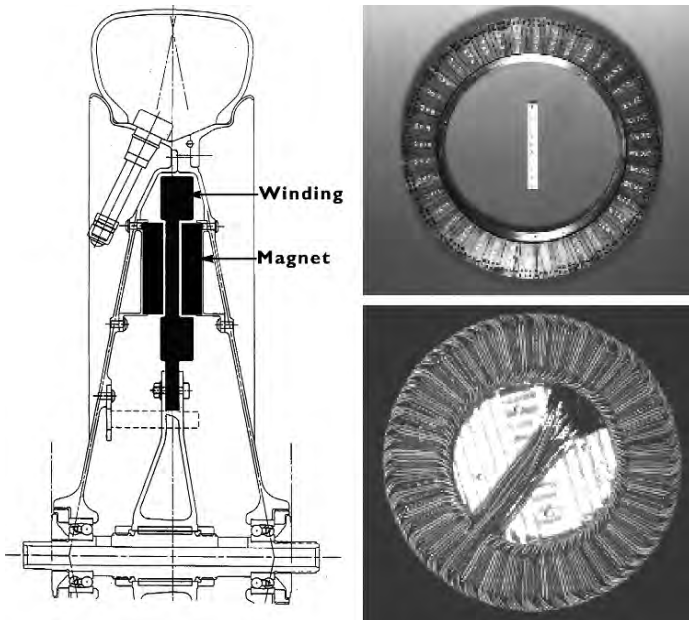


Fig. 6.8. In-wheel motor, PM ring and stator winding of a coreless AFPM brushless motor for solar-powered vehicles. Courtesy of *CSIRO*, Lindfield, NSW, Australia.

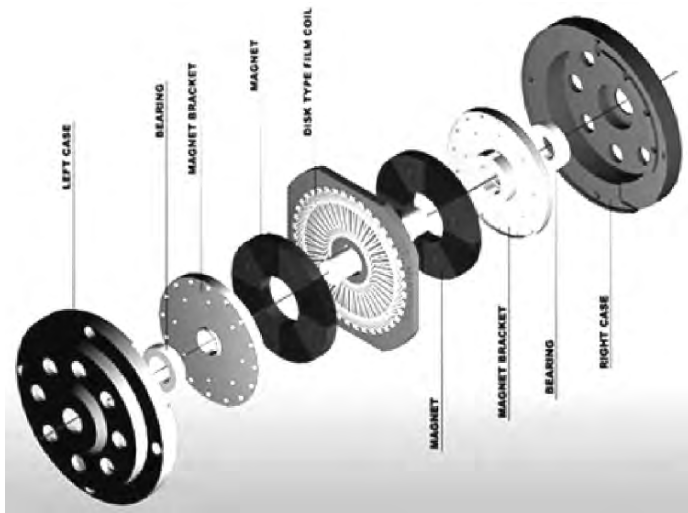


Fig. 6.9. Exploded view of the AFPM brushless motor with film coil coreless stator winding and twin PM rotor. Courtesy of *EmBest*, Soeul, South Korea.

The stator coreless and toothless design (Fig. 6.7) not only eliminates the cogging torque but also increases the area available for conductors. It also boosts peak torque capability and allows PMs to be more effectively utilized [151]. The stator coils are fully potted in high-strength, thermally conductive epoxy resin. Such construction gives the machine structural integrity and effectively damps the high frequency vibration when the motor is PWM inverted-fed. Machines with coreless stator winding perform excellently at constant speed, variable speed and at reversals.

CSIRO Telecommunications and Industrial Physics, Lindfield, NSW, Australia manufactures in-wheel coreless AFPM brushless motors for solar powered vehicles [225]. Solar-powered vehicles compete in the World Solar Challenge solar car race organized every two years in Australia between Darwin and Adelaide which has become a well-known international event.

CSIRO provides in-wheel AFPM brushless motors both with surface PMs glued to steel discs and with PMs arranged in Halbach array. The motor structure, PM disc and stator winding are shown in Fig. 6.8 [225].

Fig. 6.9 shows a coreless brushless motor with film coil stator winding and twin PM rotor manufactured by *EmBest*, Seoul, South Korea. The coreless stator has a foil winding at both sides. The 8-pole PM rotor is designed as a twin external rotor.

6.6 Case Study: Low-Speed AFPM Coreless Brushless Motor

A low speed, three-phase, Y-connected AFPM brushless motor rated at 10 kW, 750 rpm, 28.5 A has been designed and investigated. The rotor does not have any ferromagnetic core and consists of trapezoidal coils embedded in a high mechanical integrity resin. Average quality sintered NdFeB PMs with $B_r = 1.2$ T and $H_c = 950$ kA/m have been used. The double-disc construction is similar to that shown in Fig. 6.2b. The design data and calculated parameters are given in Table 6.1. All parameters except the air gap magnetic flux density B_{mg} and synchronous reactance $X_{sd} = X_{sq} = X_s$ have been calculated analytically. The magnetic flux density and synchronous reactance has been calculated using the 2D FEM. The resistance $R_1 = 0.175$ Ω and synchronous reactance $X_s = 0.609$ Ω per phase are for two stator modules connected electrically in series. The EMF constant $k_E = 5.013$ Vs and torque constant $k_T = 2.394$ Nm/A have been calculated for one stator segment.

6.6.1 Performance Characteristics

The current density 4.175 A/mm² in the stator winding at rated load is low and a totally enclosed motor can even operate without any forced air cooling. The recommended current density for enclosed a.c. motors with class F of insulation rated up to 10 kW is from 4.5 to 7 A/mm² [70].

Table 6.1. Design data and parameters of a three-phase, 10-kW, 750-rpm disc type AFPM brushless machine with ironless stator core

Design data	
Output power P_{out} , W	10,000
Speed n , rpm	750
Number of phases m_1	3 (Wye)
Input current I , A	28.4
Input frequency f , Hz	100
Number of stators (modules)	2
Number of pole pairs p	8
Number of coils (3 phases)	24
Number of turns per phase	100 (one module)
Coil pitch	1 slot
Wire diameter, mm	6×1.2
Axial thickness of PM h_M , mm	6
Axial thickness of the winding t_w , mm	10
Air gap (one side) g , mm	1
Air gap magnetic flux density B_{mg} under load, T	0.58
Current density, A/mm ²	4.175
$k_d = D_{in}/D_{out}$ ratio [5]	$1/\sqrt{3}$
Rotor outer diameter D_{out} , mm	360
Winding packing factor at $r = 0.5D_{in}$	0.65
Winding temperature, °C	75
Cooling system	natural
Class of insulation	F

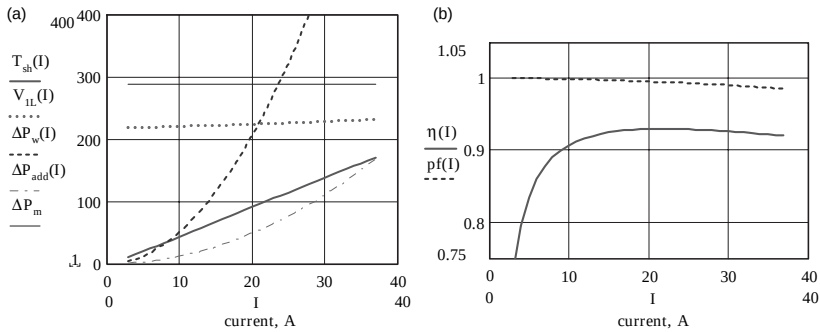


Fig. 6.10. Calculated performance characteristics of a 10 kW coreless AFPM brushless motor according to Table 6.1: (a) shaft torque T_{sh} , line-to-line voltage V_{LL} , winding losses ΔP_w , additional losses ΔP_{add} , and mechanical losses ΔP_m versus stator current at $n = 750$ rpm = const; (b) efficiency η and power factor pf versus stator current at $n = 750$ rpm = const.

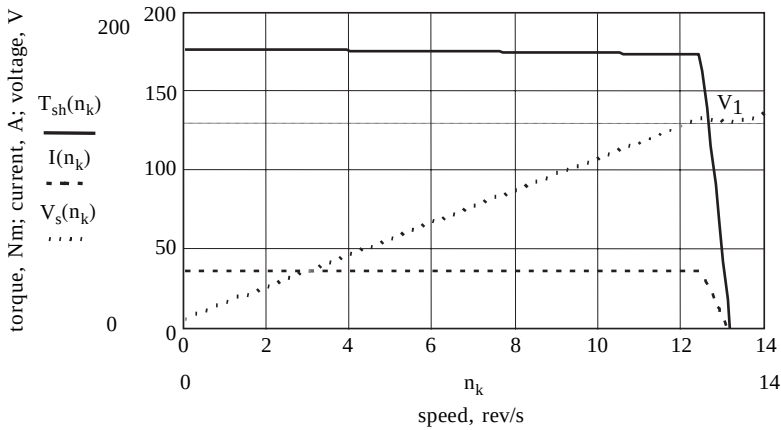


Fig. 6.11. Shaft torque, current and phase voltage versus speed of a 10 kW coreless AFPM brushless motor according to Table 6.1.

The calculated characteristics are shown in Figs. 6.10 and 6.11. It can be seen from Fig. 6.11 that the maximum-to-rated current ratio for torque-speed characteristic is 1.3. In Fig. 6.11 the voltage increases linearly up to approximately rated speed and then is kept constant.

6.6.2 Cost Analysis

The cost model has been presented in [106]. Average prices of materials for electrical machine construction are as follows (in US dollars: \$): silicon sheet steel $c_{Fe} = 1.25$ \$/kg, isolated copper conductor (magnet wire) $c_{Cu} = 5.51$ \$/kg, sintered NdFeB PMs $c_{PM} = 15$ \$/kg and steel bar for the shaft $c_{steel} = 2.6$ \$/kg. The active materials are conductors, sheet steel and PMs. The total volume of the steel bar to the volume of shaft is 1.94. The coefficient taking into account the cost of machining of the shaft is 2.15. The cost of labour involved in the assembly of active parts has not been estimated.

Table 6.2 lists the mass and cost of components of the 10 kW, 750 rpm coreless motor of disc construction. Table 6.3 lists the mass and cost of components of the equivalent cylindrical motor with laminated stator and rotor cores. Components independent of the machine shape are the encoder, terminal leads, terminal board, and name plate. The 10 kW coreless AFPM brushless motor costs \$613.05 (PMs cost \$185.66) while a classical PM brushless motor of cylindrical construction with the same ratings costs only \$510.01 (PMs cost \$23.55). The contribution of PMs to the total costs of a coreless motor is predominant.

6.6.3 Comparison With Cylindrical Motor With Laminated Stator and Rotor Cores

Table 6.4 compares the design data and performance of a 10 kW, 750 rpm coreless AFPM brushless motor and the equivalent 10 kW, 750 rpm PM brushless motor with laminated stator and rotor cylindrical cores. In both motors the number of phases is 3, number of poles is $2p = 16$, and input frequency is 100 Hz.

The analysis and the FEM simulation have shown the advantages of the coreless disc type PM brushless motors in terms of performance and mass of active materials. They have higher efficiency and power density when compared with conventional PM machines with laminated cores.

The efficiency of the 10-kW AFPM brushless motor is 1.2% higher (0.925 versus 0.914) and power density (output power to mass of active materials) is 82% higher (444.25 W/kg versus 243.55 W/kg) because the mass of active materials is 45% lower than that of an equivalent PM brushless motor with cylindrical ferromagnetic core. At zero current state the normal forces between the stator and rotor do not exist. However, the normal forces between the rotor PM discs are large.

Table 6.2. Mass and cost of components of a disc type coreless PM brushless motor rated at 10-kW, 750-rpm

Item	Mass, kg	Cost, \$
Winding	10.13	55.84
Stator plastic material	1.01	7.09
PMs	12.38	185.66
Shaft	5.48	59.45
Frame, end bells, and bearings		165.00
Components independent of the machine shape		140.00
Total	22.51 (active parts)	613.05

Table 6.3. Mass and cost of components of a cylindrical PM brushless motor with laminated core rated at 10-kW, 750-rpm

Item	Mass, kg	Cost, \$
Winding	8.51	46.9
Core	30.98	72.49
PMs	1.57	23.55
Shaft	5.72	62.07
Frame, end bells, and bearings		165.00
Components independent of the machine shape		140.00
Total	41.06 (active parts)	510.01

Table 6.4. Comparison of parameters and performance of 10-kW, 750-rpm PM brushless motors

Parameter	Coreless disc type	Cylindrical with laminated core
Input voltage (line-to-line), V	227.6	220.0
Shaft torque, Nm	131	133
Efficiency, %	92.5	91.4
Power factor	0.991	0.96
Air gap magnetic flux density, T	0.58	0.78
Stator winding current density, A/mm ²	4.175	4.247
Stator winding resistance per phase at 75°C, Ω	0.173	0.037
Synchronous reactance, Ω	0.609	0.758
Mass of PMs, kg	12.38	1.57
Mass of active materials, kg	22.51	41.06
Power density (only active materials), W/kg	444.25	243.55
Cost excluding labour, U.S.\$	613.05	510.01

These new types of machines are cogging torque free. Owing to the lack of ferromagnetic cores the hysteresis and eddy current losses in those parts do not exist.

The cost of materials and components of the disc type 10 kW coreless motor is 20% higher than that of a cylindrical motor with laminated core. On the other hand, the cost of tooling will be much lower as a coreless motor does not require stamping and stacking tools.

The high cost of PMs limits commercial applications of coreless AFPM brushless motors to small size machines, special servo drives, airborne apparatus (lightweight construction) and electromechanical drives where the cogging torque, torque ripple and normal forces must be reduced to zero.

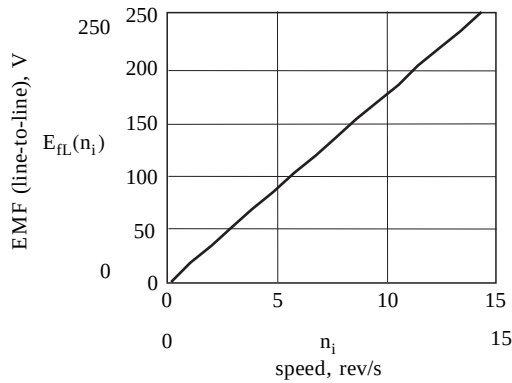


Fig. 6.12. Open circuit characteristic: line-to-line EMF versus speed.

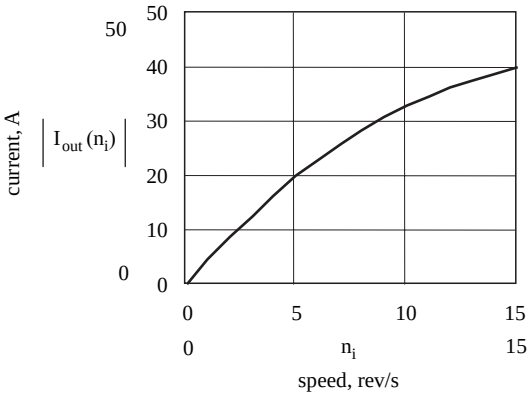


Fig. 6.13. Current versus speed for generating mode.

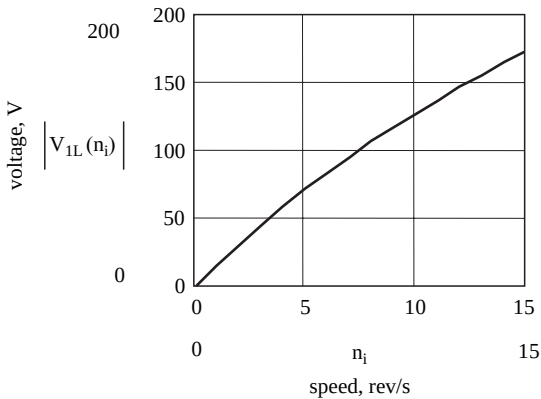


Fig. 6.14. Line voltage versus speed for generating mode.

6.7 Case Study: Low-Speed Coreless AFPM Brushless Generator

The AFPM brushless machine (Table 6.1) has been analysed as a stand alone a.c. generator. The open circuit characteristic, i.e. EMF (line-to-line) as a function of speed is shown in Fig. 6.12. This characteristic is linear.

The AFPM generator has been connected to the inductive load $\mathbf{Z}_L = R_L + jX_L = 1 + j0.628 \, \Omega$. The current–speed characteristic for generating mode is shown in Fig. 6.13.

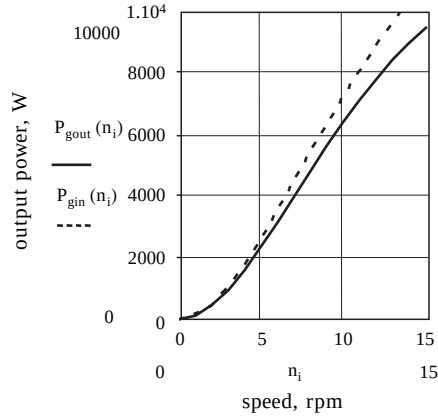


Fig. 6.15. Output power P_{gout} and input power P_{gin} for generating mode.

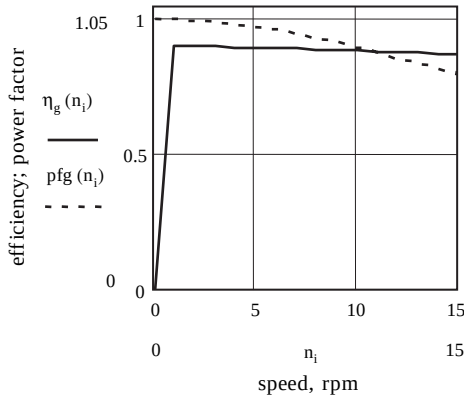


Fig. 6.16. Efficiency η_g and power factor pfg for generating mode.

The line voltage versus speed is shown in Fig.6.14, output and input active power in Fig. 6.15, and efficiency and power factor in Fig. 6.16. The voltage is a nonlinear function of speed and depends on the value of the load impedance $\mathbf{Z}_L$ and power factor. For inductive load (RL) the power factor decreases as the speed increases.

6.8 Characteristics of Coreless AFPM Machines

Table 6.5 shows specifications of coreless AFPM brushless motors for solar powered vehicles with surface PMs glued to a steel yoke and PMs arranged in Halbach array (*CSIRO*, Australia) respectively. The steady-state efficiency

curves for surface PMs and Halbach array are plotted in Fig. 6.17. The maximum efficiency is obtained when the inverter applies a sinusoidal voltage synchronized to the rotor position. The inductance of the stator winding is low and most inverters require additional series inductance.

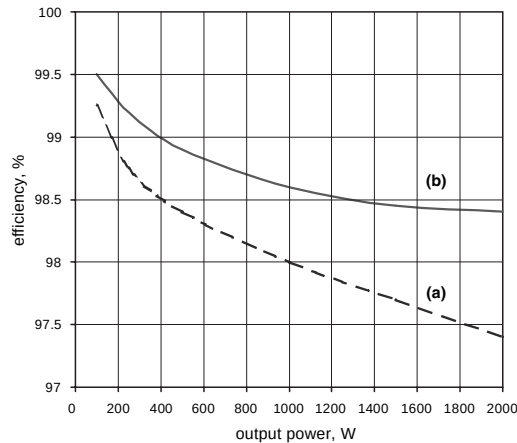


Fig. 6.17. Efficiency as a function of output power of a 3-phase, 1.8-kW coreless AFPM brushless motors for solar-powered vehicles: (a) rotor with surface PMs; (b) rotor with PMs arranged in Halbach array. Courtesy of *CSIRO*, Lindfield, NSW, Australia.

The E225 coreless AFPM motor (Table 6.6) has been developed by *Lynx Motion Technology*, Greenville, IN, U.S.A. for air-handling applications, in which system efficiency is a high priority. The family of *Lynx* 225-mm diameter motor designs offers a broad range of possible system configurations for speed and position servo applications requiring both high efficiency and high torque. Performance ratings of the E225 motor are based on air-cooled 105°C winding at 25°C ambient air temperature. The E813 coreless AFPM machines has been designed for power generation applications (distributed generation systems) where high efficiency of the system is important. Performance ratings of the E813 machine are based on 130°C winding at 40°C ambient air temperature. Both E225 and E813 machines have extremely linear torque constant k_T independent of speed. Other applications of *Lynx* coreless machines include servo applications, precision robotics, marine propulsion, generators and weapons turrets.

When coupled with a three-phase H-bridge inverter, *Lynx* motors offer a highly efficient electromechanical drive with the high peak torque capability required for the most demanding high-transient applications, e.g. fuel-cell air compressors.

Table 6.5. Comparison of three-phase 1.8-kW coreless AFPM brushless motors with surface and Halbach PM excitation manufactured by *CSIRO*, Australia

Parameter	Surface PMs and steel yoke	Halbach array
Output power, W	1800	1800
Efficiency, %	97.4	98.2
Number of poles	40	40
Nominal speed, rpm	1060	1060
Maximum speed, rpm	2865	2865
Nominal torque, Nm	16.2	16.2
Maximum continuous torque at 1060 rpm and maximum winding temperature, Nm	31	39
Absolute maximum torque, Nm	50.2	50.2
Torque constant per phase at sinusoidal excitation	0.39	0.56
Nominal phase EMF, V	43	62
EMF constant per phase at sinusoidal current, Vs/rad	0.39	0.56
Nominal phase current, A	13.9	9.6
Winding losses, W	43.9	27.6
Eddy current losses in the stator winding, W	2.6	2.7
Windage losses, W	2.1	2.1
Total losses, W	48.6	32.4
Winding temperature rise, °C	22	14
Overload winding temperature rise, °C	64	40
Maximum winding temperature, °C	110	110
Air gap at each side, mm	2	2
Resistance per phase, Ω	0.0757	0.0997
Mass of complete motor, kg	16.2	13.2
Mass of frameless motor, kg	10.7	7.7

Numerical Example 6.1

The coreless AFPM motor with the same dimensions as that described in *Numerical Example 5.1* has been redesigned to obtain the speed of 1000 rpm at $I_a = 8.2$ A ($I_{ad} = 0$). The number of pole pairs has been increased to $p = 12$, the number of coil sides to $s_1 = 72$ and PMs have been arranged in 90° Halbach array with $n_M = 4$ PM pieces per wavelength. The number of stator winding turns per phase is $N_1 = 240$, the number of parallel wires $a_w = 8$, the diameter of wire $d_w = 0.455$ mm (AWG 25) and the coil pitch measured in coils sides $w_c = 3$. Since both stator and rotor do not have any steel magnetic circuit, the magnetic saturation does not exist and $k_{sat} = 1$. All other parameters and assumptions are the same as those in *Numerical example 5.1*.

Table 6.6. Specifications of coreless AFPM motors manufactured by *Lynx Motion Technology*, Greenville, IN, U.S.A.

Parameter	E225 AFPM motor	E813 AFPM motor
Number of poles	12	28
d.c. bus voltage, V	155	850
Terminal voltage, V	117	574
	rms phase voltage H-bridge	peak line-to-line
Speed, rpm	6000	2750
Torque, Nm	9.9	450
Peak torque, Nm	97.4	900
Current, A	18.7	119
Peak current, A	180	238
Output power, kW	6.2	130
Torque constant, Nm/A	0.53	3.78
EMF constant per phase, V/krpm	19	209
Electrical time constant, μ s	0.348	1.66
Winding resistance at 100°C, Ω	0.204	0.0265
Winding inductance, μ H	71	44
Rotor moment of inertia, kgm ²	0.0184	9.16
Mass of motor, kg	8.4	295
External diameter, m	0.225	0.813

Find the motor steady state performance, i.e. output power, torque and efficiency at rated current $I_a = I_{aq} = 8.2$ A.

Solution

The following calculated parameters are the same as those in *Numerical example 5.1*: $g = 1.5$ mm, $D_{in} = 0.127$ m, $D = 0.1735$ m, $L_i = l_M = 0.0465$ m, $R'_{out} = 0.115$ m (for windage losses), $R_{sh} = 0.015$ m, $m_r = 3.4$ kg, $m_{sh} = 0.64$ kg and $m_{con} = 0.75$ kg.

The number of coils per phase for a single layer winding is $n_c = s_1/(2m_1) = 72/(2 \times 3) = 12$. The number of turns per coil is $N_{ct} = N_1/n_c = 240/12 = 20$. The number of coil sides per pole per phase (equivalent to the number of slots per pole per phase) $q_1 = s_1/(2pm_1) = 72/(24 \times 3) = 1$. The pole pitch measured in coil sides is $\tau_c = s_1/(2p) = 72/24 = 3$ and the average pole pitch $\tau = \pi D/(2p) = \pi 0.1735/24 = 0.023$ m

The input frequency at 1000 rpm

$$f = n_s p = \frac{1000}{60} \times 12 = 200 \text{ Hz}$$

The wavelength l_a at the average diameter and constant β for Halbach array is

$$l_a = 2\tau = 2 \times 0.023 = 0.046 \text{ m} \quad \beta = \frac{2\pi}{l_a} = \frac{2\pi}{0.046} = 138.3 \text{ 1/m}$$

The magnetic flux density in the air gap according to eqn (3.42) is

$$B_{mg} \approx B_{m0} = 1.2[1 - \exp(-138.3 \times 0.006)] \frac{\sin(\pi/4)}{\pi/4} = 0.609 \text{ T}$$

The magnetic flux according to eqn (2.28) is

$$\Phi_f = \frac{1}{8} \frac{2}{\pi} \frac{\pi}{12} 0.609 \times 0.22^2 \left[1 - \left(\frac{1}{\sqrt{3}} \right)^2 \right] = 0.00041 \text{ Wb}$$

The winding factor according to eqns (2.8), (2.9) and (2.10) is

$$k_{d1} = \frac{\sin \pi / (2 \times 3)}{1 \sin \pi / (2 \times 3 \times 1)} = 1; \quad \beta = \frac{w_c}{\tau_c} = \frac{7}{9}$$

$$k_{p1} = \sin \left(\frac{3}{3} \frac{\pi}{2} \right) = 1 \quad k_{w1} = 1 \times 1 = 1$$

The EMF constant according to eqn (2.37) and torque constant according to eqn (2.34) is

$$k_E = \pi \sqrt{2} \times 12 \times 240 \times 1 \times 0.00041 = 5.24 \text{ V/(rev/s)} = 0.0873 \text{ V/rpm}$$

$$k_T = k_E \frac{m_1}{2\pi} = 5.24 \frac{3}{2\pi} = 2.5 \text{ Nm/A}$$

The EMF at 1000 rpm is

$$E_f = k_E n = 0.0873 \times 1000 = 87.34 \text{ V}$$

The electromagnetic torque at $I_a = I_{aq} = 8.2 \text{ A}$ is

$$T_d = k_T I_a = 2.5 \times 8.2 = 20.5 \text{ Nm}$$

The electromagnetic power is

$$P_{elm} = 2\pi n_s T_d = 2\pi \frac{1000}{60} \times 20.5 = 2148.6 \text{ W}$$

The length of the stator shorter end connection without inner bends $l_{1emin} = (w_c/\tau_c)\pi D_{in}/(2p) = (3/3)\pi \times 0.127/24 = 0.017 \text{ m}$ and the length of the stator

longer end connection without outer bends $l_{1emax} = 0.017 \times 0.22/0.127 = 0.0288$ m (Fig. 3.16).

The average length of the stator turn with 15-mm bends is

$$\begin{aligned} l_{1av} &\approx 2L_i + l_{1emin} + l_{1emax} + 4 \times 0.015 \\ &= 2 \times 0.0465 + 0.017 + 0.0288 + 0.06 = 0.1984 \text{ m} \end{aligned}$$

The stator winding resistance at 75°C according to eqn (2.40) is

$$R_1 = \frac{240 \times 0.1984}{47 \times 10^6 \times 8 \times \pi \times (0.455 \times 10^{-3})^2/4} = 0.7789 \Omega$$

The maximum width of the coil at the diameter D_{in} is $w_w = \pi D_{in}/s_1 = \pi 0.127/72 = 0.0055$ m = 5.5 mm. The thickness of the coil is $t_w = 8$ mm. The number of conductors per coil is $N_c = a_w \times N_{ct} = 8 \times 20 = 160$. The maximum value of the coil packing factor is at D_{in} , i.e.

$$k_{fmax} = \frac{d_w^2 \times N_c}{t_w w_w} = \frac{0.455^2 \times 160}{8 \times 5.5} = 0.731$$

The stator current density is

$$j_a = \frac{8.2}{8 \times \pi 0.455^2/4} = 6.45 \text{ A/mm}^2$$

The stator winding losses according to eqn (2.49) are

$$\Delta P_w = 3 \times 8.2^2 \times 0.7789 = 157.1 \text{ W}$$

The eddy current losses in stator round conductors according to eqn (2.68) are

$$\Delta P_e = \frac{\pi^2}{4} \frac{47 \times 10^6}{8800} 200^2 \times 0.000455^2 \times 0.75 \times 0.609^2 \times 1.15^2 = 41.1 \text{ W}$$

The friction losses in bearings according to eqn (2.71) are

$$\Delta P_{fr} = 0.06 \times 1.5(3.4 + 0.64) \frac{1000}{60} = 6.1 \text{ W}$$

The windage losses according to eqn (2.74) are

$$\begin{aligned} \Delta P_{wind} &= \frac{1}{2} 0.013 \times 1.2 \left(2\pi \frac{1000}{60} \right)^3 [(0.5 \times 0.115)^5 - (0.5 \times 0.015)^5] \\ &\approx 0.18 \text{ W} \end{aligned}$$

where

$$Re = 1.2 \frac{2\pi(1000/60)}{1.8 \times 10^{-5}} \times 0.115^2 = 9.233 \times 10^4$$

$$c_f = \frac{3.87}{\sqrt{9.233 \times 10^4}} = 0.013$$

The rotational (mechanical) losses according to eqn (2.70) are

$$\Delta P_{rot} = 6.1 + 0.18 = 6.28 \text{ W}$$

The output power is

$$P_{out} = P_{elm} - \Delta P_{rot} = 2148.6 - 6.28 = 2142.3 \text{ W}$$

The shaft torque is

$$T_{sh} = \frac{2142.3}{2\pi \times 1000/60} = 20.46 \text{ Nm}$$

The input power is

$$P_{in} = P_{elm} + \Delta P_w + \Delta P_e = 2148.6 + 157.1 + 41.1 = 2346.7 \text{ W}$$

The efficiency is

$$\eta = \frac{2142.3}{2346.7} = 0.912$$

Numerical Example 6.2

A three-phase, 15-kW, 3600-rpm coreless AFPM motor with Halbach array PM twin rotor has the stator winding resistance per phase $R_1 = 0.24 \Omega$, stator winding leakage reactance $X_1 = 0.92 \Omega$ and armature reaction reactances $X_{ad} = X_{aq} = 0.88 \Omega$. The input phase voltage is $V_1 = 220 \text{ V}$, power factor $\cos \phi = 0.96$, angle between the stator current and q axis $\Psi = 5^\circ$, EMF constant $k_E = 3.363 \text{ V/(rev/s)}$, eddy current losses in stator conductors $\Delta P_e = 184 \text{ W}$ and rotational losses $\Delta P_{rot} = 79 \text{ W}$.

Find the internal voltage E_i excited by the resultant air gap flux and the stator d - and q -axis current adjusted for eddy current losses.

Solution

The phasor diagram showing the EMFs E_f and E_i is drawn in Fig. 6.18. The EMF E_f excited by the rotor magnetic flux

$$E_f = k_E n = 3.363 \times \frac{3600}{60} = 201.78 \text{ V}$$

The synchronous reactance is

$$X_{sd} = X_{sq} = X_{ad} + X_1 = X_{aq} + X_1 = 0.88 + 0.92 = 1.8 \Omega$$

The angle between the stator current and input voltage $\phi = \arccos 0.96 = 16.26^\circ$. The load angle between the voltage and EMF E_f is $\delta = \phi - \Psi = 16.26 - 5 = 11.26^\circ$. The stator currents according to eqns (2.87), (2.88) and (2.89) are:

$$I_{ad} = \frac{220(1.8 \cos 11.26^\circ - 0.24 \sin 11.26^\circ) - 201.78 \times 1.8}{1.8 \times 1.8 + 0.24^2} = 4.5 \text{ A}$$

$$I_{aq} = \frac{220(0.24 \cos 11.26^\circ + 1.8 \sin 11.26^\circ) - 201.78 \times 0.24}{1.8 \times 1.8 + 0.24^2} = 24.5 \text{ A}$$

$$I_a = \sqrt{4.5^2 + 24.5^2} = 24.9 \text{ A}$$

The stator winding losses according to eqn (2.49) are

$$\Delta P_{1w} = 3 \times 24.9^2 \times 0.24 = 445.6 \text{ W}$$

The input power is

$$P_{in} = m_1 V_1 I_a \cos \phi = 3 \times 220 \times 24.9 \times 0.96 = 15\,763.0 \text{ W}$$

The electromagnetic power according to eqn (5.1) is

$$P_{elm} = 15\,763 - 445.6 - 184 = 15\,132.3 \text{ W}$$

The output power according to eqn (5.3) is

$$P_{out} = 15\,132.3 - 79 = 15\,053.3 \text{ W}$$

The efficiency is

$$\eta = \frac{15\,053.3}{15\,763.0} = 0.955$$

The internal voltage E_i excited by the resultant air gap flux according to the phasor diagram (Fig. 6.18) is

$$E_i = \sqrt{(E_f + I_{ad}X_{ad})^2 + (I_{aq}X_{aq})^2}$$

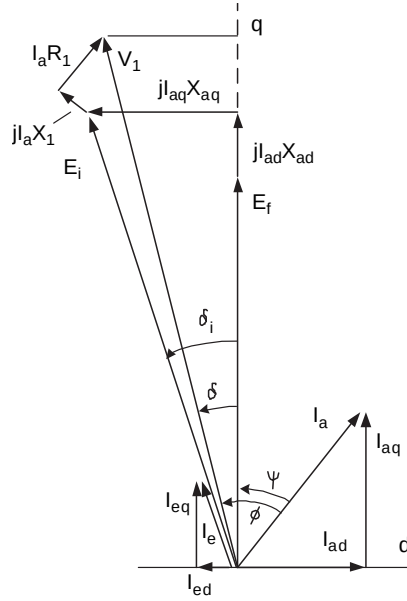


Fig. 6.18. Phasor diagram of a coreless AFPM synchronous motor (eddy current I_e is in phase with the internal voltage E_i). *Numerical example 6.2.*

$$= \sqrt{(201.78 + 4.5 \times 0.88)^2 + (24.5 \times 0.88)^2} = 206.87 \text{ V}$$

The angle δ_i between the EMF E_i and q -axis is

$$\delta_i = \arctan \left(\frac{I_{aq} X_{aq}}{E_f + I_{ad} X_{ad}} \right) = \arctan \left(\frac{24.5 \times 0.88}{201.78 + 4.5 \times 0.88} \right) = 5.97^\circ$$

The eddy current shunt resistance according to eqn (5.32) is

$$R_e = 3 \frac{206.87^2}{184} = 697.75 \text{ } \Omega$$

The current in the vertical branch of the equivalent circuit (Fig. 5.5) found on the basis of the phasor diagram (Fig. 6.18)

$$I_{ed} = -\frac{E_i}{R_e} \sin \delta_i = -\frac{206.87}{697.75} \sin 5.97^\circ = -0.031 \text{ A}$$

$$I_{eq} = \frac{E_i}{R_e} \cos \delta_i = \frac{206.87}{697.75} \cos 5.97^\circ = 0.295 \text{ A}$$

$$I_e = \sqrt{I_{ed}^2 + I_{eq}^2} = \sqrt{(-0.031)^2 + 0.295^2} = 0.296 \text{ A}$$

or

$$I_e = \frac{E_i}{R_e} = \frac{206.87}{697.75} \cos 5.97^\circ = 0.296 \text{ A}$$

The stator current adjusted for eddy current losses found on the basis of the phasor diagram (Fig. 6.18)

$$I'_{ad} = \frac{E_i \cos \delta_i - E_f}{X_{ad}} - \frac{E_i \sin \delta_i}{R_e}$$

$$= \frac{206.87 \cos 5.97^\circ - 201.78}{0.88} - \frac{206.87 \sin 5.97^\circ}{697.75} \approx 4.5 \text{ A}$$

$$I'_{aq} = \frac{E_i \sin \delta_i}{X_{aq}} - \frac{E_i \cos \delta_i}{R_e} = \frac{206.87 \sin 5.97^\circ}{0.88} - \frac{206.87 \cos 5.97^\circ}{697.75} \approx 24.8 \text{ A}$$

$$I'_a = \sqrt{(I'_{ad})^2 + (I'_{aq})^2} = \sqrt{4.5^2 + 24.8^2} \approx 25.2 \text{ A}$$

Control

In the previous chapters it was shown that there is a direct relationship between the rotor speed and the frequency of the phase voltages and currents of the AFPM machine. It was also shown that the magnitude of the induced phase voltages is directly related to the frequency and hence the rotor-speed of the AFPM machine. Therefore, for variable speed operation of the AFPM machine both the frequency and magnitude of the supply voltage must be adjustable. This requires a solid-state converter to be used between the fixed a.c. or d.c. supply and the terminals of the AFPM machine as shown in Fig. 7.1. To have good position and/or speed control of the converter-fed AFPM machine drive, the torque and therefore the current of the machine must be controlled. For current control, information of the phase current as well as the position of the rotor is necessary. Thus, both the current and rotor position of the AFPM machine is sensed and fed back to the controller as shown in Fig. 7.1. The controller in its turn controls the supply voltage and frequency of the AFPM machine via the solid-state converter (inverter).

This chapter focuses on all the control aspects of the AFPM machine drive of Fig. 7.1. The control of the two types of AFPM machines considered in previous chapters, namely the trapezoidal and sinusoidal AFPM machines, is dealt with separately. Finally, position sensorless control is briefly discussed and some numerical examples are given.

7.1 Control of Trapezoidal AFPM Machine

Trapezoidal AFPM machines are surface-mounted PM machines and are characterised by their *trapezoidal or quasi-square EMF waveforms*. Example of these EMF waveforms are shown in Figs 7.2a and 7.2b, which are the open circuit phase voltage waveforms induced by PM flux of a small 1 kW, slotted trapezoidal AFPM machine at half and rated speed respectively. It can be observed, firstly, that at rated speed the frequency and the magnitude of the

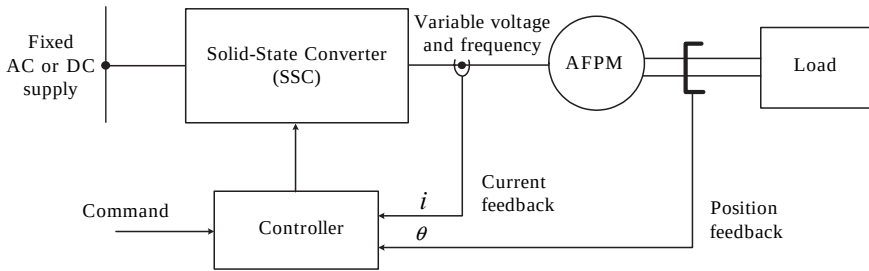


Fig. 7.1. Converter-fed AFPM machine drive.

EMF are double what they are at half rated speed. Secondly, and very important, it can be seen that the shape of the EMF waveform is not sinusoidal, but trapezoidal with a flat-topped voltage of about 120 electrical degrees. It is during this period of 120 electrical degrees that the phase is excited by means of the solid-state converter. This period is therefore called the *conduction period*, as current is flowing in that phase during that period as shown in Fig. 7.2c. Note that the time of conduction or the time of occurrence of the flat-topped voltage is directly related to the position of the rotor; the time of conduction can thus exactly be determined by means of a simple, low resolution position sensor.

During the conduction period one can consider the induced back-EMF phase voltage as a d.c. quantity due to the flat voltage waveform. With a three-phase machine the phase voltages are 120 electrical degrees out of phase, which implies that the conduction periods of two phases are always overlapping (see Fig. 7.2) and that two phases are always active. In the case of AFPM machine drives with 180 electrical degree conduction periods, e.g. [35], then all three phases are active. A different converter should be used in this case. In this section only those trapezoidal AFPM machine drives that have 120 electrical degree conduction periods are considered as they are commonly used.

7.1.1 Voltage Equations

With two phases of the Y-connected three-phase machine always active (the third phase is switched off by means of the solid-state converter), the power circuit of the trapezoidal AFPM brushless machine can be represented by the electric circuit of Fig. 7.3a. Here, phases *A* and *B* are active and phase *C* is switched off (shown in dotted lines). In this circuit R_1 and L_s are respectively the phase resistance and phase synchronous inductance. The synchronous inductances in the *d* and *q*-axis are equal, i.e. $L_{sd} = L_{sq} = L_s$. The voltages e_{fA} , e_{fB} and e_{fC} are the EMFs per phase. As phases *A* and *B* are in their conduction periods, both e_{fA} and e_{fB} are at their flat-topped values. Hence, $e_{fA} = E_f^{(tr)}$ and $e_{fB} = -E_f^{(tr)}$ where $E_f^{(tr)}$ is the flat-topped value of the

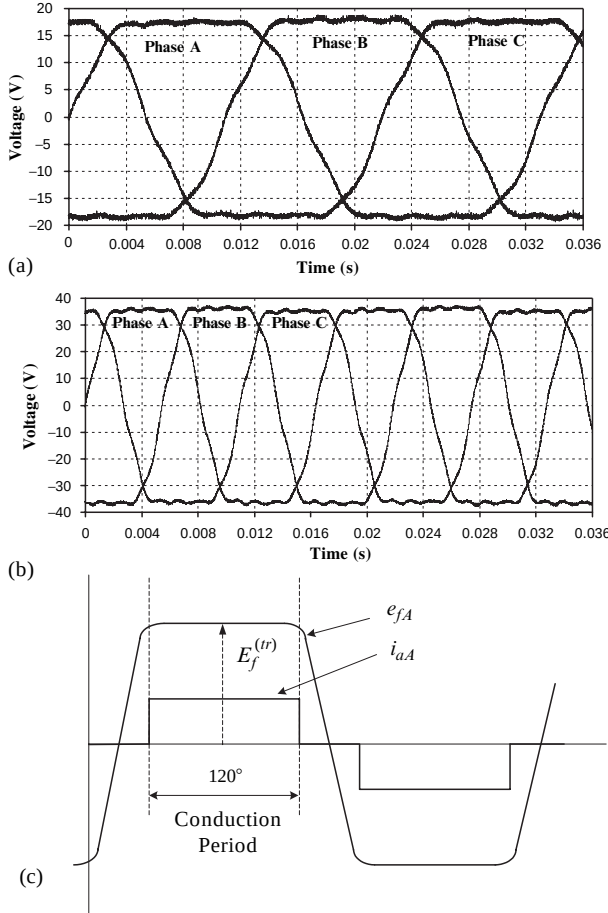


Fig. 7.2. Open-circuit induced phase voltage waveforms of a 1-kW trapezoidal AFPM machine obtained from measurements: (a) at half speed (30 Hz); (b) at rated speed (60 Hz); (c) theoretical EMF and phase current waveforms.

trapezoidal voltage waveform (Fig. 7.2c). Combining the resistances, inductances and induced voltages of Fig. 7.3a results in the equivalent circuit of Fig. 7.3b in which E_{fL-L} is taken as a d.c. voltage equal to $E_{fL-L} = 2E_f^{(tr)}$, $R_p = 2R_1$, $L_p = 2L_s$, $i_a = i_{aA} = -i_{aB}$ and $v_p = v_{1L-L}$. The dynamic voltage equation of this circuit is

$$\begin{aligned} v_p &= 2R_1 i_{aA} + 2L_s \frac{di_{aA}}{dt} + 2E_f^{(tr)} \\ &= R_p i_a + L_p \frac{di_a}{dt} + E_{fL-L} \end{aligned} \quad (7.1)$$

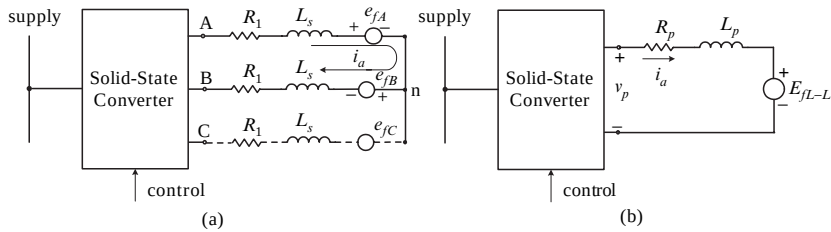


Fig. 7.3. Trapezoidal AFPM machine connected to the solid-state converter with two phases active: (a) electrical circuit; (b) combined equivalent circuit for the analysis.

Applying a d.c. voltage $v_p = V_p$ across terminals *A* and *B* by means of the solid-state converter will imply that ideally in the steady state, a d.c. current $i_a = I_a^{(sq)}$ will flow through the two phase windings according to the following steady-state voltage equation derived from eqn (7.1), i.e.

$$V_p = R_p I_a^{(sq)} + E_{fL-L} \quad (7.2)$$

The induced voltage E_{fL-L} is given by eqn (2.125), the electromagnetic power of the machine is $P_{elm} = E_{fL-L} I_a^{(sq)}$ and the developed torque is given by eqn (2.127). Note that $i_a = I_a^{(sq)}$ and E_{fL-L} are instant quantities that can change with time, similar to a brush-type d.c. machine.

7.1.2 Solid-State Converter

The complete power circuit connected to the three-phase trapezoidal AFPM brushless machine is shown in Fig. 7.4. In this circuit the fixed three-phase a.c. supply is first rectified by means of a three-phase diode rectifier to obtain a fixed d.c. voltage. The d.c. voltage is then inverted to a three-phase a.c. trapezoidal voltage. With the diode rectifier no power can flow back to the fixed a.c. supply if the AFPM machine is operating in generator mode. This means that any generated power from the machine must be dissipated at the d.c. link of the converter. This is done by switching a braking resistance at the d.c. link as shown in Fig. 7.4. If it is required that the generated power must flow back to the fixed a.c. supply, then use must be made of a so-called active rectifier. In the rest of this section only the switching and control of the d.c. to a.c. inverter of Fig. 7.4 is studied, assuming a fixed d.c.-link voltage.

