# Summary

# 6. AC brushless

## 6.1 Introduction

The expression AC brushless drives is used to indicate those drives that use a synchronous machine with permanent magnets: the excitation magnetic field, in such machines, is provided by permanent magnets.
The scope of these drives is mainly limited to low power (typically less than 50 kW) for the limitations imposed by the magnetic materials currently available to the construction of machines: the size and cost of the magnets are very high. However, this is an important field of application and in significant growth, including numerically controlled machine tools, industrial automation, robotics, light traction, heavy traction, wind generation. Furthermore, due to the virtual absence of rotor losses these machines do not require forced ventilation and are therefore suitable for applications like aerospace or contaminated environments.
The basic scheme of a servo drive (see Figure 6-1) consists of: a permanent magnet synchronous motor, a static converter (in this case a two-stage converter consists of a diode (o a IGBT/MOSFET) bridge rectifier and an inverter), a position sensor, some current and voltage sensors and a control device that operates based on information provided by the sensors.

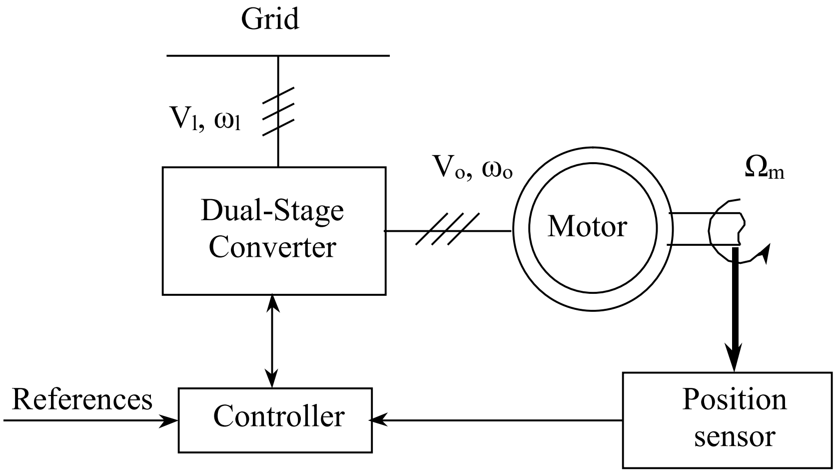


Figure 6-1: AC brushless drive scheme

The following considerations apply, mainly, to drives for axes of machine tools of small and medium power (2-10kW), but are also valid for high power drives.

## 6.2 General characteristics of the drive

In the field of drives for machine tools, a distinction is often introduced between the axes and spindles, where the first are intended solely for the motion control, while the second ones are used for rotation of the piece.
The next step is to analyze briefly the requirements of an axis drive: its task is essentially to move a tool or a piece, accelerating and decelerating from zero to a given speed, by means of a suitable and controlled torque: this task must be carried out in accordance with important requirements:

- Total bidirectionality drive with practically no dead zone around zero speed, in both static and dynamic conditions;
- The ratio between minimum and maximum speeds should be greater than 10 (at rated torque) and, passing from zero to rated load, the speed must not drop more than 1/10 of the maximum speed.

These characteristics can be summarized by saying that the drive must have a high static stiffness, meaning the static stiffness equal to the ratio between the applied external torque and the produced speed variation.
Another key element of technical opinion is the bandwidth of the speed control loop: a high performance axis drive requires a bandwidth regarding the phase not less than 40 Hz (ie the frequency at which the phase delay exceeds 45° above 40 Hz) while regarding the amplitude (frequency at which there is an attenuation of 3 dB) is not less than 70 Hz. These characteristics are valid in operation around zero speed, but they must stay within a margin of 20% also in the case of an "inertial load" (defined conventionally as a load equal to half the inertia of the machine) or of a "rated torque load" (conventionally defined as a torque equal to one half of the rated one). The magnitude of the bandwidth of the speed control loop can therefore be taken as an index of dynamic stiffness of the drive.
Finally, the acceleration provided by the machine in the presence of an inertial load or torque as defined above must be elevated, approximately, to 100 rad/s$^2$; however the maximum speeds required are not generally very high,: about 200 rad/s.
The purpose of these requirements, however, closely related to each other, is that you can get, with the help of rigid kinematic chains, a very accurate position control in order to achieve a high degree of finish of the workpiece.