The three-phase inverter of Fig. 7.4 has six power electronic transistor switches. Nowadays Insulated-Gate Bipolar Transistors (IGBTs) or MOS field effect transistors (MOSFETs) are very popular for use as switches for the inverter. The power transistor switches are switched on and off at a relatively high switching frequency (typically from 1 to 20 kHz or even higher). Either the top transistor of an inverter-leg is switched on and the bottom transistor

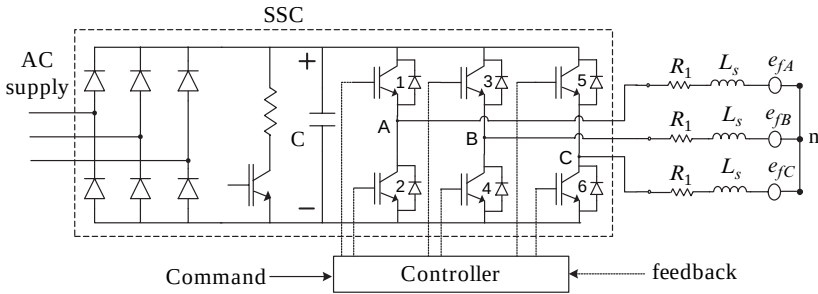


Fig. 7.4. Solid-state a.c.-a.c. converter and equivalent three-phase AFPM machine connection.

is switched off, or vice versa, or both transistors are switched off. It was explained in the previous sections that only two phases of the converter-fed trapezoidal AFPM machine drive are active at any time. The third phase is switched off by switching off the top and bottom transistors of the inverter-leg connected to that phase winding. The three-phase inverter of Fig. 7.4 also has six anti-parallel diodes to make current flow possible when a switch is turned off (this is due to inductive stored energy) and also to make the generator (braking) operation of the drive possible.

The circuit diagrams of Figs 7.3 and 7.4 show that the trapezoidal AFPM machine drive can be analysed by considering only the two active phases of the machine and the two active switching legs of the inverter. The circuit then reduces to the general circuit of Fig. 7.5.

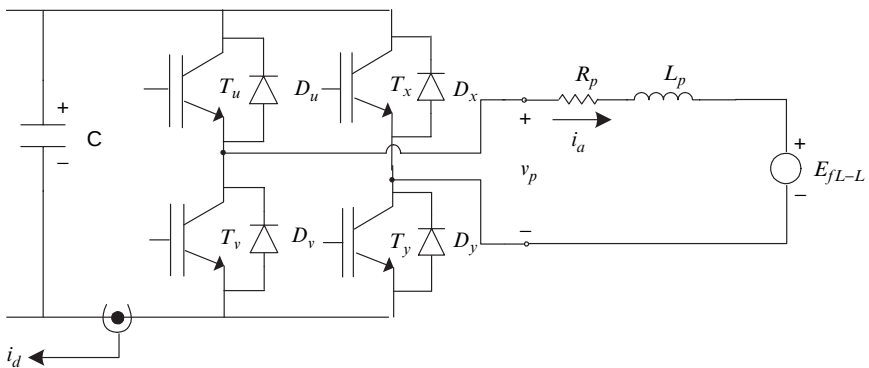


Fig. 7.5. Equivalent circuit of converter-fed trapezoidal AFPM machine (two phases active).

One method of controlling the power semiconductor switches of the inverter shown in Fig. 7.5 is explained as follows. Consider in the forward speed direction that the voltage E_{fL-L} and the average current $i_{a(avg)} = I_a^{(sq)}$ are positive, then the average supply voltage $v_{p(avg)}$ also has to be positive according to the steady-state equation, eqn (7.2), i.e. $v_{p(avg)} = V_p > E_{fL-L}$. A positive average supply voltage can be obtained by switching on transistor T_y and pulse-width-switching (PWM-switching) transistors T_u and T_v of Fig. 7.5 at duty cycles D and $(1 - D)$ respectively, such that in the steady-state

$$DV_d = v_{p(avg)} \quad (7.3)$$

where

$$D = \frac{t_{on}}{t_s} = t_{on}f_s \quad (7.4)$$

and t_{on} is the switching-on time, t_s is the switching period and f_s is the switching frequency of the transistors. This switching and the switching waveform of the supply voltage v_p across the terminals of the two phases of the machine are shown in Fig. 7.6(a). As is shown $v_{p(avg)}$ is the switching period average of the voltage v_p . Note that in Fig. 7.6 the drive is not in the steady-state as the instant average current $i_{a(avg)}$ and the average voltage $v_{p(avg)}$ change with time. Also note that the drive system only reacts on the average of the supply voltage and not on the high frequency components thereof, as the switching frequency is relatively high and the time constant of the mechanical system is relatively large.

Let us consider next the flow of current in the converter and the machine in motoring mode. When transistors T_u and T_y of Fig. 7.5 are switched on, current flows from the d.c. supply through T_u and through the machine winding and T_y , and then back to the supply. If T_v and T_y are switched on, the current then free-wheels through diode D_v and through the machine and T_y . In this case no current is flowing from or to the d.c. supply. An example of the phase winding and d.c.-link current waveforms in motoring mode is shown in Fig. 7.6a. As mentioned before $i_{a(avg)}$ is the instant average of the phase current i_a .

In generating (braking) mode with E_{fL-L} positive, the duty cycle D of eqn (7.3) is adjusted in such a way that $v_{p(avg)} < E_{fL-L}$ to allow $i_{a(avg)}$ to become negative as shown in Fig. 7.6b. In this case D_y is always conducting and the current either flows through T_v when switched on, or through D_u to the supply when switched off. An example of the d.c.-link current waveform in generating mode is shown in Fig. 7.6b.

When operating the machine in the reverse speed direction, in which case E_{fL-L} is negative with respect to the polarity given in Fig. 7.5, transistor T_v is switched on during the conduction cycle and transistors T_x and T_y are PWM-switched. In this case the supply voltage $v_{p(avg)}$ is negative and the current can be controlled again to be positive or negative in the same way as in the forward speed mode.

The advantage of the above switching method is that only one inverter-leg is switched at a high frequency at any time, keeping the switching losses of the inverter low. Which inverter-leg of Fig. 7.5 should be selected to be PWM-switched is simply a logic decision that depends on the sign (positive or negative) of the supply voltage $v_{p(avg)}$. Note that $v_{p(avg)}$ is controlled directly by the controller of the drive (Fig. 7.1).

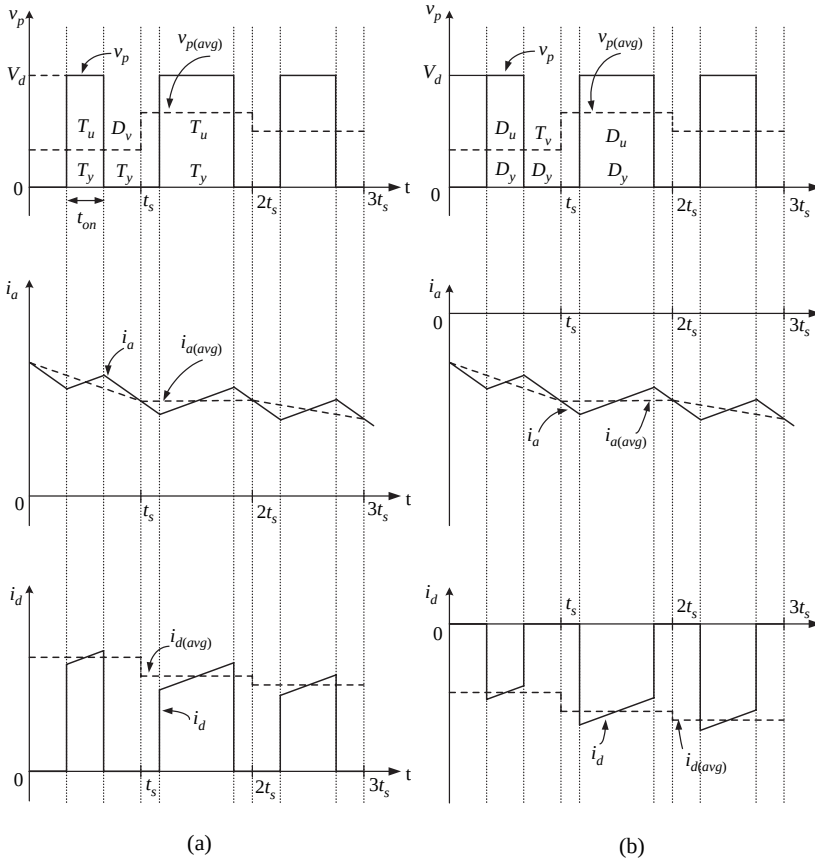


Fig. 7.6. Voltage and current waveforms of a converter-fed trapezoidal AFPM machine: (a) motoring mode; (b) generating mode (see Fig. 7.5).

7.1.3 Current Control

For good speed control of any drive a *fast torque control* is necessary. According to eqn (2.127) the torque of the trapezoidal AFPM machine drive can be controlled directly by controlling the stator current i_a and hence the instant average current $i_{a(avg)}$ explained in Fig. 7.6. Close examination of the

currents of Fig. 7.6 reveals that the required information regarding the current $i_{a(avg)}$ is also present in the average d.c.-link current $i_{d(avg)}$, i.e. if the sample-measurement of current i_d is at the right instant. The current of all the phases of the machine can be controlled by just controlling the d.c.-link current $i_{d(avg)}$. Thus, only one current sensor is necessary. One aspect of this method of current sensing, however, must be kept in mind. If the machine is operating, e.g. in forward motoring mode or reverse motoring mode, then in both cases the current $i_{d(avg)}$ will be positive. This creates a problem from a control point of view as the reference current $i_{a(avg)}^*$ will be, say, positive for forward motoring mode, but negative for reverse motoring mode. However, this problem can easily be solved in the controller by always multiplying the d.c.-link current signal with the sign (positive or negative) of the controlled supply voltage $v_{p(avg)}^*$ signal.

A block diagram of the single sensor current controller is shown in Fig. 7.7. The current controller acts on the comparison between the desired current $i_{a(avg)}^*$ and the measured d.c.-link current $i_{d(avg)}$, and then controls the supply output voltage by controlling the signal $v_{p(avg)}^*$. The PWM generator outputs the PWM signal in terms of a logic 1 or 0 to the programmable logic device (PLD). The sign (positive or negative) of $v_{p(avg)}^*$ is also an logic input (1 or 0) to the logic device. Three logic position signals s_a , s_b and s_c are also inputs to the PLD. These three signals are received from three Hall-effect sensors mounted in the machine. The signals supply information about which two phases must be active according to the rotor position. The outputs of the PLD directly control the switching of the six power semiconductor switches.

Two aspects in the block diagram of Fig. 7.7 have not been referred to yet. The first is that a speed voltage signal is added to the output of the current regulator. This is to decouple the current regulator from the effects of speed variation. The gain, K_i , represents the gain of the solid-state inverter. The second aspect is that the d.c.-link current is *sampled* at certain times to get the average value $i_{d(avg)}$ as shown in Fig. 7.6.

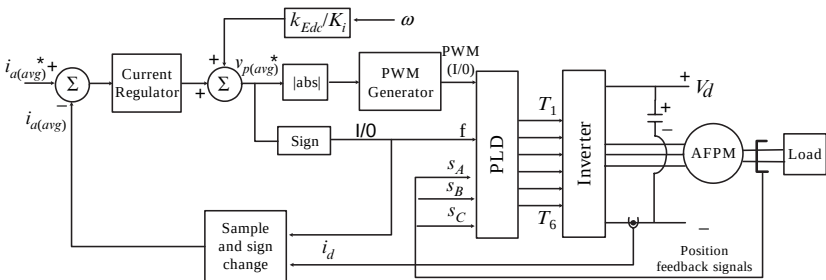


Fig. 7.7. Block diagram of single sensor current controller for trapezoidal AFPM.

To determine the parameters of the current regulator the transfer function of the current control system must be obtained. According to Fig. 7.3b and eqn (7.1), the dynamic voltage equation of the machine in terms of average components is

$$v_{p(avg)} = R_p i_{a(avg)} + L_p \frac{d}{dt} i_{a(avg)} + E_{fL-L} \quad (7.5)$$

where $v_p = v_{p(avg)} + v_r$, $i_a = i_{a(avg)} + i_r$, and v_r and i_r are the ripple voltage and ripple current respectively (see Fig. 7.6). In Laplace transform eqn (7.5) becomes

$$v_{p(avg)}(s) = (R_p + L_p s) i_{a(avg)}(s) + E_{fL-L}(s) \quad (7.6)$$

With the use of the decoupling speed voltage signal in the current control system of Fig. 7.7, the speed voltage term $E_{fL-L}(s)$ of eqn (7.6) has no effect on the current regulator's response. Hence, the control block diagram of the current control system can be represented by the block diagram shown in Fig. 7.8. In this block diagram the current regulator can be, e.g. a gain or a PI-regulator. The gain K_i represents the voltage gain of the solid-state converter (inverter) as mentioned before. The transfer function of the current control system can be determined from the block diagram of Fig. 7.8 and the parameters of the current regulator can be designed according to classical control system methods.

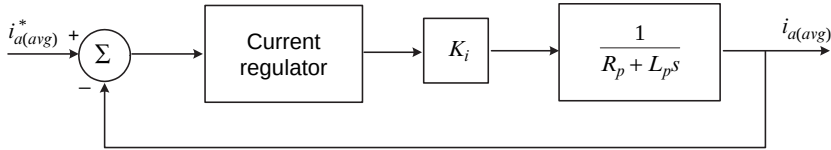


Fig. 7.8. Block diagram of decoupled current control system.

7.1.4 Speed Control

With the current control system in place, as explained in section 7.1.3, a speed control loop can be applied around the current control loop. The electrical and mechanical torque balance equations for the system are as follows:

$$T_{d(avg)}(s) = k_{Tdc} i_{a(avg)}(s) \quad (7.7)$$

$$T_{d(avg)}(s) - T_L(s) = (B + Js)\omega(s) \quad (7.8)$$

where $T_L(s)$ is the load torque, B is the equivalent total damping of the system, J is the equivalent total moment of inertia of the system and ω is the rotor speed in rad/s. For a two-pole machine the angular mechanical speed is equal to the angular frequency of the stator currents. Note that eqns (7.7) and (7.8) refer to the instant average components of current, torque and speed, and not to the high frequency ripple components due to the inverter switching. From equations (7.7) and (7.8) and from Fig. 7.8, the speed control block diagram of the system can be represented as in Fig. 7.9. Hence, the transfer function of the speed system can be determined and the parameters of the speed regulator can be designed to obtain a certain speed response. A speed signal can be obtained electronically from the signals of the three Hall-effect position sensors installed in the machine.

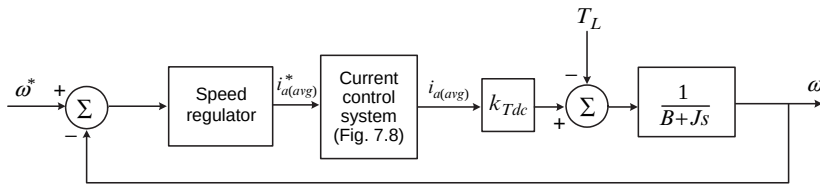


Fig. 7.9. Block diagram of speed control system.

7.1.5 High Speed Operation

Equations (2.125), (7.2) and (7.3) show that as the speed of the machine increases, the EMF E_{fL-L} , the voltage $v_{p(avg)}$ and the duty cycle D will increase till the speed is reached where D becomes unity. The speed where D becomes unity is simultaneously the speed where the output voltage of the converter (inverter) reaches a maximum. This speed is often called the *base speed* of the drive. Field or flux weakening in the high speed region of the machine to keep $E_{fL-L} = \text{constant}$ is not possible with high-energy-product surface mounted PMs. Therefore, at speeds higher than the base speed, $E_{fL-L} > V_d$, and control of the current and torque of the machine is lost. In fact, the current and torque rapidly drop to zero at speeds slightly above the base speed. This implies that the speed-range of the trapezoidal AFPM machine using surface mounted PMs is very limited. Techniques to increase the speed substantially beyond base speed are: (a) change smoothly from a 120° current conduction period, as discussed earlier in this chapter, to a 180° six-step conduction period [19, 97, 233] and (b) advance the current wave with respect to the EMF wave [136]; note that during sub-base speed operation the current wave is controlled to be in-phase with the EMF waveform as shown in Fig. 7.2c. The disadvantages of these techniques may be the following: (i) an increase in machine torque ripple, (ii) danger of converter switches being

exposed to high DC-link voltages if for some reason the converter switches are switched-off at high speeds (at this condition the high EMFs are rectified by the six diodes of the inverter to a high d.c.-link voltage) and (iii) a higher resolution position sensor is necessary for the smooth advancing of the current wave as the speed increases.

A technique to obtain high speed operation that avoid the disadvantages of the above techniques and obtain true flux weakening is to increase mechanically the air gap of the AFPM machine with increase in speed; this technique is not possible in commercial RFPM machines. However, it is questionable if this technique is economically viable.

7.2 Control of Sinusoidal AFPM Machine

Sinusoidal AFPM machines are designed to have *sinusoidal or near sinusoidal EMF waveforms*. Therefore, in variable speed drive applications where a solid-state converter is used to supply the machine (Fig. 7.1), the output voltage of the converter must be sinusoidal or sinusoidal PWM modulated. Hence, the current in the machine is sinusoidal or else the current regulators of the drive will force the phase currents to be sinusoidal.

This section focuses on the control aspects of the sinusoidal AFPM machine drive. First, the modelling and equivalent circuits of the sinusoidal AFPM machine are described. Secondly, the current and speed regulators of the drive are dealt with. Finally, the hardware of the sinusoidal AFPM machine drive is considered.

7.2.1 Mathematical Model and dq Equivalent Circuits

The phase voltage equations of any three-phase a.c. electrical machine can be written as

$$v_{1ABC} = R_1 i_{aABC} + \frac{d\psi_{ABC}}{dt} \quad (7.9)$$

where ψ_{ABC} is the total stator magnetic flux linkage of phase A, B or C – eqn (5.20). Eqn (7.9) is expressed in the so-called ABC stationary reference frame where the circuit variables (voltage, current and flux linkage) are expressed in a reference frame that is fixed to the stationary stator. For synchronous machines, like the AFPM machine, it is very convenient to transform or refer the stationary stator variables to a reference frame that is fixed to the rotor. In this transformation the stationary ABC stator windings of the machine are replaced by fictitious $dq0$ windings that rotate with the rotor.

To explain this, let us consider the cross-section of a two-pole AFPM machine as shown in Fig. 7.10a (side view) and Fig. 7.10b (front view). Fig. 7.10a shows that the PMs are embedded into the rotor steel yoke, but the magnets in the sinusoidal AFPM machine may also be on the surface of the

yoke like in a trapezoidal AFPM machine. With the magnets embedded, the AFPM machine is not a pure PM machine anymore, but a combined PM-reluctance machine because the electromagnetic torque has two components: PM synchronous torque and reluctance torque.

Fig. 7.10 explains the ABC and dq winding layout of the machine. The rotating d and q -axis windings are in quadrature of each other and their magnetic axes are aligned with the d and q -axis of the rotor. In this case the d -axis of the rotor is chosen to be aligned with the centre lines of the PMs and the q -axis between the magnets. The position of the rotor and dq -windings with respect to the stationary ABC windings is defined in Fig. 7.10 as the angle between the rotating d -axis and a defined phase A -axis. Note, that in Fig. 7.10b the direction of the flux (Φ) is into (or out of) the paper (z -direction), while the current is in the xy -plane.

To transfer the circuit variables from the stationary ABC reference frame to the $dq0$ reference frame, or vice versa, the *Park transformation*

$$[f_{dq0}] = [K_p] [f_{ABC}] \quad (7.10)$$

and inverse Park transformation

$$[f_{ABC}] = [K_p]^{-1} [f_{dq0}] \quad (7.11)$$

are used [94], where $[K_p]$ is according to eqn (5.26) and $[K_p]^{-1}$ is the inverse matrix of $[K_p]$. In the above eqns (7.10) and (7.11), f_{dq0} and f_{ABC} are the $dq0$ and ABC variables of voltage, current or flux linkage respectively. In eqn (5.26) the angle θ is the electrical angle defined in Fig. 7.10. With a constant rotor angular velocity ω the angle $\theta = \omega t$. For a balanced three-phase AFPM machine it is only necessary to consider the d and q -axis circuits as the zero-sequence circuit is playing no role in the performance of the machine. Applying eqn (7.10) to the voltage eqn (7.9) leads to the dq voltage equations (5.5) and (5.6) of the balanced AFPM machine (omitting the zero-sequence component). In eqns (5.5) and (5.6) ω is the electrical speed of the rotor, and ψ_d and ψ_q are the d and q -axis flux linkages. Let us consider the following three cases of operating conditions of the AFPM machine:

- (1) The first case is when the machine is open circuit and is mechanically driven at a certain speed ω . The dq stator currents, thus, are zero and only the magnets are active. From Fig. 7.10 it is clear that the magnet flux will only link the d -axis winding and not the q -axis winding. In this case $\psi_q = 0$ and $\psi_d = \psi_f$, and from eqns (5.5) and (5.6) $v_{1d} = 0$ and $v_{1q} = \omega\psi_f$.
- (2) The second case is when both the magnets and the d -axis winding are not active, but d.c. (steady-state) current is flowing in the q -axis stator winding. The dq stator windings and the rotor are at a rotating speed of ω . Clearly with no magnets and no d -axis current, $\psi_d = 0$. With positive d.c. q -axis current, i.e. $i_{aq} = \sqrt{2}I_{aq}$, then $\psi_q = \sqrt{2}L_{sq}I_{aq}$ where L_{sq} is the

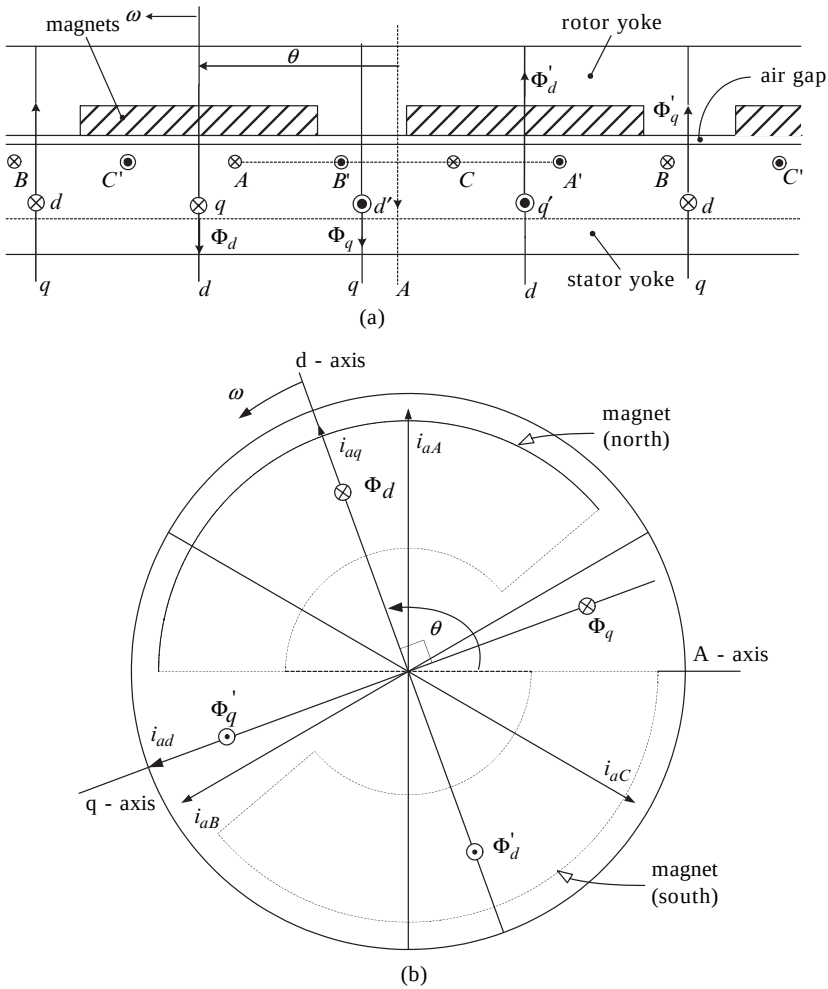


Fig. 7.10. Stationary ABC and rotating dq windings of a simple two-pole AFPM machine: (a) side view; (b) front view (PMs are shown in dotted lines).

- q -axis winding synchronous inductance. Note that the relation between ψ_q and i_{aq} is nonlinear due to magnetic saturation; L_{sq} will thus vary with saturation. From eqns (5.9) and (5.10) it follows in this case that $v_{1q} = R_1 i_{aq}$ and $v_{1d} = -\omega L_{sq} i_{aq}$.
- (3) In the third case only the d -axis winding is active in terms of a positive d.c. current flowing of $i_{ad} = \sqrt{2} I_{ad}$. Therefore $\psi_q = 0$ as the magnets are not active and $i_{aq} = 0$, and $\psi_d = \sqrt{2} L_{sd} I_{ad}$ where L_{sd} is the synchronous inductance of the d -axis winding. Again, the relation between ψ_d and i_{ad}

is nonlinear and L_{sd} will vary with saturation. In this case from eqn (5.9) and (5.10) $v_{1q} = \omega L_{sd} i_{ad}$ and $v_{1d} = R_1 i_{ad}$.

If the magnets and the dq stator windings are active, then the above three cases can be summarized by eqns (5.7) and (5.8). From eqns (5.5) to (5.10) it follows that the electromagnetic power of a three-phase machine is given by

$$P_{elm} = \frac{3}{2} (i_{aq} \omega \psi_d - i_{ad} \omega \psi_q) = \frac{3}{2} \omega (\psi_d i_{aq} - \psi_q i_{ad}) \quad (7.12)$$

The factor $3/2$ has been explained in Section 5.13. The developed torque of the machine is expressed by eqn (5.13). The first term in eqn (5.13) expresses the torque component due to PM excitation, while the second term expresses the reluctance torque component. For surface mounted AFPM machines it can be assumed that $L_{sd} = L_{sq}$, hence the torque is directly proportional to i_{aq} . In the case of an embedded AFPM machine like in Fig. 7.10a it is clear that $L_{sd} < L_{sq}$, which implies that i_{ad} in eqn (5.13) must be negative in order to generate a positive reluctance torque. Fig. 7.11a shows the phasor diagrams of the motor for $i_{ad} = 0$ while Fig. 7.11b shows the phasor diagram of the motor at unity power factor.

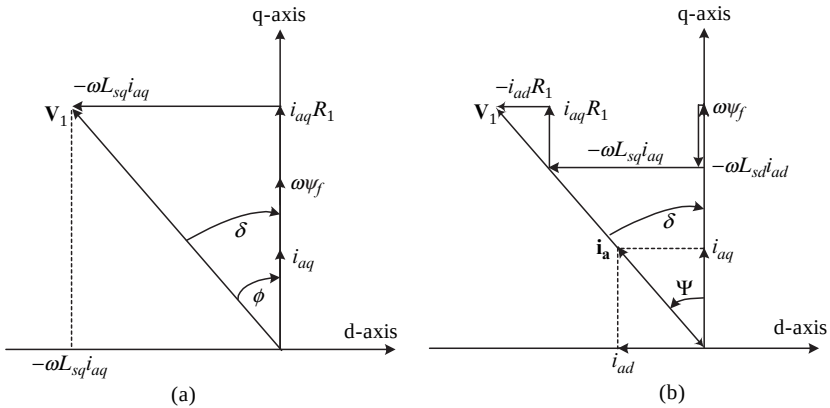


Fig. 7.11. Phasor diagrams for motoring action of overexcited AFPM machine: (a) for $i_{ad} = 0$; (b) at unity power factor. Note that δ is negative for motoring action and positive for generation action.

The inductances in the d and q -axis windings are expressed by eqns (5.17) and the magnet flux linkage ψ_f is further expressed as

$$\psi_f = L_{ad} i_f \quad (7.13)$$

where L_{ad} is the armature reaction (mutual) inductance in the d -axis and i_f is a fictitious magnetising current representing the PM excitation.

The flux linkage and mutual inductance between the d -axis winding and the fictitious permanent magnet (field) winding need further explanation. For the d -axis winding we have the following equation (ignoring cross coupling effects with the q -axis winding and ignoring any stator slotting effects):

$$\begin{aligned}\frac{d\psi_d}{dt} &= L_{sd}\frac{di_{ad}}{dt} + L_{fd}\frac{dI_f}{dt} \\ &= (L_l + L_{ad})\frac{di_{ad}}{dt} + L_{fd}\frac{dI_f}{dt}\end{aligned}\quad (7.14)$$

where L_{ad} is given by eqn (2.116) and L_l is a leakage inductance. I_f is the fictitious magnet (field) winding current. L_{fd} is the mutual inductance between the d -axis winding and the fictitious magnet (field) winding given by

$$L_{fd} = \frac{k_{wf}N_f\Phi_{df}}{I_{ad}} = \frac{k_{w1}N_1\Phi_{fd}}{I_f}\quad (7.15)$$

where $k_{wf}N_f$ and $k_{w1}N_1$ are the effective turns in series of the fictitious magnet-field and d -axis windings respectively. Φ_{df} and Φ_{fd} are the proportions of flux produced by the d -axis and magnet-field windings respectively that link with the opposite winding. Note that in this case Φ_{fd} is equal to Φ_f of eqn (2.26). From eqn (7.15) the flux linkage of the d -axis winding due to the field winding, ψ_f , is given by

$$\psi_f = k_{w1}N_1\Phi_{fd} = L_{fd}I_f\quad (7.16)$$

However, this flux linkage is also expressed by eqn (7.13) and is represented in the d -axis equivalent circuit of Fig. 7.12b. Hence, from eqns (7.13) and (7.16)

$$i_f = \frac{L_{fd}}{L_{ad}}I_f\quad (7.17)$$

Furthermore, the d -axis inductance L_{ad} is given by eqn (2.116) as

$$L_{ad} = \frac{k_{w1}N_1\Phi_{ad}}{I_{ad}}\quad (7.18)$$

where Φ_{ad} is the d -axis flux passing the air gap, linking the rotor. The total flux of the d -axis winding, Φ_d , is given by

$$\Phi_d = \Phi_l + \Phi_{ad}\quad (7.19)$$

where Φ_l is the leakage flux. Now, if Φ_{df} of eqn (7.15) [this is the proportion of the flux produced by the d -axis winding that links the field winding] is equal to $k\Phi_{ad}$, i.e. $\Phi_{df} = k\Phi_{ad}$, where $0 < k < 1$ (i.e. not all the d -axis flux linking the rotor necessarily links the fictitious magnet-field winding), then from eqns (7.15) and (7.18) it follows that

$$L_{ad} = \left[\frac{k_{w1}N_1}{k_{wf}N_fk} \right] L_{fd} \quad (7.20)$$

and furthermore from eqn (7.17) that

$$i_f = \left[\frac{k_{wf}N_fk}{k_{w1}N_1} \right] I_f \quad (7.21)$$

Hence, if the effective turns of the d -axis and field windings are the same and $k = 1$, then L_{ad} is equal to L_{fd} and i_f is equal to I_f .

Using eqns (5.7), (5.8) and (7.13) and assuming i_f is not varying with time, eqns (5.5) and (5.6) can be written as

$$v_{1d} = R_1 i_{ad} + (L_l + L_{ad}) \frac{di_{ad}}{dt} - \omega(L_l + L_{aq}) i_{aq} \quad (7.22)$$

$$v_{1q} = R_1 i_{aq} + (L_l + L_{aq}) \frac{di_{aq}}{dt} + \omega(L_l + L_{ad}) i_{ad} + \omega L_{ad} i_f \quad (7.23)$$

On the basis of eqns (7.22) and (7.23) the complete dq equivalent circuits of the sinusoidal AFPM machine can be drawn as shown in Fig. 7.12. It should be noted that the modelling according to eqns (7.22) and (7.23) is based on assumptions that

- (a) there are no core losses;
- (b) the magnetic circuit is linear, i.e. there is no saturation of the magnetic circuit;
- (c) there is no cross magnetisation or mutual coupling between the d - and q -axis circuits;
- (d) there are no stator slotting effects (smooth stator core).

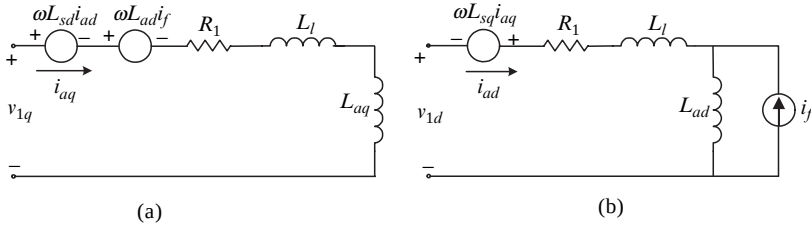


Fig. 7.12. The d and q -axis equivalent circuits of the sinusoidal AFPM brushless machine.

7.2.2 Current Control

The equivalent circuits of the sinusoidal AFPM machine shown in Fig. 7.12 can be used to design the current regulators for controlling the d and q -axis currents of the machine. Applying the principle of speed voltage decoupling, as for the trapezoidal AFPM machine (section 7.1.3), the dq current control system of the sinusoidal AFPM machine can be represented by the block diagram of Fig. 7.13. Here, the d and q -axis current regulators act on the errors between the desired and actual dq currents of the machine. The speed voltage terms are added (or subtracted) to the outputs of the regulators to determine the dq machine voltages. Within the machine model the speed voltages are subtracted (or added) again, hence the regulators act only on the RL -parts of the equivalent circuits. The block diagrams of the decoupled current controllers can therefore be simplified to those shown in Fig. 7.14. With the equivalent circuit parameters of the machine known, the transfer functions of the d and q -axis current controllers can be determined and the d and q -axis current regulators can be designed accordingly to obtain the required dq current response.

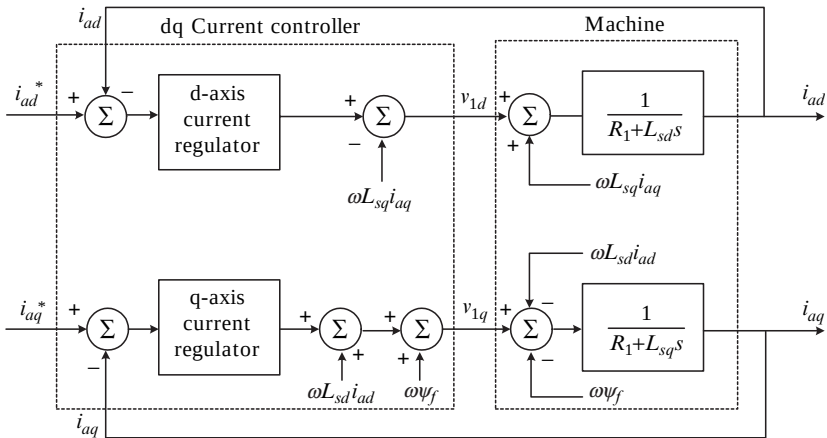


Fig. 7.13. Block diagram of d and q -axis current control system of the AFPM drive implementing speed voltage decoupling.

7.2.3 Speed Control

It is always good in terms of efficiency to control the drive at maximum torque per ampere. For the surface mounted (nonsalient) AFPM machine, where it can be taken that $L_{sd} = L_{sq}$, it is clear from eqn (5.13) that maximum torque per ampere for this machine is obtained by keeping $i_{ad} = 0$. However, for

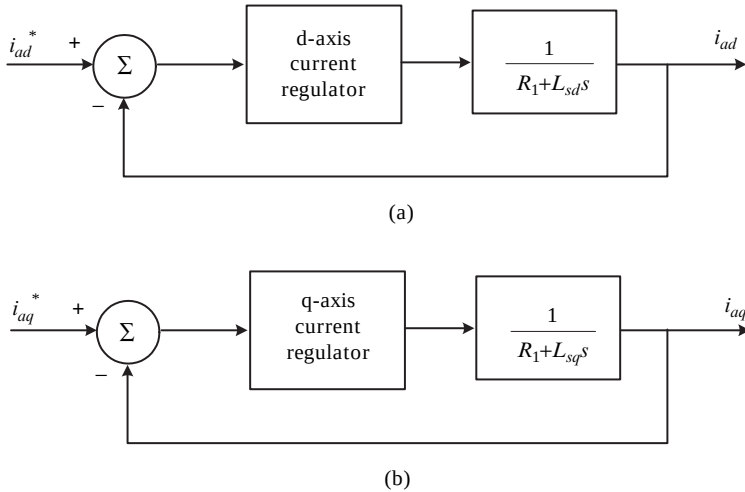


Fig. 7.14. Decoupled d and q -axis current control systems.

embedded (salient-pole) AFPM machines where $L_{sd} \neq L_{sq}$, maximum torque per ampere is obtained by making i_{ad} negative, that is by controlling the current phasor at some *current angle* Ψ as shown in Fig. 7.15 (see also Fig. 7.11b). Note from Fig. 7.15 that negative torque is obtained by keeping both i_{ad} and i_{aq} negative, or in the case of the nonsalient machine, only i_{aq} is negative. In the sub base speed region ($\omega < \omega_b$) of the drive, where there is enough voltage available so that the current regulators do not saturate, the machine is controlled to operate at maximum torque per ampere. For the nonsalient AFPM machine this implies that the current angle $\Psi = 0$, while for the salient-pole machine $\Psi = \Psi_0$ (see Fig. 7.15), which is an (average) optimum angle for all load conditions.

However, in the high speed region ($\omega > \omega_b$), the EMF voltage $\omega L_{ad} i_f = \omega \psi_f$ of eqn (7.23) (see also Fig. 7.11b) becomes larger than the possible output voltage of the inverter. The current regulators therefore start to saturate, current control is lost and the torque of the machine rapidly drops as the speed increases. To avoid this torque decrease and to ensure a wider high-speed region one has to deviate from the maximum torque per ampere control and start to inject a larger negative d -axis current to reduce the q -axis terminal voltage (Fig. 7.11b) as the speed increases. This implies that the current angle Ψ must be increased as shown in Fig. 7.15 such that always

$$2\omega \sqrt{(L_{sd}i_{ad} + L_{ad}i_f + i_{aq}R_1)^2 + (L_{sq}i_{aq} + i_{ad}R_1)^2} < V_d$$

where V_d is the dc-link voltage of the inverter, and keeping the current amplitude $I_{am} = \sqrt{i_{ad}^2 + i_{aq}^2}$ equal or less than the rated current amplitude

depending on the load. At a maximum current angle, Ψ_m , the speed range can be extended much more if the current angle is held constant at $\Psi = \Psi_m$ and the amplitude of the phase current is reduced as shown in Fig. 7.15. Note from Fig. 7.11b that as the current angle Ψ increases the apparent power of the machine decreases and the power factor improves.

The above current control strategy can be used for both salient and non-salient AFPM machines. It must be noted, however, that the speed range obtained with the above control method depends very much on the value of the stator phase inductance of the machine (see *Numerical example 7.2*). If the phase inductance is relatively small (this implies a low per unit stator phase reactance), as is the case with slotless and air-cored AFPM machines, the high speed range of the machine is very limited.

From above, thus, the current of the machine can be controlled by controlling the magnitude of the current and keeping the current angle constant at an optimum value, and advancing the current angle at higher speeds. To simplify this control method, the current angle Ψ is defined as a positive angle between 0 to 90°. The current angle is also taken as a function of the *absolute value* of the speed, $|\omega|$. Furthermore, a current i_T is defined, which controls the torque of the machine and is proportional to the current amplitude of the machine. Hence,

$$\begin{aligned} i_{ad} &= -|i_T| \sin \Psi \\ i_{aq} &= i_T \cos \Psi \quad \text{for } \Psi \geq 0 \end{aligned} \quad (7.24)$$

Substituting eqn (7.24) into eqn (5.13) and noting that $L_{sd} \leq L_{sq}$, the electromagnetic torque is

$$T_d = \frac{3}{2} p \psi_f i_T \cos(\Psi) + \frac{3}{4} p (L_{sq} - L_{sd}) i_T |i_T| \sin(2\Psi) \quad (7.25)$$

It can be seen from the first term of eqn (7.25) that there is a linear relation between the PM torque component and the current i_T for a given current angle Ψ . The second term of eqn (7.25) is the reluctance component of the torque. It is found for reluctance machines [143] that the inductance difference $\Delta L = L_{sq} - L_{sd}$ is not a constant at a certain current angle, but varies inversely with load current i_T due to saturation and cross magnetization. The cross magnetization or cross coupling is the magnetic coupling between the fictitious d and q -axis windings of the machine. It is thus a fairly good approximation to assume for this component of the torque a linear relation between reluctance torque and current i_T for a given current angle Ψ [145]. Thus, eqn (7.25) can be simplified as

$$T_d \approx k_{Tf} i_T + k_{Trel} i_T \approx k_T i_T \quad \text{for } \Psi = \text{constant} \quad (7.26)$$

where k_{Tf} and k_{Trel} are the PM torque and reluctance torque constants, respectively, and $k_T = k_{Tf} + k_{Trel}$. Note from eqn (7.26) that to generate a

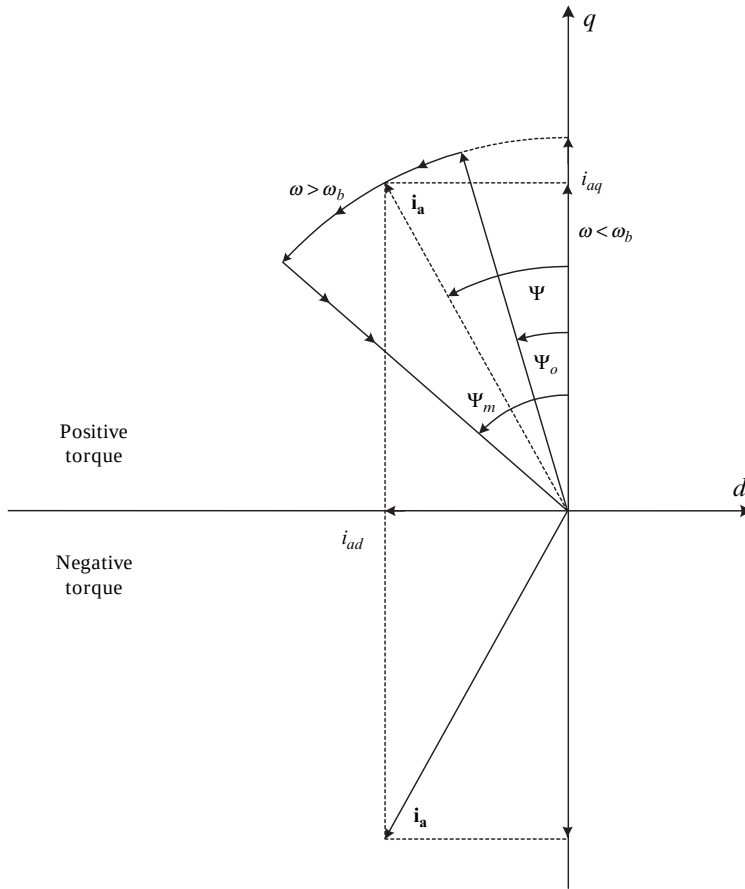


Fig. 7.15. Current phasor control of the AFPM drive in the low and high speed regions of the machine.

negative torque, i_T must be negative; Ψ is taken in the control as a positive quantity between 0 to 90° . In the case where $\Psi = 0$, then $k_{Trel} = 0$, $k_T = K_{Tf} = 3p\psi_f/2$ and $i_T = i_{aq}$.

With the current control strategy in place and with the torque as a function of current as given by eqn (7.26), a speed controller for the AFPM machine drive can be implemented around the current controller as shown in Fig. 7.16. The block diagram of Fig. 7.16 describes the basic principle of the current and speed control of the AFPM machine drive. The speed regulator acts on the error between the desired and measured speed of the system and controls the current i_T of eqn (7.26). The maximum current of the machine is limited by the limiter as shown in Fig. 7.16. The current angle Ψ is, as a positive

quantity, a function of the absolute value of the speed of the machine, $|\omega|$, as shown in Fig. 7.16. In the low speed region the angle Ψ is kept constant. In the speed region above the base speed ($\omega > \omega_b$) the angle Ψ is increased with speed. In the very high speed region the angle Ψ is kept constant again. With the current and current angle Ψ known, the desired dq currents, i_{ad}^* and i_{aq}^* , are determined by eqn (7.24). With the measured speed and measured dq currents as the other inputs, the current controller outputs the necessary dq voltages according to the block diagram of Fig. 7.13. These voltages are transformed to ABC voltages using the inverse transformation given by eqn (7.11), and are then put through the PWM generator to control the inverter and hence the terminal voltages of the machine.

To design the speed regulator of the system of Fig. 7.16 a few approximations can be made to simplify the model of the speed control drive. Firstly, the mechanical time constant of drive systems is in general much longer than the electrical time constant of the drive so that the latter can be ignored. Together with this, a good approximation is to assume that the current controllers are forcing the actual currents equal to the command (desired) currents, i.e. $i_{ad} = i_{ad}^*$, $i_{aq} = i_{aq}^*$, and $i_T = i_T^*$. Applying further eqn (7.26), the speed control system can be simplified to that shown in Fig. 7.17. From the simplified control system the approximate transfer function of the system can be determined and the speed regulator can be designed to obtain the required speed response.

It must be noted that the speed control method of Fig. 7.16 is only one way of controlling the speed of the PM brushless motor and is recognized as only a basic method; the method e.g. has the disadvantage that in the high speed regions under low-load conditions the drive is not necessarily operating at best (highest efficiency) operating conditions. Several other techniques and algorithms for wide-range speed control of PM motor drives have been proposed [24, 57, 137, 164, 179, 196, 280] and can be applied to the speed control of the sinusoidal AFPM brushless motor drive. The important aspect is the prevention of saturation of the dq current regulators in the high speed region of the motor.

7.2.4 Hardware of Sinusoidal AFPM Machine Drive

For the sinusoidal AFPM brushless motor the same solid-state converter as for the trapezoidal AFPM brushless motor can be used (Fig. 7.4). A more detailed diagram of the hardware of the sinusoidal AFPM brushless drive system is shown in Fig. 7.18. The fixed a.c. supply voltage is rectified by means of a diode rectifier or voltage-source controlled (active) PWM rectifier to obtain a fixed d.c. voltage. The d.c. voltage is then inverted by means of a voltage-source PWM inverter to a three-phase a.c. voltage of variable frequency and magnitude. The difference between the trapezoidal and sinusoidal drives lies in the PWM switching-control of the inverter bridge. In the case of the trapezoidal AFPM motor drive, quasi-square PWM voltages are applied, whereas

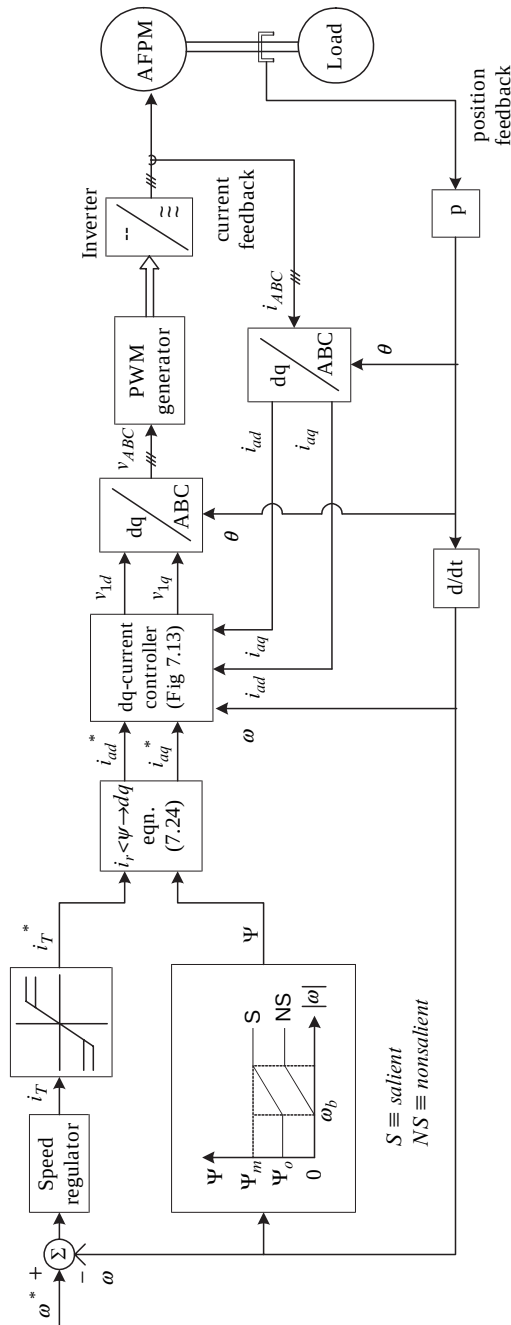


Fig. 7.16. Basic current and speed control system of the converter-fed sinusoidal AFPM drive.

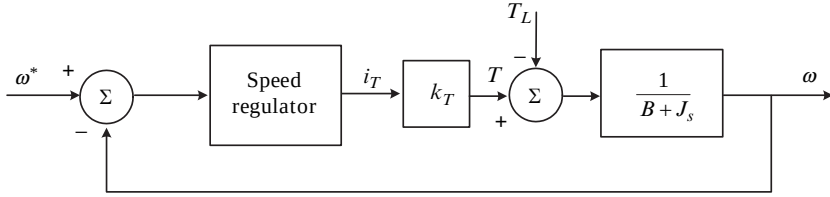


Fig. 7.17. Block diagram of simplified speed control system.

in the case of the sinusoidal AFPM motor drive, sinusoidal PWM voltages are applied and the phase currents of the stator are sinusoidally regulated as described in the previous sections.

The peak value of the fundamental frequency voltage component of the inverter-legs of Fig. 7.18 with respect to the neutral point N is clearly $\frac{1}{2}V_d$. As the fundamental frequency voltage components of the inverter-legs are 120° out of phase, the maximum *rms* line-to-line voltage of the inverter at the fundamental frequency with sinusoidal PWM is

$$V_{1L-Lmax} = \frac{\sqrt{3}}{\sqrt{2}} \frac{V_d}{2} = 0.612V_d \quad (7.27)$$

This result shows the limitation on the maximum *rms* output voltage of the inverter. This affects, amongst other things, the maximum speed at which rated torque can be delivered by the drive (see *Numerical example 7.2*).

The currents in the three phases of the AFPM machine drive are sinusoidally regulated. There is, however, a ripple in the phase currents due to the PWM switching. It can be shown that the peak-to-peak ripple in the phase current, Δi_{p-p} , is directly proportional to

$$\Delta i_{p-p} \propto \frac{V_d}{L_{ph}f_s} \quad (7.28)$$

where f_s is the switching frequency of the inverter and L_{ph} the total load inductance per phase of the inverter. The ripple current is thus inversely proportional to L_{ph} and f_s . Note that L_{ph} can be increased by the addition of an external inductor in series with the stator (armature) of the AFPM machine as shown in Fig. 7.18. Careful consideration should be given to the selection of L_{ph} and f_s . As the air gaps of AFPM machines, specifically those AFPM machines with coreless and slotless iron-cored stators, are relatively large, the internal phase inductances of these machines are relatively small. Thus to keep the ripple current in these machines within limits, the switching frequency f_s is required to be high. With power MOSFETs the switching frequency can be made high (say $f_s \leq 50$ kHz), but with IGBTs in the medium-to-large power range the switching frequency is very much limited to $f_s \leq 10$ kHz. In such a case the phase inductance must be increased. The high ripple current that

can occur in most of the AFPM machine drives is often overlooked. A high ripple current has serious consequences on the current rating of the inverter and the efficiency, stator winding temperature, torque quality and acoustic noise of the AFPM brushless motor.

For rotor position feedback of the sinusoidal AFPM drive a position sensor with high resolution is necessary. Resolvers are often used for this as they are robust and can be used in harsh environments. Integrated and analog circuitry are necessary to interface the resolver with the digital controller (see Fig. 7.18). Digital interface outputs of up to 16 bit resolution are possible. Normally 12 bit resolution gives more than enough accuracy for position feedback.

All the measured information of voltage, current and rotor position are fed to the digital controller as shown in Fig. 7.18. The controller outputs PWM signals via fibre optic links to the inverter bridge. Digital signal processors (DSPs) are nowadays frequently used as drive controllers [254]. Examples of dedicated fixed point DSPs from *Texas Instruments* include TMS320F240 (20 MHz), TMS320LF2407A (40 MHz), TMS320F2812 (150 MHz). These DSPs have 12 or 16 PWM channels, 16 A/D channels (10 or 12 bit resolution) and operate at frequencies of between 20 to 150 MHz. In another recent development on drive controllers *International Rectifier* introduced their motion control chipset IRACO210 [109]. This chipset uses a high speed configurable field programmable gate array (FPGA), tightly coupled to gate driver and current sense interface chips. In this way flexible drive control algorithms are implemented in FPGA hardware, rather than in software.

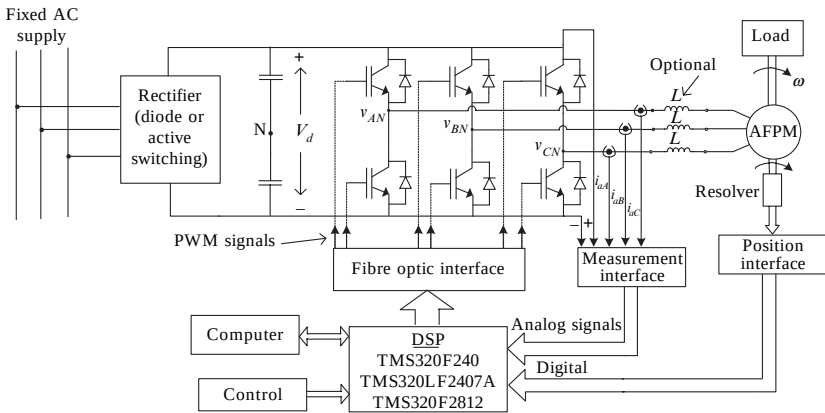


Fig. 7.18. Hardware of sinusoidal AFPM drive system.

7.3 Sensorless Position Control

It is obvious that information of the rotor position of the AFPM brushless motor is absolutely necessary for proper current and speed control. For trapezoidal AFPM motors a *low resolution position sensor* may be used, but for sinusoidal AFPM motors a *high resolution position sensor* is necessary. The use of a position sensor that is mechanically connected to the machine and electrically connected to the controller reduces the electromechanical drive system reliability and increases its cost. Moreover, in some applications insufficient space may make the use of a position sensor difficult. *Sensorless position control* techniques, in which an accurate information of the rotor position can be obtained without the need of a mechanical sensor, are superior.

Various methods and techniques have been proposed the last two decades for the sensorless position control of PM brushless motors. Most of these methods are based on the measurement of the voltage and/or current of the machine. Thus, instead of using a mechanical sensor, electrical sensors (not mounted in the machine but placed at the solid-state converter) are used to obtain the rotor position signal. In this way, the position of the rotor is determined indirectly, as measured electrical quantities together with the model of the machine are used. State observers, filters and other calculations are used to estimate the mechanical position of the rotor.

The methods of sensorless control can be classified into two types: (i) methods that are based on the back EMF estimation [23, 90, 271], and (ii) methods that make use of the magnetic saliency present in the machine [138, 167, 168, 234]. The EMF measurement is not suitable for low and zero speed operation of the motor as the induced voltage is proportional to the rotor speed. At stand still, the induced voltage is zero and there is no position-information anymore in the stator quantities. The second method, which makes use of the magnetic saliency, is more suitable for the detection of the rotor position at low speeds and standstill. With this method either test voltage pulses are applied to the machine or additional high-frequency signals superimposed on the fundamental voltages (or currents) are injected. The injection of the revolving high-frequency voltage vector is commonly used. All these methods can be used successfully for the position sensorless control of AFPM motor drives [205]. However, it is still questionable if the high-frequency injection sensorless control method can be applied to coreless type AFPM brushless motors as these machines have practically no magnetic saliency.

Fig. 7.19 explains the basic principle of the *high frequency voltage injection sensorless position control* technique. High frequency voltage signals are injected to the system by adding them to the dq voltage signals v_{1d} and v_{1q} of the drive. The current response of the drive system to these voltage signals is monitored by band pass filtering (BPF) the dq currents, so as to obtain the high frequency current components i_{adh} and i_{aqh} . These high frequency current components contain information about the rotor position as

they are affected by the magnetic saliency of the machine. These currents, together with the high frequency injected signals are used by an observer or position estimator to estimate the rotor position θ_m . A low pass filter (LPF) is furthermore used to extract from the dq currents the fundamental current components i_{ad} and i_{aq} , which are necessary for the fundamental current controller of the drive (Fig. 7.13). The frequency of the injected voltage signals must be much higher than the fundamental frequency, but also much lower than the switching frequency of the inverter, and is typically between 500 Hz and 2 kHz.

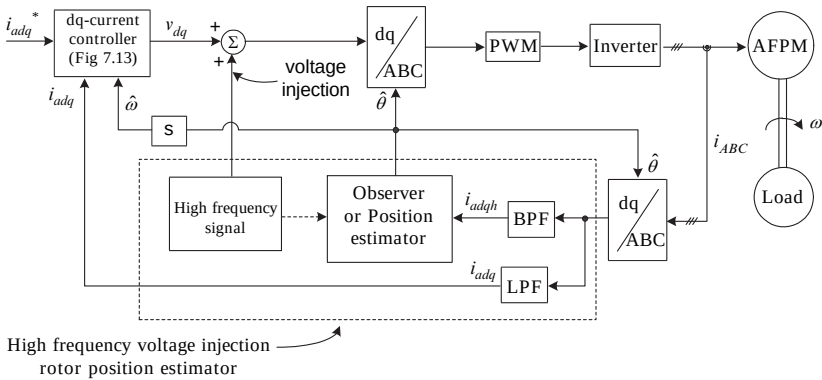


Fig. 7.19. High frequency voltage injection sensorless position control.

Numerical Example 7.1

A small, 35-W, 4300 r/min, 12 pole, Y-connected air-cored, trapezoidal AFPM brushless motor is used as part of a reaction wheel for a micro-satellite (see Fig. 7.20). Reaction or momentum wheels are used on satellites to keep the satellite steady and to orientate it. Orientation control is necessary as, e.g. the photovoltaic panels must be pointed at all times to the sun for maximum power generation. The motor has a stator phase inductance of $L_p = 3.2 \mu\text{H}$ and a phase resistance of $R_1 = 22 \text{ m}\Omega$. The motor is fed by a 3-phase MOSFET inverter with $V_d = 14 \text{ V}$, and is under square-wave current control as described in Section 7.1.3. Assuming the voltage drop across a MOSFET switches during switching-on as 1 V, calculate:

- the switching frequency f_s of the inverter;
- the external phase inductance required for a maximum peak-to-peak ripple current less than $\Delta i_{rmax} = 0.5 \text{ A}$ if $f_s = 40 \text{ kHz}$.



Fig. 7.20. Small coreless AFPM brushless motor. Photo courtesy of the *University of Stellenbosch*, South Africa.

Solution

First, an approximate equation for the maximum peak-to-peak phase current ripple of the motor must be derived. For the circuit shown in Fig. 7.5 during switching-on, i.e. with the transistors T_u and T_y switched on and assuming ideal transistor switches and drive in the steady-state, the voltage equation (7.1) can be written as

$$V_d = R_p I_a + R_p i_r + L_p \frac{di_r}{dt} + E_{fL-L} \quad (7.29)$$

where the phase current $i_a = I_a + i_r$, I_a is the steady-state average current and i_r is the ripple current. Putting eqn (7.3) and neglecting the voltage drop $R_p i_r$, eqn (7.29) can be brought to

$$V_d - DV_d \approx L_p \frac{di_r}{dt} \quad (7.30)$$

From eqns (7.4) and (7.30) the peak-to-peak ripple current Δi_r is

$$\Delta i_r \approx \frac{V_d(1-D)t_{on}}{L_p} = \frac{V_d(1-D)D}{L_p f_s} \quad (7.31)$$

The maximum peak-to-peak ripple current will occur at a duty cycle of $D = 0.5$, hence

$$\Delta i_{rmax} \approx \frac{V_d}{4L_p f_s} \quad (7.32)$$

- (a) Switching frequency. Using eqn (7.32) and taking into account the voltage drop across the transistor switches, the switching frequency for a maximum ripple current of 0.5 A is

$$f_s \approx \frac{V_d - 2}{4L_p \Delta i_{rmax}} = \frac{14 - 2}{4 \times (2 \times 3.2 \times 10^{-6}) \times 0.5} = 937.5 \text{ kHz}$$

- (b) External inductance. Eqn (7.32) gives the following additional phase inductance required to keep the ripple current less than 0.5 A at $f_s = 40$ kHz

$$L_{ph(add)} \approx \frac{V_d - 2}{4f_s \Delta i_{rmax}} - L_s = \frac{14 - 2}{4 \times 40\,000 \times 0.5} - 3.2 \times 10^{-6} = 71.8 \text{ } \mu\text{H} \quad (7.33)$$

These results show the low phase inductance problem of coreless AFPM brushless motors when fed from solid-state inverters. Either the switching frequency of the inverter must be high or relatively large external inductors must be added in series with the phase windings of the machine to keep the ripple current within limits. Usually, external inductances are used, as a high switching frequency causes the power losses in the inverter to be high.

Numerical Example 7.2

Consider a 75-kW, 460-V, Y-connected, 100-Hz, 1500 r/min, 8-pole AFPM motor with single PM rotor disc and double slotted stators connected in series. The equivalent circuit parameters and flux linkage of this machine are: $R_1 = 0.034 \text{ } \Omega$, $L_{sd} = L_{sq} = 2.13 \text{ mH}$, $\psi_f = 0.538 \text{ Wb}$. The machine is under dq current control by using amongst other things a three-phase IGBT voltage-source inverter with d.c.-link voltage $V_d = 755 \text{ V}$. Find:

- the maximum speed and the developed power of the drive with the machine at rated current and the machine controlled at maximum torque per ampere;
- the per unit high speed of the drive developing the same power as in (a) with the drive at rated current, rated voltage and using an advanced current angle.

Solution

- (a) Maximum speed and electromagnetic power

With a surface-mounted AFPM machine maximum torque per ampere is obtained with $i_{ad} = 0 \text{ A}$. Thus, at rated stator current $i_{aq} = \sqrt{2} \times 105 = 148.5 \text{ A}$. To obtain maximum speed the output voltage of the inverter must be at a maximum, i.e. from eqn (7.27) the *rms* phase voltage $V_1 = 0.612 \times 755/\sqrt{3} = 266.8 \text{ V}$. Thus, the peak phase voltage, $V_{1m} = \sqrt{2} \times 266.8 = 377.3 \text{ V}$ or

simply $1/2V_d$, and note that $V_{1m} = \sqrt{v_{1d}^2 + v_{1q}^2}$. By considering the drive in the steady state and ignoring the resistive voltage drop, from eqn (7.22), eqn (7.23) and Fig. 7.11 we have

$$v_{1d} = V_{1m} \sin \delta \approx -\omega L_{sq} i_{aq} \quad (7.34)$$

$$v_{1q} = V_{1m} \cos \delta \approx \omega \psi_f \quad (7.35)$$

or else

$$\tan \delta \approx \frac{-L_{sq} i_{aq}}{\Psi_f} = \frac{-2.13 \times 10^{-3} \times 148.5}{0.538} \quad \text{and} \quad \delta \approx -30.45^\circ$$

Using eqn (7.34) the angular frequency is

$$\omega \approx \frac{V_{1m} \sin \delta}{-L_{sq} i_{aq}} = \frac{377.2 \times \sin(-30.45)}{-2.13 \times 10^{-3} \times 148.5} = 604.5 \text{ rad/s}$$

and the mechanical speed $n = 30\omega/p\pi = 1443 \text{ rpm}$. The maximum speed with maximum torque per ampere control is thus just less than the rated speed of 1500 r/min of the machine. Using eqn (7.34), an approximate equation for the steady-state electromagnetic power (7.12) can be found as follows:

$$P_{elm} = \Omega T_d = \frac{3}{2} \psi_f \omega i_{aq} \approx \frac{3}{2} \left(\frac{\psi_f V_{1m}}{-L_{sq}} \right) \sin \delta \quad (7.36)$$

Thus,

$$= \frac{3}{2} \times \frac{0.538 \times 377.3}{-2.13 \times 10^{-3}} \times \sin(-30.45) \approx 72.45 \text{ kW}$$

which is also just less than the rated power of 75 kW of the machine.

(b) Per unit high speed.

The speed of the AFPM brushless motor can be increased beyond base speed by increasing the current angle Ψ , as shown in Fig. 7.11(b), keeping the peak value of the sinusoidal phase current $I_{am} = \text{constant}$. Note that $I_{am} = \sqrt{i_{ad}^2 + i_{aq}^2}$.

At rated conditions $I_{am} = 148.5 \text{ A}$ and $V_{1m} = 377.3 \text{ V}$. The same power as in (a), i.e. 72.45 kW must be obtained. Hence, it is clear from eqn (7.36) that δ will be the same as in (a), i.e. $\delta = -30.45^\circ$. With the machine considered as 100% efficient, the electromagnetic power is equal to the input power

$$P_{elm} \approx \frac{3}{2} V_{1m} I_{am} \cos \phi$$

where

$$\cos \phi \approx \frac{72\,450}{(3/2) \times 377.3 \times 148.5} = 0.862 \quad \text{and} \quad \phi \approx 30.45^\circ \text{ (leading)}$$

The q -axis current i_{aq} can now be calculated as

$$i_{aq} = I_{am} \cos(|\delta| + \phi) \approx 148.5 \cos(30.45 + 30.45) = 72.2 \text{ A}$$

$$\Psi = |\delta| + \phi = 30.45 + 30.45 = 60.9^\circ$$

Using eqn (7.34) that is still valid, the angular frequency

$$\omega \approx \frac{V_{1m} \sin \delta}{-L_{sq} i_{aq}} = \frac{377.3 \times \sin(-30.45)}{-2.13 \times 10^{-3} \times 72.2} = 1243 \text{ rad/s}$$

and the mechanical speed $n = 30\omega/p\pi = 2967 \text{ rpm}$. Taking the mechanical speed 1443 rpm in (a) as the base speed, the per unit speed is $2972/1443 = 2.0$ p.u. Hence, the same power can be developed by the drive at two times base speed by advancing the current angle Ψ from zero to almost 61° . It must be noted that this increase in speed is very much dependent on the q -axis synchronous inductance L_{sq} as can be seen clearly from all the equations used above. If L_{sq} becomes relatively small, the speed increase will be much less. The per unit phase reactance of the above iron-cored, slotted AFPM machine is $x_s = 0.53$ p.u. For slotless and coreless AFPM machines the per unit phase reactance is much less (typically 0.1 p.u.) and the speed increase will be little.

Numerical Example 7.3

Simulate the dq and ABC current response of the drive system of *Numerical example 7.2*. Assume that

- (a) the speed is constant at 1000 r/min;
- (b) the drive is operating at maximum torque per ampere;
- (c) the switching frequency of the inverter is 1.5 kHz;
- (d) the dq current regulators consist only of a proportional gain $K_c = 3 \text{ V/A}$.

Solution

Various software, such as PSpice and Simpler, can be used for simulating converter-fed electrical machine drives. Assuming the switching of the inverter as ideal, Matlab-Simulink has been used in this example to simulate the current response of the drive.

A block diagram of the simulation set-up is shown in Fig. 7.21. The simulation is done as “continuous”, i.e. without any sampling and so forth like in a digital control system. The current control is done in the dq reference frame

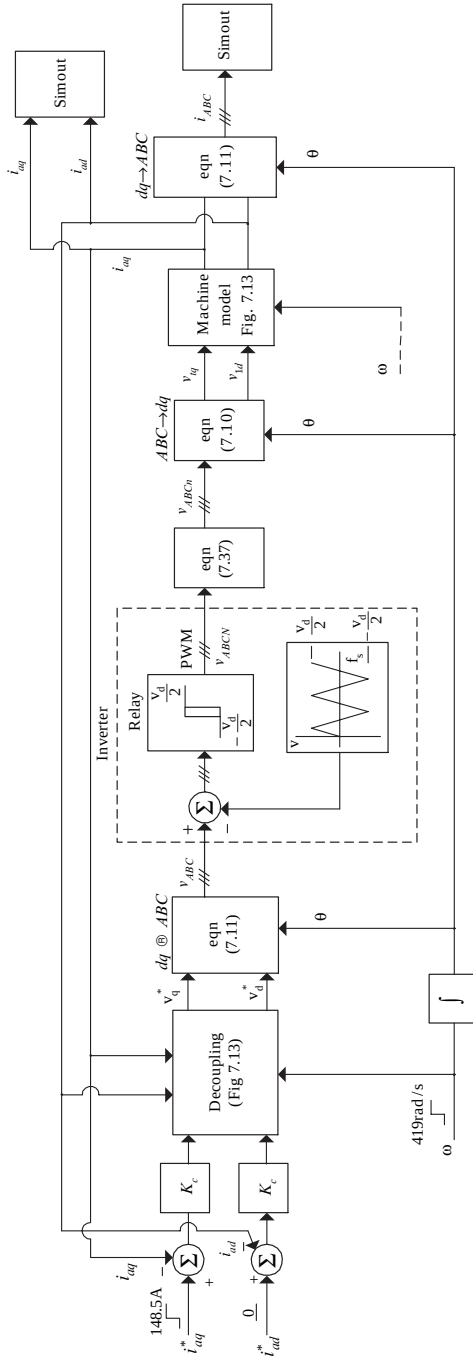


Fig. 7.21. Simulation block diagram of current controlled voltage source inverter drive system. Numerical example 7.3.

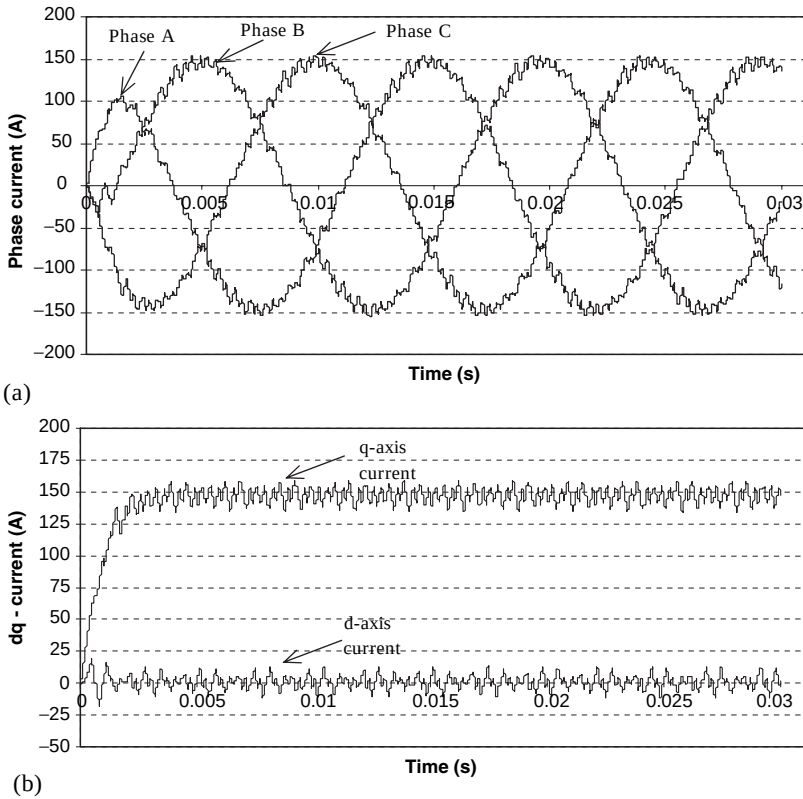


Fig. 7.22. Simulated current response for step input in the q -axis current i_{aq} : (a) phase currents i_{aA} , i_{aB} and i_{aC} ; (b) dq currents i_{ad} and i_{aq} . Numerical example 7.1.

using the current controller of Fig. 7.13. For both the d and q -axis regulators of Fig. 7.13 a gain K_c is used as shown in Fig. 7.21. The decoupling is done as shown in Fig. 7.13 with the actual dq currents and speed as inputs. For simulation of the inverter, the dq voltages are transformed to ABC voltages by using eqn (7.11). To generate the PWM phase voltages, the ABC voltages are compared with a 1.5 kHz triangle wave (amplitude of $V_d/2$), by using a subtractor (sum) and comparator (relay). The relay outputs $\pm V_d/2$. The generated PWM voltages, v_{ABCN} , are with respect to the d.c.-link midpoint N of the inverter (see Fig. 7.18), and have to be converted to true phase voltages, v_{ABCn} , with respect to the floating neutral point n of the machine as follows:

$$\begin{aligned} v_{An} &= \frac{1}{3}(2v_{AN} - v_{BN} - v_{CN}) \\ v_{Bn} &= \frac{1}{3}(2v_{BN} - v_{CN} - v_{AN}) \end{aligned} \quad (7.37)$$

$$v_{Cn} = \frac{1}{3}(2v_{CN} - v_{AN} - v_{BN})$$

The next step is to either solve the machine equations in the ABC or in the dq reference frame. Using the dq reference frame, the true PWM phase voltages must be converted back to dq voltages as shown in Fig. 7.21. Using the machine model of Fig. 7.13, the i_{ad} and i_{aq} currents can be determined and fed back to be compared with the input command currents as shown. To obtain information of the actual phase currents, the i_{ad} and i_{aq} currents are transformed to ABC currents.

The whole block diagram of Fig. 7.21 can be set up in Matlab-Simulink using the Simulink blocks. For the machine model, eqns (7.22) and (7.23) can directly be set-up in Simulink using integrators. To simulate the inverter switching, the relay block can be used. All the other source and mathematical function blocks are available in Simulink. The simulated results can be output in various ways, in this case to the Matlab workspace.

All the machine parameters are known from *Numerical example 7.2*. The mechanical speed is taken as constant at 419 rad/s (1000 rpm). For maximum torque per ampere the dq command currents $i_{ad}^* = 0$ and $i_{aq}^* = 148.5$ A are given as step inputs to the system (Fig. 7.21). The dq and ABC current responses of the drive for a step input in i_{aq}^* are shown in Fig. 7.22. The effect of the switching of the inverter on the current waveforms is clear. In this case, the q -axis current builds up to the rated current in less than 5 ms. This depends amongst others on the value of the proportional gain K_c . The effect of the decoupling on the current response can be self-studied by removing the decoupling signals in the simulation.

Cooling and Heat Transfer

8.1 Importance of Thermal Analysis

During the operation of an electrical machine, heat is generated due to power losses in electric and magnetic circuits and mechanical (rotational) losses. To ensure a long operational life for the machine, these losses must be removed as far as possible from the machine so that the temperature limitations established for the machine materials, such as insulating materials, lubricants and PMs are complied with. In addition to the consideration of the machine's operational life, a lower operating temperature reduces extra winding losses introduced by the temperature coefficient of the electric resistance — eqn (3.47).

Whereas extensive research has been devoted to the thermal studies of conventional electrical machines, AFPM machines have received very little attention [128, 245, 260]. Owing to the fact that AFPM machines possess a relatively large air gap volume and quite often have multi-gaps, the general perception is that AFPM machines have better ventilation capacity than their radial field counterparts [52, 106].

Since the external diameter increases rather slowly with the increase of output power, i.e. $D_{out} \propto \sqrt[3]{P_{out}}$ (Fig. 2.16), the existing heat dissipation capacity may be insufficient to cope with excessive heat at certain power ratings, so that more effective means of cooling have to be enforced. Thus, quantitative studies of the heat dissipation potential of AFPM machines with vastly different topologies is important.

8.2 Heat Transfer Modes

Heat transfer is a complex phenomenon presenting formidable analytical difficulties. Heat is removed from an electrical machine by a combination of *conduction*, *radiation* and *convection* processes to the ambient air and surroundings.

8.2.1 Conduction

When a temperature gradient exists in a solid body, such as in the copper, steel, PMs or the insulation of an electrical machine, heat is transferred from the high-temperature region of temperature ϑ_{hot} to the low-temperature region of temperature ϑ_{cold} according to *Fourier's law*, which is given as:

$$\Delta P_c = -kA \frac{\partial \vartheta}{\partial x} = \frac{kA}{l} (\vartheta_{hot} - \vartheta_{cold}) \tag{8.1}$$

where ΔP_c is the rate of heat conduction, A is the area of the flow path, l is the length of the flow path and k is the thermal conductivity of the material. The latter is experimentally determined and is relatively insensitive to temperature changes. Thermal properties of typical materials used for AFPM machines are given in Table 8.1, where c_p is the specific heat of material at constant pressure.

Table 8.1. Selected thermal properties of materials

Material (20°C)	Grade	ρ kg/m ³	c_p J/(kg °C)	k W/(m °C)
Air	-	1.177	1005	0.0267
Water	-	1000	4184	0.63
Mica	-	3000	813	0.33
Epoxy resin	-	1400	1700	0.5
Copper	-	8950	380	360
Aluminum	Pure	2700	903	237
	Alloy (cast)	2790	883	168
Steel	1% Carbon	7850	450	52
	Silicon	7700	490	20-30
Permanent magnet	Sintered NdFeB	7600 to 7700	420	9

8.2.2 Radiation

The net *radiant energy* interchange between two surfaces with a temperature difference is a function of the absolute temperature, the emissivity and the geometry of each surface. If heat is transferred by radiation between two gray surfaces of finite size, A_1 and A_2 , and temperature, ϑ_1 and ϑ_2 (in Celsius degree), the rate of heat transfer, ΔP_r , may be written as

$$\Delta P_r = \sigma \frac{(\vartheta_1 + 273)^4 - (\vartheta_2 + 273)^4}{\frac{1-\varepsilon_1}{\varepsilon_1 A_1} + \frac{1}{A_1 F_{12}} + \frac{1-\varepsilon_2}{\varepsilon_2 A_2}} \tag{8.2}$$

where σ is the *Stefan-Boltzmann* constant, F_{12} is the shape factor which takes into account the relative orientation of the two surfaces and ε_1 and ε_2 are their respective emissivities which depend on the surfaces and their treatment. An ideal surface or body that absorbs and emits energy at maximum rate is called *black surface* or *body*. Real surfaces or bodies are normally approximated as gray surfaces or bodies where $0 < \varepsilon < 1$. Some selected emissivities related to AFPM machines are given in Table 8.2.

Table 8.2. Selected emissivities relevant to AFPM machines

Material	Surface condition	Emissivity, ε
Copper	Polished	0.025
Epoxy	Black	0.87
	White	0.85
Mild steel	-	0.2-0.3
Cast iron	Oxidized	0.57
Stainless steel	-	0.2-0.7
Permanent magnet, NdFeB	Uncoated	0.9

8.2.3 Convection

Convection is the term describing heat transfer from a surface to a moving fluid. The rate of convective heat transfer, ΔP_v , is given according to *Newton's law of cooling* as:

$$\Delta P_v = hA(\vartheta_{hot} - \vartheta_{cold}) \quad (8.3)$$

where h is the convection heat transfer coefficient, which is a rather complex function of the surface finish and orientation, fluid properties, velocity and temperature, and is usually experimentally determined. The coefficient h increases with the velocity of the cooling medium relative to the cooled surface. For a surface with forced ventilation, the following empirical relation may be used [159], i.e.

$$h_f = h_n(1 + c_h\sqrt{v}) \quad (8.4)$$

where h_f and h_n are the coefficients of heat transfer for the forced and natural convection respectively, v is the linear velocity of cooling medium and $c_h \approx 0.5$ to 1.3 is an empirical coefficient.

Some important formulae for evaluating convective heat transfer coefficients of AFPM machines are discussed in the following sections.

Convection Heat Transfer in Disc Systems

The rotating disc system plays a major role in the cooling and ventilation of the AFPM machine. Accurately determining the convection heat transfer coefficients needs thorough theoretical and experimental investigation because of the complexity of flow regimes.

In this section the convection heat transfer coefficients in different parts of the AFPM machine are evaluated, exploiting a number of existing models.

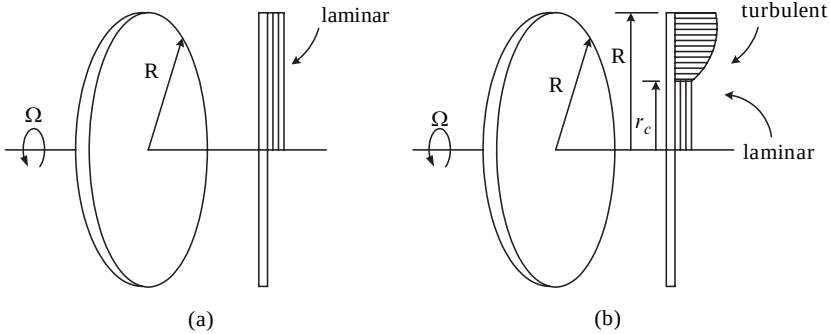


Fig. 8.1. Free rotating disc: (a) in laminar flow, (b) transition from laminar to turbulent flow.

Free rotating disc

The average heat transfer coefficient at the outside surface of a rotating disc may be evaluated using the formula developed for a free rotating disc [270], i.e.

$$\bar{h} = \frac{k}{R} \overline{Nu} \quad (8.5)$$

where R is the radius of the disc and the average Nusselt number $\overline{Nu}$ is given according to the different flow conditions as follows:

(i) For combined effects of free convection and rotation in laminar flow (Fig. 8.1a) [270]

$$\overline{Nu} = \frac{2}{5} (Re^2 + Gr)^{\frac{1}{4}} \quad (8.6)$$

$$Gr = \frac{\beta g R^3 \pi^{3/2} \Delta\vartheta}{\nu^2} \quad (8.7)$$

where Re is the Reynolds number according to eqn (2.72), β is the coefficient of thermal expansion, ν is the kinematic viscosity of the fluid (m^2/s) and $\Delta\vartheta$

is the temperature difference between the disc surface and surrounding air.

(ii) For a combination of laminar and turbulent flow with the transition at a radius r_c (Fig. 8.1b) [270]

$$\overline{Nu} = 0.015 Re^{\frac{4}{5}} - 100 \left(\frac{r_c}{R} \right)^2 \quad (8.8)$$

where

$$r_c = (2.5 \times 10^5 \nu / \Omega)^{1/2} \quad (8.9)$$

The angular speed $\Omega = 2\pi n$ when n is the rotational speed in rev/s. It is instructive to compare the heat transfer capabilities between a rotation disc and a stationary disc. If we consider a steel disc, which has a diameter of 0.4 m and rotates at 1260 rpm., the convection heat transfer coefficient may be calculated as 41 W/(m² °C), which is about ten times that of the same disc at standstill. Alternatively, one can say that the effective heat dissipation area of the same disc can be increased by a factor of 10 when the disc rotates at the specific speed.

Rotor radial peripheral edge

The heat transfer correlations for the radial periphery of the rotor disc are similar to those of a rotating cylinder in air. In this case the average heat transfer coefficient is given as

$$\bar{h}_p = (k/D_{out}) \overline{Nu} \quad (8.10)$$

where D_{out} is the outer diameter of the rotor disc, the average Nusselt number is given by

$$\overline{Nu} = 0.133 Re_D^{2/3} Pr^{1/3} \quad (8.11)$$

and Reynolds number at the disc periphery is

$$Re_D = \Omega D_{out}^2 / \nu \quad (8.12)$$

Note that a uniform temperature distribution in the cylinder is normally assumed when eqn (8.10) is used. Since $\bar{h}_p$ is proportional to the angular speed Ω , it may be concluded that the rotor periphery plays an increasingly important role in the heat dissipation as Ω increases.

Rotor-stator system

As seen in Fig. 8.2, an AFPM machine consists of a number of rotating and stationary discs. The heat transfer relations between a rotating and a stationary disc are of paramount importance in the thermal calculations. Due

to centrifugal effects, there is a forced flow between the two discs, which increases the local heat transfer rate compared with that of a free disc. The relative increase will depend on the gap ratio, $G = g/R$, where g is the clearance between the rotor and the stator and R is the radius of the disc, the mass flow rate and the rotational speed of the system [209].

Having radial channels and thick impellers, an air-cooled AFPM machine may be regarded as a poorly designed fan from a fluid flow perspective. Its tangential velocity component is much larger than the radial component. Thus, the heat transfer rate near the rotating disc shows more dependence on the rotational Reynolds number, Re_r , given by eqn (2.72).

Owen [207] provided an approximate solution for the flow between a rotating and a stationary disc, which relates the average Nusselt number to the moment coefficient of the stator-side rotor face, C_{mo} , by the following equation:

$$\begin{cases} \overline{Nu} = Re_r C_{mo} / \pi \\ C_{mo} Re_r^{1/5} = 0.333 \lambda_T \end{cases} \quad (8.13)$$

where λ_T is a *turbulence parameter* given as a function of volumetric flow rate, Q , as follows

$$\lambda_T = \frac{Q}{\nu R} Re_r^{-\frac{4}{5}} \quad (8.14)$$

By replacing λ_T in eqn (8.13) with eqn (8.14), the average Nusselt number becomes

$$\overline{Nu} = 0.333 \frac{Q}{\pi \nu R} \quad (8.15)$$

As discussed in [208], it has been shown that for a small gap ratio ($G < 0.1$) the flow in the air-gap space between the rotor and stator can be treated as a boundary layer. Whilst it is not absolutely true that the convective heat transfer coefficient from the stator to the air flow is close to that of the air flow to a rotating disc, the same heat transfer coefficient may be assumed in the thermal circuit simulation.

8.3 Cooling of AFPM Machines

Depending on the size of the machine and the type of enclosures, different arrangements for cooling may be used. From a cooling perspective, AFPM machines may be classified into two categories as follows:

- *machines with self-ventilation*, in which cooling air is generated by a rotating disc, PM channels or other fan-alike devices incorporated with the rotating part of the machine, and

- *machines with external ventilation*, in which the cooling medium is circulated with the aid of external devices, for an example, a fan or a pump.

8.3.1 AFPM Machines With Self-Ventilation

The majority of AFPM machines are air-cooled. Compared with conventional electrical machines, a particularly advantageous feature of disc-type AFPM machines from a cooling perspective is that they possess inherent self-ventilation capability. Fig. 8.2 shows the layout and active components of a typical AFPM machine. A close examination of the machine structure reveals that an air stream will be drawn through the air inlet holes into the machine and then forced outwards into the radial channel as the rotor discs rotate. The PMs in fact act as impeller blades. The fluid behaviour of the AFPM machine is much like that of a centrifugal fan or compressor.

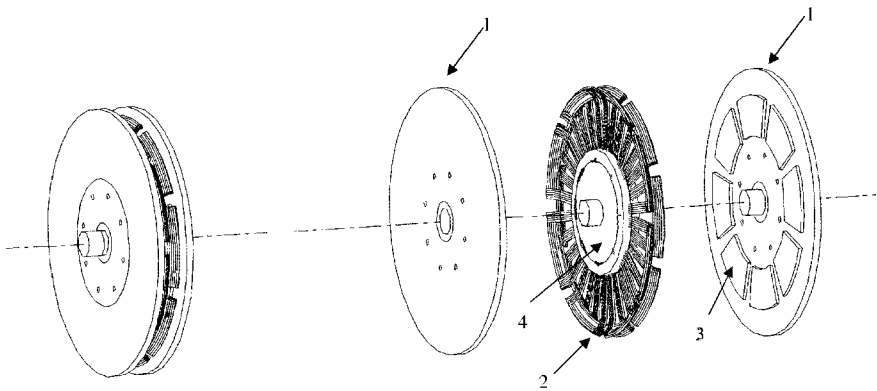


Fig. 8.2. Exploded view of an AFPM machine: 1 — rotor disc, 2 — stator winding, 3 — PM, and 4 — epoxy core.

The Ideal Radial Channel

According to the theory of an ideal impeller, a number of assumptions have to be made to establish the one-dimensional model of the ideal radial channel [78, 232]:

- there are no tangential components in the flow through the channel;
- the velocity variation across the width or depth of the channel is zero;
- the inlet flow is radial, which means that air enters the impeller without pre-whirl;

- (d) the pressure across the blades can be replaced by tangential forces acting on the fluid;
 (e) the flow is treated as incompressible and frictionless.

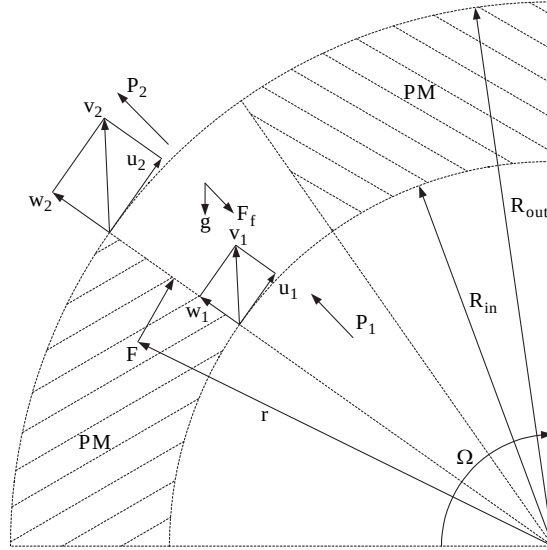


Fig. 8.3. Velocity triangles for a PM channel.

Figure 8.3 shows a radial channel with the velocity triangles drawn at the inlet and the outlet. It can be observed that the pressures at the inlet p_1 and the outlet p_2 , and friction F_{fr} make no contribution to the sum of the momentum, $\sum M_0$. If gravity is ignored, the general representation of conservation of momentum takes the following form [266]:

$$\sum M_0 = \frac{\partial}{\partial t} \left[\int_{cv} (\mathbf{r} \times \mathbf{v}) \rho dV \right] + \int_{cs} (\mathbf{r} \times \mathbf{v}) \rho (\mathbf{v} \cdot \mathbf{n}) dA \quad (8.16)$$

where $\mathbf{r}$ is the position vector from 0 to the elemental control volume dV and $\mathbf{v}$ is the velocity of the element.

For steady-state, one-dimensional air flowing between the entrance and exit of the channel, eqn (8.16) may be simplified as:

$$\sum M_0 = T_0 = (\mathbf{R}_{out} \times \mathbf{u}_2) \dot{m}_2 - (\mathbf{R}_{in} \times \mathbf{u}_1) \dot{m}_1 \quad (8.17)$$

where $\dot{m}_2 = \dot{m}_1 = \rho Q$, $u_1 = \Omega R_{in}$ and $u_2 = \Omega R_{out}$. The input shaft power P_{in} is then given by:

$$P_{in} = T_0 \Omega = \rho Q \Omega^2 (R_{out}^2 - R_{in}^2) \quad (8.18)$$

Re-arranging the above equation gives:

$$\frac{P_{in}}{Q} = \rho \Omega^2 (R_{out}^2 - R_{in}^2) \quad (8.19)$$

Based on the principle of conservation of energy, the input shaft power may be given as:

$$P_{in} = \dot{m} \left(\frac{p_2 - p_1}{\rho} + \frac{w_2^2 - w_1^2}{2} + z_2 - z_1 + U_2 - U_1 \right) \quad (8.20)$$

If the potential $(z_2 - z_1)$ and internal energy $(U_2 - U_1)$ (friction) are ignored, eqn (8.20) may be written in the same units as eqn (8.19) as:

$$\frac{P_{in}}{Q} = (p_2 - p_1) + \rho \frac{w_2^2 - w_1^2}{2} \quad (8.21)$$

If equations (8.19) and (8.21) are equated and noting that $w_1 = Q/A_1$ and $w_2 = Q/A_2$, where A_1 and A_2 are the cross-section areas of the inlet and outlet of the channel respectively, the pressure difference Δp between the entrance and exit of the radial channel (shown in Fig. 8.3) may be expressed as:

$$\Delta p = p_2 - p_1 = \rho \Omega^2 (R_{out}^2 - R_{in}^2) - \frac{\rho}{2} \left(\frac{1}{A_2^2} - \frac{1}{A_1^2} \right) Q^2 \quad (8.22)$$

Eqn (8.22) may be termed the *ideal equation* describing the air flow through the radial channel.