## 6.3 Synchronous machine model

In the following, the mathematical model of the synchronous machine is briefly introduced. This model has implied some simplifications regarding the constructive nature of the machine, but it appears to be sufficiently adequate to its control.

Consider the machine shown schematically in Figure 6-2

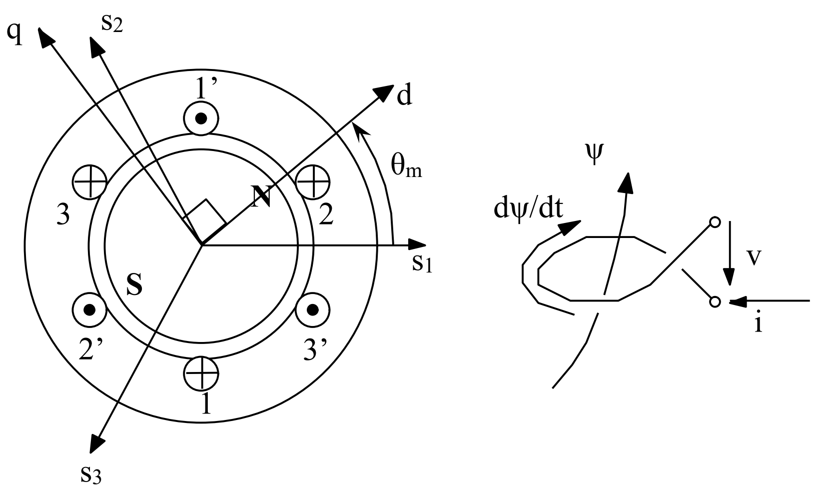


Figure 6-2: Schematic diagram of the brushless motor

It consists of a cylindrical stator with a symmetrical three-phase winding so as to generate a distribution of sinusoidal magnetomotive force in the air-gap, and a rotor flux created by permanent magnets (for the windings apply the conventions of Figure 6-2 and without damper cages. We also admits that the stator and the rotor are made of laminated material with infinite permeability. Saturation, hysteresis of iron and anisotropy of the machine due to the slots are neglected.

There are two main solutions for the rotor: with permanent magnets glued to the rotor surface (SMPM Surface Mounted Permanent Magnet, Figure 6-3) and interior magnets (IPM Interior Permanent Magnet, Figure 6-4).

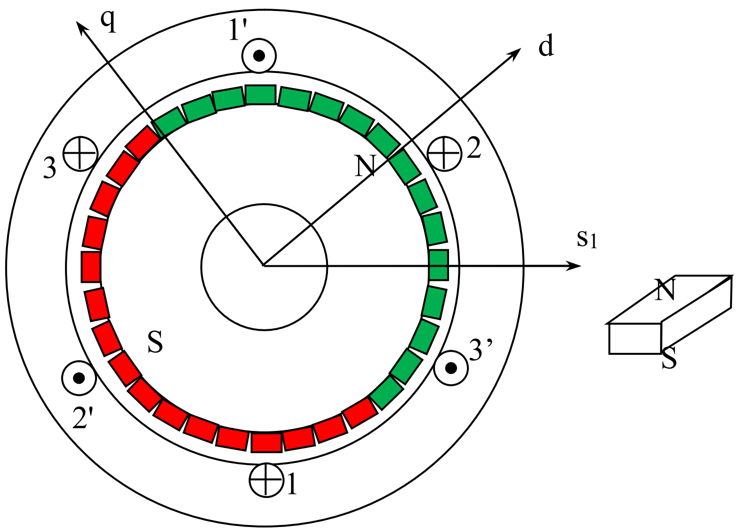


Figure 6-3: Example of a SMPM synchronous machine, 2 poles (green magnets have North outward while the South towards the rotor; red ones vice versa)