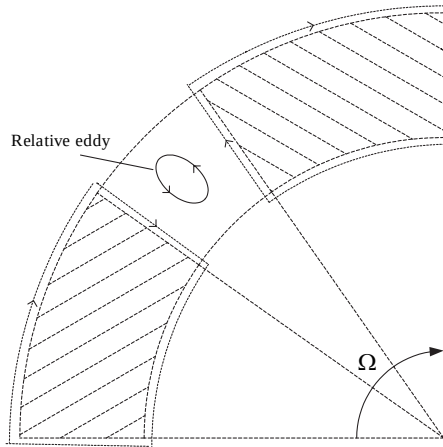


Fig. 8.4. The relative eddy in the PM channel.

The Actual Radial Channel

The actual characteristics of a hydraulic machine differ from the ideal case owing to two reasons: (i) the uneven spatial distribution of velocities in the blade passages, and (ii) the leakage and recirculation of flow and hydraulic losses such as friction and shock losses. These are completely different issues [78] and shall be dealt with separately.

Slip factor

As a result of the unbalanced velocity distribution of the leading and trailing edges of a PM channel and the rotation effects [232], there exists, according to Stodola [198, 232], a *relative eddy* within the blade passage shown in Fig. 8.4. This results in the reduction of the tangential velocity components and is called *slip*, which is usually accounted for using a *slip factor*. For approximately radial blades, the *Stanitz slip factor* k_s ($80^\circ < \beta_2 < 90^\circ$) is

$$k_s = 1 - 0.63\pi/n_b \quad (8.23)$$

where β_2 is the blade angle at exit and n_b is the number of the blades. When applying a slip factor, the pressure relation (8.22) becomes

$$\Delta p = \rho\Omega^2(k_s R_{out}^2 - R_{in}^2) + \frac{\rho}{2}\left(\frac{1}{A_1^2} - \frac{1}{A_2^2}\right)Q^2 \quad (8.24)$$

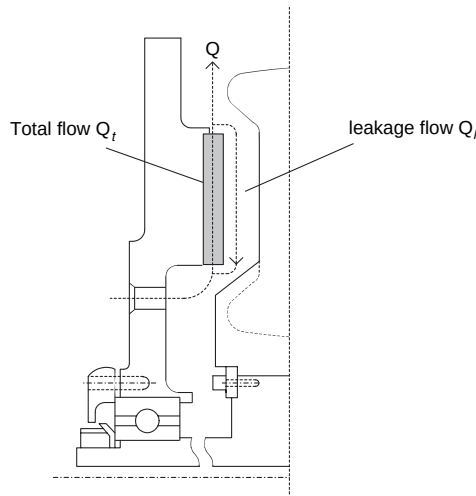


Fig. 8.5. Leakage flow in an AFPM machine (not to scale).

Shock, leakage and friction

Energy losses due to friction, separation of the boundary layer (shock loss) and leakage should also be considered in the flow analysis. As illustrated in Fig. 8.5, if the total volumetric flow rate through the PM channel is Q_t , the pressure difference between the PM exit and the entrance will cause a *leakage* or recirculation of a volume of fluid Q_l , thus reducing the flow rate at outlet to $Q = Q_t - Q_l$. The Q_l is a function of mass flow rate and discharge and leakage path resistances. The leakage flow reaches its maximum when the main outlet flow is shut.

These losses can be accounted for by introducing a pressure loss term Δp_l in eqn (8.24) as follows [232]

$$\Delta p = \rho \Omega^2 (k_s R_{out}^2 - R_{in}^2) + \frac{\rho}{2} \left(\frac{1}{A_1^2} - \frac{1}{A_2^2} \right) Q^2 - \Delta p_l \quad (8.25)$$

System losses

As the air passes through the AFPM machine, the system pressure loss due to friction must be taken into account. The sum of these losses is given by:

$$\Delta p_{fr} = \frac{\rho Q^2}{2} \sum_{i=1}^n \frac{k_i}{A_i^2} \quad (8.26)$$

where k_i and A_i are the loss coefficient and the cross section area of the flow path i respectively.

There are a number of sections through which the air flows in the AFPM machine (see Fig. 8.6). They are:

- (1) entry into the rotor air inlet holes;
- (2) passage through rotation short tube;
- (3) bending of air from pipe to annulus (90°);
- (4) round bend air passage (90°);
- (5) contraction of air in the tapering passage;
- (6) bending of air through the 90° elbow;
- (7) entry to the permanent magnet channels;
- (8) expansion of air through the tapering enlargement;
- (9) abrupt expansion of the air on the exit of the channel;
- (10) expansion as the air leaves the opening of the parallel rotor discs.

The loss coefficients associated with each section in eqn (8.26) are given in [61, 188, 198]. When the section is not circular, use is made of the hydraulic diameter to characterise the cross section. The hydraulic diameter is defined as $D_h = 4A/\wp$ where A is the cross-sectional area of the flow path and $\wp$ is the wetted perimeter.

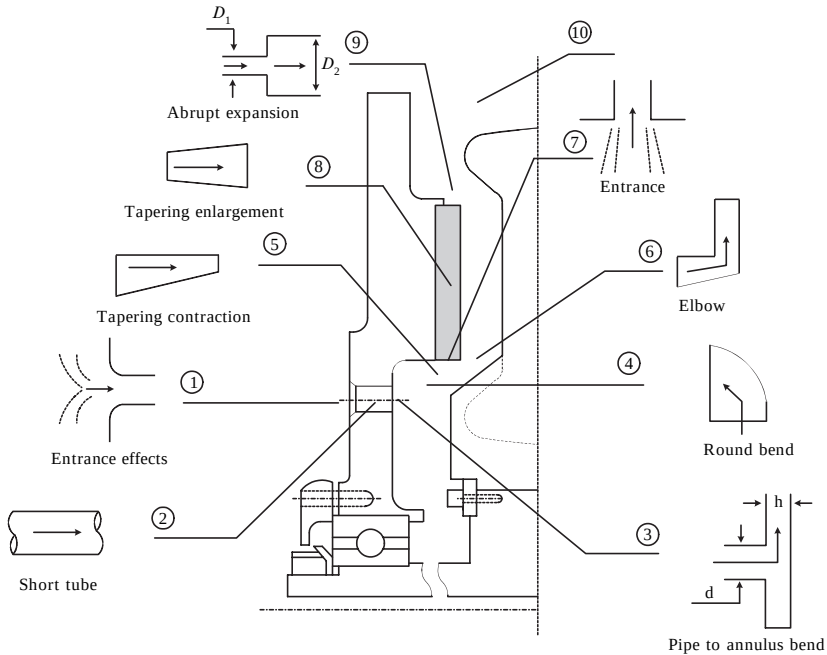


Fig. 8.6. System losses of an AFPM machine.

The loss coefficient for a pipe is given by $\lambda L/d$ where λ is a friction factor obtained as a function of Reynolds number Re and surface roughness from a *Moody diagram* [190]. To facilitate numeric calculations, the *Moody diagram* may be represented by [61]:

$$\begin{cases} \lambda = 8\{(8/Re)^{12} + (X + Y)^{-\frac{3}{2}}\}^{\frac{1}{12}} \\ X = \{2.457 \ln\{(7/Re)^{0.9} + 0.27\gamma/D\}^{-1}\}^{16} \\ Y = \{37530/Re\}^{16} \end{cases} \quad (8.27)$$

where $Re = \frac{\rho D_h Q}{\mu A}$ and where γ is the equivalent sand grain roughness [61].

Characteristics

It is now possible to relate the theoretical prediction obtained from the ideal flow model to the actual characteristic by accounting for the various losses discussed above.

Assuming that the AFPM machine (shown in Fig. 8.1) operates at a constant speed of 1200 rpm, the ideal developed pressure characteristic for a radial channel is a function described by eqn (8.22) as shown in Fig. 8.6. After introducing the slip factor, the resultant curve is shown as a dotted line as

eqn (8.24). It was not possible to obtain a suitable correlation in the literature [248] for the pressure loss due to shock and leakage as was the case for the slip.

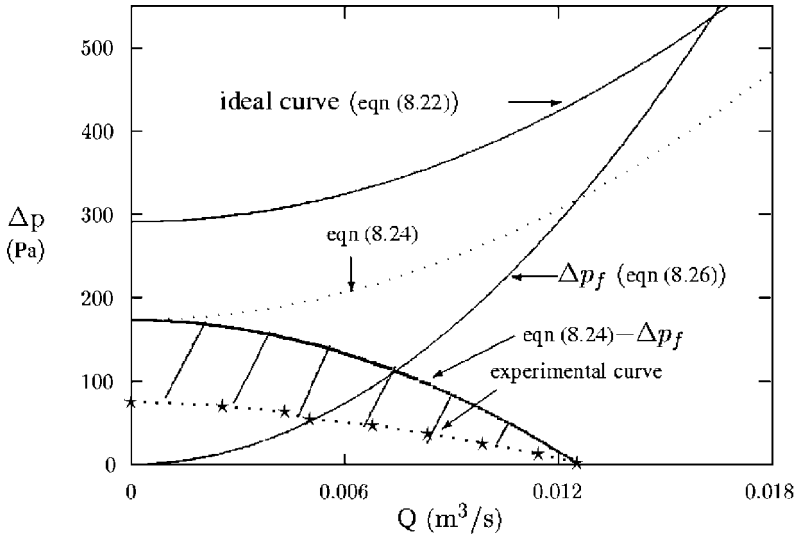


Fig. 8.7. Losses and characteristic curves at 1200 rpm.

The calculated characteristic curve without considering shock and leakage losses, i.e. eqn (8.24)– Δp_{fr} shown in Fig. 8.7, is significantly higher than the experimental one. The shaded area in Fig. 8.7 represents the shock and leakage losses. It can be seen that at low flow rates the shock and leakage losses are greater but tend to zero at the maximum flow rate. This has been discussed and experimentally validated in [260].

The derived characteristics describes the pressure relations of a single rotating PM disc facing a stator. For a double-sided AFPM machine with two identical coaxial rotating discs (Fig. 8.2) operating on the same stator, the characteristic curve presented in Fig. 8.7 represents only half of the AFPM machine. The characteristic curve of the whole machine may be obtained by adding flow rate at the same pressure, which is similar to two identical fans in parallel.

Flow and Pressure Measurements

Due to the nature and complexity of thermofluid analysis, the form of the system characteristics curve can at best be established by test. Depending on the machine topologies and size, the measurements may be taken either

at the air inlet or outlet. The AFPM machine under test is normally driven by another motor. Fig. 8.8 shows the experimental arrangements of the flow measurements at the machine outlet, in which a discharge duct is set up to provide good conditions for observing the flow. Along one side of the duct, several tapping points were made for measuring the static pressure with a pressure gauge (manometer). Near the outlet of the duct, provision was made for measuring velocity using a hot-wire anemometer probe.

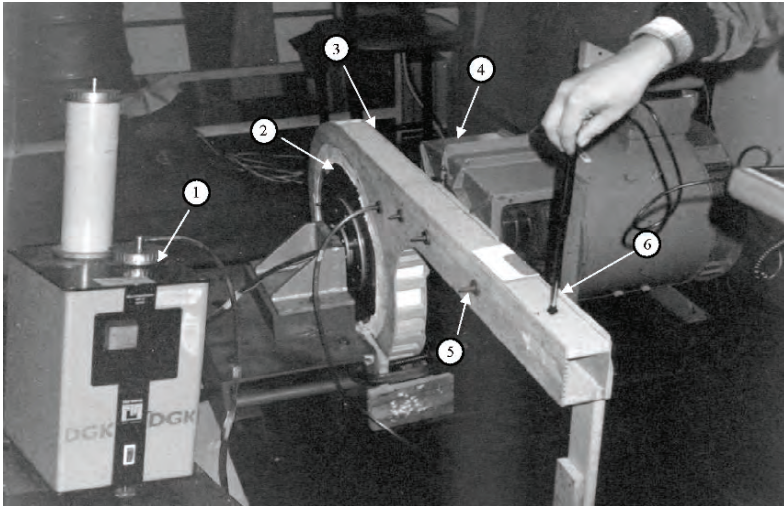


Fig. 8.8. The experimental set up. 1 — manometer, 2 — AFPM machine, 3 — discharge duct, 4 — prime mover (drive machine), 5 — pressure tapping point, 6 — wind speed probe. Photo courtesy of the *University of Stellenbosch*, South Africa.

To vary the flow rate, the test duct was fitted at its outer end with an obstruction. The test was started with no obstruction at the end of the discharge duct. The only resistance was then the duct friction, which was small and could be readily computed out of results. As the end of the duct was obstructed progressively, the flow was reduced and the static pressure increased to a maximum at zero volumetric flow rate. The static pressure difference Δp is measured as a function of volumetric flow rate $Q = A \times v$ for different motor speeds, where v is the linear speed.

The air flow-rate measurement can also be carried out by measuring inlet air pressure difference Δp , which is then used for calculating mass flow-rate $\dot{m}$ according to the following equation:

$$\dot{m} = \sqrt{2\rho\Delta p} A_d \quad (8.28)$$

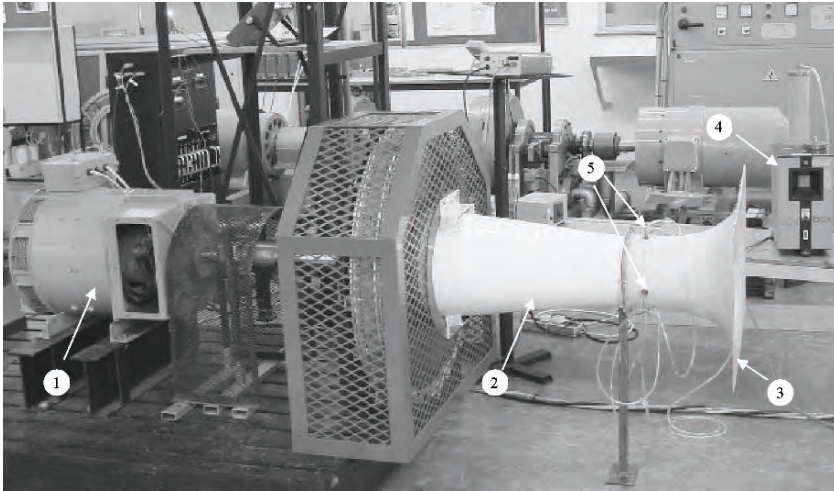


Fig. 8.9. Flow measurement at the air intake of AFPM machine. 1 — Prime mover (drive machine), 2 — inlet duct, 3 — bell mouth, 4 — manometer, and 5 — pressure tapping point. Photo courtesy of the *University of Stellenbosch*, South Africa.

where A_d is the cross section area of the inlet duct. Fig. 8.9 shows the setup of the flow measurement. A specially designed inlet duct with a bell mouth was mounted to the inlet (hub) of the AFPM machine. A few tapping points were made on the inlet duct for pressure measurements. The pressure drop through the bell mouth and inlet duct may often be assumed negligible.

8.3.2 AFPM Machines With External Ventilation

For medium to large power AFPM machines, the loss per unit heat dissipation area increases almost linearly with the power ratings. Thus the forced-cooling with the aid of external devices may be necessary. Some common techniques are described as follows.

External Fans

Large AFPM machines may require a substantial amount of air flow per unit time in order to bring out the heat generated in the stator windings. Depending on the operation conditions obtained on site, either an air-blast or a suction fan may be used as shown in Fig. 8.10. In both cases, intake and/or discharge ducts are needed to direct and condition the air flow. Since the inlet air temperature, for a given volumetric flow rate, has a significant effect on the machine temperature, this cooling arrangement can also help prevent recirculation of hot air should the machine operate in a confined space (e.g. small machine room). For high speed AFPM machines, a shaft-integral fan may be

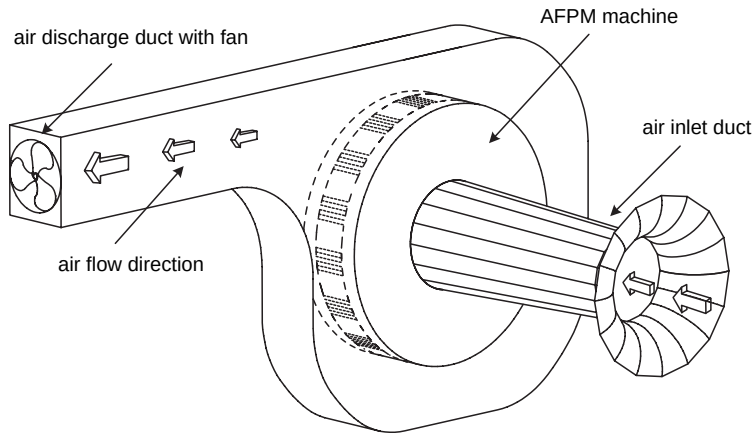


Fig. 8.10. AFPM machine with external air cooling.

a good option. Fig. 8.11 shows the assembly of a large power AFPM machine developed in the Department of Electrical and Electronic Engineering at the University of Stellenbosch, South Africa, in which the rotor hub part serves as both cooling fan and supporting structure for the rotor discs. It can be seen that the “blades” of the hub are not curved as the machine may operate in both directions of rotation.

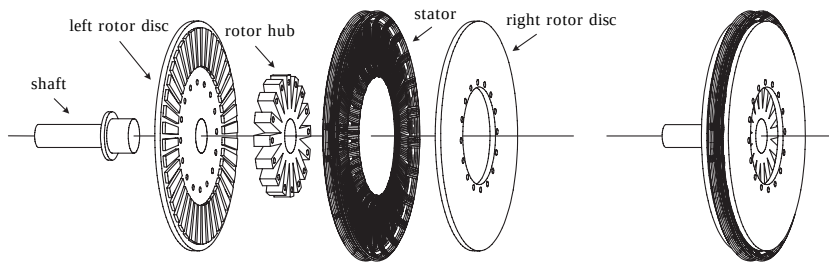


Fig. 8.11. Configuration of the AFPM machine with shaft-integral fan.

Heat Pipes

The concept of a passive two-phase heat transfer device capable of transferring large amount of heat with a minimal temperature drop was introduced by R.S.

Gaugler of the General Motors Corporation in 1942. This device received little attention until 1964, when Grover and his colleagues at Los Alamos National Laboratory, NM, U.S.A., published the results of an independent investigation and first used the term *heat pipe*. Since then, heat pipes have been employed in many applications ranging from temperature control of the permafrost layer under the Alaska pipeline to the thermal control of optical surfaces in spacecraft.

A typical *heat pipe* consists of a sealed container with wicking material. The container is evacuated and filled with just enough liquid to fully saturate the wick. A heat pipe has three different regions, namely (i) *an evaporator* or heat addition region of the container, (ii) *a condenser* or heat rejection region, and (iii) *an adiabatic* or isothermal region. If the evaporator region is exposed to a high temperature, heat is added and the working fluid in the wicking structure is heated until it evaporates. The high temperature and the corresponding high pressure in this region cause the vapour to flow to the cooler condenser region, where the vapour condenses, dissipating its latent heat of vaporisation. The capillary forces existing in the wicking structure then pump the liquid back to the evaporator. The wick structure thus ensures that the heat pipe can transfer heat whether the heat source is below the cooled end or above the cooled end.

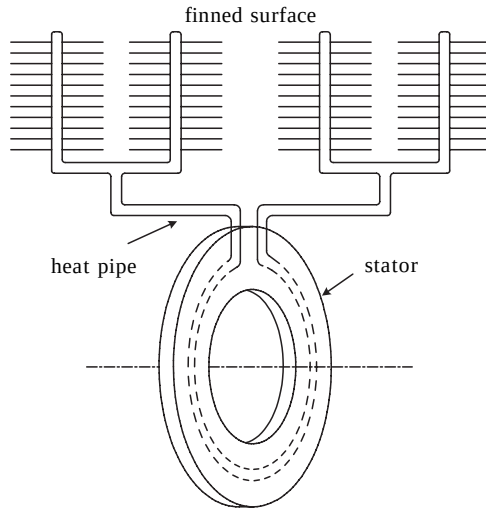


Fig. 8.12. AFPM machine cooled by heat pipes.

A heat pipe presents an alternative means for removing the heat from the AFPM machine. The heat pipe in an AFPM machine may be configured as shown in Fig. 8.12. Heat is transferred into the atmosphere through the finned

surface. The finned surface is cooled by air moving over the fins. The heat loss removed by the heat pipe, ΔP_{hp} , is given by [235]:

$$\Delta P_{hp} = \frac{\vartheta_{hot} - \vartheta_{cold}}{\frac{1}{h_{hot}A_{hot}} + \frac{1}{h_{cold}A_{cold}} + \frac{1}{\eta_{fin}h_{fin}A_{fin}}} \quad (8.29)$$

where ϑ_{hot} is the average temperature of the elements that surround the heat pipe in the stator, ϑ_{cold} is the average temperature of the air cooling the finned surface, h_{hot} is the convective heat transfer coefficient on the inside wall of the heat pipe in the stator, A_{hot} is the exposed area of the heat pipe in the stator, h_{cold} is the convection heat transfer coefficient on the inside wall of the heat pipe in the finned area, A_{cold} is the exposed area of the heat pipe at the finned surface, η_{fin} is the efficiency of the finned surface, h_{fin} is the convection heat transfer coefficient on the surface of the fins and A_{fin} is the total exposed area of the finned surface.

Direct Water Cooling

Depending on the conditions at the site of operation, it is often necessary to use forced water circulation to cool the stator windings directly, especially for large power AFPM machines. An external water pump is required for forcing water circulation. A longitudinal section of a water cooled double-disc motor has been shown in Fig. 2.7. For AFPM machines with an internal iron core stator, an ideal location of placing cooling channel is around the outer periphery of the stator disc, where the effective heat transfer area is the largest. For coreless winding AFPM machines, the winding coils may have a rhomboidal shape so that the space between the two active sides of each coil may be utilised for placing a cooling water duct [45].

The heat removed through cooling pipes can be calculated by using eqn (8.29). However, the heat transfer coefficients, h_{hot} and h_{cold} , are calculated using the following relationships:

- (i) for laminar flow, i.e. $Re_d = \frac{\rho v d}{\mu} < 2000$, where v is the flow velocity and d is the diameter of the water pipe, the Nusselt number may be obtained using following empirical relation [124]

$$Nu_d = 1.86(Re_d Pr)^{\frac{1}{3}} \left(\frac{d}{L_p} \right)^{\frac{1}{3}} \left(\frac{\mu}{\mu_w} \right)^{0.14} \quad (8.30)$$

where L_p is the length of the water pipe, μ and μ_w are the dynamic viscosity of water at inlet and wall temperature respectively.

- (ii) for turbulent flow, i.e. $Re_d = \frac{\rho v d}{\mu} > 2000$, the Nusselt number may be calculated as [134]

$$Nu_d = 0.023 Re_d^{0.8} Pr^n \quad (8.31)$$

where

$$n = \begin{cases} 0.4 & \text{for heating of the water,} \\ 0.3 & \text{for cooling of the water.} \end{cases}$$

8.4 Lumped Parameter Thermal Model

Lumped-parameter circuits, consisting of a network of thermal resistances, thermal capacitances, nodal temperatures and heat sources, have been used extensively to represent the complex distributed thermal parameters of electrical machines [84, 159, 245].

8.4.1 Thermal Equivalent Circuit

A *thermal equivalent circuit* is essentially an analogy of an electrical circuit, in which the heat (analogous to *current*) flowing in each path of the circuit is given by a temperature difference (analogous to *voltage*) divided by a thermal resistance (analogous to *electrical resistance*). For conduction, the thermal resistance depends on the thermal conductivity of the material, k , and the length, l , and cross-sectional area, A_d , of the heat flow path and may be expressed as

$$R_d = \frac{l}{A_d k} \quad (8.32)$$

Thermal resistances for convection is defined as:

$$R_c = \frac{1}{A_c h} \quad (8.33)$$

where A_c is the surface area of convective heat transfer between two regions and h is the convection coefficient.

The thermal resistance for radiation between two surfaces is

$$R_r = \frac{\frac{1-\varepsilon_1}{\varepsilon_1 A_1} + \frac{1}{A_1 F_{12}} + \frac{1-\varepsilon_2}{\varepsilon_2 A_2}}{\sigma[(\vartheta_1 + 273) + (\vartheta_2 + 273)][(\vartheta_1 + 273)^2 + (\vartheta_2 + 273)^2]} \quad (8.34)$$

It is seen that the radiation thermal resistance in eqn (8.34) depends on the difference of the third power of the temperature, the surface spectral property ε and the surface orientation taken into account by a form factor F .

The thermal circuit in the steady state consists of thermal resistances and heat sources connected between motor component nodes. For transient analysis, the thermal capacitances are used additionally to account for the change of internal energy in the various parts of the machine with time. The heat capacitance is defined as:

$$C = \rho V c_v = m c_v \quad (8.35)$$

where c_v is the heat capacity of the material, ρ is the density, and V and m are the volume and mass of the material respectively. Fig. 8.13a shows a sectional view of an AFPM machine with a coreless stator. It can be observed that the AFPM stator is symmetrical from a heat transfer perspective and each half of the machine from the centre line mirrors the other half. It is therefore reasonable to model only half of the machine as shown in Fig. 8.13b.

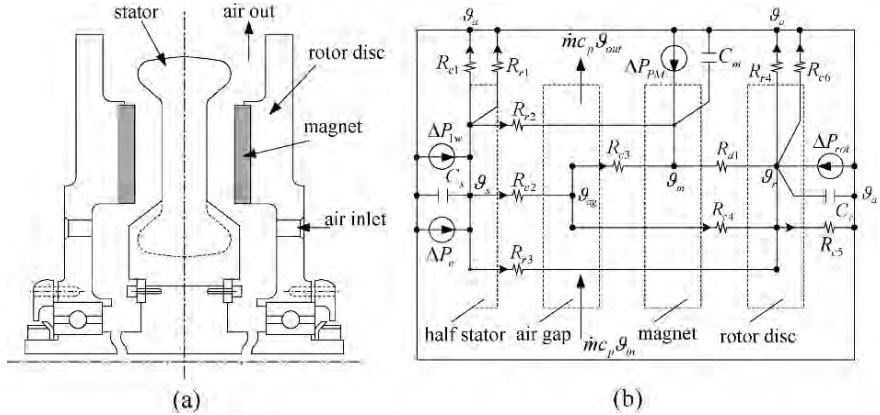


Fig. 8.13. The thermal resistance circuit of an AFPM brushless machine with coreless stator.

The heat source terms ΔP_{1w} , ΔP_e , ΔP_{PM} and ΔP_{rot} stand for winding losses (2.49), eddy current losses in one half of the stator winding (2.68), losses in PMs (2.61) and rotational losses (2.70) per one rotor disc respectively. C_s , C_m and C_r are the thermal capacitances of stator, PMs and rotor steel disc respectively. The heat resistances used in the circuit are described in Table 8.3.

In the lumped parameter thermal circuit analysis, it is often assumed that the temperature gradients within certain part of the machine is negligible. This assumption can only be made if the internal resistance to heat transfer is small compared with the external resistance [124]. The *Biot number*, B_i , is often used for determining the validity of this assumption. In the case that internal conduction resistance is compared with external convection resistance, B_i is defined as:

$$B_i = \frac{\bar{h}_c L}{k_s} \quad (8.36)$$

where k_s is the thermal conductivity of the solid material, L is the characteristic length of a solid body and $\bar{h}_c$ is the convection heat transfer coefficient. The criterion $B_i < 0.1$ ensures that the internal temperature will not differ

Table 8.3. Definition of thermal resistances

Symbols	Definition
R_{c1}	Convection resistance from stator end-winding to open air
R_{c2}	Convection resistance from stator to air-gap
R_{c3}	Convection resistance from air-gap to permanent magnets
R_{c4}	Convection resistance from air-gap to rotor disc plate
R_{c5}	Convection resistance from rotor disc to open air
R_{c6}	Convection resistance from rotor radial periphery to open air
R_{r1}	Radiation resistance from stator end-winding to environment
R_{r2}	Radiation resistance from stator to permanent magnets
R_{r3}	Radiation resistance from stator to rotor disc
R_{r4}	Radiation resistance from rotor radial periphery to environment
R_{d1}	Conduction resistance from PMs to rotor disc

from that at the surface of a solid body by more than 5% [190], and thus is frequently used.

8.4.2 Conservation of Energy

If conservation of energy is applied, the rate of internal energy change in each part of a machine (also called *control volume*) may be written as follows:

$$\frac{\Delta U}{\Delta t} = C \frac{\Delta \vartheta}{\Delta t} = \Delta P_{in} - \Delta P_{out} + \dot{m}_{in} i_{in} - \dot{m}_{out} i_{out} \quad (8.37)$$

where U is the internal energy, $\dot{m}$ is the mass flow rate and i is the *enthalpy* and C is the thermal capacitance of a control volume.

For steady-state conditions, $\frac{\Delta U}{\Delta t} = 0$ and therefore,

$$0 = \Delta P_{in} - \Delta P_{out} + \dot{m}_{in} i_{in} - \dot{m}_{out} i_{out} \quad (8.38)$$

These equations are applied to each part (the stator, air gap, PM and rotor disc) of the AFPM machine to obtain a set of equations with the temperatures of the parts being the only unknowns. This set of equations is rather complex but is readily solved using for example, the *Gauss-Seidel* iteration. It should be noted that the term $(\dot{m}_{in} i_{in} - \dot{m}_{out} i_{out})$ in the above equations represents internal heat removed due to the air flow pumped through the machine. This air flow is of paramount importance to the cooling of the machine. The determination of the air gap temperature in the thermal equivalent circuit is only possible if the mass flow rate through the air gap can be somehow predicted, hence the necessity of using the fluid-flow model described in section 8.3.1.

8.5 Machine Duties

Depending on the load conditions, there are mainly three types of duties or operation modes for all the electrical machines, i.e. continuous duty, short-time duty and intermittent duty.

8.5.1 Continuous Duty

When an electrical machine operates for a long period so that the temperatures of various parts of the machine reach steady-state at a given ambient temperature, this operation mode is called *continuous duty*. Due to the different physical properties, the final stabilised temperatures in various parts of the machine may vary greatly. The machine can be continuously operated for an infinitely long time without exceeding the temperature limits specified for each component of the machine. When only the solid parts of the machine are considered, the air flow term in the eqn (8.37) is ignored. The temperature rise versus time relationship in a control volume of the machine may be derived based on the theory of solid body heating as [159]:

$$\vartheta_c = \Delta P R(1 - e^{-\frac{t}{\tau}}) + \vartheta_o e^{-\frac{t}{\tau}} \quad (8.39)$$

where R is the thermal resistance, $\tau = RC$ is the thermal time constant, C is the thermal capacitance, ΔP is the heat loss flow, and ϑ_o is the initial temperature rise of the control volume. In a case where $\vartheta_o = 0$, the above equation is simply:

$$\vartheta_c = \Delta P R(1 - e^{-\frac{t}{\tau}}) \quad (8.40)$$

According to the properties of an exponential function, the final temperature rise $\vartheta_{cf} = \Delta P \times R$. Fig. 8.14a shows the typical transient temperature response of an AFPM machine operated continuously.

8.5.2 Short-Time Duty

Short-time duty means that machine operates only within a short specified period followed by a long period of rest or no-load condition. The operation time is so short that the machine does not attain its steady-state temperature, and the machine practically returns to its cold state after a long period of rest. Given the operation time t_s , the temperature rise of the machine can be found by using eqn (8.40) as

$$\vartheta_s = \Delta P R(1 - e^{-\frac{t_s}{\tau}}) = \vartheta_{cf} (1 - e^{-\frac{t_s}{\tau}}) \quad (8.41)$$

Compared with continuous duty, it is obvious that $\vartheta_s < \vartheta_{cf}$. This implies that the permissible load for the same machine in short time duty can be $1/(1 - e^{-\frac{t_s}{\tau}})$ times greater than that in continuous duty. Fig. 8.14b shows the typical temperature rise curve for short time duty machines.

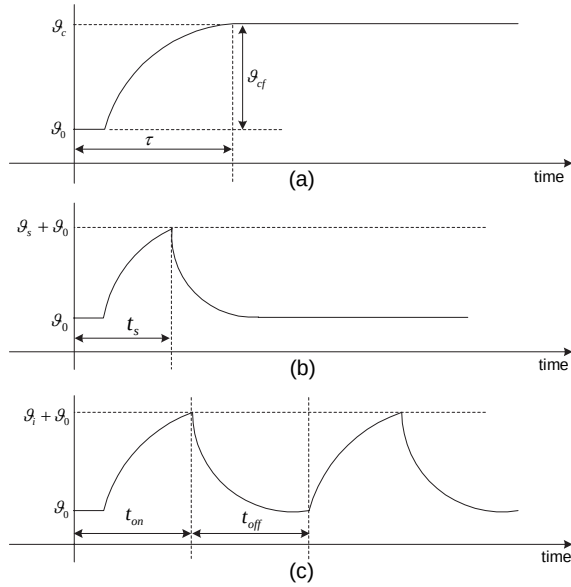


Fig. 8.14. Characteristic temperature curves of AFPM machines for different operation modes: (a) continuous duty, (b) short-time duty, (c) intermittent duty.

8.5.3 Intermittent Duty

The *intermittent duty* is characterised by short-time operations alternated with short-time pause intervals. Suppose a machine operates for a short period of t_{on} and then stops for a period of t_{off} , the cycle t_{cy} is then $t_{cy} = t_{on} + t_{off}$. The duty cycle d_{cy} may be defined as:

$$d_{cy} = \frac{t_{on}}{t_{on} + t_{off}} \quad (8.42)$$

The temperature rise of the machine ϑ_i during the t_{on} period can be calculated by using eqn (8.39) provided that thermal time constant and the steady-state temperature rise for continuous operation duty are known. During the t_{off} period, the machine loss $\Delta P = 0$ and the machine's temperature decreases according to an exponential function, i.e. $\vartheta_i e^{-t_{off}/\tau}$ before the second cycle sets in.

After many subsequent cycles, the machine's temperature variation becomes uniform and tends to be within a certain limited range (see Fig. 8.14c).

Under the same load and cooling condition, the maximum stabilised temperature of a machine in intermittent operation duty is smaller than that in continuous duty. Hence, similar to short-time operation duty, a machine operated in intermittent duty has overload capacity.

Numerical Example 8.1

A self-cooled 8-pole, 16-kW AFPM generator with an ironless stator as shown in Fig. 8.2 will be considered. The outer diameter is $D_{out} = 0.4$ m and the inner diameter is $D_{in} = 0.23$ m. The magnet width-to-pole pitch ratio $\alpha_i = 0.8$ and thickness of a rotor disc $d = 0.014$ m. The measured flow characteristic curves are shown in Fig. 8.15. At rated speed 1260 rpm, the total losses are 1569 W, of which (i) rotational losses $\Delta P_{rot} = 106$ W; (ii) eddy current losses in the stator $\Delta P_e = 23$ W; (iii) stator winding losses (rated) $\Delta P_{1w} = 1440$ W. Find:

- (a) Convective heat transfer coefficients of the disc system
- (b) Steady-state temperatures at different parts of the machine.

Solution

(a) Convective heat transfer coefficients of the disc system

The dynamic viscosity, density and thermal conductivity of air are assumed to be $\mu = 1.8467 \times 10^{-5}$ Pa s, $\rho = 1.177$ kg/m³ and $k = 0.02624$ W/(m °C) respectively in the following convective coefficients calculations.

Convection coefficient: outside rotor disc surface

At rated speed, the Reynolds number according to eqn (2.72) is

$$Re = \rho \frac{\Omega D_{out}^2}{4\mu} = 1.177 \times \frac{2\pi \times 1260/60 \times 0.4^2}{4 \times 1.846 \times 10^{-5}} = 336384.7$$

According to eqn (8.9) the transition between laminar and turbulent flow takes place at

$$r_c = \sqrt{\frac{2.5 \times 10^5 \nu}{\Omega}} = \sqrt{\frac{2.5 \times 10^5 \times 1.569 \times 10^{-5}}{2\pi \times 1260/60}} = 0.172 < \frac{D_{out}}{2} \text{ m}$$

The average Nusselt number of the disc according to eqn (8.8) is

$$\overline{Nu} = 0.015 \times Re^{\frac{4}{5}} - 100 \times \left(\frac{2r_c}{D_{out}}\right)^2$$

$$= 0.015 \times 336384.7^{\frac{4}{5}} - 100 \times \left(\frac{2 \times 0.172}{0.4} \right)^2 = 321.9$$

The average heat transfer coefficient at the outer surface of the disc according to eqn (8.5) is

$$\bar{h}_{fr} = \frac{k}{D_{out}/2} \times \overline{Nu} = \frac{0.02624}{0.4/2} \times 321.9 = 42.2 \text{ W/(m}^2 \text{ }^\circ\text{C)}$$

Convection coefficient: rotor disc peripheral edge

The Prandtl number is taken as $Pr = 0.7$ (Atmospheric pressure, 25 °C). The Reynolds number at disc periphery according to eqn (8.12) is

$$Re_D = \Omega \frac{D_{out}^2}{\nu} = \frac{2\pi \times 1260}{60} \frac{0.4^2}{1.569 \times 10^{-5}} = 1345538.8$$

The average Nusselt number according to eqn (8.11) is

$$\overline{Nu} = 0.133 \times Re_D^{\frac{2}{3}} \times Pr^{\frac{1}{3}} = 0.133 \times 1345538.8^{\frac{2}{3}} \times 0.7^{\frac{1}{3}} = 1439.3$$

The average heat transfer coefficient around the radial periphery according to eqn (8.10) is

$$\bar{h}_p = \frac{k}{D_{out}} \times \overline{Nu} = \frac{0.02624}{0.4} \times 1439.3 = 94.4 \text{ W/(m}^2 \text{ }^\circ\text{C)}$$

Convection coefficient: rotor-stator system

The volumetric flow rate of the machine at rated speed can be looked up from Fig. 8.15. Assuming equal flow on both sides of stator, the equivalent flow rate is taken as $Q = 0.013 \text{ m}^3/\text{s}$. The average Nusselt number according to eqn (8.15) is

$$\overline{Nu} = 0.333 \frac{Q}{\pi \nu (D_{out}/2)} = 0.333 \times \frac{0.013}{\pi 1.566 \times 10^{-5} \times (0.4/2)} = 440$$

The average heat transfer coefficient between two discs according to eqn (8.5) is

$$\bar{h}_{rs} = \frac{2k}{D_{out}} \times \overline{Nu} = \frac{2 \times 0.02624}{0.4} \times 440 = 57.7 \text{ W/(m}^2 \text{ }^\circ\text{C)}$$

(b) Steady-state temperatures at different parts of the machine

If the radiation from the rotor disc to ambient, the convection from the stator overhang to air flow, and the conduction resistance between magnets and rotor are ignored, the generic thermal equivalent circuit given in Fig. 8.13 is simplified as shown in Fig. 8.16.

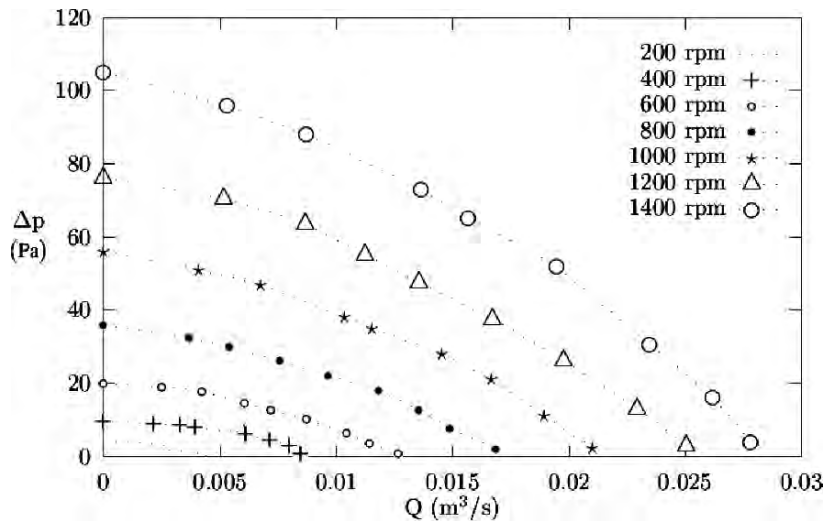


Fig. 8.15. Measured characteristic curves of the AFPM machine. *Numerical example 8.1.*

Control volume 1 (half of the stator)

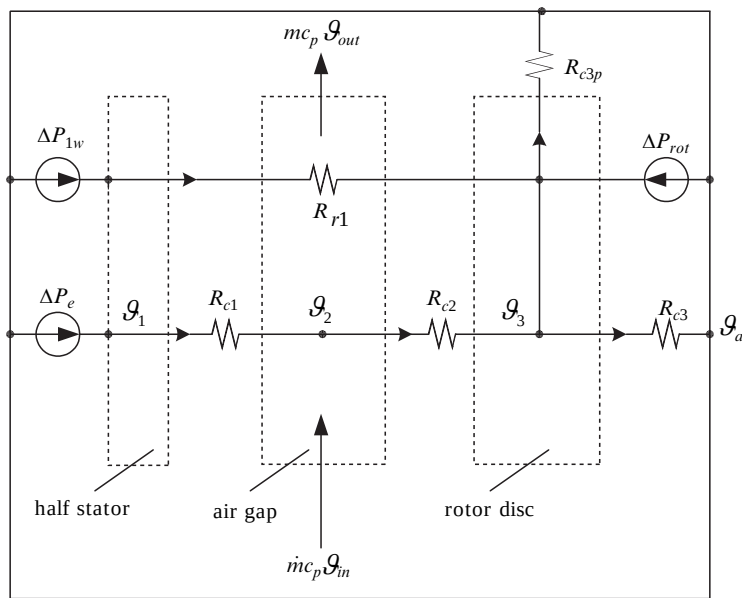


Fig. 8.16. Simplified thermal equivalent circuit. *Numerical example 8.1.*

The convection heat transfer resistance between the stator and air flow in the air gap is

$$R_{c1} = \frac{1}{\bar{h}_{rs} \times \frac{\pi}{4}(D_{out}^2 - D_{in}^2)} = \frac{1}{57.7 \times \frac{\pi}{4}(0.4^2 - 0.23^2)} = 0.223 \text{ } ^\circ\text{C/W}$$

The radiation heat transfer resistance between the stator and rotor discs is

$$R_{r1} = \frac{\frac{1-\varepsilon_1}{\varepsilon_1 A_1} + \frac{1}{A_1 F_{12}} + \frac{1-\varepsilon_2}{\varepsilon_2 A_2}}{\sigma[(\vartheta_1 + 273) + (\vartheta_2 + 273)][(\vartheta_1 + 273)^2 + (\vartheta_2 + 273)^2]}$$

in which the areas of both discs can be taken as the same, i.e. $A_1 = A_2 = \frac{\pi}{4}(D_{out}^2 - D_{in}^2) = 0.084 \text{ m}^2$, the shape factor $F_{12} = 1$, the *Stefan-Boltzmann* constant $\sigma = 5.67 \times 10^{-8} \text{ W/(m}^2 \text{ K}^4)$, the emissivity of epoxy encapsulated stator $\varepsilon_1 = 0.85$. Since part of the rotor disc is covered with PMs, the emissivity of the rotor disc is defined based on the proportion of different materials, i.e.

$$\varepsilon_2 = \varepsilon_{fe}\alpha_i + \varepsilon_{pm}(1 - \alpha_i) = 0.3 \times 0.8 + 0.9 \times (1 - 0.8) = 0.42$$

Apparently R_{r1} is a function of ϑ_1 and ϑ_2 , an iterative approach has to be used to solve R_{r1} for different temperatures.

According to conservation of energy, the steady-state energy equation for *control volume 1* can be written as

$$\frac{1}{2}(\Delta P_{1w} + \Delta P_e) - \frac{\vartheta_1 - \vartheta_2}{R_{c1}} - \frac{\vartheta_1 - \vartheta_3}{R_{r1}} = 0 \quad (8.43)$$

Control volume 2 (*air gap*)

The convection heat transfer resistance from air gap to the rotor disc may be taken as the same as that from stator to air gap, i.e. $R_{c2} = R_{c1}$. The mass flow rate $\dot{m} = \rho Q = 1.177 \times 0.013 = 0.0153 \text{ kg}$. Assume that the air temperature at the machine inlet is ambient temperature, i.e. $\vartheta_{in} = \vartheta_a$ and the air gap average temperature $\vartheta_2 = \frac{1}{2}(\vartheta_{out} + \vartheta_{in})$. The heat dissipated due to the air flow is

$$\begin{aligned} \dot{m}_{out}i_{out} - \dot{m}_{in}i_{in} &= \dot{m} c_p (\vartheta_{out} - \vartheta_{in}) = 2\dot{m} c_p (\vartheta_2 - \vartheta_a) \\ &= 2 \times 0.0153 \times 1005 \times (\vartheta_2 - 20) \end{aligned}$$

The steady-state energy equation for *control volume 2* is

$$\frac{\vartheta_1 - \vartheta_2}{R_{c1}} - \frac{\vartheta_2 - \vartheta_3}{R_{c2}} - 2 \times 0.0153 \times 1005 \times (\vartheta_2 - 20) = 0 \quad (8.44)$$

Control volume 3 (rotor disc)

The convection heat transfer resistance at the outside surface of the disc is

$$R_{c3} = \frac{4}{\bar{h}_{fr}\pi D_{out}^2} = \frac{4}{42.4 \times \pi \times 0.4^2} = 0.1877 \text{ }^\circ\text{C/W}$$

The convection heat transfer resistance at the periphery of the disc is

$$R_{c3p} = \frac{1}{\bar{h}_p \pi D_{out} d} = \frac{1}{94.4 \times \pi \times 0.4 \times 0.014} = 0.602 \text{ }^\circ\text{C/W}$$

The steady-state energy equation for *control volume 3* is

$$\frac{\vartheta_2 - \vartheta_3}{R_{c2}} + \frac{\vartheta_1 - \vartheta_3}{R_{r1}} + \frac{1}{2}\Delta P_{rot} - \frac{\vartheta_3 - \vartheta_a}{R_{c3}} - \frac{\vartheta_3 - \vartheta_a}{R_{c3p}} = 0 \quad (8.45)$$

Having established the energy equations (8.43), (8.44) and (8.45) for each part of the machine, the steady-state temperatures can be found by solving these equations. Due to the temperature dependency of R_{r1} , a simple computer program using *Gauss Seidel* iteration has been created to find the solutions of the equations. The results are given in Table 8.4.