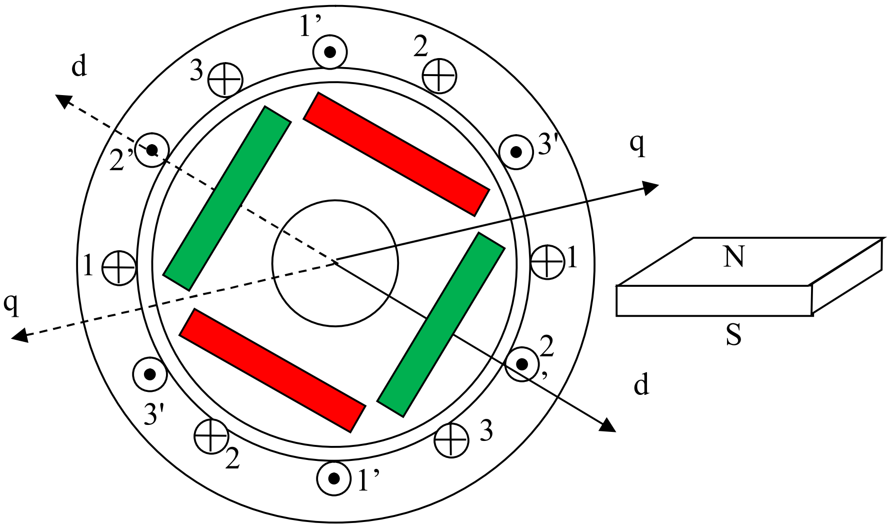


Figure 6-4: Example of IPM synchronous machine, 4 poles (green magnets have North outward while the South towards the rotor; red ones vice versa)

In the first case the machine can be considered isotropic (the self and mutual inductances do not depend on the angular position); in the second case the machine is anisotropic.

Starting from the stator equations and the current-flux relationship (for which, given the above assumptions, the superposition principle is valid)

$$v_{s1} = R_s i_{s1} + p\psi_{s1}$$
$$v_{s2} = R_s i_{s2} + p\psi_{s2}$$
$$v_{s3} = R_s i_{s3} + p\psi_{s3}$$
$$\psi_{s1} = L_{ss}(\theta_m) i_{s1} + M_{ss}(\theta_m) i_{s2} + M_{ss}(\theta_m - \frac{2}{3}\pi) i_{s3} + \psi_{pm}(\theta_m)$$
$$\psi_{s2} = L_{ss}(\theta_m - \frac{2}{3}\pi) i_{s2} + M_{ss}(\theta_m) i_{s1} + M_{ss}(\theta_m + \frac{2}{3}\pi) i_{s3} + \psi_{pm}(\theta_m - \frac{2}{3}\pi)$$
$$\psi_{s3} = L_{ss}(\theta_m - \frac{4}{3}\pi) i_{s3} + M_{ss}(\theta_m - \frac{2}{3}\pi) i_{s1} + M_{ss}(\theta_m + \frac{2}{3}\pi) i_{s2} + \psi_{pm}(\theta_m - \frac{4}{3}\pi)$$