Table 8.4. Predicted temperature rises

Machine parts	Temperature rise, $^\circ\text{C}$
Stator winding, ϑ_1	114.9
Air-gap flow, ϑ_2	21.35
Rotor disc, ϑ_3	18.32

Numerical Example 8.2

A totally enclosed AFPM brushless machine has a power loss of $\Delta P_{1W} = 2500$ W in the stator winding at continuous duty. The machine's outer and inner diameters are $D_{out} = 0.72$ m and $D_{in} = 0.5$ m respectively. To remove the heat from the stator, use is made of heat pipes for direct cooling of the motor as shown in Fig. 8.12. The convection heat transfer coefficients on the inside wall of heat pipes in the stator and in the finned area are assumed to be $h = 1000$ W/(m² °C). The average convective heat transfer coefficient on the fin surface is taken as $h_{fin} = 50$ W/(m² °C). The finned surface has an overall area of $A_{fin} = 1.8$ m² and an efficiency of $\eta_{fin} = 92\%$. The length of heat pipe embedded in the finned surface is $l_{fin} = 1.5$ m. Find:

- (a) Steady-state temperature of the stator winding if the heat pipe with a diameter $D_{hp} = 9$ mm is placed along the average radius of the stator

- (b) Steady-state temperatures of the stator winding if the heat pipe is replaced by a $d = 9$ mm water cooling pipe, in which water (with an average temperature of 60°C) flows at $v_w = 0.5$ m/s.

Solution

- (a) Steady-state temperature of the stator winding if the heat pipe with a diameter $D_{hp} = 9$ mm is placed along the average radius of the stator

Assuming the outside wall of the heat pipe is in full contact with the stator winding and finned surface, the exposed area of the heat pipe in the stator will be

$$A_{hot} = \pi D_{hp} \frac{\pi(D_{out} + D_{in})}{2} = 0.009\pi \times \frac{\pi(0.72 + 0.5)}{2} = 0.0542 \text{ m}^2$$

The exposed area of the heat pipe embedded in the finned surface will be

$$A_{cold} = \pi D_{hp} l_{fin} = 0.009\pi \times 1.5 = 0.0424 \text{ m}^2$$

Assuming the temperature of cooling air over the finned area to be $\vartheta_{cold} = 30^\circ\text{C}$, the steady-state temperature of the stator may be obtained by using eqn (8.29)

$$\begin{aligned} \vartheta_{stator} &= \vartheta_{cold} + \Delta P_{hp} \left(\frac{1}{h_{hot} A_{hot}} + \frac{1}{h_{cold} A_{cold}} + \frac{1}{\eta_{fin} h_{fin} A_{fin}} \right) \\ &= 30 + 2500 \left(\frac{1}{1000 \times 0.0542} + \frac{1}{1000 \times 0.0424} + \frac{1}{0.92 \times 50 \times 1.8} \right) = 165.3^\circ\text{C} \end{aligned}$$

- (b) Steady-state temperatures of the stator winding if the heat pipe is replaced by a water cooling pipe of diameter $d = 9$ mm, in which water flows at 0.5 m/s.

The Reynolds number is first calculated to determine the flow regime. The properties of water at 60°C are $\rho = 983.3 \text{ kg/m}^3$, $c_p = 4179 \text{ J/(kg }^\circ\text{C)}$, $\mu = 4.7 \times 10^{-4} \text{ Pa s}$, $k = 0.654 \text{ W/(m }^\circ\text{C)}$, $Pr = \mu c_p / k = 4.7 \times 10^{-4} \times 4179 / 0.654 = 3$, $Re_d = \rho v_w d / (2\mu) = 983.3 \times 0.5 \times 0.009 / 4.7 \times 10^{-4} = 9414.6 > 2000$, so that the flow is turbulent. Thus, the Nusselt number for the heating of the water is

$$Nu_{dh} = 0.023 Re_d^{0.8} Pr^{0.4} = 0.023 \times 9414.6^{0.8} \times 3^{0.4} = 53.9$$

while the Nusselt number for the cooling of the water is

$$Nu_{dh} = 0.023 Re_d^{0.8} Pr^{0.3} = 0.023 \times 9414.6^{0.8} \times 3^{0.3} = 48.3$$

The convection heat transfer coefficients of the water pipe inside the stator h_{hot} and in the finned area h_{cold} are calculated as

$$h_{hot} = \frac{k Nu_{dh}}{d} = \frac{0.654 \times 53.9}{0.009} = 3916.7 \text{ W/(m}^2 \text{ } ^\circ\text{C)}$$

$$h_{cold} = \frac{k Nu_{dc}}{d} = \frac{0.654 \times 48.3}{0.009} = 3509.8 \text{ W/(m}^2 \text{ } ^\circ\text{C)}$$

The steady-state temperature in the stator can be calculated as in (a) with the exception that the heat transfer coefficients are different

$$\begin{aligned} \vartheta_{stator} &= \vartheta_{cold} + \Delta P_{hp} \left(\frac{1}{h_{hot} A_{hot}} + \frac{1}{h_{cold} A_{cold}} + \frac{1}{\eta_{fin} h_{fin} A_{fin}} \right) \\ &= 30 + 2500 \times \left(\frac{1}{3916.7 \times 0.0542} + \frac{1}{3509.8 \times 0.0424} + \frac{1}{0.92 \times 50 \times 1.8} \right) = 88.8 \text{ } ^\circ\text{C} \end{aligned}$$

Applications

9.1 Power Generation

Brushless PM electrical machines are the primary generators for *distributed generation systems*. They are compact, high efficient and reliable self-excited generators. The distributed generation is any electric power production technology that is integrated within a distribution system. Distributed generation technologies are categorized as *renewable* and *nonrenewable*. Renewable technologies include solar, photovoltaic, thermal, wind, geothermal and ocean as sources of energy. Nonrenewable technologies include internal combustion engines, combined cycles, combustion turbines, microturbines and fuel cells.

AFPM brushless generators can be used both as high speed and low speed generators. Their advantages are high power density, modular construction, high efficiency and easy integration with other mechanical components like turbine rotors or flywheels. The output power is usually rectified and then inverted to match the utility grid frequency or only rectified. A method for active output voltage regulation in an AFPM automotive alternator by means of mechanical flux weakening is proposed in [34].

9.1.1 High Speed Generators

Although the minimization of windage losses of high speed generators requires rotors with small diameters, a multidisc design has the advantages of modular and compact construction, low synchronous reactance, good voltage regulation and very high efficiency in the case of coreless stators.

A multidisc AFPM *high speed generator* is usually driven by a gas microturbine. The turbine rotor and PM rotors are mounted on the same shaft. Such a generator has compact construction, low mass and very high efficiency.

According to *TurboGenset Company*, U.K., a 100-kW, 60 000 rpm multidisc generator has an outer diameter of 180 mm, length of 300 mm and weighs only 12 kg (Fig. 9.1)¹. The high frequency output is rectified to d.c.

¹ ULEV-TAP Newsletter, No. 2, 2000, www.ulev-tap.org

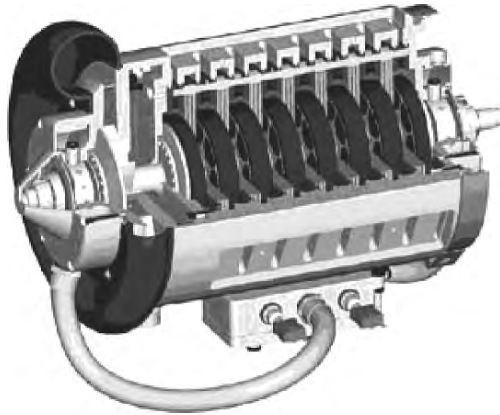


Fig. 9.1. 100-kW, 8-disc AFPM synchronous generator. Courtesy of *TurboGenset*, London, U.K.

and then inverted to 50, 60 or 400 Hz a.c. (Fig. 2.20). The generator is totally air cooled.

Armed forces are interested in applications of microturbine driven PM synchronous generators to battery chargers for soldiers. The low energy density of portable backpack batteries imposes a major constraint and challenge to future infantry operations. With recent advancement in PM brushless machine technologies, a lightweight *miniature generator set* can minimize the

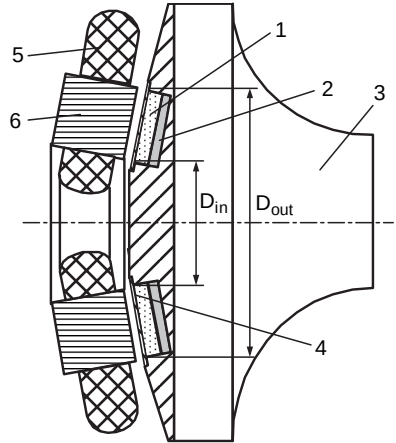


Fig. 9.2. High speed AFPM synchronous generator integrated with the rotor of a microturbine. 1 — SmCo PM, 2 — backing steel ring, 3 — rotor of microturbine, 4 — nonmagnetic retaining ring, 5 — stator winding, 6 — stator core.

mass of heavy batteries and charge them during field operations. A miniature generator can deliver electric power for a long period of time limited only by the fuel availability. Easy and rapid refueling can be done in the field by using, e.g. kerosene. Fig. 9.2 shows a tiny microturbine integrated with a miniature AFPM synchronous generator. At speed 150 000 to 250 000 rpm and outer diameter of PM ring $D_{out} \approx 50$ mm, the generator can deliver about 1 kW electric power.

Table 9.1. Specifications of five phase Kestrel AFPM synchronous generators			
Type of generator	Kestrel 600	Kestrel 800	Kestrel 2000
Rated power, W	400 at 12.5m/s	800 at 11.5m/s	2000 at 10.5m/s
Maximum power, W	600 at 14.5m/s	850 at 12m/s	2200 at 11m/s
Number of poles $2p$	48	48	200
Generation steps	2	2	4
Rated wind speed, m/s	12.5	11.5	10.5
Cut in wind speed, m/s	2.5	2.5	2.2
Rated rotational speed, rpm	1100	1010	925
Rotor diameter, m	1.2	2.1	3.6
Number of blades	6	3	3
Type of blade	Basic aerofoil		Full aerofoil
Tower top weight, kg	23	35	80
Lateral thrust, N	250	300	450
Speed control	Self stall		Dynamic tail
Standard rectified d.c. voltage, V	12, 24, 36, 48		48
Protection	IP55		

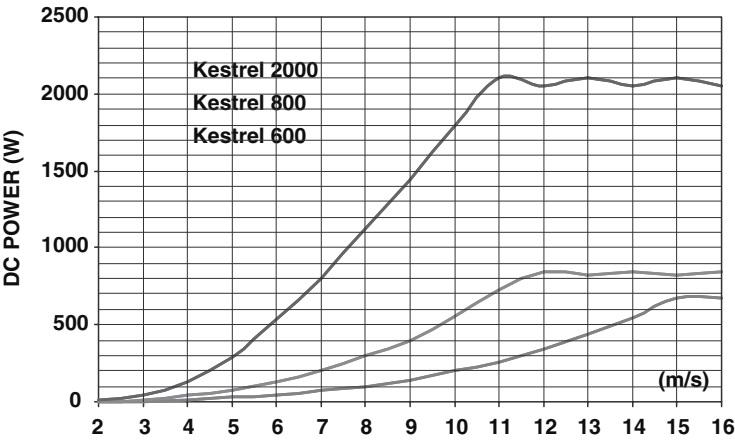


Fig. 9.3. Output power versus wind speed of Kestrel AFPM synchronous generators. Courtesy of Kestrel Wind Turbines, South Africa.

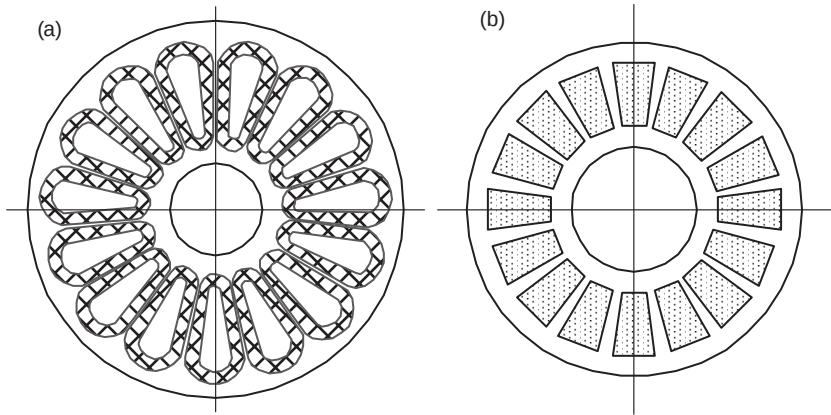


Fig. 9.4. Stator and rotor of a medium power AFPM generator with inner coreless stator and twin external rotor: (a) stator coils, (b) rotor disc with NdFeB PMs.

9.1.2 Low Speed Generators

A *low speed AFPM generator* is usually driven by a *wind turbine*. With wind power rapidly becoming one of the most desirable alternative energy sources world-wide, AFPM generators offer the ultimate low cost solution as compared with e.g. solar panels. Table 9.1 shows specifications of five-phase AFPM synchronous generators manufactured by *Kestrel Wind Turbines (Pty) Ltd*, Gauteng, South Africa. All three types of Kestrel generators use laminated cores. The Kestrel 2000 is a double-sided AFPM generator with twin rotors. The Kestrel 600 generator rated at 400 W is in fact capable of delivering power in excess of 600 W. Controlled by a shunt regulator the maximum charging current is 50 A into a 12 V d.c. battery bank. At wind speeds of 52.2 km/h, which translates into 14.5 m/s (28.2 knots) and rotational speed of 1100 rpm, the power output is 600 W (Fig. 9.3).

Figure 9.4 shows a medium-power AFPM generator with twin external rotor and inner coreless stator for a 5-m blade diameter wind turbine. The stator system consists of 15 non-overlapping concentrated coils, one coil per phase. The thickness of coil is 12 mm and diameter of wire 0.6 mm. Each coil is individually rectified to d.c., to reduce the commutation torque ripple effect and provide better control over the voltage. The rotor has $2p = 16$ poles. Performance characteristics are shown in Fig. 9.5.

Figure 9.6 shows the construction detail of a 2 kW 500 rpm AFPM air-cored wind turbine-driven generator with concentrated coil topology developed by Durham University, UK [27]. The stator is located between two rotor discs and contains 12 circular bobbin armature coils while each rotor disc has 16 circular NdFeB magnets. The rotor discs are mechanically joined by an outer location ring and through bolts. The magnets are secured and positioned

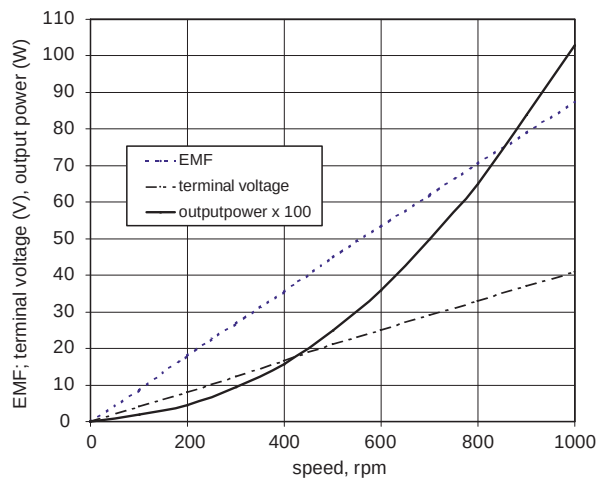


Fig. 9.5. Steady-state performance characteristics of a 10-kW AFPM generator shown in Fig. 9.4.

Table 9.2. Specifications of the AFPM wind generator [27]

Rated speed, rpm	500
Rated power, W	2000
Number of phases	3
Number of magnets	16
Number of armature coils	12
Frequency, Hz	66.67
Magnet type	NdFeB
Magnet diameter, mm	65
Magnet thickness, mm	8
Pole pitch at mean diameter, mm	70
Pole pitch, deg	22.5
Rotor thickness, mm	10.0
Generator overall diameter, mm	495
Generator axial length, mm	67
Stator thickness, mm	20
Pitch of armature coil, deg	30
Outer diameter of armature coil, mm	75.0
Inner diameter of armature coil, mm	30.0
Coil height, mm	14
Wire diameter, mm	1.0
Number of turns/coil	276
Running clearance, mm	3

in a retainer ring. The stator is supported from the inside using a double-shaft arrangement. The generator is of an overall diameter 495 mm and axial length 67 mm. The main generator parameters are given in Table 9.2. The advantages associated with this type of generator are as follows [27]:

- circular coils offer the shortest turns and lowest resistance;
- simple manufacture and low production costs;
- easy assembly as no magnetic forces between rotor and stator;
- no cogging torque and thus presents little mechanical resistance to turbine starting.

A prototype generator has been manufactured and integrated with a vertical axis wind turbine as shown in Fig. 9.7.

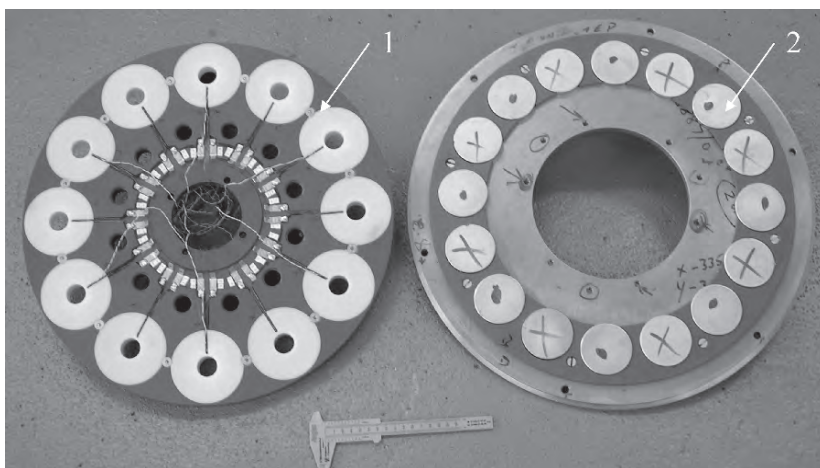


Fig. 9.6. The stator and rotor layout of the 2 kW 500 rpm air-cored AFPM wind generator with concentrated coil topology developed by Durham University: 1 — bobbin wound air-cored stator, 2 — rotor disc. Courtesy of *Durham University*, UK.

9.2 Electric Vehicles

Electric vehicles are divided into three general categories: *hybrid electric vehicles* (HEVs), *battery electric vehicles* (EVs) and *fuel cell electric vehicles* (FCEVs). The switch from gasoline to HEVs and EVs/FCEVs would reduce the total primary energy consumed for personal transportation. Traction electric motors for HEVs and EVs/FCEVs should meet the following requirements:



Fig. 9.7. Generator integrated with a vertical axis wind turbine. Courtesy of *Durham University* and *Carbon Concepts*, UK.

- power rating: high instant power, high power density;
- torque–speed characteristics: high torque at low speed for starting and climbing, high speed at low torque for cruising, wide speed range including constant torque region and constant power region, fast torque response;
- high efficiency over wide speed and torque ranges;
- high reliability and robustness under various operating conditions, e.g. at high and low temperature, rain, snow, vibration, etc.;
- low cost.

9.2.1 Hybrid Electric Vehicles

HEVs are now at the forefront of transportation technology development. *HEVs* combine the internal combustion engine of a conventional vehicle with the electric motor of an EV, resulting in twice the fuel economy of conventional vehicles. The electric motor/generator is usually located between the combustion engine and gear box. One end of the rotor shaft of the electric motor/generator is bolted to the combustion engine crankshaft, while the opposite end can be bolted to the flywheel or gearbox via the clutch (Fig. 9.8). The electric motor serves a number of functions, i.e.:

- assisting in vehicle propulsion when needed, allowing the use of a smaller internal combustion engine;

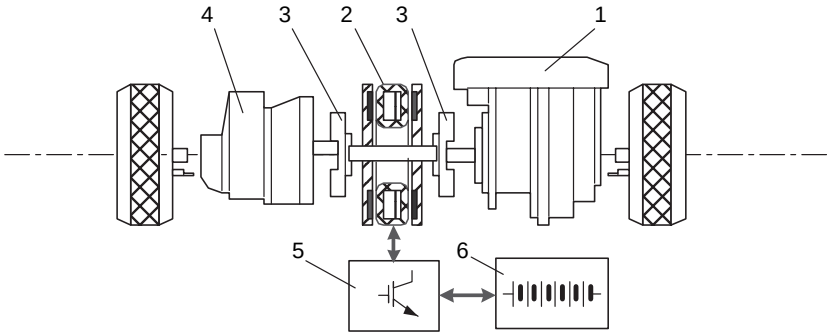


Fig. 9.8. Hybrid electric gasoline car: 1 - gasoline combustion engine, 2 - integrated motor-generator, 3 - cranking clutch, 4 - gearbox, 5 - inverter, 6 - battery.

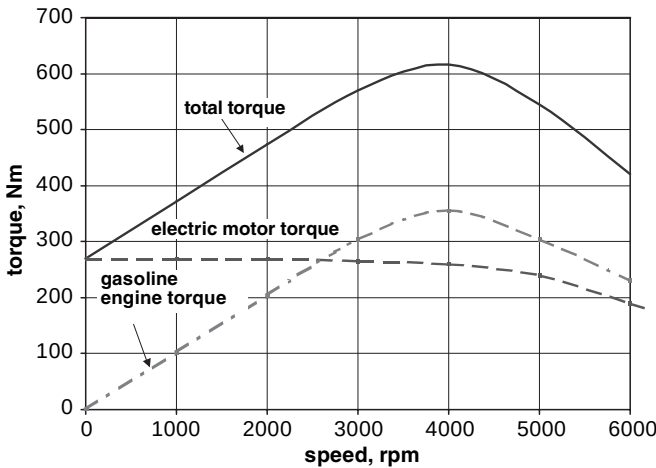


Fig. 9.9. Torque-speed characteristics of an electric motor and gasoline engine. The electric motor assists the gasoline engine at low speeds.

- operating as a generator, allowing excess energy (during braking) to be used to recharge the battery;
- replacing the conventional alternator, providing energy that ultimately feeds the conventional low voltage, e.g. 12 V electrical system;
- starting the internal combustion engine very quickly and quietly. This allows the internal combustion engine to be turned off when not needed, without any delay in restarting on demand;
- damping crankshaft speed variations, leading to smoother idle.

In a hybrid electric gasoline car the electric motor assists the gasoline engine in the low speed range by utilizing the high torque of the electric motor, as shown in Fig. 9.9. Currently manufactured hybrid electric gasoline cars

(Fig. 9.8) are equipped either with cage induction motors or PM brushless motors. A PM brushless motor can increase the overall torque by more than 50%. In most applications, the rated power of electric motors is from 10 to 75 kW. Because of limited space between the combustion engine and gear box as well as the need to increase the flywheel effect, electric motors for HEVs are short and have large diameters. AFPM brushless machines are pancake-type high torque density motors and fit perfectly the HEV requirements. An AFPM brushless machine can be liquid cooled and integrated with a power electronics converter.

In the HEV shown in Fig. 9.10 the energy is stored in a flywheel integrated with the AFPM machine. The vehicle may operate without an electrochemical battery. If the electromagnetic conversion process is included within the flywheel, there is no external shaft end, because the power flow to and from the machine is via electric connections to the stationary armature [3]. Thus, complex vacuum seals are avoided.

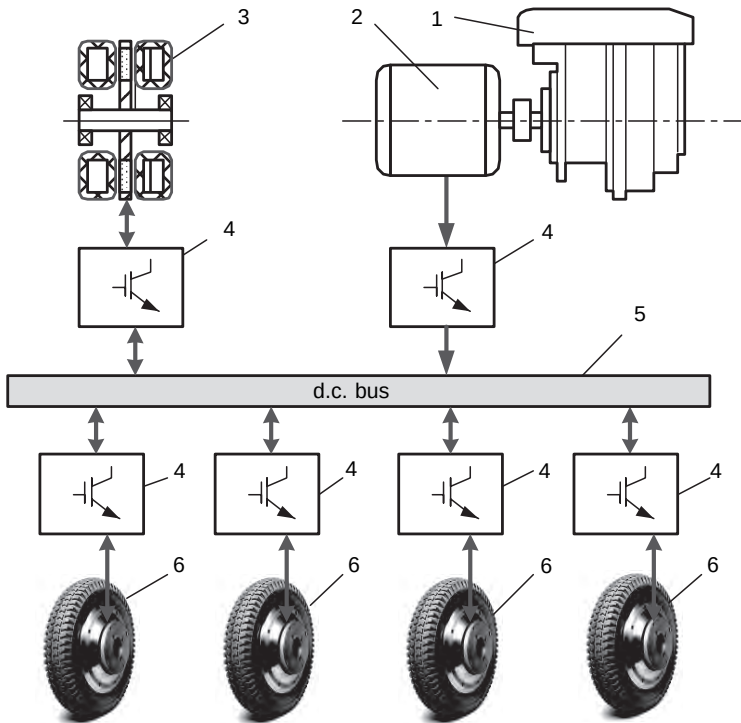


Fig. 9.10. HEV with flywheel energy storage: 1 — gasoline combustion engine, 2 — brushless generator, 3 — integrated flywheel-motor-generator, 4 — solid state converter, 5 — d.c. bus, 6 — electrical motorized wheel.

The moment of inertia and tensile stress are respectively

- for a rotating disc

$$J = \frac{1}{2}mR_{out}^2, \quad \sigma = \frac{\rho\Omega^2 R_{out}^2}{3} \quad (9.1)$$

- for a rotating ring

$$J = \frac{1}{2}m(R_{in}^2 + R_{out}^2), \quad \sigma = \frac{\rho\Omega^2}{3}(R_{in}^2 + R_{in}R_{out} + R_{out}^2) \quad (9.2)$$

where ρ is the specific mass density of the rotating body, m is the mass, $\Omega = 2\pi n$ is the angular speed, R_{out} is the outer radius and R_{in} is the inner radius. The kinetic energy is

- for a rotating disc

$$E_k = \frac{1}{2}J\Omega^2 = \pi^3 \rho d R_{out}^4 n^2 \quad \text{J} \quad (9.3)$$

- for a rotating ring

$$E_k = \pi^3 \rho d (R_{out}^4 - R_{in}^4) n^2 \quad \text{J} \quad (9.4)$$

where d is the thickness of the flywheel (disc or ring).

Combining together pairs of eqns (9.1), (9.3) and (9.2), (9.4), the stored energy density in a flywheel can be expressed as

- for a rotating disc

$$e_k = \frac{E_k}{m} = \frac{3}{4} \frac{\sigma}{\rho} \quad \text{J/kg} \quad (9.5)$$

- for a rotating ring

$$e_k = \frac{E_k}{m} = \frac{3}{4} \frac{R_{in}^2 + R_{out}^2}{R_{in}^2 + R_{in}R_{out} + R_{out}^2} \frac{\sigma}{\rho} \quad \text{J/kg} \quad (9.6)$$

For a very high strength steel, e.g. alloy steel (3% CrMoC) with $\sigma \approx 1000$ MPa and $\rho = 7800$ kg/m³ [3], the stored energy density of a disc-type flywheel is $e_k = 96.1$ kJ/m³. If we need to store a kinetic energy $E_k = 500$ kJ, the mass of the disc will be 5.2 kg.

9.2.2 Battery Electric Vehicles

Battery EVs are vehicles that use secondary batteries (rechargeable batteries or storage batteries) as their only source of energy. An EV power train can convert energy stored in a battery into vehicle motion and it can also reverse direction and convert the kinetic energy of the vehicle back into the storage battery through regenerative braking.

AFPM brushless motors are used in EVs as in-wheel motors (Figs 6.8 and 9.11) [215, 225]. The pancake shape of an AFPM motor allows for designing a compact motorized wheel.

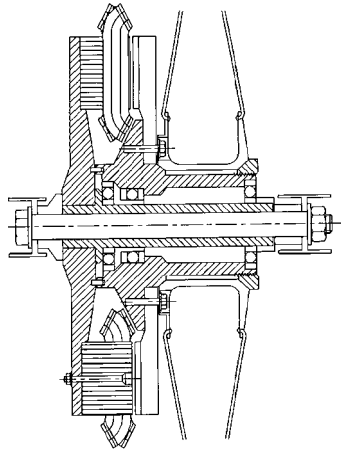


Fig. 9.11. Single-sided AFPM brushless motor fitted to spoked wheel [215].

With the aid of the AFPM brushless motor the differential mechanism can be replaced by an electronic differential system [101]. The configuration shown in Fig. 9.12a illustrates the use of a pair of electric motors mounted on the chassis to drive a pair of wheels through drive shafts, which incorporate constant velocity joints. In the configuration shown in Fig. 9.12b, the motors forming the electronic differential are mounted directly in the wheels of the vehicle. The electromechanical drive system is considerably simplified, when the motor is mounted in the wheel, because the drive shafts and constant velocity joints are now no longer needed. However, the resultant “unsprung” wheel mass of the vehicle is increased by the mass of the motor. Wheel motors of this direct drive configuration also suffer overloading because the speed of the rotor is lower than would be the case with a geared arrangement. This leads to an increase of the active materials volume required in the motor.

The disadvantages of conventional wheel motors can be overcome by using the arrangement shown in Fig. 9.13. The two stators are directly attached to

the vehicle body whilst the PM rotor is free to move in radial directions. It can be observed that in this case the wheel and disc rotor form the unsprung mass, whilst the stators of the motor become sprung mass supported on the chassis [101].

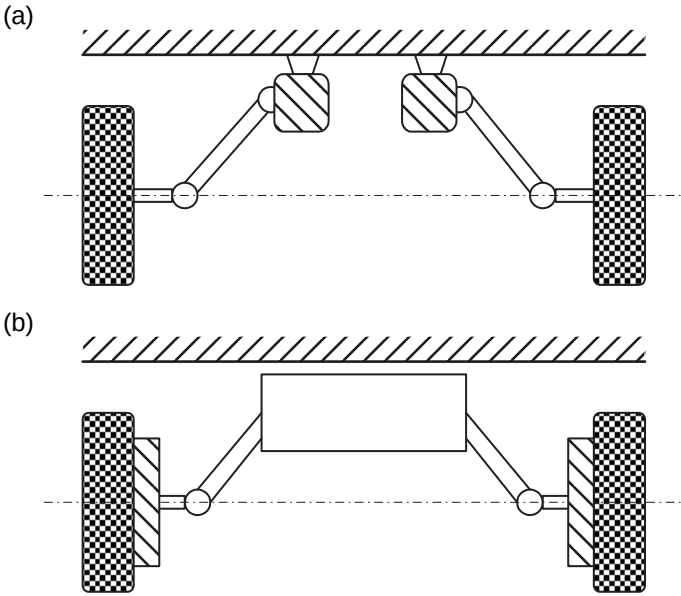


Fig. 9.12. Alternative forms of "electronic differential" drive schemes: (a) onboard motor, (b) wheel mounted motor.

Given the advantages of using wheel motor in electric vehicle drive system such as more onboard space, better vehicle control and modular drive train topology, there is a growing interest in automotive industry to develop wheel motors for future electric vehicles. Figure 9.14 shows the AFPM wheel motor developed at General Motors Advanced Technology Center, Torrance, CA, U.S.A. The developed wheel hub motor has a single stator located between two PM rotor discs. This topology features higher torque density design as it utilizes copper from both surfaces of the stator for torque production [66, 199, 223]. To remove heat from inside stator winding, provision is made of an aluminum ring with internal liquid coolant passages along end winding of the outer stator winding [P149]. The design specification of the GM's wheel motor is summarised in Table 9.3. To demonstrate the performance and robustness of the AFPM wheel motor technology, two wheel motors were fitted on GM's S-10 show truck and underwent several thousand miles road tests.

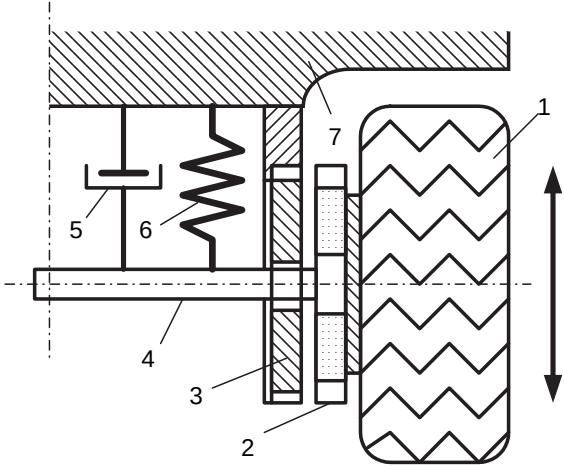


Fig. 9.13. Schematic representation of the disc-type machine with lower unsprung mass: 1 — wheel, 2 — disc rotor, 3 — stator, 4 — shaft, 5 — damper, 6 — spring, 7 — chassis.

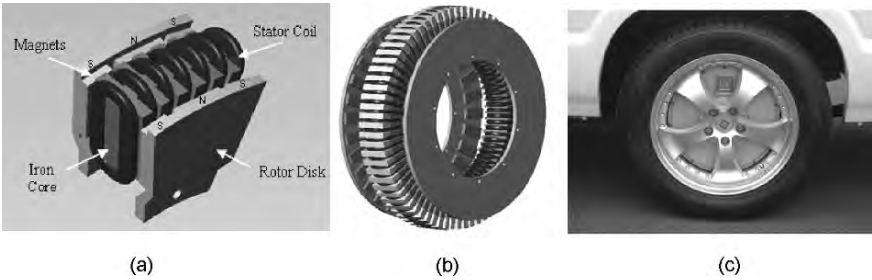


Fig. 9.14. GM's AFPM wheel motor: (a) Section of the AFPM motor showing details of slotted stator, (b) Basic layout of the AFPM motor showing complete stator located between two PM rotor discs, (c) AFPM wheel motor mounted on GM S-10 show truck. Courtesy of *General Motors Corporation*, U.S.A.

Table 9.3. GM's wheel motor design specification [66]

Number of phases	3
Number of poles	24
Peak torque, Nm	500
Stator outer diameter, m	340
Maximum axial length, m	75
Peak power, kW	25
Nominal bus voltage, V	280
Peak current, A	150
Maximum speed, RPM	1200

9.2.3 Fuel Cell Electric Vehicles

The fuel cell principle was first discovered by William Robert Grove in 1839 [272, 273]. Although fuel cells are well-proven technologies, they are presently still at a pre-commercial stage. Being widely regarded as a transforming technology, fuel cells offer the prospects of great improvements in energy efficiency for both mobile and stationary applications, and of a long-term transition to the hydrogen economy. In transport sector, the light-duty automotive applications are potentially the largest market opportunity available to fuel cell technology and have been the focus of intense development effort [76, 98]. Most major car manufacturers have on-going fuel cell vehicle programmes.

Fuel cells can be used with hydrogen or hydrocarbon fuels. The high cost of producing hydrogen, as compared to hydrocarbon fuels, has been a major barrier to its commercialisation, which would also require major infrastructural changes such as build-up of a hydrogen production and distribution infrastructure. However, with the continuing technology development in this fast-moving field and increasing possibility for introduction of renewable energy sources into fuel supply path, costs are expected to decline to an extent within the next decade.

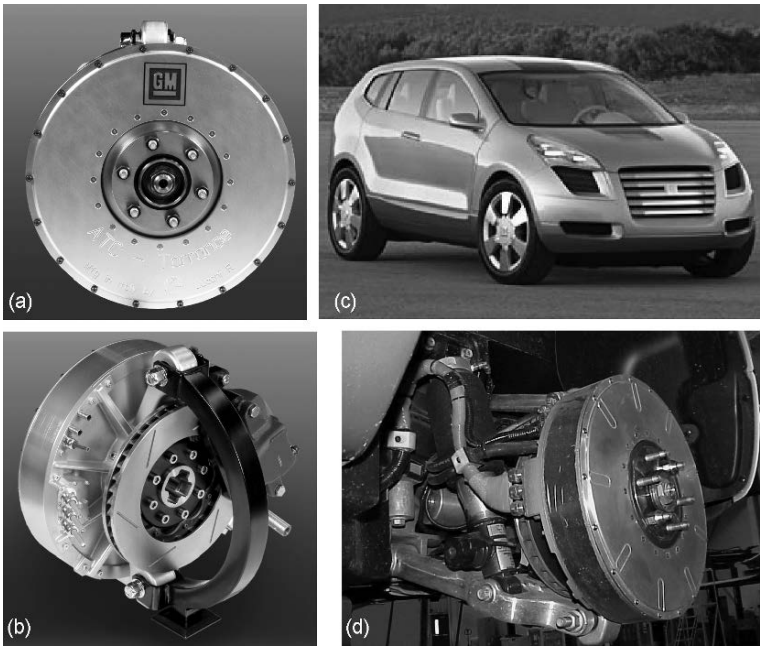


Fig. 9.15. GM's Sequel fuel cell concept electric vehicle: (a) Front view of the AFPM wheel motor, (b) Back view of the AFPM wheel motor, (c) Sequel FCEV, (d) Rear AFPM wheel motor with electromechanical brake, rear steering actuator and integrated controlled damper. Courtesy of General Motors Corporation, U.S.A.

Recently developed GM Sequel FCEV is powered by an induction motor located in the front compartment and two AFPM wheel motor on the rear. Figure 9.15 shows front and back view of the AFPM wheel motor and the mechanical integration detail. This powertrain arrangement demonstrated excellent vehicle performance, driveability and safety as well as potential of axial flux motor technology. GM Sequel FCEV achieved a record-breaking range of 300 mile on a single tank of hydrogen.

9.3 Ship Propulsion

9.3.1 Large AFPM Motors

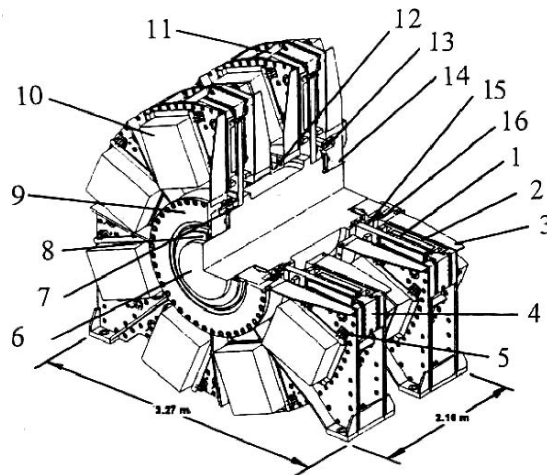


Fig. 9.16. Large power double disc AFPM brushless motor. 1 — PMs, 2 — stator assembly, 3 — housing, 4 — shock snubber, 5 — shock mount, 6 — rotor shaft, 7 — rotor disc clamp, 8 — shaft seal assembly, 9 — bearing retainer, 10 — stator segment, 11 — centre frame housing, 12 — spacer housing, 13 — rotor disc, 14 — bearing assembly, 15 — rotor seal runner, 16 — rotor seal assembly. Courtesy of Kaman Aerospace, EDC, Hudson, MA, U.S.A.

Stators of large AFPM brushless motors with disc type rotors usually have three basic parts [48]:

- aluminum cold plate;
- bolted ferromagnetic core;
- polyphase winding.

Table 9.4. Design data of large power, three phase AFPM brushless motors manufactured by *Kaman Aerospace*, EDC, Hudson, MA, U.S.A.

Quantity	PA44-5W-002	PA44-5W-001	PA57-2W-001
Number of poles, $2p$	28	28	36
Number of windings per phase		2	
Output power P_{out} , kW	336	445	746
Peak phase voltage, V	700	530	735
Rated speed, rpm	2860	5200	3600
Maximum speed, rpm	3600	6000	4000
Efficiency			
at rated speed	0.95	0.96	0.96
Torque			
at rated speed, Nm	1120	822	1980
Stall torque, Nm	1627	1288	2712
Continuous current			
(six step waveform), A	370	370	290
Maximum current, A	500	500	365
Peak EMF constant			
per phase, V/rpm	0.24	0.10	0.20
Winding resistance			
per phase at 500 Hz, Ω	0.044	0.022	0.030
Winding inductance			
per phase at 500 Hz, μH	120	60	100
Moment of inertia, kgm^2	0.9	0.9	2.065
Cooling	Water and glycol mixture		
Maximum allowable			
motor temperature, $^{\circ}\text{C}$		150	
Mass, kg	195	195	340
Power density, kW/kg	1.723	2.282	2.194
Torque density, Nm/kg	5.743	4.215	5.823
Diameter of frame, m	0.648	0.648	0.787
Length of frame, m	0.224	0.224	0.259
Application	Traction, Drilling industry		General purpose

The *cold plate* is a part of the frame and transfers heat from the stator to the heat exchange surface. The slots are machined into a laminated core wound of steel ribbon in a continuous spiral in the circumferential direction. The copper winding, frequently a Litz wire, is placed in slots and then impregnated with a potting compound. The construction of a double disc AFPM motor developed by *Kaman Aerospace*, EDC, Hudson, MA, U.S.A. is shown in Fig. 9.16 [48]. Specifications of large axial flux motors manufactured by *Kaman* are given in Table 9.4.

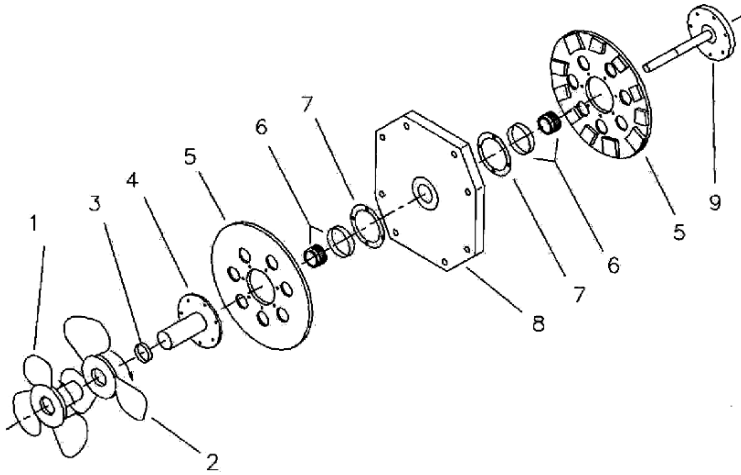


Fig. 9.17. Exploded view of the axial flux PM motor with counter-rotating rotor: 1 — main propeller, 2 — counter-rotating propeller, 3 — radial bearing, 4 — outer shaft, 5 — PM rotor, 6 — motor bearing, 7 — assembly ring, 8 — stator, 9 — inner shaft.

9.3.2 Propulsion of Unmanned Submarines

An electric *propulsion system for submarines* requires high output power, high efficiency, very salient and compact motors [64, 195]. Disc-type brushless motors can meet these requirements and run for over 100 000 h without a failure, cooled only by ambient sea water. These motors are virtually silent and operate with minimum vibration level. The output power at rated operating conditions can exceed 2.2 kW/kg and torque density 5.5 Nm/kg. The typical rotor linear speed of large marine propulsion motors is 20 to 30 m/s [195].

9.3.3 Counterrotating Rotor Marine Propulsion System

An AFPM brushless motor can be designed with the *counterrotation of two rotors* [39]. This machine topology can find applications in marine propulsion systems, which use an additional counter-rotating propeller in order to recover energy from the rotational flow of the main propeller stream. In this case the use of an AFPM motor having counter-rotating rotors allows the elimination of the motion reversal epicyclical gear.

The stator winding coils have a rectangular shape, which depends on the cross section of the toroidal core [39] (also see Fig. 2.4). Each coil has two active surfaces and each coil surface interacts with the facing PM rotor. In order to achieve the opposite motion of the two rotors, the stator winding

coils have to be arranged in such a manner that counter-rotating magnetic fields are produced in the machine’s annular air gaps. The stator is positioned between two rotors that consist of mild steel discs and axially magnetized NdFeB PMs. The magnets are mounted on the disc’s surface from the stator sides. Each rotor has its own shaft that drives a propeller, i.e. the motor has two coaxial shafts that are separated by a radial bearing. The arrangement is shown in Fig. 9.17 [39].

Table 9.5. EMALS requirements [79]

End speed, m/s	103
Maximum peak-to-mean thrust force ratio	1.05
Cycle time, s	45
End speed variation, m/s	1.5
Maximum acceleration, m/s ²	56
Launch energy, MJ	122
Maximum thrust, MN	1.3
Maximum power, MW	133
Volume, m ³	<425
Mass, kg	<225 000

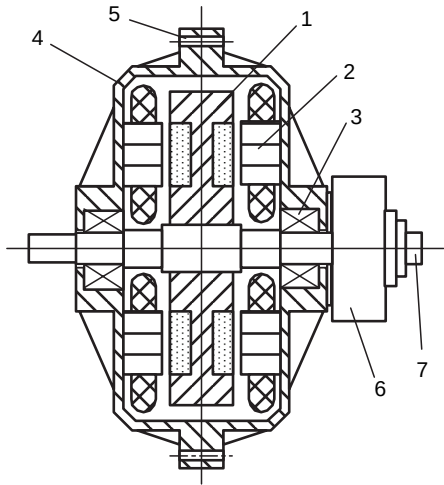


Fig. 9.18. AFPM synchronous machine for EMALS. 1 — PM rotor assembly, 2 — stator assembly, 3 — bearing, 4 — enclosure, 5 — mounting flange, 6 — brake, 7 — shaft encoder.