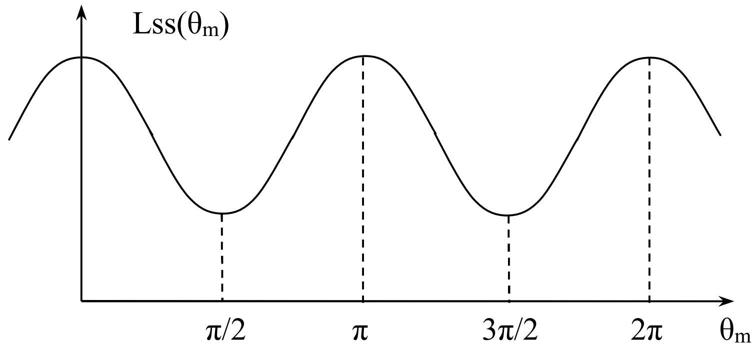


Figure 6-5: Self inductance behaviour of the first winding

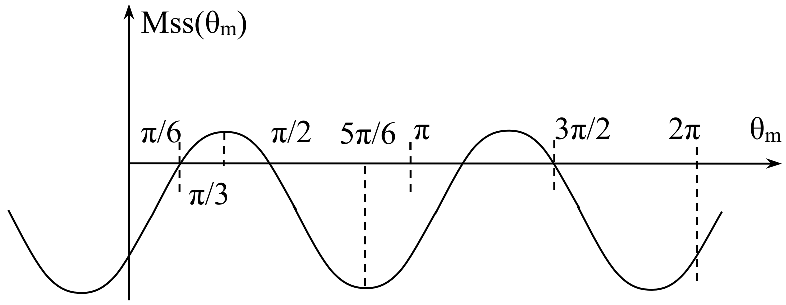


Figure 6-6: Behaviour of the mutual inductance between the first and the second winding

Consider now, for the sake of simplicity, an isotropic machine (the case of anisotropic machine will be taken up after). Recalling that the sum of the three phase currents is zero due to the connection (in isolated star or delta connection), we have:

$$\psi_{s1} = L_s i_{s1} + \psi_{pm}(\theta_m)$$
$$\psi_{s2} = L_s i_{s2} + \psi_{pm}(\theta_m - \frac{2}{3}\pi)$$
$$\psi_{s3} = L_s i_{s3} + \psi_{pm}(\theta_m - \frac{4}{3}\pi)$$

where $L_s = L_{ss} - M_{ss}$ (synchronous inductance).

The synchronous PM machine (AC brushless) is characterized by a flux linked with the stator windings, due to the permanent magnets with a sinusoidal waveform (due to a suitable distribution of the windings in the stator slots and/or a shaping of the magnets).

$$\psi_{pm}(\theta_m) = \hat{\psi}_{pm}\cos(\theta_m)$$

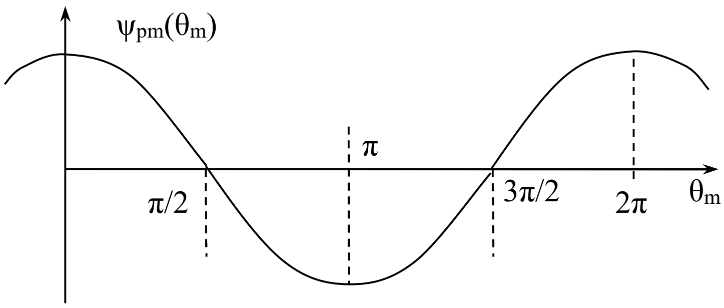


Figure 6-7: Waveform of the flux linked with the stator windings due to the PM

Applying now the space phasor formula in a stationary reference frame (α axis has the same direction of the magnetic axis of the first winding), we obtain the following relationships:

$$\begin{aligned}\bar{v}_s^{\alpha\beta} &= R_s \bar{i}_s^{\alpha\beta} + p\bar{\psi}_s^{\alpha\beta} \\ \bar{\psi}_s^{\alpha\beta} &= L_s \bar{i}_s^{\alpha\beta} + \sqrt{\frac{3}{2}}\hat{\psi}_{pm} e^{j\theta m}\end{aligned}$$

This procedure has transformed the three windings machine into a two-phase machine, equipped with two windings, fixed with the stator and orthogonal each other.

Defined $\psi_{pm}$ as

$$\psi_{pm} = \sqrt{\frac{3}{2}}\hat{\psi}_{pm}$$

and passing to a reference frame fixed with the PM flux (axis " $d$ " fixed with the North of the permanent magnets), we obtain the following relations ( $\overline{F}^{\alpha\beta} = \overline{F}^{dq} e^{j\theta m}$ )

$$\begin{aligned}\bar{v}_s &= R_s \bar{i}_s + p\bar{\psi}_s + j\dot{\theta}_m \bar{\psi}_s \\ \psi_{sd} &= L_s i_{sd} + \psi_{pm} \\ \psi_{sq} &= L_s i_{sq}\end{aligned}$$

where $\bar{v}_s = v_d + jv_q$ and $v_d$, $v_q$ are the projections of the voltage space phasor on the axes $d$ and $q$ (fixed with rotor).