Table 9.6. Specifications of AFPM synchronous machine for EMALS [79]

Number of phases	6
Number of poles	$2p = 40$
Number of active stator slots	$s_1 = 240$
Type of stator	double-sided slotted
Number of stator windings	2
Generator winding	air gap side
Motor winding	bottom of slots
Maximum speed, rpm	6400
Maximum frequency, Hz	2133
Output power at maximum speed, MW	81.6
Maximum line-to-line EMF for motoring, V	1122
Maximum line-to-line peak voltage, V	1700
Maximum phase peak current, A	6400
Power losses, kW	127
Efficiency	0.893
Resistance per phase, m Ω	8.6
Inductance per phase, μ H	10.4
Stator cooling	cold plate
Coolant	water and glycol mixture
Flow rate, l/min	151
Average temperature of winding, $^{\circ}$ C	84
Average temperature of stator core, $^{\circ}$ C	61
Type of PMs	sintered NdFeB
Remanence at 40° C, T	1.05
Air gap magnetic flux density, T	0.976
Tooth magnetic flux density, T	1.7
Maximum energy storage capacity, MJ	121
Energy density, kJ/kg	18.1
Mass, kg	6685

9.4 Electromagnetic Aircraft Launch System

Military aircrafts are catapulted from the decks of aircraft carriers with the aid of steam catapults. A steam catapult uses two parallel rows of slotted cylinders with pistons connected to the shuttle that tows the aircraft. The pistons under the steam pressure accelerate the shuttle until the aircraft will takeoff. Steam catapults have many drawbacks [79], i.e.:

- operation without feedback control;
- large volume (over 1100 m³) and mass (up to 500 t);
- occupation of the "prime" real estate on the carrier and negative effect on the stability of the ship;
- low efficiency (4 to 6%);
- operational energy limit, approximately 95 MJ;
- need for frequent maintenance.

The *electromagnetic aircraft launch system* (EMALS) is free of the drawbacks listed above. The EMALS technology employs linear induction or synchronous motors fed from disc-type alternators through cycloconverters. The electrical energy obtained from the carrier power plant is stored kinetically in the rotors of AFPM synchronous generators. The energy is then released as a few-second pulses to accelerate and launch an aircraft. The cycloconverter between the AFPM generator and linear motor raises the voltage and frequency. The EMALS operates in a “real time” closed loop control [79]. The requirements are given in Table 9.5.

Specifications of the AFPM synchronous machine shown in Fig. 9.18 are given in Table 9.6 [79]. The EMALS uses four AFPM machines mounted in a torque frame and grouped in counter-rotating pairs to reduce the torque and gyroscopic effect. The rotor of the AFPM machine serves both as the kinetic energy storage when operating as a motor and a field excitation system when operating as a generator. The electric power from the on-board generators is fed to the AFPM machine via rectifier–inverters. There are two separate stator windings for motoring and generating mode. The motor winding is located at the bottom of slots to provide better thermal conductivity between the winding and enclosure.

Table 9.7. Specifications of ECI drilling systems with AFPM synchronous motors manufactured by *Tesco Corporation*, Calgary, Alberta, Canada

ECE drilling system	670	1007
Drilling unit without swivel		
Mass, kg	5,909	6,270
Operating length, m		
(including 2.8-m links and elevators)	4.34	4.34
Make-up Torque (typical), Nm	5,965	9,490
Breakout torque, Nm	7,592	11,389
Max. Drill Torque, Nm	5,098	7,860
Max. Speed, rpm	187	187
Power system (mechanical module)		
Approximate mass, kg	3,340	3,730
Length, m	6.4	6.4
Width, m	2.3	2.3
Power system (power module)		
Approximate mass, kg	3,545	4,320
Length, m	6.35 mm	6.30
Width, m	2.311	2.30

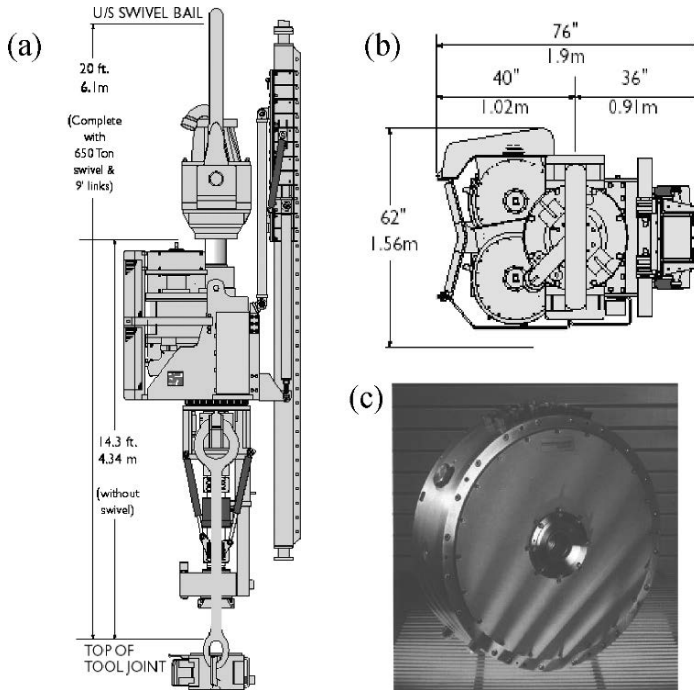


Fig. 9.19. ECI drilling system: (a) general view, (b) power unit with AFPM synchronous motors, (c) PA-44 Kaman Aerospace AFPM motor. Courtesy of *Tesco Corporation*, Calgary, Alberta, Canada.

9.5 Mobile Drill Rigs

Drill rigs vary from small electric motor driven tools to large oil field rigs. The basic elements of an aboveground drill rig are the power unit or motor, pump or air compressor for circulating drilling fluid to the bit and cleaning the borehole, drill head, hoisting drums and cables, derrick, mounting platform or deck and assorted equipment such as hammers for driving and removing casing, portable mud pit, racks for stacking the drill rods and samples, small tools for coupling or uncoupling and hoisting the drill string, etc.

The power unit (motor) performs the following functions:

- operates a drive weight mechanism for percussion or churn drilling or provides rotary motion to turn augers and coring equipment for rotary drilling operations;
- operates a winch for raising and lowering the drilling and sampling equipment;
- provides downward pressure for pushing boring and sampling equipment, or lifting and dropping a hammer to drive casing or sampling equipment.

For most drilling and sampling operations, the power source is the power take-off from the truck motor on which the drilling machine is mounted or from a separate engine, which is assigned or attached as an integral component of the drilling rig. It is estimated that approximately 90% of the motors are gasoline or diesel engines and 10% are compressed air or electric motors. A drive train that consists of gears or hydraulic pumps is used to convert the power supply to speed and torque for hoisting and rotating the drilling equipment. Most units have a transmission that allows 4 to 8 speeds for hoisting and drilling. In general, the hoisting capacity of the drill rig governs the depth of the borehole. A rule of thumb for selecting the power source is that the power required to hoist the drill rods should be about three times the power that is required to turn the drill string. For high elevations, the power loss is about 3% for each 300 m.

Large power AFPM brushless motors (Table 9.4) are especially suitable for portable drilling equipment because of their compact design, light weight, precise speed control, high efficiency and high reliability.

In the ECI lightweight *top drive drilling system* (Fig. 9.19) manufactured by *Tesco Corporation*, Calgary, Alberta, Canada, the traditional induction or brush type d.c. motors have been replaced with high performance liquid cooled AFPM synchronous motors. This system is compatible with most 600 V a.c. rig power systems. The product highlights include:

- *Kaman* PA44 AFPM synchronous motors that (Table 9.4) have been successfully tested to 60g triaxial loading, making them the ideal choice for rough drilling environments;
- AFPM synchronous motors that allow a high level of precision in speed and torque control not available with other motors;
- a modular design that allows drilling to continue at reduced power with one motor;
- direct connection to the rigs a.c. bus and interface with the existing power supply with minimal degradation to the rig power;
- the entire ECI system can be transported in three standard 6-m sea containers.

Specifications of the ECI drilling system are given in Table 9.7.

9.6 Oil Beam Pumps

An oil beam pump (Fig. 9.20) consists of an overground electromechanical drive and underground reciprocating piston pump. An electric motor drives one end of a heavy beam via pair of cranks with counterweights and transmission. The other end of the beam is equipped with the so called horse head with attached vertical cable. The cable is connected to a polished rod of a pump at the bottom of the well. The cranks move the beam up and down and

actuates the pump. Typically, a beam pump delivers 5 to 40 liters of a crude oil-water mixture per each stroke.

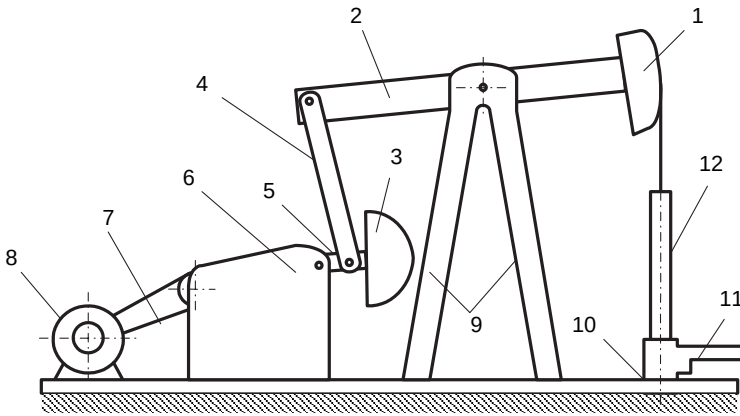


Fig. 9.20. Oil beam pump: 1 - horse head, 2 - walking beam, 3 - counterweights, 4 - Pitmans arm, 5 - cranks, 6 - transmission, 7 - sheaves and belts, 8 - electric motor or other prime mover, 9 - Samson beams, 10 - casing, 11- flow line, 12 - polishing rod.

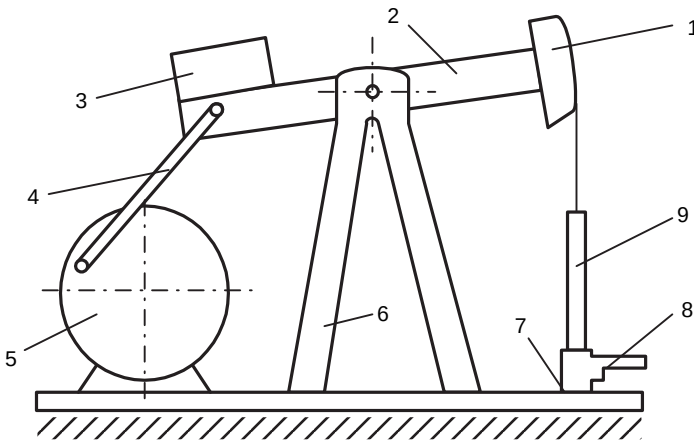


Fig. 9.21. Novel oil beam pump with AFPM brushless motor: 1 - horse head, 2 - walking beam, 3 - counterweights, 4 - Pitmans arm, 5 - disc type motor 6 - Samson beams, 7 - casing, 8- flow line, 9 - polishing rod.

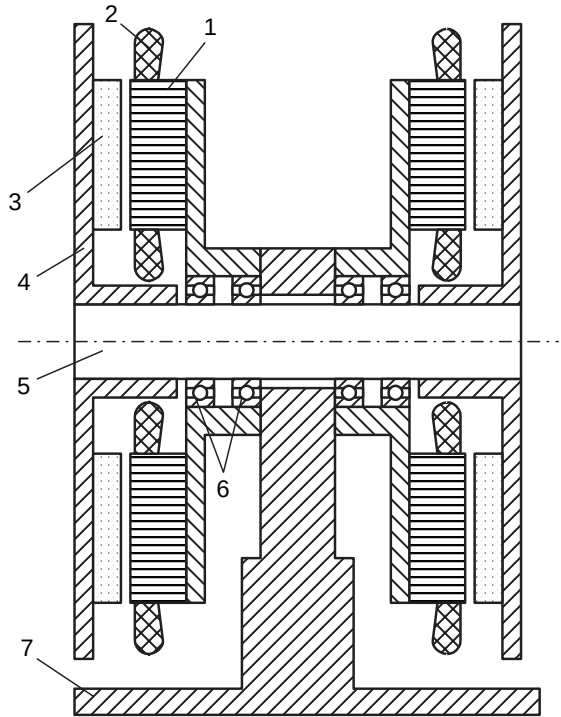


Fig. 9.22. AFPM brushless motor for oil beam pumps: 1 - stator core, 2 - stator winding, 3 - PM, 4 - rotor, 5 - shaft, 6 - bearing, 7 - base.

Table 9.8. Specifications of a prototype of AFPM brushless motor for beam pumps [130]

Rated power, kW	9.0
Rated torque, Nm	4820
Rated speed, rpm	20
Rated frequency, Hz	8
Rated voltage, V	380
Rated current, A	16
Number of rotor poles	48
Number of stator slots	144
Remanent magnetic flux density of PM, T	1.2
Thickness of PM, mm	10.0
Air gap, mm	4.0
Stator inner diameter, m	0.6
Stator outer diameter, m	0.9

The novel beam oil pump (Fig. 9.21) uses a low speed AFPM brushless motor, which replaces an induction motor, sheaves, belts and transmission [130]. There is a counterweight on the motor side of the beam that has an approximate weight as the horse head. Applications of an AFPM brushless motors simplifies the construction and maintenance of a beam pump, reduces its cost and improves the overall efficiency of the machine. Longitudinal section of the prototype of AFPM brushless motor is shown in Fig. 9.22 and specifications are given in Table 9.8 [130].

9.7 Elevators

The concept of *gearless electromechanical drive for elevators* was introduced in 1992 by *Kone Corporation* in Hyvinkää, Finland [113]. With the aid of a disc type low speed compact AFPM brushless motor (Table 9.10), the penthouse machinery room can be replaced by a space-saving direct electromechanical drive. In comparison with a low speed axial flux cage induction motor of similar diameter, the AFPM brushless motor has a doubled efficiency and a three times higher power factor.

Fig. 9.23a shows the propulsion system of the elevator while Fig. 9.23b shows how the AFPM brushless motor is installed between the guide rails of the car and the hoistway wall.

Table 9.9 contains key parameters for the comparison of different hoisting technologies [113]. The disc type AFPM brushless motor is a clear winner.

Specifications of single-sided AFPM brushless motors for gearless passenger elevators are given in Table 9.10 [113]. Laminated stators have from 96 to 120 slots with three phase short pitch winding and class F insulation. For example, the MX05 motor rated at 2.8 kW, 280 V, 18.7 Hz has the stator

Table 9.9. Comparison of hoisting technologies for 630 kg elevators [113]

Quantity	Hydraulic elevator	Warm gear elevator	Direct PM brushless motor elevator
Elevator speed, m/s	0.63	1.0	1.0
Motor shaft power, kW	11.0	5.5	3.7
Speed of motor, rpm	1500	1500	95
Motor fuses, A	50	35	16
Annual energy consumption, kWh	7200	6000	3000
Hoisting efficiency	0.3	0.4	0.6
Oil requirements, l	200	3.5	0
Mass, kg	350	430	170
Noise level, dB(A)	60..65	70...75	50...55

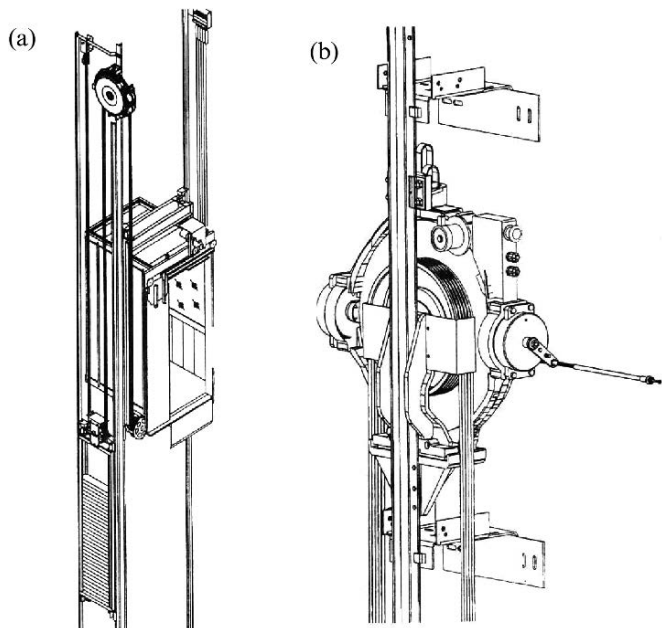


Fig. 9.23. MonoSpaceTM elevator: (a) elevator propulsion system; (b) EcodiskTM motor. Courtesy of *Kone*, Hyvinkää, Finland.

Table 9.10. Specifications of single-sided AFPM brushless motors for gearless elevators manufactured by *Kone*, Hyvinkää, Finland

Quantity	MX05	MX06	MX10	MX18
Rated output power, kW	2.8	3.7	6.7	46.0
Rated torque, Nm	240	360	800	1800
Rated speed, rpm	113	96	80	235
Rated current, A	7.7	10	18	138
Efficiency	0.83	0.85	0.86	0.92
Power factor	0.9	0.9	0.91	0.92
Cooling	natural	natural	natural	forced
Diameter of sheave, m	0.34	0.40	0.48	0.65
Elevator load, kg	480	630	1000	1800
Elevator speed, m/s	1	1	1	4
Location	hoistway	hoistway	hoistway	machine room

winding resistance $R_1 = 3.5 \, \Omega$, stator winding reactance $X_1 = 10 \, \Omega$, $2p = 20$, sheave diameter 340 mm and weighs 180 kg.

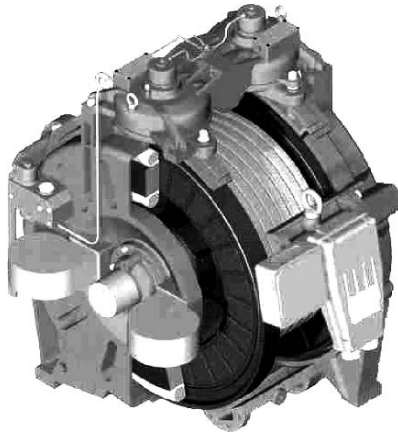


Fig. 9.24. Double-disc AFPM brushless motor for gearless elevators. Courtesy of *Kone*, Hyvinkää, Finland.

Table 9.11. Specifications of double disc AFPM brushless motors manufactured by *Kone*, Hyvinkää, Finland

Quantity	MX32	MX40	MX100
Rated output power, kW	58	92	315
Rated torque, Nm	3600	5700	14,000
Rated speed, rpm	153	153	214
Rated current, A	122	262	1060
Efficiency	0.92	0.93	0.95
Power factor	0.93	0.93	0.96
Elevator load, kg	1600	2000	4500
Elevator speed, m/s	6	8	13.5

A double disc AFPM brushless motor for gearless elevators is shown in Fig. 9.24 [113]. Table 9.11 lists specifications data of double disc AFPM brushless motors rated from 58 to 315 kW [113]. The double disc configuration offers high output torque and compensation of axial forces [P105].



Fig. 9.25. Penny-motor. Photo courtesy of *Mymotors & Actuators GmbH*, Wendelsheim, Germany.

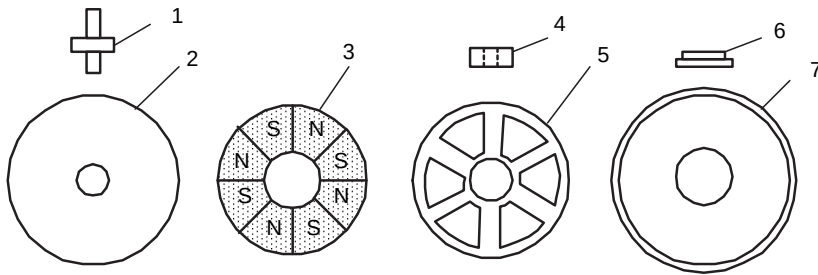


Fig. 9.26. Construction of penny-motor: 1 — shaft, 2 — soft steel yoke cover, 3 — PM ring, 4 — ball bearing, 5 — stator winding, 6 — flange, 7 — bottom steel yoke.

9.8 Miniature AFPM Brushless Motors

Ultra-flat PM micromotor, the so called *penny-motor* is shown in Figs 9.25 and 9.26 [154]. The thickness is 1.4 to 3.0 mm, outer diameter about 12 mm, torque constant up to $0.4 \mu\text{Nm}/\text{mA}$ and speed up to 60,000 rpm. A $400\text{-}\mu\text{m}$ eight pole PM and a three-strand, $110\text{-}\mu\text{m}$ disc shaped, lithographically produced stator winding have been used [154]. Plastic bound NdFeB magnets are a cost effective solution. However, the maximum torque is achieved with sintered NdFeB magnets. A miniature ball bearing has a diameter of 3 mm. Penny-motors find applications in miniaturized hard disc drives, cellular phones as vibration motors, mobile scanners and consumer electronics.

Moving Magnet Technologies (MMT) two or three phase, *miniature AFPM brushless motors* (Fig. 9.27) are well adapted to high volume low cost production. In order to obtain rigidity the stator coils are overmoulded. The stator is mounted on a single-sided printed circuit board. Each part can be manufactured using standard moulding, stamping techniques and automatic coil winding with a specific design for simple and efficient assembly. Three position

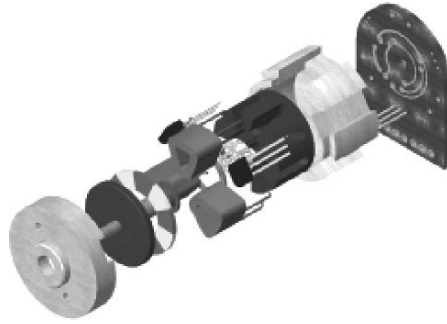


Fig. 9.27. Two-phase, four-coil, single-sided 15-mm AFPM brushless motor. Photo courtesy of *Moving Magnet Technologies, SA*, Besancon, France.

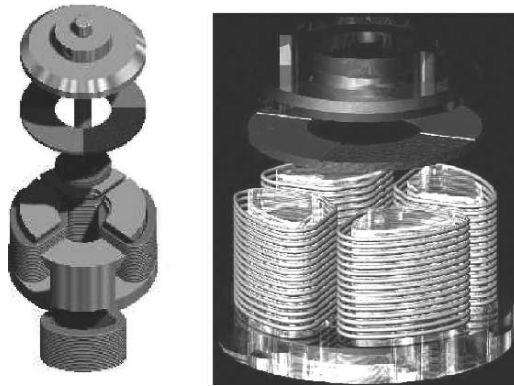


Fig. 9.28. Miniature two-phase, four-coil, 4-pole, single-sided AFPM rotary actuator. Photo courtesy of *Moving Magnet Technologies, SA*, Besancon, France.

sensing methods for the closed-loop control are used: (a) digital Hall probes located in the stator, (b) EMF signal (sensorless mode) and (c) absolute analogue position sensor (position sensor mode).

MMT AFPM miniature rotary actuators (Fig. 9.28) have been designed for automotive applications to provide an efficient, contactless, direct drive rotary motion on a limited stroke. Ring, disc and tile shaped PMs have been used. The main features of this family of actuators are:

- contactless actuation principle;
- constant torque independent of the angular position for a given current;
- linear torque–current characteristic;
- two directions of rotation;
- high torque density.

Additional functions, such as a magnetic return spring or analogue contactless position sensing can also be implemented. Owing to the linearity of the torque-current characteristic and its independence of position, it is possible to operate actuators in an open-loop against a spring or in a simple closed-loop mode by using a position sensor.

9.9 Vibration Motors

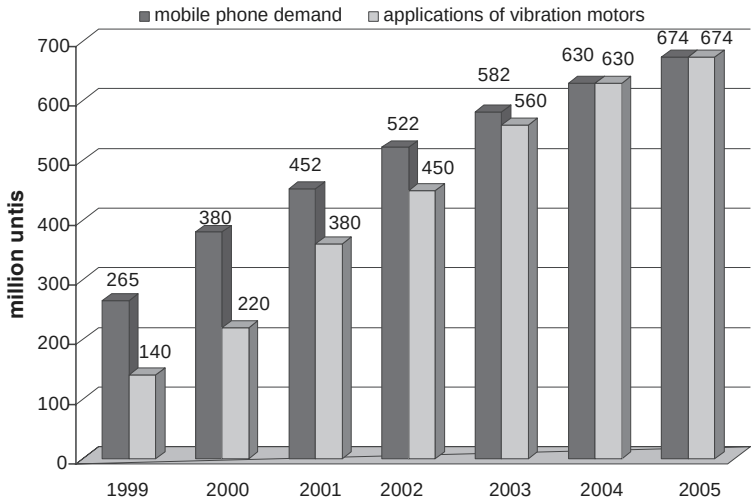


Fig. 9.29. Worldwide demand estimate for mobile phones and vibration motors.

Advances in cellular telecommunication have made *mobile phones* to be highly popular communication tool in modern society. Since a mobile phone is now a gadget as necessary as a watch or wallet, small vibration motors with diameters from 6 to 14 mm are manufactured in unbelievable large quantities (Fig. 9.29). The trends in *vibration motors* for mobile phones include reduction of mass and size, minimization of energy consumption and guaranteed stable vibration alarming in any circumstances [62].

There are two types of brushless vibration motors for mobile phones: *cylindrical* or RFPM motor and *coin type* or AFPM motor (Fig. 9.30). The *unbalanced exciting force* generated by an eccentric rotor is

$$F = m\epsilon\Omega = 2\pi m\epsilon n^2 \tag{9.7}$$

where m , ϵ and n denote the rotor mass, eccentricity and rotational speed respectively. The rotational speed is the most effective design parameter to increase the unbalanced exciting force [62].

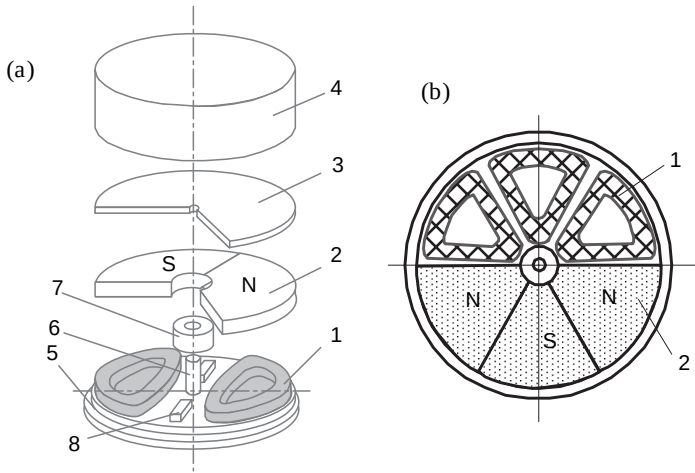


Fig. 9.30. Coin type AFPM vibration motor for mobile phones [62]: (a) two coil motor; (b) multicoil motor. 1 — phase coil, 2 — PM (mechanically unbalanced system), 3 — ferromagnetic yoke, 4 — cover, 5 — base plate, 6 — shaft, 7 — bearing, 8 — detent iron [62].



Fig. 9.31. Cost effective AFPM brushless vibration motor. Photo courtesy of *Moving Magnet Technologies, SA*, Besancon, France.

Moving Magnet Technologies single-phase AFPM brushless vibration motors provide a strong vibration feeling and silent alert with a very simple design (Fig. 9.31). Fabrication of *MMT* vibration motors is cost effective due to contactless design, scalable size and slim shape.

9.10 Computer Hard Disc Drives

The data storage capacity of a *computer hard disc drive* (HDD) is determined by the aerial recording density and number of discs. It is expected that the aerial density will soon increase from $6 \text{ Gb/cm}^2 = 38.7 \text{ Gb/in}^2$ to $15.5 \text{ Gb/cm}^2 = 100 \text{ Gb/in}^2$. The mass of the rotor, moment of inertia and vibration increase with the number of discs.

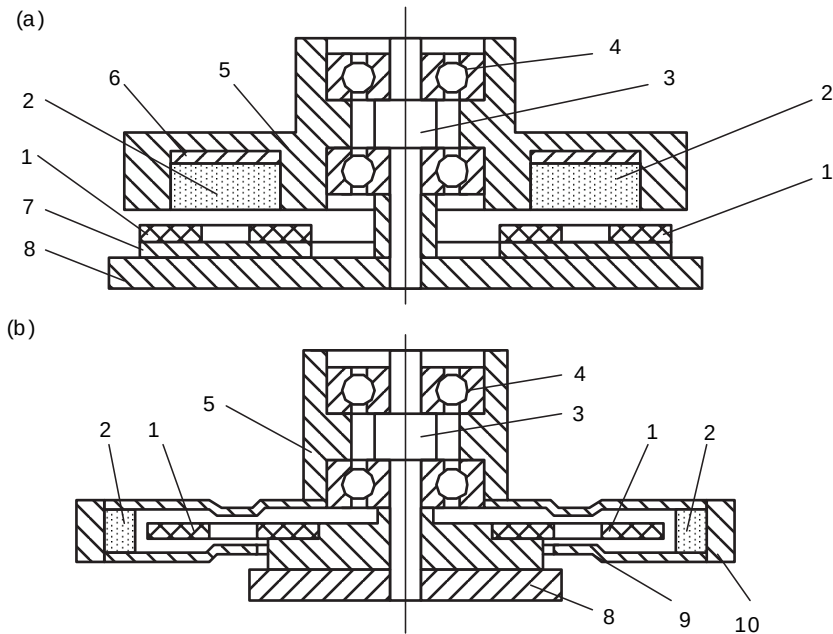


Fig. 9.32. Construction of computer HDDs with AFPM brushless motors: (a) single-sided motor; (b) double-sided motor. 1 — stator coil, 2 — PM, 3 — shaft, 4 — bearing, 5 — hub, 6 — rotor ferromagnetic yoke, 7 — stator ferromagnetic yoke, 8 — base plate, 9 — pole fingers, 10 — nonmagnetic ring.

Special design features of computer HDD motors are their high starting torque, limited current supply, low vibration and noise, physical constraints on volume and shape, protection against contamination and scaling problems. A high starting torque, 10 to 20 times the running torque is required, since

the read/write head tends to stick to the disc when not moving. The starting current is limited by the computer power supply unit, which in turns, can severely limit the starting torque.

AFPM brushless motors (Fig. 9.32) can develop higher starting torque than their radial field counterparts. The additional advantage is zero cogging torque. The drawback of the single-sided AFPM HDD motor shown in Fig. 9.32a is the high normal attractive force between the stator with ferromagnetic yoke and rotor mounted PMs. In double-sided HDD AFPM motors the stator does not have any ferromagnetic core and no normal attractive force is produced at zero-current state. The stator has a three-phase winding with coreless coils fabricated with the aid of a lithographic method. In a typical design, there are two coils per phase for $2p = 8$ PM poles and three coils per phase for $2p = 12$ PM poles. To reduce the air gap and increase the magnetic flux density, the so called *pole fingers* are created in the lower part of the hub and the rotor magnetic circuit is bended by pressing it towards the stator coil centres (Fig. 9.32b). According to [140], for a HDD AFPM brushless motor with a 51-mm stator outer diameter, $N_1 = 180$ turns per phase and torque constant $k_T = 6.59 \times 10^{-3}$ Nm/A the current consumption is 0.09 A at 13 900 rpm and no load condition.

The acoustic noise of HDD brushless motors with ball bearings is usually below 30 dB(A) and projected mean time between failures MTBF = 100 000 hours. HDD spindle motors are now changing from ball bearing to fluid dynamic bearing (FDB) motors. Contact-free FDBs produce less noise and are serviceable for an extended period of time.

9.11 Ventricular Assist Devices

When electricity was new, people had high hopes that it had curative powers. For example, electropathy (electrodes between patient's hands and ailing body part), very popular from 1850 to 1900, promised to cure most diseases and conditions, including mental illness. The 21st century biomedical engineering community has resurrected magnetic fields to treat depression, e.g. magnetic seizure therapy (high frequency, powerful electromagnets) or transcranial magnetic stimulation (strong pulse magnetic fields).

At present, many medical devices use small permanent magnet (PM) electric motors as, for example, high-quality pumps, centrifuges, infusion pumps, hemodialysis machines, precision handpieces and implantable devices (ventricular assist devices, pacemakers, defibrillators, nerve stimulators, etc). Since the reliability of medical products is critical, the electric motor is considered a precision component rather than a commodity device [203]. This section focuses on very small PM brushless motors for implantable devices, in particular, motors for rotary blood pumps [102].

A left ventricular assist device (LVAD) is an electromechanical pump implanted inside the body and intended to assist a weak heart that cannot

efficiently pump blood on its own. It is used by end-stage heart failure patients who are unable to receive a heart transplant due to donor availability, eligibility, or other factors.

Motor driven pumps implanted in the human body must be free of shaft seals. This problem can be solved by embedding PMs in the pump rotor placed in a special enclosure and driven directly by the stator magnetic field. In this case the nonmagnetic air gap is large and high energy PMs are required.

Electromagnetic pumps for LVADs can be classified into three categories:

- 1st generation (1G), i.e. electromagnetic pulse pumps;
- 2nd generation (2G), i.e. electromagnetic rotary pumps;
- 3rd generation (3G), i.e. electromagnetic pumps with magnetic or hydrodynamic bearings.

Electromagnetic 1G pumps were driven by electromagnets, linear oscillating motors or linear short-stroke actuators. The pump, integrated with a linear actuator, is heavy, large and noisy.

DuraHeart^R 3G LVAD developed by *Terumo Corporation*, Tokyo, Japan, combines centrifugal pump with magnetic levitation technologies (Figs 9.33 and 9.34) [102]. Magnetic levitation allows the impeller to be suspended within the blood chamber by electromagnets and position sensors. The three-phase, 8-pole, axial flux PM brushless motor with slotless stator resembles a floppy disc drive spindle motor (Fig. 9.33). NdFeB PMs are integrated with the impeller. The output power of the motor is 4.5 W, speed 2000 rpm and torque 0.0215 Nm [226].

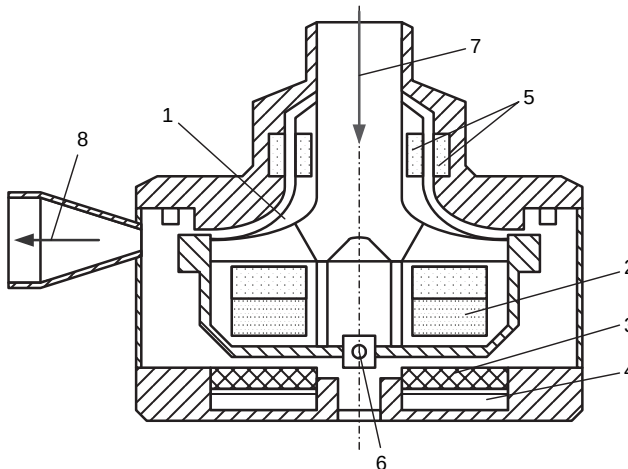


Fig. 9.33. *DuraHeart^R* centrifugal rotary pump with axial flux PM brushless motor and magnetic levitation bearings: 1 - impeller, 2 - PM, 3 - stator winding, 4 - stator core, 5 - magnetic suspension, 6 - ceramic pivot, 7 - inlet, 8 - outlet.



Fig. 9.34. Exploded view of *DuraHeart^R* centrifugal rotary pump. Courtesy of *Terumo Corporation*, Tokyo, Japan.

An axial flux slotless motor integrated with a centrifugal blood pump is shown in Figs. 9.35 and 9.36. The so-called *VentrAssistTM* manufactured now by Australian company *Ventricor* is a new cardiac LVAD, which has only one moving part - a hydrodynamically suspended impeller integrated with a PM rotor. The hydrodynamic forces act on tapered edges of the four blades. The stator of the brushless electric motor is of slotless type and has only upper and lower coils. The three coil winding and four pole rotor use the second harmonic of the magnetic field wave to produce the torque. To provide redundancy, body coils and cover coils are connected in parallel so that the motor still can run even if one coil is damaged.

The housing and impeller shell are made of titanium alloy Ti-6Al-4V. Vacodym 510 HR NdFeB PMs are embedded into impeller. To reduce the reluctance for the magnetic flux, laminated silicon steel return paths (yokes) are designed.

The 2D FEM simulation of the magnetic field distribution is shown in Figs 9.37 and 9.38 [102]. The measured performance characteristics are shown in Figs 9.39a and 9.39b [226]. For output power between 3 and 7 W and speed between 2000 and 2500 rpm, the efficiency is from 45 to 48% (Fig. 9.39a). At 3 W and 2250 rpm the winding losses are 1.7 W, eddy current losses in titanium

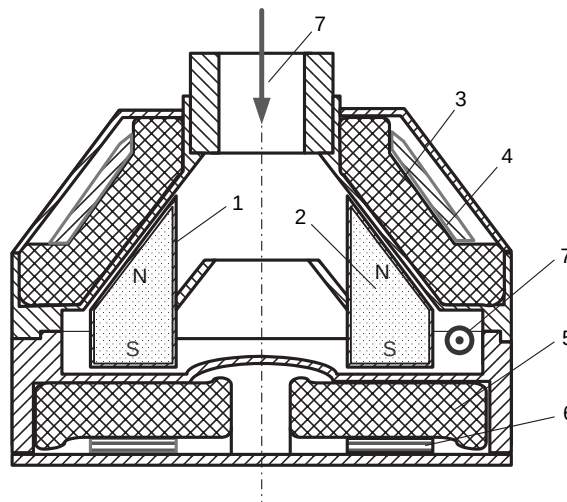


Fig. 9.35. Longitudinal section of *VentrAssistTM* hydronamically levitated centrifugal blood pump: 1 - impeller, 2 - PM, 3 - body coil, 4 - body yoke, 5 - cover coil, 6 - cover yoke, 7 - inlet, 8 - outlet.



Fig. 9.36. Computer generated 3D image of *VentrAssistTM* centrifugal blood pump. Photo Courtesy of *Ventracor Global Company*, Australia, www.ventracor.com.

1.0 W and losses in laminated yokes are 0.7 W [226]. At load torque 0.03 Nm the fundamental phase current is 0.72 A (Fig. 9.39b).

The device weighs 298 g and measures 60 mm in diameter, making it suitable for both children and adults.

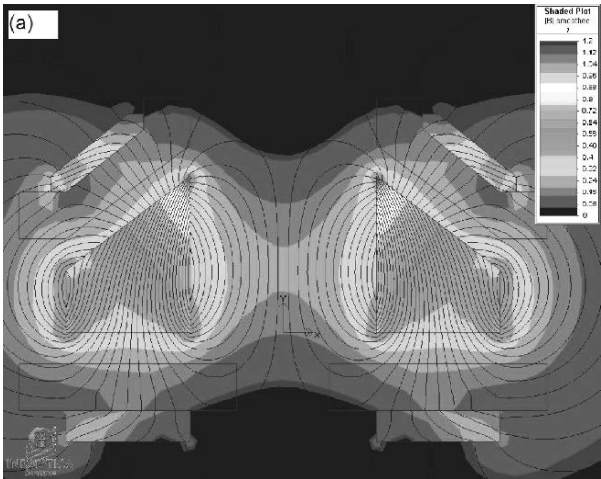


Fig. 9.37. Magnetic flux (contour plot) and magnetic flux density (shaded plot) excited by NdFeB PMs at no load [102].

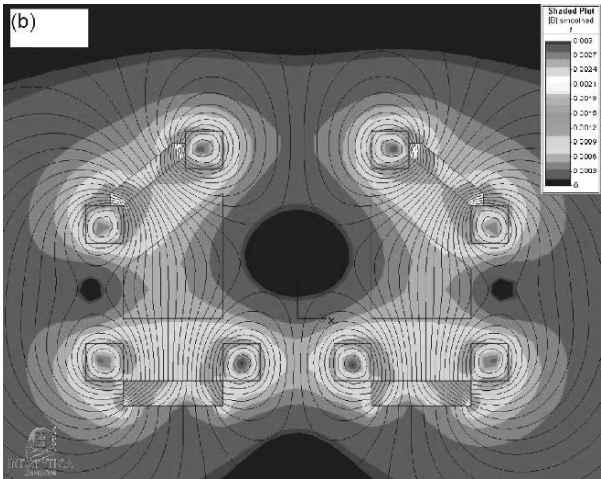


Fig. 9.38. Magnetic flux (contour plot) and magnetic flux density (shaded plot) excited by armature (stator) coils [102].

9.12 Axial Flux Machines with Superconducting Field Excitation System

Superconducting excitation system can provide stronger magnetic field in the air gap than the best rare-earth PMs. A superconducting axial flux machines is at least 30% smaller in volume than an equivalent PM machine.

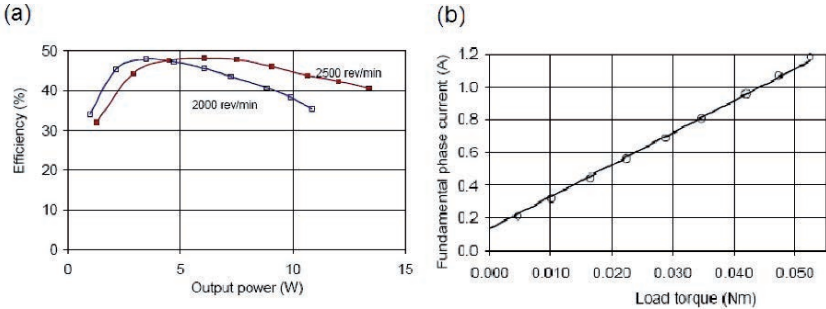


Fig. 9.39. Steady state performance characteristics of *VentrAssist™* BA2-4 pump driven by a six-step sensorless inverter: (a) efficiency; (b) fundamental phase current calculated on the basis of EMF (solid line) and obtained from laboratory tests (circles) [226].

A Japanese industry-academia group, called *frontier group*, which includes *Sumitomo Electric Industries, Ltd.*² (SEI), the manufacturer of HTS wires, has developed a 12.5 kW synchronous motor with HTS BiSrCaCuO (BSCCO) field excitation winding [249, 252]. This motor was integrated with a pod propulsor and completed in 2003.

The pod propulsor has outer diameter of 0.8 m and is 2 m long. The diameter of the propeller is 1 m. When a large conventional motor is used with a pod propulsor, the outside diameter of the propulsor becomes also large.

This HTS motor is an axial flux (disc type) three phase, eight-pole machine. The stationary field excitation system and rotating armature simplify the cooling system. Technical specifications of the HTS motor are given in Table 9.12. The expanded view of the motor is shown in Fig. 9.40. The main features include:

- machine is cooled by low-cost, easy-to-handle liquid nitrogen;
- uses the bismuth-based 1G BSCCO HTS wires that were fabricated in the lengths of several kilometers with the aid of SEI innovative manufacturing process;
- very low noise;
- very low leakage magnetic flux;
- the surface of the machine is kept at a room temperature and, therefore, can be installed in used in any place.

² *Sumitomo Electric Industries, Ltd., Fuji Electric Systems Co., Ltd., Hitachi, Ltd., Ishikawajima-Harima Heavy Industries Co., Ltd., Nakashima Propeller Co., Ltd., Taiyo Nippon Sanso Corporation, University of Fukui (H. Sugimoto)*

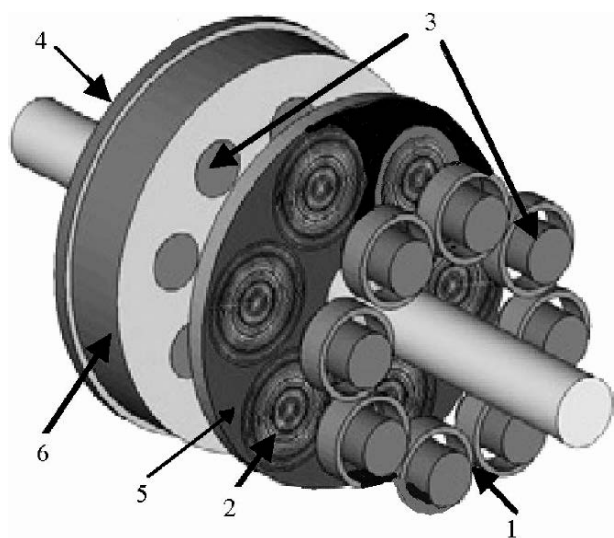


Fig. 9.40. Computer created 3D image of the axial flux 12.5 kW HTS motor for pod propulsor: 1 — HTS field excitation coils, 2 — armature excitation coil, 3 — cores, 4 — back yoke, 5 — fiberglass reinforced plate of cryostat, 6 — cryostat.

Table 9.12. Specifications of Japanese 12.5 kW HTS axial flux motor

Rated output power, kW	12.5
Rated torque, Nm	1194
Short time duty power, kW	62.5
Speed, rpm	100
Armature phase rms current, A	30
Armature current density, A/mm ²	5.0
Liquid nitrogen temperature, K	66
Diameter, m	0.65
Length, m	0.36
Number of turns of HTS field excitation coil	330
Dimensions of HTS BSC wire, mm	W = 4.3, T = 0.22
Diameters of HTS coil, mm	ID = 136, OD = 160
Filed excitation pole core diameter, mm	96
Number of turns of armature coil	850
Diameter of armature coil, mm	208
Diameter of armature copper wire, mm	2

The rotating armature is located in the center of the motor between twin stators. The coreless armature winding consists of 6 pancake coils wound with a copper wire. Coreless winding does not produce any cogging torque. The motor is furnished with slip rings and brushes to supply the heavy electric current to each of the armature phase windings, which is the main disadvantage.

The stationary field excitation system is placed in fiberglass reinforced plastic (FRP) cryostats. The FRP cryostat is lightweight, strong and does not conduct electric current (no eddy-current losses). The temperature of liquid nitrogen is 66 K in order to increase the field current. There are two field excitation systems (one on each side of the armature) with eight ferromagnetic poles and eight HTS coils each. The pole cores are concentric with respect to the HTS coils, which results in limited penetration of the magnetic flux into the coils. The pole cores protrude outside the cryostat where they are mechanically fixed to the disc-shaped ferromagnetic yokes, which create a back path for the magnetic flux. The cores are wound with ferromagnetic ribbon to reduce the eddy current losses. Since cores are outside the cryostat, the loss to cool cores is zero.

The maximum magnetic flux density in the core is only 1.8 T, so that the magnetic circuit is unsaturated. The maximum normal component of the magnetic flux density penetrating to HTS coils is 0.29T, i.e. below the value corresponding to critical current.