This procedure has transformed the two-phase machine, equipped with two windings, fixed with the stator and orthogonal each other, into a two-phase machine, with two windings equally distributed, whose magnetic axes (orthogonal each other) are moving fixed with the $d$ and $q$ axes.

If the machine is anisotropic, the winding *sd*, with magnetic axis in the direction $d$, will be crossed by the current $i_{sd}$ and will support the flux $\psi_{sd}$ through a self inductance $L_d$ that will be, now, constant (no more function of the mechanical angle) and different from the self inductance $L_q$ of the winding *sq*.

$$\begin{aligned}\bar{v}_s &= R_s \bar{i}_s + p\bar{\psi}_s + j\dot{\theta}_m \bar{\psi}_s \\ \psi_{sd} &= L_d i_{sd} + \psi_{pm} \\ \psi_{sq} &= L_q i_{sq}\end{aligned}$$

In the case of Figure 6-4, since the permanent magnet, by applying the principle of superposition and considering only the effect of the currents, has a behaviour similar to air, the flux along the direct axis $\psi_{sd}$ must pass through more air than the quadrature axis: in this case is $L_d<L_q$.

The energy balance says that:

$$\mathrm{Re}(\overline{v}_s \underline{i_s}) = R_s i_s^{\,2} + \mathrm{Re}(\underline{i_s} p \overline{\psi}_s) + \mathrm{Re}(j\dot{\theta}_m \overline{\psi}_s \underline{i_s})$$

The term at the left is the total power entering the machine, the first term on the right is the power losses (due to Joule effect), the second term is the variation of the internal energy (magnetic energy stored in the inductances $L_d$ e $L_q$: $i_{sd} p \psi_{sd}$ e $i_{sq} p \psi_{sq}$),) while the third is the mechanical power.

$$\begin{aligned} P_m &= \mathrm{Re}(j\dot{\theta}_m \overline{\psi}_s \underline{i_s}) = -\dot{\theta}_m \,\mathrm{Im}(\overline{\psi}_s \underline{i_s}) = -\dot{\theta}_m \,\mathrm{Im}\left[\left(L_d i_{sd} + \psi_{pm} + jL_q i_{sq}\right)\left(i_{sd} - j i_{sq}\right)\right] = \\ &= \dot{\theta}_m \left[\left(L_d - L_q\right) i_{sd} i_{sq} + \psi_{pm} i_{sq}\right] \end{aligned}$$

The expression of the torque is:

$$T_e = \frac{P_m}{\Omega_m} = \frac{P_m}{\Omega_m / n_p} = n_p \left[(L_d - L_q) i_{sd} i_{sq} + \psi_{pm} i_{sq}\right]$$

The first term is called "reluctance torque" (null if the machine is isotropic) and the second one "excitation torque".

Finally, there is the equation of the mechanical energy balance:

$$p\dot{\theta}_m = \frac{n_p}{J}(T_e - T_r)$$

## 6.4 Vector control

The vector control of synchronous machine is based on a suitable choice of reference axes $d$ and $q$, used by the controller of the power converter so that a component of the space phasor of the stator current acts only on the flux while the other one on the electromagnetic torque (in the air-gap). In this way, the synchronous motor is controlled like a DC machine where the regulator acts separately on the excitation current (flux) and the armature current (torque).

To illustrate how the vector control of synchronous machine, recall the above mentioned model of the machine:

$$\begin{aligned} \overline{v}_s &= R_s \overline{i}_s + p\overline{\psi}_s + j\dot{\theta}_m \overline{\psi}_s \\ \psi_{sd} &= L_d i_{sd} + \psi_{pm} \\ \psi_{sq} &= L_q i_{sq} \end{aligned}$$

$$p\dot{\theta}_m = \frac{n_p}{J}(T_e - T_r)$$

$$T_e = n_p \,\mathrm{Im}(\underline{i_s \overline{\psi}_s}) = n_p ((L_d - L_q) i_{sd} i_{sq} + \psi_{pm} i_{sq})$$

Substituting the stator flux components with their dependence on the stator currents and the permanent magnets flux (with the assumption that the machine parameters are time invariant) we have:

$$v_{sd} = R_s i_{sd} + L_d p i_{sd} - \dot{\theta}_m L_q i_{sq}$$
$$v_{sq} = R_s i_{sq} + L_q p i_{sq} + \dot{\theta}_m \psi_{pm} + \dot{\theta}_m L_d i_{sd}$$

and finally:

$$T_e = n_p \left[ (L_d - L_q) i_{sd} i_{sq} + \psi_{pm} i_{sq} \right]$$

## 6.5 Operating regions

By analyzing the expression of the torque you can see that the anisotropy of the rotor, represented by the difference ($L_d$ - $L_q$), produces a torque of anisotropy (or "reluctance torque"); in the following, unless otherwise specified, we will consider only the isotropic case of synchronous machine, so it results $L_d = L_q = L_s$.

In this case we have:

$$T_e = n_p \psi_{pm} i_{sq}$$

Looking at the last expression, it is evident therefore that, for the torque birth, only the quadrature component of stator current is effective, while direct component has no effect on the torque. So in order to minimize the magnitude of the current space phasor (and consequently the losses), vector control should operate on the power converter in such a way that is, at any time:

$$i_{sd} = 0$$

From the point of view of the torque control, therefore, the behaviour of a synchronous motor with vector control is similar to that of a DC machine: in this analogy, the quadrature component of stator current assumes the role of the armature current. Unlike a DC machine, however, with an $i_{sd}$ equal to zero, it is not possible to act on the excitation (weakening) because the excitation field is provided by permanent magnets.

Of course, in this case, $i_{sd}$ and $i_{sq}$ are currents flowing in fictitious windings (not real windings), but they are the components, on a reference frame fixed with rotor, of a current space phasor created by a three-phase system of currents: the transition from one system to the other is obtained by the space phasor formulas.

It is clear now the important function performed by the position sensor that appears in Figure 6-1, as it is the device that provides, to the control system, the instant value of the angular position of the rotor (and therefore the position of the permanent magnets flux $\psi_{pm}$), needed to realize the above mentioned transformation.

In steady state operation and with the machine powered by a three-phase symmetrical direct voltage (with a pulsating $\omega$ frequency), the mechanical speed $\dot{\theta}_m$ (also indicated by $\omega_m$) coincides with the frequency $\omega$. In the reference frame fixed with the rotor, therefore, the voltage phasor is stationary and constant: accordingly, all the electric quantities are constant and it results ($i_{sd}$ is set to 0 and the derivatives are null):

$$\dot{\theta}_m = \omega_m = \omega$$
$$v_{sq} = R_s i_{sq} + \omega_m \psi_{pm} = R_s i_{sq} + E$$
$$v_{sd} = -\omega_m L_s i_{sq}$$

One can thus draw the following vector diagram:

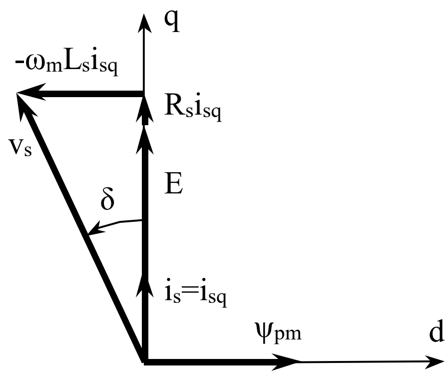


Figure 6-8: steady state vector diagram of an AC brushless

In the case in which the voltage drop on the resistance $R_s$ is negligible with respect to E, it is noted that, during a the variation of the mechanical speed, the angle δ remains constant (both catheti are proportional to the mechanical speed) and that the amplitude of the voltage vector increases linearly with the same speed. The speed to which the vector voltage reaches the maximum value given by the power supply (compatibly with the degree of insulation of the stator windings) is called base speed $\omega_{base}$. This limit, in the case of three-phase inverter supply, is represented by a circle whose radius is slightly smaller (necessary to maintain a voltage margin for the dynamic control of the current) than the radius of the circle inscribed in the hexagon of the field of operation of the inverter.