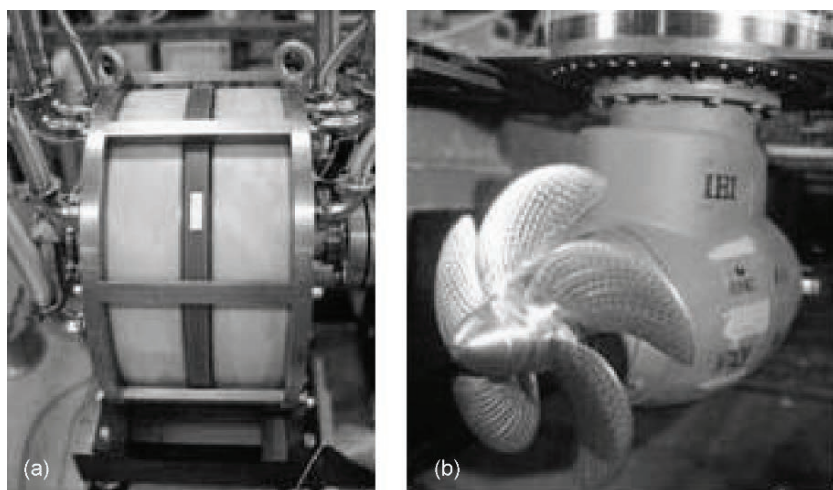


Fig. 9.41. Japanese 12.5 kW HTS motor for pod propulsor: (a) motor; (b) pod propulsor integrated with the motor.

The 12.5 kW HTS motor and pod propulsor are shown in Fig. 9.41. Performance characteristics obtained from laboratory tests are shown in Fig. 9.42. The maximum efficiency is 97.7 % and it remains practically constant for the load from 60 to 210 Nm. However, the cooling loss is not taken into account. The motor was loaded far below the rated torque, which is 1194 Nm (Table 9.12). The maximum shaft torque at laboratory tests was only 210 Nm, which corresponds to 2.2 kW shaft power at 100 rpm. The frontier group justifies it by the limit on the load equipment.

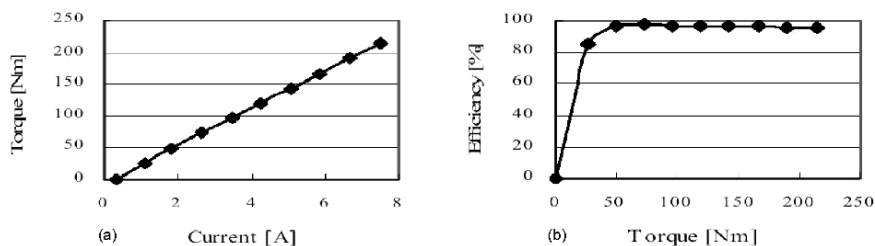


Fig. 9.42. Performance characteristics of the 12.5 kW HTS machine obtained from laboratory tests: (a) voltage versus speed (generating mode); (b) efficiency versus shaft torque (motoring mode).

The axial flux HTS motor has a small size and is characterized by low energy consumption. It is expected to find potential applications not only in pod propulsors, but also in railcar motors, wind generators and large industrial electromechanical drives, e.g. steel rolling mills.

The group has a plan to develop a commercial 400 kW, 250 rpm HTS motor. The development will include the test at a wharf the contra-rotating propeller pod propulsion system incorporating 2 units of a 400 kW full superconducting motor. The 400 kW motor is proposed for a coasting vessel. Further plans include development of larger motors 2.5 MW and 12.5 MW. Instead of BSCCO, coated YBCO wire will be the mainstream HTS wire. With YBCO wire the current density is considerably improved.

Another Japanese industry-academia group³ has developed a 15 kW, 720 rpm, coreless axial flux HTS motor, similar to that shown in Figs 9.40 and 9.41. *Kitano Seiki* is marketing this motor [153].

The axial gap superconductive motor uses GdBaCuO bulk HTS for rotor excitation system and HTS tapes for the stator (armature) winding. The rotor system is in the central position of the machine and the twin stator (armature) is on both sides of the rotor (Fig. 9.43). The rotor has 8 poles 26-mm diameter each. The armature winding has 6 poles. The rotor assembly is 0.3 m long and is 0.5 m in diameter. The magnetic flux density in the air gap is 3 T at 77K under field cool magnetization. The bulk HTS poles are magnetized inside the machine using a pulsed-copper coils. The armature system is stationary so that there are no slip rings, only armature leads. The rotor field excitation system with bulk HTS does not require any leads. No electric power supply is needed to maintain the magnetic field. The machine can operate both as a motor and generator. The prototype has been designed for marine pod propulsors. Potential application to medical, biomedical, and environmental protection are under consideration.

³ *Fuji Electric Systems, Fukui University, Hitachi, Ishikawa-hajima Heavy Industry (IHI), Kitano Seiki, Nakashima Propeller Co., Ltd., Niigata Motors, SEI, Taiyou Nissan and Tokyo University of Marine Science and Technology.*

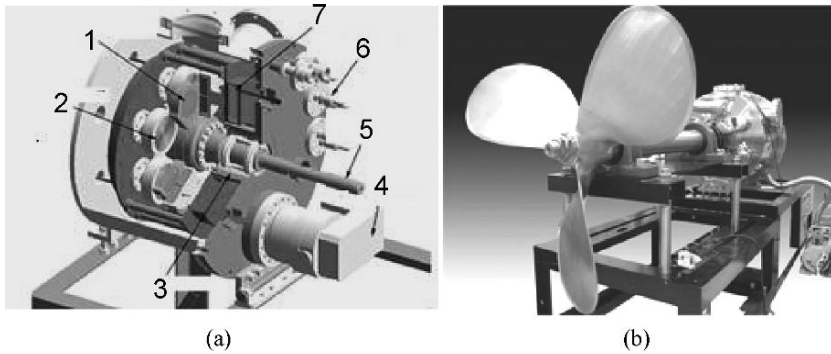


Fig. 9.43. Axial flux brushless machine with bulk HTS rotor developed by *Kitano Seiki*, Japan: (a) computer 3D image; (b) motor and propeller on test bed. 1 — HTS bulk magnet, 2 — armature coil, 3 — magnetic seal unit, 4 — vacuum pump, 5 — liquid nitrogen inlet, 6 — armature winding terminal, 7 — coil cooling layer.

Numerical Example 9.1

Find the kinetic energy and tensile stress of the ring with dimensions $R_{out} = 0.142$ m, $R_{in} = 0.082$ m, thickness $d = 0.022$ m and density $\rho = 7800$ kg/m³ rotating at the speed of 30 000 rpm.

Solution

The mass of the ring is

$$m = \pi \rho d (R_{out}^2 - R_{in}^2) = \pi 7800 \times 0.022 (0.142^2 - 0.082^2) = 7.25 \text{ kg}$$

The moment of inertia according to eqn (9.2) is

$$J = \frac{1}{2} 7.25 (0.142^2 + 0.082^2) = 0.097 \text{ kgm}^2$$

The rotational speed $n = 30\,000/60 = 500$ rev/s and angular speed $\Omega = 2\pi 500 = 3141.6$ rad/s. Thus the kinetic energy according to eqn (9.4) is

$$E_k = \pi^3 7800 \times 0.022 (0.142^4 - 0.082^4) \times 500^2 = 480.74 \text{ kJ}$$

The tensile stress according to eqn (9.2) is

$$\sigma = \frac{7800 \times 3141.6^2}{3} (0.082^2 + 0.082 \times 0.142 + 0.142^2) = 988.6 \text{ MPa}$$

The energy density according to eqn (9.6) is

$$e_k = \frac{480\,740}{7.25} = 66.3 \text{ kJ/kg}$$

The shape factor of the ring [3] is

$$k_{sh} = e_k \frac{\rho}{\sigma} = 66\,300 \frac{7800}{988.6 \times 10^6} = 0.523$$

Numerical Example 9.2

Find the main dimensions of a three-phase, 1200 W, 4-pole, 200 000 rpm AFPM synchronous generator for a mobile battery charger. The total nonmagnetic air gap including magnet retaining ring (Fig. 9.2) should be $g = 2$ mm, number of slots $s_1 = 12$, air gap magnetic flux density $B_g = 0.4$ T, line current density $A_m = 16\,000$ A/m, PM inner-to-outer diameter ratio $k_d = 0.5$, PM temperature 350°C . Vacumax 225 SmCo PMs (*Vaccumschmelze*, Hanau, Germany) with $B_{r20} = 1.04$ T and $H_c = 760$ kA/m at 20°C are recommended. The temperature coefficient for B_r is $\alpha_B = -0.035\, \%/^\circ\text{C}$, leakage flux coefficient $\sigma_{lM} = 1.3$, EMF-to-voltage ratio $\epsilon = 1.3$ ($E_f > V_1$), pole shoe width-to-pole pitch ratio $\alpha_i = 0.72$, efficiency $\eta = 0.8$ and power factor $\cos \phi = 0.8$.

Solution

The winding coefficients (2.8), (2.9) and (2.10) are $k_{d1} = 1$, $k_{p1} = 1$ and $k_{w1} = 1$ since $q_1 = 12/(4 \times 3) = 1$. The coefficient k_D according to eqn (2.27) is

$$k_D = \frac{1}{8}(1 + 0.5)(1 - 0.5^2) = 0.141$$

The outer diameter of PMs according to eqn (2.100) is

$$D_{out} = \sqrt[3]{\frac{1.3 \times 1200}{\pi^2 0.141 \times 1 \times (200\,000/60) \times 0.4 \times 16\,000 \times 0.8 \times 0.8}} = 0.044 \text{ m}$$

The remanent magnetic flux density at 350°C according to eqns (3.2) is

$$B_r = 1.04 \left[1 + \frac{-0.035}{100}(350 - 20) \right] = 0.92 \text{ T}$$

Since demagnetization curves for different temperatures are parallel lines, the relative recoil magnetic permeability will be approximately the same as for 20°C (linear magnetization curve), i.e.

$$\mu_{rec} = \frac{0.92}{0.4\pi \times 10^{-6} \times 760\,000} = 0.963$$

The PM thickness with the leakage coefficient σ_{lm} included, can be calculated on the basis of eqns (3.13) and (3.9), i.e.

$$h_M \approx \mu_{rec} \frac{\sigma_{lm} B_g}{B_r - \sigma_{lm} B_g} g = 1.089 \frac{1.3 \times 0.4}{0.92 - 1.3 \times 0.4} 0.002 = 0.0026 \text{ m}$$

The inner diameter of PMs is $D_{in} = k_d D_{out} = 0.5 \times 44 = 22$ mm, radial length of PM $l_M = 0.5(44.0 - 22.0) = 11.0$ mm, average diameter of PMs $D = 0.5(44.0 + 22.0) = 33$ mm, average pole pitch $\tau = \pi 33/4 = 26$ and average PM circumferential width $b_p = \alpha_i \tau = 0.72 \times 26 = 18$ mm.

Numerical Example 9.3

A 10-kW, 2200-rpm electric motor operates at almost constant speed according to the torque profile given in Fig. 9.44. The overload capacity factor $k_{ocf} = T_{max}/T_{shr} = 1.8$. Find the thermal utilisation coefficient of the motor.

Solution

The required rated shaft torque is

$$T_{shr} = \frac{P_{out}}{2\pi n} = \frac{10,000}{2\pi \times (2200/60)} = 43.4 \text{ Nm}$$

The *rms* torque based on the given duty cycle is

$$T_{rms}^2(t_1 + t_2 + \dots + t_n) = T_1^2 t_1 + T_2^2 t_2 + \dots + T_n^2 t_n \quad \text{or} \quad T_{rms}^2 \sum t_i = \sum T_i^2 t_i$$

Thus

$$\begin{aligned} T_{rms} &= \sqrt{\frac{\sum T_i^2 t_i}{\sum t_i}} \\ &= \sqrt{\frac{25^2 \times 3 + 40^2 \times 8 + 65^2 \times 5 + 38^2 \times 30 + 15^2 \times 10}{3 + 8 + 5 + 30 + 10}} \\ &= 38.1 \text{ Nm} \end{aligned}$$

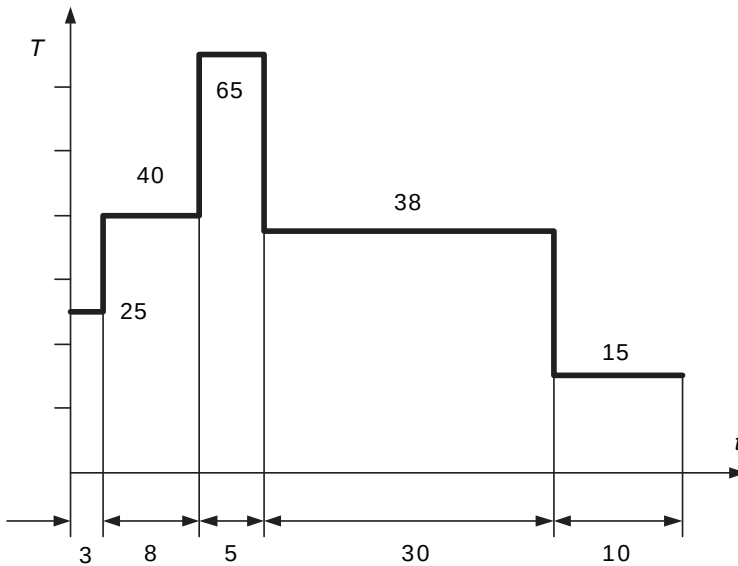


Fig. 9.44. Torque profile of the electric motor according to *Numerical example 9.3*.

The maximum torque in Fig. 9.44 cannot exceed the rated shaft torque times the overload capacity factor $k_{ocf}T_{shr} = 1.8 \times 43.4 = 78.12$ Nm. Also, the required T_{shr} should be greater than or equal to T_{rms} .

The coefficient of thermal utilisation of the motor is

$$\frac{T_{rms}}{T_{shr}} \times 100\% = \frac{38.1}{43.4} \times 100\% = 87.8\%$$

Symbols and Abbreviations

A	magnetic vector potential
<i>A</i>	line current density; cross-section area
<i>a</i>	number of parallel current paths of the stator (armature) winding
B	vector magnetic flux density
<i>B</i>	magnetic flux density; damping of the system
<i>b</i>	instantaneous value of the magnetic flux density; width of slot
<i>b_p</i>	pole shoe width
<i>C_f</i>	cost of frame
<i>C_{ins}</i>	cost of insulation
<i>C₀</i>	cost of all other components independent of the shape of the machine
<i>C_{PM}</i>	cost of PMs
<i>C_{rc}</i>	cost of the rotor core
<i>C_{sh}</i>	cost of shaft
<i>C_w</i>	cost of winding
<i>c_{Cu}</i>	cost of copper conductor per kg
<i>c_E</i>	armature constant (EMF constant)
<i>c_{Fe}</i>	cost of ferromagnetic core per kg
<i>c_{ins}</i>	cost of insulation per kg
<i>c_p</i>	specific heat at constant pressure
<i>c_{PM}</i>	cost of PMs per kg
<i>c_{steel}</i>	cost of steel per kg
<i>c_v</i>	heat capacity
<i>D</i>	diameter; duty cycle of power semiconductor switches
<i>D_{in}</i>	inner diameter of PMs equal to the inner diameter of stator bars
<i>D_{out}</i>	outer diameter of PMs equal to the outer diameter of stator bars
<i>E</i>	EMF (<i>rms</i> value); electric field intensity
<i>E_f</i>	EMF per phase induced by the rotor without armature reaction
<i>E_i</i>	internal EMF per phase
<i>e</i>	instantaneous EMF; eccentricity
<i>F</i>	force; number of winding sections
<i>F₁₂</i>	shape factor of two surfaces involved in radiation

$\mathcal{F}$	space and/or time distribution of the MMF
$\mathcal{F}_a$	armature reaction MMF
$\mathcal{F}_{exc}$	MMF of the rotor excitation system
f	frequency; friction factor
G	permeance; gap ratio g/R
g	air gap (mechanical clearance); gravitational acceleration
Gr	Grashof number
g'	equivalent air gap
$\mathbf{H}$	vector magnetic field intensity
H	magnetic field intensity
h	height; heat transfer coefficient
h_M	height of the PM
I	electric current
I_a	stator (armature) current
i	instantaneous value of current; enthalpy
$\mathbf{J}$	vector electric current density
J	moment of inertia
J_a	current density in the stator (armature) winding
K_c	current regulator gain
K_i	inverter gain
k	coefficient, general symbol; thermal conductivity
k_{1R}	skin effect coefficient for the stator conductor resistance
k_{ad}	reaction factor in d -axis
k_{aq}	reaction factor in q -axis
k_C	Carter's coefficient
k_d	inner-to-outer diameter ratio $k_d = D_{in}/D_{out}$
k_{d1}	distribution factor for fundamental
k_E	EMF constant $k_E = c_E \Phi_f$
k_f	form factor of the field excitation $k_f = B_{mg1}/B_{mg}$
k_i	stacking factor of laminations
k_{ocf}	overload capacity factor $k_{ocf} = T_{max}/T_{shr}$
k_{p1}	pitch factor for fundamental
k_{sat}	saturation factor of the magnetic circuit due to the main (linkage) magnetic flux
k_T	torque constant $k_T = c_T \Phi_f$
k_{w1}	winding factor $k_{w1} = k_{d1} k_{p1}$ for fundamental
L	inductance; length
l_{1e}	length of the one-sided end connection
L_i	armature stack effective length
l_M	axial length of PM
M	mutual inductance
M_o	momentum
m	number of phases; mass
$\dot{m}$	mass flow rate
m_a	amplitude modulation ratio

m_f	frequency modulation ratio
N	number of turns per phase; number of machines
Nu	Nusselt number
n	rotational speed in rpm; independent variables
n_0	no-load speed
P	active power
P_{elm}	electromagnetic power
P_{in}	input power
P_{out}	output power
Pr	Prandtl number
ΔP	active power losses
ΔP_{1Fe}	stator core losses
ΔP_{1w}	stator winding losses
ΔP_{2Fe}	rotor core losses
ΔP_e	eddy current losses in stator conductors
ΔP_{fr}	friction losses
ΔP_{PM}	losses in PMs
ΔP_{rot}	rotational (mechanical) losses
ΔP_{wind}	windage losses
Δp	specific core loss
p	number of pole pairs; pressure
p_r	radial force per unit area
$\wp$	wetted perimeter
Q	electric charge; reactive power; volumetric flow rate
Q_1	number of slots per pole
Q_c	number of stator coils
Q_{en}	enclosed electric charge
q_1	number of slots per pole per phase
R	radius; resistance
R_1	armature winding resistance of a.c. motors
R_{in}	inner radius of PMs equal to the inner radius of stator bars
R_{out}	outer radius of PMs equal to the outer radius of stator bars
Re	Reynolds number
$\mathfrak{R}_{\mu g}$	air gap reluctance
$\mathfrak{R}_{\mu la}$	external armature leakage reluctance
$\mathfrak{R}_{\mu M}$	permanent magnet reluctance
S	apparent power; surface
S_M	cross section area of PM; $S_M = w_M L_M$ or $S_M = b_p L_M$
s	cross section area of stator conductor
s_1	number of stator slots equal to the number of stator teeth
T	torque
T_d	electromagnetic torque developed by the machine
T_{drel}	reluctance torque
T_{dsyn}	synchronous or synchronizing torque
T_m	mechanical time constant

T_{sh}	shaft torque (output or load torque)
t	time; slot pitch
U	internal energy
u	tangential velocity
V	electric voltage; volume
v	instantaneous value of electric voltage; linear velocity
W	energy produced in outer space of PM; rate of change of the air gap energy
W_m	stored magnetic energy
w	energy per volume, J/m ³ ; radial velocity
w_M	width of PM
X	reactance
X_1	stator winding leakage reactance
X_{ad}	d -axis armature reaction (mutual) reactance
X_{aq}	q -axis armature reaction (mutual) reactance
X_{sd}	d -axis synchronous reactance; $X_{sd} = X_1 + X_{ad}$
X_{sq}	q -axis synchronous reactance; $X_{sq} = X_1 + X_{aq}$
$\mathbf{Z}$	impedance $\mathbf{Z} = R + jX$; $ \mathbf{Z} = Z = \sqrt{R^2 + X^2}$
z	number of coils in a phase group
α	complex attenuation constant of electromagnetic field
α_i	effective pole arc coefficient $\alpha_i = b_p/\tau$
γ	form factor of demagnetization curve of PM material
δ	power (load) angle
δ_i	inner torque angle
ϵ	eccentricity
ε	emissivity; surface spectral property
η	efficiency
γ	equivalent sand grain roughness
θ	rotor angular position for brushless motors
θ_m	slot pitch angle
θ_{re}	electrical angle corresponding to the coil layer width
ϑ	temperature; angle between $\mathbf{I}_a$ and $\mathbf{I}_{ad}$
λ	coefficient of leakage permeance (specific leakage permeance)
λ_T	turbulent parameter
μ	dynamic viscosity
μ_o	magnetic permeability of free space $\mu_o = 0.4\pi \times 10^{-6}$ H/m
μ_r	relative magnetic permeability
μ_{rec}	recoil magnetic permeability
μ_{rrec}	relative recoil permeability $\mu_{rrec} = \mu_{rec}/\mu_o$
ν	number of the stator ν th harmonic; kinematic viscosity
ξ	reduced height of the stator conductor
ρ	specific mass density
σ	electric conductivity; Stefan-Boltzmann constant
σ_f	form factor to include the saturation effect
σ_p	output coefficient

σ_r	radiation factor
τ	average pole pitch; thermal time constant
Φ	magnetic flux
Φ_f	excitation magnetic flux
Φ_l	leakage flux
ϕ	power factor angle
Ψ	flux linkage $\Psi = N\Phi$; angle between $\mathbf{I}_a$ and $\mathbf{E}_f$
ψ	flux linkage
Ω	angular speed $\Omega = 2\pi n$
ω	angular frequency $\omega = 2\pi f$

Subscripts

<i>a</i>	armature (stator)
<i>avg</i>	average
<i>c</i>	conduction
<i>cv</i>	control volume
<i>Cu</i>	copper
<i>d</i>	direct axis; differential; developed
<i>e</i>	end connection; eddy-current
<i>elm</i>	electromagnetic
<i>eq</i>	equivalent
<i>exc</i>	excitation
<i>ext</i>	external
<i>Fe</i>	ferromagnetic
<i>f</i>	field; forced
<i>fr</i>	friction; free
<i>g</i>	air gap
<i>h</i>	hydraulic; hysteresis
<i>in</i>	inner
<i>l</i>	leakage
<i>M</i>	magnet
<i>m</i>	peak value (amplitude)
<i>n, t</i>	normal and tangential components
<i>out</i>	output, outer
<i>q</i>	quadrature axis
<i>r</i>	rated; remanent; radiation; rotor
<i>r, θ, z</i>	cylindrical coordinate system
<i>rel</i>	reluctance
<i>rot</i>	rotational
<i>s</i>	slot; synchronous; system; stator
<i>sat</i>	saturation
<i>sh</i>	shaft
<i>st</i>	starting

<i>syn</i>	synchronous or synchronizing
<i>t</i>	teeth; total
<i>u</i>	useful
<i>v</i>	convection
<i>vent</i>	ventilation
<i>wind</i>	windage
<i>y</i>	yoke
<i>x, y, z</i>	cartesian coordinate system
1	stator; fundamental harmonic; inlet
2	rotor; exit

Superscripts

inc	incremental
(sq)	square wave
(tr)	trapezoidal

Abbreviations

A/D	analog to digital
AFPM	axial flux permanent magnet
AIFI	American Iron and Steel Industry
a.c.	alternating current
AWG	American Wire Gauge
BPF	band pass filtering
BSCCO	bismuth-strontium-calcium-copper-oxygen
CAD	computer-aided design
CPU	central processor unit
DSP	digital signal processor
d.c.	direct current
EDM	electro-discharge machining
EMALS	electro-magnetic aircraft launch system
EMF	electromotive force
EMI	electromagnetic interference
EV	electric vehicle
FCEV	fuel cell electric vehicle
FDB	fluid dynamic bearing
FEM	finite element method
FPGA	field programmable gate array
FRP	fiberglass reinforced plastic
GCD	greatest common divisor
HDD	hard disk drive
HEV	hybrid electric vehicle
HTS	high temperature superconducting

IC	integrated circuit
IGBT	insulated-gate bipolar transistor
ISG	integrated starter-generator
LPF	low pass filter
LVAD	left ventricular assist device
MMF	magnetomotive force
MMT	moving magnet technologies
MOSFET	metal oxide semiconductor (MOS) field effect transistor
MTBF	mean time between failures
MVD	magnetic voltage drop
NdFeB	neodymium iron boron
PFM	pulse frequency modulation
PLD	programmable logic device
PM	permanent magnet
PWM	pulse width modulation
RFI	radio frequency interference
RFPM	radial flux permanent magnet
SEMA	segmented electro-magnetic array
SMC	soft magnetic composite
SmCo	samarium cobalt
SSC	solid state converter
YBCO	yttrium-barium-copper oxide ($\text{YBa}_2\text{Cu}_3\text{O}_{7-x}$)

References

1. Abdel-Razek A.A., Coulomb J.L., Feliachi M., and Sabonnadiere J.C. (1981). The calculation of electromagnetic torque in saturated electric machines within combined numerical and analytical solutions of the field equations, IEEE Trans. MAG-17(6):3250–3252.
2. Abdel-Razek A.A., Coulomb J.L., Feliachi M., and Sabonnadiere J.C. (1982). Conception of an air-gap element for the dynamic analysis of the electromagnetic field in electric machines, IEEE Trans. MAG-18(2):655–659.
3. Acarnley P.P., Mecrow B.C., Burdess J.S., Fawcett J.N., Kelly J.G., and Dickinson P.G. (1996). Design principles for a flywheel energy store for road vehicles, IEEE Trans. IA-32(6):1402–1408.
4. Accucore, TSC Ferrite International, Wadsworth, IL, U.S.A., (2000), www.tscinternational.com
5. Afonin A.A., and Cierzniewski P. (1999). Electronically commutated disc-type permanent magnet motors (in Russian), Int. Conf. on Unconventional Electromechanical and Electr Systems UEES'99. St Petersburg, Russia, pp. 271–276.
6. Afonin A.A., Kramarz W., and Cierzniewski P. (2000) *Electromechanical Energy Converters with Electronic Commutation (in Polish)*. Szczecin: Wyd Ucz PS.
7. Ahmed A.B., and de Cachan L.E. (1994). Comparison of two multidisc configurations of PM synchronous machines using an elementary approach, Int. Conf. on Electr. Machines ICEM'94, vol. 1, Paris, France, pp. 175–180.
8. Amaratunga G.A.J., Acarnley P.P., and McLaren P.G. (1985). Optimum magnetic circuit configurations for PM aerospace generators, IEEE Trans on AES, 21(2):230–255.
9. Anderson H.H. (1972). *Centrifugal pumps*. The Trade and Technical Press.
10. Angelo J.D., Chari M.V.K., and Campbell P. (1983). 3-D FE solution for a PM axial field machine, IEEE Trans. PAS-102(1):83–90.
11. Armensky E.V., and Falk G.B. (1978). *Fractional-Horsepower Electrical Machines*. Moscow: Mir Publishers.
12. Arnold D.P., Zana I., Herrault F., Galle P., Park J.W., Das, S., Lang, J. H., and Allen, M. G. (2005). Optimization of a microscale, axial-flux, permanent-magnet generator, Proc. 5th Int. Workshop Micro Nanotechnology for Power

- Generation and Energy Conversion Apps. (PowerMEMS 2005), Tokyo, Japan, pp. 165–168.
13. Arnold D.P., Das S., Park, J.W., Zana, I., Lang, J.H., and Allen, M.G. (2006). Microfabricated high-speed axial-flux multiwatt permanent magnet generators Part II: Design, fabrication, and testing, *Journal of Microelectromechanical Systems*, 15(5):1351–1363.
 14. Atallah K., Zhu Z.Q., Howe D., and Birch T.S. (1998). Armature Reaction Field and Winding Inductances of Slotless Permanent-Magnet Brushless Machines, *IEEE Trans. MAG*-34(8):3737–3744.
 15. Balagurov V.A., Galtiev F.F., and Larionov A.N. (1964). *Permanent Magnet Electrical Machines (in Russian)*. Moscow: Energia.
 16. Baluja S. (1994). *Population-based incremental learning: A method for integrating genetic search based function optimization and competitive learning*. Tech. report No. CMU-CS-94-163, Carnegie Mellon University, Pittsburgh, U.S.A.
 17. Barakat G., El-Meslouhi T., and Dakyo B. (2001). Analysis of the cogging torque behavior of a two-phase axial flux permanent magnet synchronous machine. *IEEE Trans. MAG*-37(4):2803–2805.
 18. Baudot J.H. (1967). *Les Machines Électriques en Automatique Appliquée (in French)*. Paris: Dunod.
 19. Becerra R.C., and Ehsani M. (1988). High-speed torque control of brushless PM motors, *IEEE Trans. IE*-35(3):402–405.
 20. Berry C.H. (1954). *Flow and fan - principle of moving air through ducts*. The Industrial Press, New York.
 21. Bertotti G., Boglietti A., Champi M., Chiarabaglio D., Fiorillo D., and Lazari M. (1991). An improved estimation of iron losses in rotating electrical machines, *IEEE Trans. MAG*-27(6):5007–5009.
 22. Biwersi, S., Billet, L., Gandel, P. and Prudham, D. (2002). Low cost – high speed small size disk magnet synchronous motor, 8th Int. Conf. Actuator'2002, Bremen, Germany, pp. 196–200.
 23. Bolognani S., Oboe R., and Zigliotto M. (1999). Sensorless full-digital PMSM drive with EKF estimation of speed and rotor position, *IEEE Trans. IE*-46(1): 184–191.
 24. Bose B.K. (1988). A high-performance inverter-fed drive system of an interior PM synchronous machine, *IEEE Trans. IA*-24(6):987–998.
 25. Braga G., Farini A., and Manigrasso R. (1991). Synchronous drive for motorized wheels without gearbox for light rail systems and electric cars, 3rd European Power Electronic Conf. EPE'91, vol. 4, Florence, Italy, pp. 78–81.
 26. Brauer J.R., (ed.) (1988). *What Every Engineer Should Know about Finite Element Analysis*. New York: Marcel Dekker.
 27. Bumby J.R., and Martin R. (2005). Axial-flux permanent-magnet air-cored generator for small-scale wind turbines *IEE Proc.-Electr. Power Appl.*, 152(5):1065–1075
 28. Campbell P. (1972). A new wheel motor for commuter cars, *Electrical Review*, (190), pp. 332–333.
 29. Campbell P. (1974). Principle of a PM axial field DC machine, *Proceedings of IEE*, vol.121, no.1, pp. 1489–1494.
 30. Campbell P. (1975). The magnetic circuit of an axial flux DC electrical machine, *IEEE Trans. MAG*-11(5):1541–1543.

31. Campbell P. (1979). Performance of a permanent magnet axial-field d.c. machine, IEE Proc Pt B 2(4):139–144.
32. Campbell P., Rosenberg D.J., and Stanton D.P. (1981). The computer design and optimization of axial field PM motor, IEEE Trans. PAS-100(4):1490–1497.
33. Campbell P. (1994). *Permanent magnet materials and their application*. Cambridge University Press, Cambridge, UK.
34. Capponi F., Terrigi R., Caricchi F., Del Ferraro L. (2007). Active output voltage regulation for an ironless axial-flux PM automotive alternator with electromechanical flux weakening, IEEE-IAS Annual Meeting, New Orleans, Paper no. 48P2.
35. Caricchi F., Crescimbinì F., di Napoli A., Honorati O., Lipo T.A., Noia G., and Santini E. (1991). Development of a IGBT inverter driven axial-flux PM synchronous motor drive, European Power Electronics Conf. EPE'91, Firenze, Italy, vol.3, pp. 482–487.
36. Caricchi F., Crescimbinì F., Honorati O., and Santini E. (1992). Performance evaluation of an axial-flux PM generator, Int. Conf. on Electr. Machines ICEM'92, Manchester, U.K., vol. 2, pp. 761–765.
37. Caricchi F., Crescimbinì F., di Napoli A., and Santini E. (1992). Optimum CAD-CAE design of axial flux permanent magnets motors, Int. Conf. on Electr. Machines ICEM'92, Manchester, U.K., vol. 2, pp. 637–641.
38. Caricchi F., Crescimbinì F., Fedeli E., and Noia G. (1994). Design and construction of a wheel-directly-coupled axial-flux PM motor prototype for EVs, IEEE-IAS Annual Meeting, IAS-29, part 1, pp. 254–261.
39. Caricchi F., Crescimbinì F., and Santini E. (1995). Basic principle and design criteria of axial-flux PM machines having counterrotating rotors, IEEE Trans. IA-31(5):1062–1068.
40. Caricchi F., Crescimbinì F., Mezzetti F., and Santini E. (1996). Multistage axial-flux PM machines for wheel-direct-drive, IEEE Trans. IA-32(4):882–888.
41. Caricchi F., Crescimbinì F., di Napoli A., and Marcheggiani M. (1996). Prototype of electric-vehicle-drive with twin water-cooled wheel direct-drive motors, IEEE Annual Power Electronics Specialists Conf. PESC'96, Part 2, pp. 1926–1932.
42. Caricchi F., Crescimbinì F., Santini E., and Santucci C. (1997). Influence of the radial variation of the magnet pitches in slot-less PM axial flux motors, IEEE-IAS Annual Meeting, vol. 1, pp. 18–23.
43. Caricchi F., Crescimbinì F., Santini E., and Santucci C. (1998). FEM evaluation of performance of axial flux slotted PM machines, IEEE IAS Annual Meeting, vol. 1, pp. 12–18.
44. Caricchi F., Crescimbinì F., and Honorati O. (1998). Low-cost compact permanent magnet machine for adjustable-speed pump application, IEEE Trans. IA-34(1):109–116.
45. Caricchi F., Crescimbinì F., Honorati O., Bianco G.L., and Santini E. (1998). Performance of core-less winding axial-flux PM generator with power output at 400Hz – 3000 rpm, IEEE Trans. IA-34(6):1263–1269.
46. Caricchi F., Santini E., Crescimbinì F., and Solero L. (2000). High efficiency low volume starter/alternator for automotive applications, IEEE-IAS Annual Meeting, Rome, vol. 1, pp. 215–222.
47. Carter G.W. (1954). *Electromagnetic field in its engineering aspects*. Longmans.

48. Cascio A.M. (1997). Modeling, analysis and testing of orthotropic stator structures, Naval Symp. on Electr. Machines, Newport, RI, USA, pp. 91–99.
49. Chalmers B.J., Hamed S.A., and Baines G.D. (1985). Parameters and performance of a high-field permanent magnet synchronous motor for variable-frequency operation, *Proc IEE Pt B* 132(3):117–124.
50. Chalmers B.J., Spooner E., Honorati O., Crescimbin F., and Caricch F. (1997). Compact permanent magnet machines, *Electr. Machines and Power Systems*, 25(6): 635–648.
51. Chalmers B.J., Wu W., and Spooner E. (1999). An axial-flux permanent-magnet generator for a gearless wind energy system, *IEEE Trans. EC-14*(2): 251–257.
52. Chan C.C. (1982). *Axial-field electrical machines with yokeless armature core*. PhD Thesis, University of Hong Kong.
53. Chan C.C. (1987). Axial-field electrical machines: design and application, *IEEE Trans. EC-2*(2): 294–300.
54. Chandler P.L., and Patterson D.J. (1999). Counting the losses in very high efficiency machine design, World Renewable Energy Congress, Perth, Australia.
55. Chang L. (1994). Comparison of a.c. drives for electric vehicles — a report on experts' opinion survey, *IEEE AES Systems Magazine* 8: 7–11.
56. Chari M.V.K., and Silvester P.P. (ed.) (1980). *Finite element in electrical and magnetic field problems*. John Wiley & Sons, New York.
57. Chen J., and Chin K. (2003). Minimum copper loss flux-weakening control of surface mounted permanent magnet synchronous motors, *IEEE Trans. PE-18*(4): 929–936.
58. Chen S.X., Low T.S., Lin H., and Liu Z.J. (1996). Design trends of spindle motors for high performance hard disk drives. *IEEE Trans. MAG-32*(5): 3848–3850.
59. Chiliet C., Brissonneau P., and Yonnet J.P. (1991). Development of a water cooled permanent magnet synchronous machine. *Int. Conf. on Synchronous Machines SM100*, vol 3, Zürich, Switzerland, pp. 1094–1097.
60. Chin Y.K., Nordlund E., and Staton D.A. (2003). Thermal analysis - lumped circuit model and finite element analysis, *The 6th Int. Power Engineering Conf.*, Singapore, pp.1067–1072.
61. Chisholm D. (1983). *Two-phase flow in pipelines and heat exchangers*. George Godwin, New York.
62. Chung S.U., Hwang, G.Y., Hwang, S.M., Kang, B.S., and Kim H.G. (2002) Development of brushless and sensorless vibration motor used for mobile phones, *IEEE Trans. MAG-18*(5): 3000–3002.
63. Cistelecan M.V., Popescu M., and Popescu M. (2007). Study Of the number of slots / pole combinations for low speed permanent magnet synchronous generators, *IEEE-IEMDC Conference*, Turkey, pp. 1616–1620.
64. Coilgun research spawns mighty motors and more. *Machine Design* 9(Sept 24):24–25, (1993).
65. Coulomb J.L., and Meunier G. (1984). Finite element implementation of virtual work principle for magnetic or electric force and torque computation, *IEEE Trans. MAG-20*(5):1894–1896.
66. Crescimbin F., and Solero L. (2005). Advances in Propulsion Systems of Hybrid and Electric Vehicles, *The 8th Brazilian Power Electronics Conference COBEP'05*, Recife, Brazil.

67. Cros J., and Viarouge P. (2002). Synthesis of high performance PM motors with concentrated windings, *IEEE Trans. EC-17(2)*:248–253.
68. Cvetkovski G., Petkovska L., Cundev M., and Gair S. (1998). Mathematical model of a PM axial field synchronous motor for a genetic algorithm optimisation, *Int. Conf. on Electr. Machines ICEM'98, Istanbul*, vol. 2, pp. 1172–1177.
69. Dąbrowski M., and Gieras J.F. (1977). *Induction motors with solid rotors (in Polish)*. PWN, Warsaw-Poznan.
70. Dąbrowski M. (1977). *Construction of Electrical Machines (in Polish)*. Warsaw: WNT.
71. Dąbrowski M. (1988). *Magnetic fields and circuits of electrical machines (in Polish)*. Warsaw, WNT.
72. Dąbrowski M. (1980). Joint action of permanent magnets in an electrical machine (in Polish), *Zeszyty Nauk. Polit. Pozn. Elektryka* 21:7–17.
73. Day A.J., and Hirzel A. (2002). Redefining power generation, *Gorham Conference*, Cincinnati, OH, U.S.A. www.lightengineering.com
74. De Angelo C., Bossio G., Solsona J., García G., and Valla M. (2002). Sensorless speed control of permanent magnet motors with torque ripple minimization, *28th Annual Conf. of the IEEE Industrial Electronics Society (IECON'02)*, Sevilla, Spain, vol. 1, pp. 680–685.
75. De Angelo C., Bossio G., Solsona J., García G., and Valla M. (2002). A rotor position and speed observer for permanent magnet motors with non-sinusoidal EMF waveform, *28th Annual Conf. of the IEEE Industrial Electronics Society (IECON'02)*, Sevilla, Spain, vol. 1, pp. 756–761.
76. De Bruijn F. (2005). The current status of fuel cell technology for mobile and stationary applications, *Green Chemistry*, 2005, 7, 132–150.
77. Dote Y., and Kinoshita S. (1990). *Brushless Servomotors. Fundamentals and Applications*. Oxford: Clarendon Press.
78. Douglas J.F., Gasiorek J.M., and Swaffield J.A. (1995). *Fluid mechanics*. 3rd ed., Longman Scientific & Technical.
79. Doyle M.R., Samuel D.J., Conway T., and Klimowski R.R. (2001). Electromagnetic aircraft launch system, *Int. Electr. Machines and Drives Conf. (IEMDC'2001)*, Boston, MA, U.S.A.
80. Dunkerley S. (1893). On the whirling and vibration of shafts, *Proc. of the Royal Society of London*, vol. 54, pp. 365–370.
81. Eastham, J.F., Profumo, F., Tenconi, A., Hill-Cottingham R., Coles, P., and Gianolio, G. (2002). Novel axial flux machine for aircraft drive: design and modeling, *IEEE Trans. MAG-38(5)*:3003–3005
82. El-Hasan T.S., Luk, P.C.K., Bhinder, F.S., and Ebaid M.S. (2000). Modular design of high-speed permanent-magnet axial-flux generators. *IEEE Trans. MAG-36(5)*:3558–3561.
83. El-Hasan T., and Luk P.C.K. (2003). Magnet topology optimization to reduce harmonics in high speed axial flux generators, *IEEE Trans. MAG-39(5)*:3340–3342.
84. Engelmann R.H., and Middelndorf W.H. (ed.) (1995). *Handbook of electric motors*. Marcel Dekker, Inc., New York.
85. Engstrom J. (2000). Inductance of slotless machines, in *Proc. of the IEEE Nordic Workshop on Power and Industrial Electronics*, Aalborg, Denmark.
86. Evans P.D., and Easham J.F. (1980). Disc-geometry homopolar synchronous machine, *Proc. of IEE*, pt. B, 127(6): 299–307.

87. Evans P.D., and Easham J.F. (1983). Slot-less alternator with ac-side excitation, *Proc. of IEE*, 130(6): 399–406.
88. Evans P.D., and Easham J.F. (1983). Double-disc alternator with ac-side excitation, *IEE Proc. Part B*, EPA-130(2), pp. 95–102.
89. Ertugrul N., and Acarnley P.P. (1992). Analytical solution of the system equations of the axial field permanent magnet synchronous motor drive, *Proceedings of ICEM'92*, vol. 2, pp. 785–789.
90. Ertugrul N., and Acarnley P. (1994). A new algorithm for sensorless operation of permanent magnet motors, *IEEE Trans. IA*-30(1):126–133.
91. Favre E., Cardoletti L., and Jufer M. (1993). Permanent magnet synchronous motors: a comprehensive approach to cogging torque suppression, *IEEE Trans. IA*-29(6):1141–1149.
92. Ficheux R., Caricchi F., Honorati O., and Crescimbeni F. (2000). Axial flux permanent magnet motor for direct drive elevator systems without machine room, *IEEE-IAS Annual Meeting, Rome*, vol. 1, pp. 190–197.
93. *Film coil motor*. EmBest, Seoul, Korea, (2001), www.embest.com
94. Fitzgerald A.E., and Kingsley C. (1961). *Electric Machinery*. 2nd ed., New York: McGraw-Hill.
95. Flack T.J., and Volschenk A.F. (1994). Computational aspects of time-stepping finite element analysis using an air-gap element, *Int. Conf. on Electr. Machines ICEM'94*, Paris, France, vol. 3, pp. 158–163.
96. Fracchia M., and Sciotto G. (1994). Cycloconverter Drives for Ship Propulsion, *Symp. on Power Electronics, Electr. Drives, Advanced Electr. Motors SPEEDAM'94*, Taormina, Italy, pp. 255–260.
97. Fratta A., Villata F., and Vagati A. (1991). Extending the voltage saturated performance of a DC brushless drive, *European Power Electronic Conf. EPE'91*, vol. 4, Florence, Italy, pp. 134–138.
98. World Fuel Cell Council, www.fuelcellworld.org.
99. Furlani E.P. (1992). A method for predicting the field in axial field motors, *IEEE Trans. MAG*-28(5):2061–2066.
100. Furlani E.P. (1994). Computing the field in permanent magnet axial-field motors, *IEEE Trans. MAG*-30(5):3660–3663.
101. Gair S., Eastham J.F., and Profumo F. (1995). Permanent magnet brushless d.c. drives for electric vehicles, *Int. Aeagean Conf. on Electr. Machines and Power Electronics ACEMP'95*, Kuşadasi, Turkey, pp. 638–643.
102. Gieras I.A., and Gieras J.F. (2006). Recent advancements in permanent magnet motors technology for medical applications, (invited paper), 25th Int. Symp. on Micromachines and Servo Drives MIS'06, Białowieża, Poland, pp.7–13.
103. Gieras J.F. (1981). Electrodynamic levitation forces – theory and small-scale test results, *Acta Technica CSAV*, No. 4, pp. 389–414.
104. Gieras J.F. (1983). Simplified theory of double-sided linear induction motor with squirrel-cage elastic secondary, *IEE Proc. Part B* 130(6):424–430.
105. Gieras J.F., Santini E., and Wing M. (1998). Calculation of synchronous reactances of small permanent magnet alternating-current motors: comparison of analytical approach and finite element method with measurements, *IEEE Trans. MAG*-34(5):3712–3720.
106. Gieras J.F., and Wing M. (2002). *Permanent magnet motor technology: design and applications*. 2nd ed., Marcel Dekker, New York.

107. Gieras J.F., and Gieras I.A. (2002). Performance analysis of a coreless permanent magnet brushless motor, IEEE 37th IAS Meeting, Pittsburgh, PA, U.S.A.
108. Glinka T. (1995). *Electrical Micromachines with Permanent Magnet Excitation (in Polish)*. Gliwice (Poland): Silesian Techn University, Gliwice.
109. Goetz J., and Takahashi, T. (2003). A design platform optimized for inner loop control, presented at 24th Int. Exhibition and Conf. on Power Electronics, Intelligent Motion and Power Quality (PCIM 2003), Nuremburg, Germany.
110. Gottvald A., Preis K., Magele C., Biro O., and Savini A. (1992). Global optimization methods for computational electromagnetics, IEEE Trans. MAG-28(2):1537–1540.
111. Grover L. (1962). *Inductance calculations - working formulas and tables*. Dover, New York.
112. Gu C., Wu W., and Shao K. (1994). Magnetic field analysis and optimal design of d.c. permanent magnet core-less disk machines, IEEE Trans. MAG-30(5) Pt 2, pp. 3668–3671.
113. Hakala H. (2000). Integration of motor and hoisting machine changes the elevator business, Int. Conf. on Electr. Machines ICEM'2000, vol 3, Espoo, Finland, pp. 1242–1245.
114. Halbach K. (1980). Design of permanent multipole magnets with oriented rare earth cobalt material, Nuclear Instruments and Methods, vol. 169, pp. 1–10.
115. Halbach K. (1981). Physical and optical properties of rare earth cobalt magnets, Nuclear Instruments and Methods, vol. 187, pp. 109–117.
116. Halbach K. (1985). Application of permanent magnets in accelerators and electron storage rings, J. Appl. Physics, vol. 57, pp. 3605–3608.
117. Hamarat S., Leblebicioglu K., and Ertan H.B. (1998). Comparison of deterministic and non-deterministic optimization algorithms for design optimization of electrical machines, Int. Conf. on Electr. Machines ICEM'98, Istanbul, Turkey, vol.3, pp. 1477–1482.
118. Hanitsch R., Belmans R., and Stephan R. (1994). Small axial flux motor with permanent magnet excitation and etched air gap winding, IEEE Trans. MAG-30(2):592–594.
119. Hanselman D.C. (2003). *Brushless permanent-magnet motor design*. Cranston, RI: The Writers' Collective.
120. *Hardware interfacing to the TMS320C25*. Texas Instruments.
121. Heller B., and Hamata V. (1977). *Harmonic Field Effect in Induction Machines*. Prague: Academia (Czechoslovak Academy of Sciences).
122. Hendershot J.H., and Miller T.J.E. (1994). *Design of Brushless Permanent Magnet Motors*. Oxford: Clarendon Press.
123. Holland J.H. (1994). *Adaption in nature and artificial systems*. Bradford Books, U.S.A.
124. Holman J.P. (1992). *Heat transfer*. 7th ed., McGraw-Hill (UK), London.
125. Honorati O., Solero L., Caricchi F., and Crescimbin F. (1998). Comparison of motor drive arrangements for single-phase PM motors, Int. Conf. on Electr. Machines, ICEM'98, vol. 2, Istanbul, Turkey, pp. 1261–1266.
126. Honsinger V.B. (1980). Performance of polyphase permanent magnet machines, IEEE Trans. PAS-99(4):1510–1516.

127. Hoole S.R. (1989). *Computer-aided analysis and design of electromagnetic devices*. Elsevier Science Publishing, New York.
128. Hrabovcová V., and Bršlica V. (1990). Disk synchronous machines with permanent magnets – electric and thermal equivalent circuits, *Electr. Drives Symp.*, Capri, Italy, pp. 163–169.
129. Hredzak B., and Gair S. (1996). Elimination of torque pulsation in a direct drive EV wheel motor, *IEEE Trans. MAG-32(5) Pt 2*, pp. 5010–5012.
130. Huang M., Ye Y., Fan C., Clancy T. (2007). Application of disc permanent magnet linear synchronous motor in novel oil-pumping machines, 6th Int. Symp. on Linear Drives for Industrial Applications, LDIA'07, Lille France, paper number: 122.
131. Huang S., Luo J., Leonardi F., and Lipo T.A. (1996). A general approach to sizing and power density equations for comparison of electrical machines, *IEEE-IAS Annual Meeting*, San Diego, CA, U.S.A., pp. 836–842.
132. Huang S., Luo J., Leonardi F., and Lipo T.A. (1999). A comparison of power density for axial flux machines based on general purpose sizing equations, *IEEE Trans. EC-14(2)*:185–192.
133. Hughes A., and Miller T.J. (1977). Analysis of fields and inductances in air-cored and iron-cored synchronous machines, *Proceedings of IEE*, 124(2): 121–126.
134. Incropera F.P., and DeWitt D.P. (2001). *Fundamentals of heat and mass transfer*. 5th ed., John Wiley & Sons, New York.
135. Ivanuskin V.A., Sarapulov F.N., and Szymczak P. (2000). *Structural Simulation of Electromechanical Systems and Their Elements (in Russian)*. Szczecin: Wyd Ucz PS.
136. Jahns T.M. (1984). Torque production in permanent magnet synchronous motor drives with rectangular current excitation, *IEEE Trans. IA-20(4)*:803–813.
137. Jahns T.M. (1987). Flux weakening regime operation of an interior PM synchronous motor drive, *IEEE Trans. IA-23(4)*:681–689.
138. Jang J., Sul S.K., Ha J., Ide K., and Sawamura M. (2003). Sensorless drive of surface-mounted permanent-magnet motor by high-frequency signal injection based on magnetic saliency, *IEEE Trans. IA-39(4)*: 1031–1039.
139. Jang G.H., and Chang J.H. (1999). Development of dual air gap printed coil BLDC motor, *IEEE Trans. MAG-35(3)*: 1789–1792.
140. Jang G.H., and Chang J.H. (2002). Development of an axial-gap spindle motor for computer hard disk drives using PCB winding and dual air gaps, *IEEE Trans. MAG-38(5)*: 3297–3299.
141. Jensen C.C., Profumo F., and Lipo T.A. (1992). A low loss permanent magnet brushless d.c. motor utilizing tape wound amorphous iron, *IEEE Trans. IA-28(3)*:646–651.
142. Jones B.L., and Brown J.E. (1987). Electrical variable-speed drives, *IEE Proc. Pt A* 131(7):516–558.
143. Kamper M.J., and Mackay A.T. (1995). Optimum control of the reluctance synchronous machine with a cageless flux barrier rotor, *Trans. of SAIEE*, 86(2): 49–56.
144. Kamper M.J., Van der Merwe F.S., and Williamson S. (1996). Direct finite element design optimisation of cageless reluctance synchronous machine, *IEEE Trans. EC-11(3)*:547–555.

145. Kamper M.J., and Jackson S. (1998). Performance of small and medium power flux barrier rotor reluctance synchronous machine drives, Proc. of ICEM'98, Istanbul, Turkey, vol. 1, pp. 95-99.
146. Kamper M.J., Wang R.-J., and Rossouw F.G. (2007). Analysis and performance evaluation of axial flux air-cored stator permanent magnet machine with concentrated coils, IEEE-IEMDC Conference, Turkey, pp. 13-20.
147. Kenjo T., and Nagamori S. (1985). *Permanent magnet and brushless d.c. motors*. Oxford: Clarendon Press.
148. Kenjo T. (1990). *Power electronics for the microprocessor era*. Oxford: OUP.
149. Kenjo T. (1991). *Electric motors and their control*. Oxford: OUP.
150. Kessinger R.L., and Robinson S. (1997). SEMA-based permanent magnet motors for high-torque, high-performance, Naval Symp. on Electr. Machines, Newport, RI, U.S.A., pp. 151-155.
151. Kessinger R.L. (2002). *Introduction to SEMA motor technology*. Kinetic Art and Technology, Greenville, IN, U.S.A.
152. King R.D., Haefner K.B., Salasoo L., and Koegl R.A. (1995). Hybrid electric transit bus pollutes less, conserves fuel, IEEE Spectrum 32(7): 26-31.
153. Kitano Seiki Co., Ltd., www.kitano-seiki.co.jp
154. Kleen S., Ehrfeld W., Michel F., Nienhaus M., and Stölting H.D. (2000). Penny-motor: A family of novel ultraflat electromagnetic micromotors, Int. Conf. Actuator'2000, Bremen, Germany, pp. 193-196.
155. Klug L. (1990). Axial field a.c. servomotor, Electr. Drives and Power Electronics Symp. EDPE'90, Košice, Slovakia, pp. 154-159.
156. Klug L. (1991). Synchronous servo motor with a disc rotor (in Czech), Elektrotechnický Obzor 80(1-2):13-17.
157. Klug L., and Guba R. (1992). Disc rotor a.c. servo motor drive, Electr. Drives and Power Electronics Symp. EDPE'92, Košice, Slovakia, pp. 341-344.
158. Komęza K., Pelikant A., Tegopoulos J. and Wiak S. (1994). Comparative computation of forces and torques of electromagnetic devices by means of different formulae, IEEE Trans. MAG-30(5):3475-3478.
159. Kostenko M., and Piotrovsky L. (1974). *Electrical Machines*. vol.1: Direct Current Machines and Transformers. Moscow: Mir Publishers.
160. Kubzdela S., and Węgliński B. (1988). Magnetodielectrics in induction motors with disk rotors, IEEE Trans. MAG-24(1):635-638.
161. Lammeraner J., and Štafl M. (1964). *Eddy Currents*. London: Iliffe Books.
162. Lange A., Canders W.R., Laube F., and Mosebach H. (2000). Comparison of different drive systems for a 75 kW electrical vehicles drive, Int. Conf. on Electr. Machines ICEM'2000, vol. 3, Espoo, Finland, pp. 1308-1312.
163. Lange A., Canders W.R., and Mosebach H. (2000). Investigation of iron losses of soft magnetic powder components for electrical machines, Int. Conf. on Electr. Machines ICEM'2000, vol. 3, Espoo, Finland, pp. 1521-1525.
164. Leibold R., Bossio G., Garcia G., and Valla M. (2001). A new strategy to extend the speed range of a permanent magnet a.c. motor, The 6th Brazilian Power Electronics Conference (COBEP'2001), Florianópolis, SC, Brazil.
165. Leung W.S., and Chan C.C. (1980). A new design approach for axial-field electrical machine, IEEE Trans. PAS-99(4):1679-1685.
166. Libert F., and Soulard J. (2004). Investigation on pole-slot combinations for permanent magnet machines with concentrated windings, Proceedings of ICEM, Poland, pp. 530-535.

167. Linke M., Kennel R., and Holtz J. (2002). Sensorless position control of permanent magnet synchronous machine without limitation at zero speed, IEEE-IECON Annual Conf., Spain, pp. 674–679.
168. Linke M., Kennel R., and Holtz J. (2003). Sensorless speed and position control of synchronous machines using alternating carrier injection, IEEE Electr. Mach. and Drives Conf. (IEMDC'03) (Madison/Wisconsin, USA), pp. 1211–1217.
169. Liu C.T., Chiang T.S., Zamora J.F., and Lin S.C. (2003). Field-oriented control evaluations of a single-sided permanent magnet axial-flux motor for an electric vehicle, IEEE Trans. MAG-39(5):3280–3282.
170. Lombard N.F., and Kamper M.J. (1999). Analysis and performance of an ironless stator axial flux PM machine, IEEE Trans. EC-14(4):1051–1056.
171. Lovatt H.C., Ramdenand V.S., and Mecrow B.C. (1998). Design of an in-wheel motor for a solar-powered electric vehicle, Proc. of IEE: Electric Power Applications, vol. 145, No. 5, pp. 402–408.
172. Lowther D.A., and Silvester P.P. (1986). *Computer-aided design in magnetics*. Berlin: Springer Verlag.
173. Lukaniszyn M., Wróbel R., Mendrela A., and Drzewoski R. (2000). Towards optimisation of the disc-type brushless d.c. motor by changing the stator core structure, Int. Conf. on Electr. Machines ICEM'2000, Vol. 3, Espoo, Finland, pp. 1357–1360.
174. Lukaniszyn M., Mendrela E., Jagiello M., and Wróbel R. (2002). Integral parameters of a disc-type motor with axial stator flux, Zesz. Nauk.Polit. Slaskiej, vol. 200, Elektryka no. 177, pp. 255–262.
175. Magnetfabrik Schramberg GmbH & Co, Schramberg-Sulgen, (1989).
176. Magnussen F., and Sadarangani C. (2003). Winding factors and joule losses of permanent magnet machines with concentrated windings, Proc. IEEE International Electrical machines and drives conference (IEMDC), Madison (USA), vol. 1, pp. 333–339.
177. Magnussen F., Thelin P., and Sadarangani C. (2004). Performance evaluation of permanent magnet synchronous machines with concentrated and distributed windings including the effect of field-weakening, Proc. IEE International Conference on Power Electronics, Machines and Drives (PEMD), Edinburgh (UK), vol. 2, pp. 679–685.
178. Mangan J., and Warner A. (1998). *Magnet wire bonding*. Joyal Product Inc., Linden, NJ, U.S.A., www.joyalusa.com
179. Maric D.S., Hiti S., Stancu C.C., Nagashima J.M., and Rutledge D.B. (1999). Two flux weakening schemes for surface-mounted permanent-magnet synchronous drives – design and transient response consideration. IEEE-IAS 34th Annual Meeting, Phoenix, AZ, pp. 673–678.
180. Marignetti F., and Scarano M. (2000). Mathematical modeling of an axial-flux PM motor wheel, Int. Conf. on Electr. Machines ICEM'2000, vol. 3, Espoo, Finland, pp. 1275–1279.
181. *Maxon Motor*. Sachseln, Switzerland: Interelectric AG, (1991/92).
182. Mbidi D.N., Van der Westhuizen K., Wang R-J, Kamper M.J., and Blom J. (2000). Mechanical design considerations of double stage axial-flux permanent magnet machine, IEEE-IAS 35th Annual Meeting, Rome, vol. 1, pp. 198–201.
183. McFee S., and Lowther D.A. (1987). Towards accurate and consistent force calculation in finite element based computational magnetostatics, IEEE Trans. MAG-23(5):3771–3773.

184. Mellara B., and Santini E. (1994). FEM computation and optimization of L_d and L_q in disc PM machines," 2nd Int. Workshop on Electr. and Mag. Fields, Leuven, Belgium, paper No. 89.
185. Mendrela E., Lukaniszyn M., and Macek-Kaminska K. (2002). *Electronically commutated d.c. brushless disc motors (in Polish)*. Warsaw: Gnome.
186. Metzger D.E., and Afgan N.H. (1984). *Heat and mass transfer in rotating machinery*. Hemisphere Publishing Corporation, Washington DC, U.S.A.
187. Miller T.J.E. (1989). *Brushless Permanent-Magnet and Reluctance Motor Drives*. Oxford: Clarendon Press.
188. Miller D.S. (1990). *Internal flow systems*. 2nd ed., BHRA Information Services, The Fluid Eng. Centre, Cranfield, Bedford, U.K.
189. Millner A.R. (1994). Multi-hundred horsepower permanent magnet brushless disk motors, IEEE Appl. Power Electronics Conf. and Exp. APEC'94, pp. 351–355.
190. Mills A.F. (1995). *Basic heat and mass transfer*. Richard D. Irwin, U.S.A.
191. Mishler W.R. (1981). Test results on a low amorphous iron induction motor, IEEE Trans. PAS-100(6):860–866.
192. Miti G.K., and Renfrew A.C. (1998). Field weakening performance of the TORUS motor with rectangular current excitation, Int. Conf. on Electr. Machines IECM'98, Istanbul, Vol. 1, pp. 630–633.
193. Mizia J., Adamiak K., Eastham A.R., and Dawson G.E. (1988). Finite element force calculation: comparison of methods for electric machines, IEEE Trans. MAG-24(1):447–450.
194. Mohan N., Undeland T.M., Robbins W.P. (1989). *Power Electronics Converters, Applications, and Design*. New York: J Wiley & Sons.
195. Mongeau P. (1997). High torque/high power density permanent magnet motors, Naval Symp. on Electr. Machines, Newport, RI, USA, pp. 9–16.
196. Morimoto S., Sanada, M., Takeda, Y. (1994). Wide-speed operation of interior permanent magnet synchronous motors with high performance current regulator, IEEE Trans. IA-30(4):920–926.
197. Muljadi E., Butterfield C.P., and Wan Y.H. (1999). Axial flux, modular, permanent magnet generator with a toroidal winding for wind turbine applications, IEEE Trans. IA-35(4):831–836.
198. Munson B.R., Young D.F., and Okiishi T.H. (1994). *Fundamentals of fluid mechanics*. 2nd ed., New York, John Wiley & Sons.
199. Nagashima J. (2005). Wheel Hub Motors for Automotive Applications, Proceedings of the 21st Electric Vehicles Symposium, Monaco, April 2-5.
200. Nasar S.A., Boldea I., and Unnewehr L.E. (1993). *Permanent magnet, reluctance, and self-synchronous motors*. Boca Raton: CRC Press.
201. Neyman L.R. (1949). *Skin effect in ferromagnetic bodies (in Russian)*. GEI, Leningrad–Moskva.
202. Ofori-Tenkorang J., and Lang J.H. (1995). A comparative analysis of torque production in Halbach and conventional surface-mounted permanent magnet synchronous motors, IEEE-IAS Annual Meeting, Orlando, CA, U.S.A., pp. 657–663.
203. O'Neil S.J. (1997). Advances in motor technology for the medical industry, Medical Device and Diagnostic Journal Ind. Magazine, No. 5.
204. Osin I.L., Kolesnikov V.P., and Yuferov F.M. (1976). *Permanent Magnet Synchronous Micromotors (in Russian)*. Moscow: Energia.

205. Ovrebo S. (2002). Comparison of excitation signals for low and zero speed estimation of the rotor position in a axial flux PMSM, Nordic Workshop on Power and Industrial Electronics NORPIE 2002, Stockholm, Sweden.
206. Owen J.M. (1988). Air cooled gas turbine discs: a review of recent research, *Int. Journal of Heat and Fluid Flow*, vol. 9, No. 4, pp. 354–365.
207. Owen J.M. (1989). An approximate solution for the flow between a rotating and a stationary disk, *ASME Journal of Turbomachinery*, Vol. 111, No. 4, pp. 323–332.
208. Owen J.M. (1971). The Reynolds analogy applied to flow between a rotating and a stationary disc, *Int. Journal of Heat and Mass Transfer*, Vol.14, pp.451–460.
209. Owen J.M. (1971). The effect of forced flow on heat transfer from a disc rotating near a stator, *Int. Journal of Heat and Mass Transfer*, vol. 14, pp. 1135–1147.
210. Owen J.M., and Rogers R.H. (1989). *Flow and heat transfer in rotating-disc system, Vol. 1: Rotor-stator systems*. Research Studies Press, Taunton, UK.
211. Owen J.M., and Rogers R.H. (1995). *Flow and heat transfer in rotating-disc system, Vol. 2: Rotating cavities*. Research Studies Press, Taunton, UK.
212. Parker R.J. (1990). *Advances in Permanent Magnetism*. New York: J Wiley & Sons.
213. Parviainen A., Pyrhonen J., and Kontkanen P. (2005). Axial flux permanent magnet generator with concentrated winding for small wind power applications, *IEEE Int'l Conf. on Electric Machines and Drives*, pp. 1187–1191.
214. Parviainen A., Niemela M., and Pyrhonen J. (2004). Modeling of Axial Flux Permanent-Magnet Machines, *IEEE Trans. IA-40(5)*:1333–1340.
215. Patterson D., and Spée R. (1995). The design and development of an axial flux permanent magnet brushless d.c. motor for wheel drive in a solar powered vehicle, *IEEE Trans. IA-31(5)*:1054–1061.
216. Perho J., and Ritchie E. (1996). A motor design using reluctance network, 2nd Int. Conf. on Unconventional Electr. and Electromechanical Systems, UEES'96, Alushta, Ukraine, pp. 309–314.
217. Plat D. (1989). Permanent magnet synchronous motor with axial flux geometry, *IEEE Trans. MAG-25(4)*:3076–3079.
218. Powell M.J.D. (1964). An efficient method for finding the minimum of a function of several variables without calculating derivatives, *Computer Journal*, Vol. 7, pp. 155–162.
219. Profumo F., Zhang Z., and Tenconi A. (1997). Axial flux machines drives: a new viable solution for electric cars, *IEEE Trans. IE-44(1)*:39–45.
220. Profumo F., Tenconi A., Zhang Z., and Cavagnino A. (1998). Design and realization of a novel axial flux interior PM synchronous motor for wheel-motors applications, *Int. Conf. on Electr. Machines ICEM'98, Istanbul*, Vol. 3, pp. 1791–1796.
221. Profumo F., Tenconi A., Zhang Z., and Cavagnino A. (1998). Novel axial flux interior PM synchronous motor realized with powdered soft magnetic materials, *IEEE-IAS Annual Meeting*, vol. 1, pp. 152–159.
222. Pullen K., Etemad M.R., and Fenocchi A. (1996). The high speed axial flux disc generator unlocking the potential of the automotive gas turbine, in *Proc. IEE Colloq. Machines and Drives for Electric and Hybrid Vehicles, IEE Colloq. Dig.*, London, U.K., pp. 8/1-4.

223. Rahman K., Ward T., Patel N., Nagashima J., Caricchi F., and Crescimbin F. (2006). Application of Direct Drive Wheel Motor for Fuel Cell Electric and Hybrid Electric Vehicle Propulsion Systems, *IEEE Trans. IA*-42(5):1185–1192.
224. Rajashekara K., Kawamura A., and Matsuse K., (ed.) (1996). *Sensorless Control of AC Motor Drives*. New York: IEEE Press.
225. Ramsden V.S., Mecrow B.C., and Lovatt H.C. (1997). Design of an in wheel motor for a solar-powered electric vehicle, *Proc. of EMD'97*, pp. 192–197.
226. Ramsden V.S., Watterson P.A., Holliday W.M., Tansley G.D., Reizes J.A., and Woodard J.C. (2000). A rotary blood pump, *Journal IEEE Australia*, vol. 20, pp.17–22
227. Rao J.S. (1983). *Rotor dynamics*, 3rd edition, New Age Int. Publishers, New Delhi.
228. Ratnajeevan S., Hoole H., and Haldar M.K. (1995). Optimisation of electromagnetic devices: circuit models, neural networks and gradient methods in concept, *IEEE Trans. MAG*-31(3):2016–2019.
229. Reichert, K. (2004). Motors with concentrated, non overlapping windings, some characteristics, *Int. Conf. on Electr. Machines ICEM'2004*, Cracow, Poland, Paper No. 541.
230. Salon S.J. (1995). *Finite element analysis of electrical machines*. Kluwer Academic Publishers, Norwell, MA, U.S.A.
231. Say M.G. (1992). *Alternating Current Machines*. Singapore: ELBS with Longman.
232. Sayers A.T. (1990). *Hydraulic and compressible flow turbomachines*. McGraw-Hill (U.K.), London.
233. Schaefer G. (1991). Field weakening of brushless PM servomotors with rectangular current, *European Power Electronic Conf. EPE'91*, vol. 3, Florence, Italy, pp. 429–434.
234. Schroedl M. (1996). Sensorless control of AC machines at low speed and standstill based on the INFORM method, *IEEE-IAS 31st Annual Meeting*, San Diego, CA, pp. 270–277.
235. Scowby S.T., Dobson R.T., and Kamper M.J. (2004). Thermal modelling of an axial flux permanent magnet machine, *Applied Thermal Engineering*, vol. 24, No. 2-3, pp. 193–207.
236. SGS–Thomson Motion Control Applications Manual (1991).
237. Sidelnikov B., and Szymczak P. (2000). Areas of application and appraisal of control methods for converter-fed disc motors (in Polish), *Prace Nauk. IMNiPE*, Technical University of Wroclaw, 48: Studies and Research 20:182–191.
238. Silvester P.P., and Ferrari R.L. (1990). *Finite Elements for Electrical Engineers*. 2nd ed. Cambridge: Cambridge University Press.
239. Sitapati K., and Krishnan R. (2001). Performance comparisons of radial and axial field permanent magnet brushless machines, *IEEE Trans. IA*-37(5):1219–1226.
240. Skaar S.E., Krovel O., and Nilssen R. (2006). Distribution, coil span and winding factors for PM machines with concentrated windings, *Int. Conf. on Electr. Machines ICEM-2006*, Chania, Greece, paper 346.
241. Soderlund L., Koski A., Vihriala H., Eriksson J.T., and Perala R. (1997). Design of an axial flux PM wind power generator, *Int. Conf. on Electr. Machine and Drives ICEMD'97*, Milwaukee, WI, U.S.A., pp. 224–228.

242. SomaloyTM 500, Höganäs, Höganäs, Sweden, (1997), www.hoganas.com
243. Spooner E., and Chalmers B.J. (1988). Toroidally-wound, slotless, axial-flux PM brushless d.c. motor, Int. Conf. on Electr. Machines ICEM'88, Pisa, Italy, Vol. 3, pp. 81–86.
244. Spooner E., Chalmers B., and El-Missiry M.M. (1990). “A compact brushless d.c. machine,” Electr. Drives Symp. EDS'90, Capri, Italy, pp. 239–243.
245. Spooner E., and Chalmers B.J. (1992). TORUS: a slotless, toroidal-stator, permanent-magnet generator, Proc. of IEE, Pt. B EPA vol. 139, pp. 497–506.
246. Stec T.F. (1994). Amorphous magnetic materials Metgkass 2605S-2 and 2605TCA in application to rotating electrical machines, NATO ASI Modern Electrical Drives, Antalya, Turkey.
247. Stefani P., and Zandla G. (1992). Cruise liners diesel electric propulsion. Cyclo- or synchroconverter? The shipyard opinion, Int. Symp. on Ship and Shipping Research vol. 2, Genoa, Italy, pp. 6.5.1–6.5.32.
248. Strachan P.J., Reynaud F.P., and von Backström T.W. (1992), The hydrodynamic modeling of torque converters, R&D Journal, South African Inst. of Mech. Eng., vol. 8, No. 1, pp. 21–29.
249. Sugimoto H., Nishikawa T., Tsuda T., Hondou Y., Akita Y., Takeda T., Okazaki T., Ohashi S., and Yoshida Y. (2006). Trial manufacture of liquid nitrogen cooling high temperature superconductivity rotor, Journal of Physics: Conference Series, (43):780–783
250. Sullivan C.R. (2001). Computationally efficient winding loss calculation with multiple windings, arbitrary waveforms, and two-dimensional or three-dimensional field geometry, IEEE Trans. PE-16(1):142–150.
251. Takano H., Itoh T., Mori K., Sakuta A., and Hirasa T. (1992). Optimum values for magnet and armature winding thickness for axial-field PM brushless DC motor, IEEE Trans. IA-28(2):350–357.
252. Takeda T., Togawa H., and Oota T. (2006). Development of liquid nitrogen-cooled full superconducting motor, IHI Engineering Review, 39(2):89–94
253. Timar P.L., Fazekas A., Kiss J., Miklos A., and Yang S.J. (1989). *Noise and Vibration of Electrical Machines*. Amsterdam: Elsevier.
254. Toliyat H.A., and Campbell S. (2003). *DSP-based electromechanical motion control*. Boca Raton, FL, CRC Press.
255. Tomassi G., Topor M., Marignetti F., and Boldea I. (2006). Characterization of an axial-flux machine with non-overlapping windings as a generator, Electromotion, vol. 13, pp. 73–79.
256. Upadhyay P.R., and Rajagopal K.R. (2005). Comparison of performance of the axial-field and radial-field permanent magnet brushless direct current motors using computer aided design and finite element methods, Journal of Applied Physics 97, 10Q506, pp. 1–3.
257. Varga J.S. (1992). A breakthrough in axial induction and synchronous machines, Int. Conf. on Electr. Machines ICEM'1992, vol. 3, Manchester, UK, pp. 1107–1111.
258. Voldek A.I. (1974). *Electrical Machines (in Russian)*. St Petersburg: Energia.
259. Wallace R., Lipo T.A., Moran L.A., and Tapia J.A. (1997). Design and construction of a PM axial flux synchronous generator, Int. Conf. on Electr. Machines and Drives ICEDM'97, Milwaukee, WI, U.S.A., pp. MA1 4.1–4.3.
260. Wang R-J, Kamper M.J., and Dobson R.T. (2005). Development of a thermofluid model for axial field permanent magnet machines, IEEE Trans EC-20(1):80–87.

261. Wang R-J, Mohellebi H., Flack T., Kamper M., Buys J., and Feliachi M. (2002). Two-dimensional Cartesian air-gap element (CAGE) for dynamic finite element modeling of electrical machines with a flat air gap, *IEEE Trans. MAG*-38(2):1357–1360.
262. Wang R-J, and Kamper M.J. (2002). Evaluation of eddy current losses in axial flux permanent magnet (AFPM) machine with an ironless stator, 37th IEEE-IAS Meeting, Pittsburgh, USA, 2:1289–1294.
263. Wang R-J, and Kamper M.J. (2004). Calculation of eddy current loss in axial field permanent magnet machine with coreless stator, *IEEE Trans EC*-19(3):532–538.
264. Wang R-J, Kamper M.J., Van der Westhuizen K., and Gieras J.F. (2005). Optimal design of a coreless stator axial field permanent magnet generator, *IEEE Trans MAG*-41(1):55–64.
265. Węgliński, B. (1990). Soft magnetic powder composites — dielectromagnetics and magnetodielectrics, *Reviews on Powder Metallurgy and Physical Ceramics*, Vol. 4, No. 2, Freund Publ. House Ltd., London, UK.
266. White F.M. (1994). *Fluid mechanics*. McGraw-Hill Book Company, New York.
267. Wiak S., and Welfle H. (2001). *Disc type motors for light electric vehicles (in Polish)*. Lodz: Technical University of Lodz.
268. Wijenayake A.H., Bailey J.M., and McCleer P.J. (1995). Design optimization of an axial gap PM brushless d.c. motor for electric vehicle applications, *IEEE-IAS Annual Meeting*, pp. 685–691.
269. Williamson S., and Smith J.R. (1980). The application of minimisation algorithms in electrical engineering, *Proc. of IEE*, vol.127, Pt. A, No. 8, pp. 528–530.
270. Wong W.Y. (1977). *Heat transfer for engineers*. Longmans.
271. Wu R., and Slemon G.R. (1991). A permanent magnet motor drive without a shaft sensor, *IEEE Trans. IA*-27(5): 1005–1011.
272. Wurster R. (1999). PEM fuel cells in stationary and mobile applications: pathways to commercialization, 6th International Technical Congress (BIEL'99), 13-19 September.
273. Wurster R. (2005). Status quo fuel cell development in europe, *Fuel Cell Congress (F-Cell 2005)*, Workshop: 'Germany meets Canada', Stuttgart, 28 September.
274. Xu L., Xu X., Lipo T.A., and Novotny D.W. (1991). Vector control of a synchronous reluctance motor including saturation and iron loss, *IEEE Trans. IA*-27(5):977–987.
275. Zangwill W.I. (1967). Nonlinear programming via penalty functions, *Management Science*, Vol. 13, pp. 344–358.
276. Zhang Z., Profumo F., and Tenconi A. (1994). Axial flux interior PM synchronous motors for electric vehicle drives, *Symp. on Power Electronics, Electr. Drives, Advanced Electr. Motors SPEEDAM'94*, Taormina, Italy, pp. 323–328.
277. Zhang Z., Profumo F., and Tonconi A. (1996). Axial flux wheel machines for electric vehicles, *Electr. Machines and Power Systems*, vol.24, no.8, pp. 883–896.
278. Zhang Z., Profumo F., and Tonconi A. (1996). Axial flux versus radial flux permanent magnet motors, *Electromotion*, vol. 3, pp. 134–140.

- 279. Zhang Z., Profumo F., and Tenconi A. (1997). Analysis and experimental validation of performance for an axial flux PM brushless d.c. motor with powder iron metallurgy cores, *IEEE Trans. MAG-33*(5):4194–4196.
- 280. Zhu Z.Q, Chen Y.S., and Howe D. (2000). Online optimal flux-weakening control of permanent-magnet brushless AC drives, *IEEE Trans. IA-36*(6):1661–1668.
- 281. Zhilichev Y.N. (1996). Calculation of 3D magnetic field of disk-type micromotors by integral transformation method,, *IEEE Trans. MAG-32*(1):248–253.
- 282. Zhilichev Y.N. (1998). Three-dimensional analytic model of permanent magnet axial flux machine, *IEEE Trans. MAG-34*(6):3897–3901.

Patents

- [P1] US132. Davenport T. (1837). Improvement in propelling machinery by magnetism and electro-magnetism.
- [P2] US405858. Tesla N. (1889). Electro-Magnetic Motor.
- [P3] US3069577. Lee R. (1962). Disc rotor for induction motor.
- [P4] US3230406. Henry-Baudot J. (1966). High frequency electromechanical generator.
- [P5] US3231774. Henry-Baudot J. (1966). A.C. rotating electric machines with printed circuit armatures.
- [P6] US3239702. Van de Graaff R.J. (1966). Multi-disk electromagnetic power machinery.
- [P7] US3280353. Haydon A.W., et al (1966). Electric rotating machine.
- [P8] US3293466. Henry-Baudot J. (1966). Axial airgap electric rotary machines.
- [P9] US3315106. Reynst M.F. (1967). Disk shaped electric motor.
- [P10] US3375386. Hayner P.F., et al (1968). Printed circuit motor.
- [P11] US3383535. Lohr T.E. (1968). Electric motor.
- [P12] US3407320. McLean W.B. (1968). Wafer type submersible motor for underwater device.
- [P13] US3440464. Tolmie R.J. (1969). Electromagnetic apparatus.
- [P14] US3443133. Henry-Baudot J. (1969). Stator for axial airgap electric rotary machines.
- [P15] US3487246. Long B.E. (1969). Electric machine.
- [P16] US3524250. Burr R.P. (1970). Method of manufacturing electrical wire wound machines.
- [P17] US3524251. Burr R.P. (1970). Method of manufacturing disc-type wire wound electrical machines.
- [P18] US3525007. Henry-Baudot J. (1970). Axial airgap machines having means to reduce eddy current losses in the armature windings.
- [P19] US3529191. Henry-Baudot J. (1970). Homopolar machine having disklike rotor.
- [P20] US3534469. Keogh R.J. (1970). Method of manufacturing windings for disc-type dc machine armatures.
- [P21] US3538704. Kawanaka H., et al (1970). Balance wheel motor in a timepiece.
- [P22] US3550645. Keogh R.J. (1970). Wire wound armature, method and apparatus for making same.

- [P23] US3555321. Gruener R., et al (1971). Eddy current brake for use in business machines or the like.
- [P24] US3558947. Burr R.P. (1971). Discoidal wire wound armatures.
- [P25] US3575624. Keogh R.J. (1971). Wire wound disc armature.
- [P26] US3678314. Carter A.H. (1972). Discoidal electric motor.
- [P27] US3745388. Frederick D.M. (1973). Axial air gap motor.
- [P28] US3922574. Whiteley E. (1975). Permanent magnet hermetic synchronous motor.
- [P29] US3979619. Whiteley E. (1976). Permanent magnet field structure for dynamoelectric machines.
- [P30] US3999092. Whiteley E. (1976). Permanent magnet synchronous dynamoelectric machine.
- [P31] US4007387. Rustecki R.Z. (1977). Electrical synchronous machines.
- [P32] US4020372. Whiteley E. (1977). Cooling of discoidal dynamoelectric machines.
- [P33] US4059777. Whiteley E. (1977). Cooling of discoidal dynamoelectric machines.
- [P34] US4068143. Whiteley E. (1978). Discoidal winding for dynamoelectric machines.
- [P35] US4076340. Meinke P., et al (1978). Drive and bearing support for a disc-shaped rotor.
- [P36] US4100443. Kuwako T. (1978). Electrical rotary machine.
- [P37] US4187441. Oney W.R. (1980). High power density brushless dc motor.
- [P38] US4188556. Hahn J.H. (1980). Electro-mechanical machine.
- [P39] US4237396. Blenkinsop P.T., et al (1980). Electromagnetic machines with permanent magnet excitation.
- [P40] US4253031. Frister M. (1981). Directly driven dynamo electric machine-gas turbine generator structure.
- [P41] US4297604. Tawse I.S. (1981). Axial air gap alternators/generators of modular construction.
- [P42] US4319152. van Gils A.W. (1982). Laminated winding for electric machines.
- [P43] US4363988. Kliman G.B. (1982). Induction disk motor with metal tape components.
- [P44] US4371801. Richter E. (1983). Method and apparatus for output regulation of multiple disk permanent magnet machines.
- [P45] US4390805. Hahn J.H. (1983). Electromechanical machine.
- [P46] US4398112. van Gils A.W. (1983). Aminated winding for electric machines.
- [P47] US4435662. Tawse I.S. (1984). Axial air gap alternators/generators of modular construction.
- [P48] US4508998. Hahn J.H. (1985). Brushless disc-type dc motor or generator.
- [P49] US4536672. Kanayama K., et al (1985). Flat type rotary electric machine.
- [P50] US4567391. Tucker H.F., et al (1986). Permanent magnet disc rotor machine.
- [P51] US4578610. Kliman G.B., et al (1986). Synchronous disk motor with amorphous metal stator and permanent magnet rotor and flywheel.
- [P52] US4605873. Hahn J.H. (1986). Electromechanical machine.
- [P53] US4644207. Catterfeld F.C., et al (1987). Integrated dual pump system.
- [P54] US4701655. Schmider F. (1987). D.C. machine, with mechanical and electrical connections among collector segments.
- [P55] US4710667. Whiteley E. (1987). Brushless d.c. dynamoelectric machine with decreased magnitude of pulsations of air gap flux.

- [P56] US4720640. Anderson B.M., et al (1988). Fluid powered electrical generator.
- [P57] US4823039. Lynch C. (1989). Electrical machines.
- [P58] US4841393. MacLeod D.J., et al (1989). Spindle motor for a disc drive.
- [P59] US4864175. Rossi L. (1989). Rotor for an electric motor.
- [P60] US4866321. Blanchard H.J., et al (1989). Brushless electrical machine for use as motor or generator.
- [P61] US4868443. Rossi L. (1989). Tachogenerator for electric machines.
- [P62] US4959578. Varga J.S. (1990). Dual rotor axial air gap induction motor.
- [P63] US4978878. Dijken R.H. (1990). Electric multipolar machine.
- [P64] US4996457. Hawsey R.A., et al (1991). Ultra-high speed permanent magnet axial gap alternator with multiple stators.
- [P65] US5001412. Carter I.W., et al (1991). Alternator starter.
- [P66] US5021698. Pullen K.R., et al (1991). Axial field electrical generator.
- [P67] US5097140. Crall F.W. (1992). Alternator starter.
- [P68] US5130595. Arora R.S. (1992). Multiple magnetic paths machine.
- [P69] US5194773. Clarke P.W. (1993). Adjustable speed split stator rotary machine.
- [P70] US5200659. Clarke P.W. (1993). Axial and radial field electric rotating machines having relatively rotatable first and second stators.
- [P71] US5334898. Skybyk D. (1994). Polyphase brushless dc and ac synchronous machines.
- [P72] US5334899. Skybyk D. (1994). Polyphase brushless dc and ac synchronous machines.
- [P73] US5394321. McCleer P.J., et al (1995). Quasi square-wave back-emf permanent magnet ac machines with five or more phases.
- [P74] US5396140. Goldie J.H., et al (1995). Parallel air gap serial flux a.c. electrical machine.
- [P75] US5535582. Paweletz A. (1996). Drive for a shaftless spinning rotor of an open end spinning machine.
- [P76] US5548950. Paweletz A. (1996). Motor for a shaftless spinning rotor for an open-end spinning machine.
- [P77] US5615618. Berdud E. (1997). Orbital and modular motors using permanent magnets and interleaved iron or steel magnetically permeable members.
- [P78] US5619087. Sakai K. (1997). Axial-gap rotary-electric machine.
- [P79] US5637941. Paweletz A. (1997). Shaftless spinning rotor for an open-end spinning machine.
- [P80] US5642009. McCleer P.J., et al (1997). Quasi square-wave back-emf permanent magnet ac machines with five or more phases.
- [P81] US5710476. Ampela M.J. (1998). Armature design for an axial-gap rotary electric machine.
- [P82] US5731645. Clifton D.B., et al (1998). Integrated motor/generator/flywheel utilizing a solid steel rotor.
- [P83] US5744896. Kessinger R.L., Jr., et al (1998). Interlocking segmented coil array.
- [P84] US5777421. Woodward R.C., Jr. (1998). Disc-type electrical machine.
- [P85] US5801473. Helwig A. (1998). Open stator axial flux electric motor.
- [P86] US5877578. Mitcham A.J., et al (1999). Rotor disc construction for use in an electrical machine.
- [P87] US5905321. Clifton D.B., et al (1999). Energy storage flywheel apparatus and methods.

- [P88] US5920138. Clifton D.B., et al (1999). Motor/generator and axial magnetic bearing utilizing common magnetic circuit.
- [P89] US5925965. Li Y., et al (1999). Axial flux reluctance machine with two stators driving a rotor.
- [P90] US5932935. Clifton D.B., et al (1999). Energy storage flywheel emergency power source and methods.
- [P91] US5955808. Hill W. (1999). Multi-phase electric machine with offset multi-polar electric pole units.
- [P92] US5955809. Shah M.J. (1999). Permanent magnet generator with auxiliary winding.
- [P93] US5955816. Clifton D.B., et al (1999). Energy storage flywheel apparatus and methods.
- [P94] US5982069. Rao D.K. (1999). Axial gap machine phase coil having tapered conductors with increasing width in radial direction.
- [P95] US5982074. Smith S.H., et al (1999). Axial field motor/generator.
- [P96] US6002193. Canini J-M, et al (1999). Basic module for a discoidal electric machine, and corresponding electric machine.
- [P97] US6037696. Sromin A., et al (2000). Permanent magnet axial air gap electric machine.
- [P98] US6040650. Rao D.K. (2000). Stator with coplanar tapered conductors.
- [P99] US6046518. Williams M.R. (2000). Axial gap electrical machine.
- [P100] US6064135. Hahn J.H. (2000). Electromechanical machine and armature structure therefore.
- [P101] US6155969. Schima H., et al (2000). Centrifugal pump for pumping blood and other shear-sensitive liquids.
- [P102] US6163097. Smith S.H., et al (2000). Motor generator including interconnected stators and stator laminations.
- [P103] US6166472. Pinkerton J.F., et al (2000). Airgap armature coils and electric machines using same.
- [P104] US6194802B1. Rao D.K. (2001). Axial gap motor with radially movable magnets to increase speed capability.
- [P105] US6199666B1. Aulanko E., et al (2001). Elevator drive machine.
- [P106] US6404097B1. Pullen K.R. (2002). Rotary electrical machines.
- [P107] US6445105B1. Kliman G.B., et al (2002). Axial flux machine and method of fabrication.
- [P108] US6476534B1. Vanderbeck W.E., et al (2002). Permanent magnet phase-control motor.
- [P109] US6515390B1. Lopatinsky E.L., et al (2003). Electric drive apparatus with a rotor having two magnetized disks.
- [P110] US6633106B1. Swett D.W. (2003). Axial gap motor-generator for high speed operation.
- [P111] US6674214B1. Knorzer K-H, et al (2004). Electric axial flow machine.
- [P112] US6700271B2. Detela A. (2004). Hybrid synchronous motor equipped with toroidal winding.
- [P113] US6700288B2. Smith J.S. (2004). High speed rotor.
- [P114] US6707221B2. Carl R.J. (2004). Axial flux machine, stator and fabrication method.
- [P115] US6750584B2. Smith J.S. (2004). High speed rotor.
- [P116] US6750588B1. Gabrys C.W. (2004). High performance axial gap alternator motor.

- [P117] US6762523B1. Lisowski L. (2004). Continuously variable electromagnetic transmission.
- [P118] US6768239B1. Kelecý P., et al (2004). Electromotive devices using notched ribbon windings.
- [P119] US6794791B2. Ben Ahmed A.H., et al (2004). Motor/generator with energized reluctance and coil in the air gap.
- [P120] US6803694B2. Decristofaro N.J., et al (2004). Unitary amorphous metal component for an axial flux electric machine.
- [P121] US6828710B1. Gabrys C.W. (2004). Airgap armature.
- [P122] US6833647B2. Saint-Michel J., et al (2004). Discoid machine.
- [P123] US6844656B1. Larsen K.D., et al (2005). Electric multipole motor/generator with axial magnetic flux.
- [P124] US6891301B1. Hsu J.S. (2005). Simplified hybrid-secondary uncluttered machine and method.
- [P125] US6940200B2. Lopatinsky E., et al (2005). Electric drive.
- [P126] US6967554B2. Eydelie A., et al (2005). Coil for a rotary electric machine.
- [P127] US6977454B2. Hsu J.S. (2005). Hybrid-secondary uncluttered permanent magnet machine and method.
- [P128] US6995494B2. Haugan O., et al (2006). Axial gap brushless dc motor.
- [P129] US7034422B2. Ramu K. (2006). Radial-axial electromagnetic flux electric motor, coaxial electromagnetic flux electric motor, and rotor for same.
- [P130] US7034425B2. Detela A. (2006). Hybrid synchronous electric machine.
- [P131] US7034427B2. Hirzel A.D. (2006). Selective alignment of stators in axial airgap electric devices comprising low-loss materials.
- [P132] US7042109B2. Gabrys C.W. (2006). Wind turbine.
- [P133] US7064469B2. Jack A., et al (2006). Core back of an electrical machine and method for making the same.
- [P134] US7067950B2. Hirzel A.D., et al (2006). Efficient high-speed electric devices using low-loss materials.
- [P135] US7084548B1. Gabrys C.W. (2006). Low cost high speed electrical machine.
- [P136] US7098566B2. Rajasingham A.I. (2006). Axial gap electrical machine.
- [P137] US7105979B1. Gabrys C.W. (2006). Compact heteropolar hybrid alternator-motor.
- [P138] US7109625B1. Jore L.M., et al (2006). Conductor optimized axial field rotary energy device.
- [P139] US7144468B2. Decristofaro N.J., et al (2006). Method of constructing a unitary amorphous metal component for an electric machine.
- [P140] US7148594B2. Rajasingham A.I (2006). Axial gap electrical machine.
- [P141] US7157826B2. Rajasingham A.I (2006). Axial gap electrical machine.
- [P142] US7157829B2. Van Tichelen P., et al (2007). Axial flux permanent magnet generator/motor.
- [P143] US7170212B2. Balson J.C., et al (2007). Synchronous axial field electrical machine.
- [P144] US7187098B2. Hasebe M., et al (2007). Axial gap rotating electrical machine.
- [P145] US7190101B2. Hirzel A.D. (2007). Stator coil arrangement for an axial airgap electric device including low-loss materials.
- [P146] US7226018B2. Sullivan S. (2007). Landing gear method and apparatus for braking and maneuvering.
- [P147] US7230361B2. Hirzel A.D. (2007). Efficient high-speed electric devices using low-loss materials.

- [P148] US7237748B2. Sullivan S. (2007). Landing gear method and apparatus for braking and maneuvering.
- [P149] US7262536B2. Rahman K.M., et al (2007). Gearless wheel motor drive system.

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