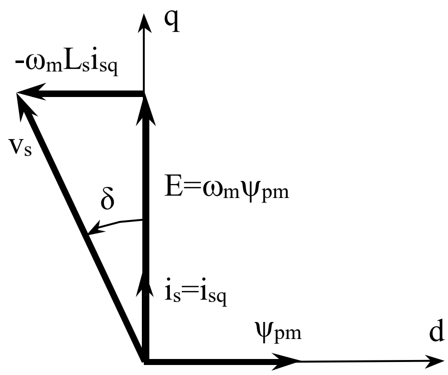


Figure 6-9: steady state vector diagram of an AC brushless (neglecting the resistive drop)

Suppose now that the machine is fed so as to produce a negative direct axis component of the current; in this case $i_{sd}$ <0 (no more $i_{sd}$ = 0); then, at steady state, the following relations are valid:

$$v_{sd} = R_s i_{sd} - \omega_m L_s i_{sq}$$
$$v_{sq} = R_s i_{sq} + \omega_m L_s i_{sd} + E$$

which implies the following diagram (neglecting the resistive drop):

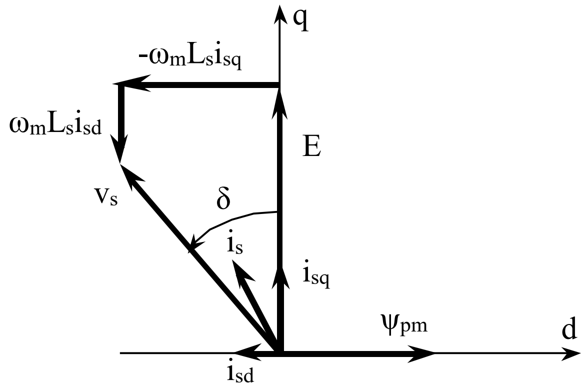


Figure 6-10: steady state vector diagram of an AC brushless when $i_{sd}<0$

From the diagram, it is evident that, for the same value of $i_{sq}$ (and then the torque), the addition of a negative component $i_{sd}$ led to a reduction of the stator voltage and an increase of the stator current. Having to keep the stator current within a predetermined maximum value imposed by the characteristics of the components of the power supply and by thermal effects, thus leads to the need to reduce the $i_{sq}$ (ie the torque) when $i_{sd}$ increases in its absolute value.

These observations suggest the following procedure:

- for speeds below the base one ($\omega_{base}$), it is suitable to put $i_{sd} = 0$ and the component $i_{sq}$ equal to the maximum allowed by the thermal constraints ($i_{max}$) so to operate in a region with constant torque, equal to the maximum possible (Figure 6-11 (a); increasing supply frequency ω, the emf E and the stator voltage increase and therefore the power converter has to supply this voltage. Once at the base speed, the voltage is equal to the maximum voltage supplied from power converter except for a certain margin necessary to control the current (Figure 6-11 (b));
- for speeds higher than the base one, it is not possible to maintain $i_{sq} = i_{max}$ because the further increase of E would require a stator voltage higher than the maximum available by the power converter: in this region, therefore, $i_{sq}$ has to be reduced to leave place to a negative value of $i_{sd}$ in order to validate the following inequality:

$$i_s = \sqrt{i_{sd}^2 + i_{sq}^2} \leq i_{\max}$$

Doing so it is verified that $v_s = v_{max}$ e $i_s \leq i_{max}$, but with an $i_{sq}$ (and then the torque) less than in cases where it is $i_{sd} = 0$.

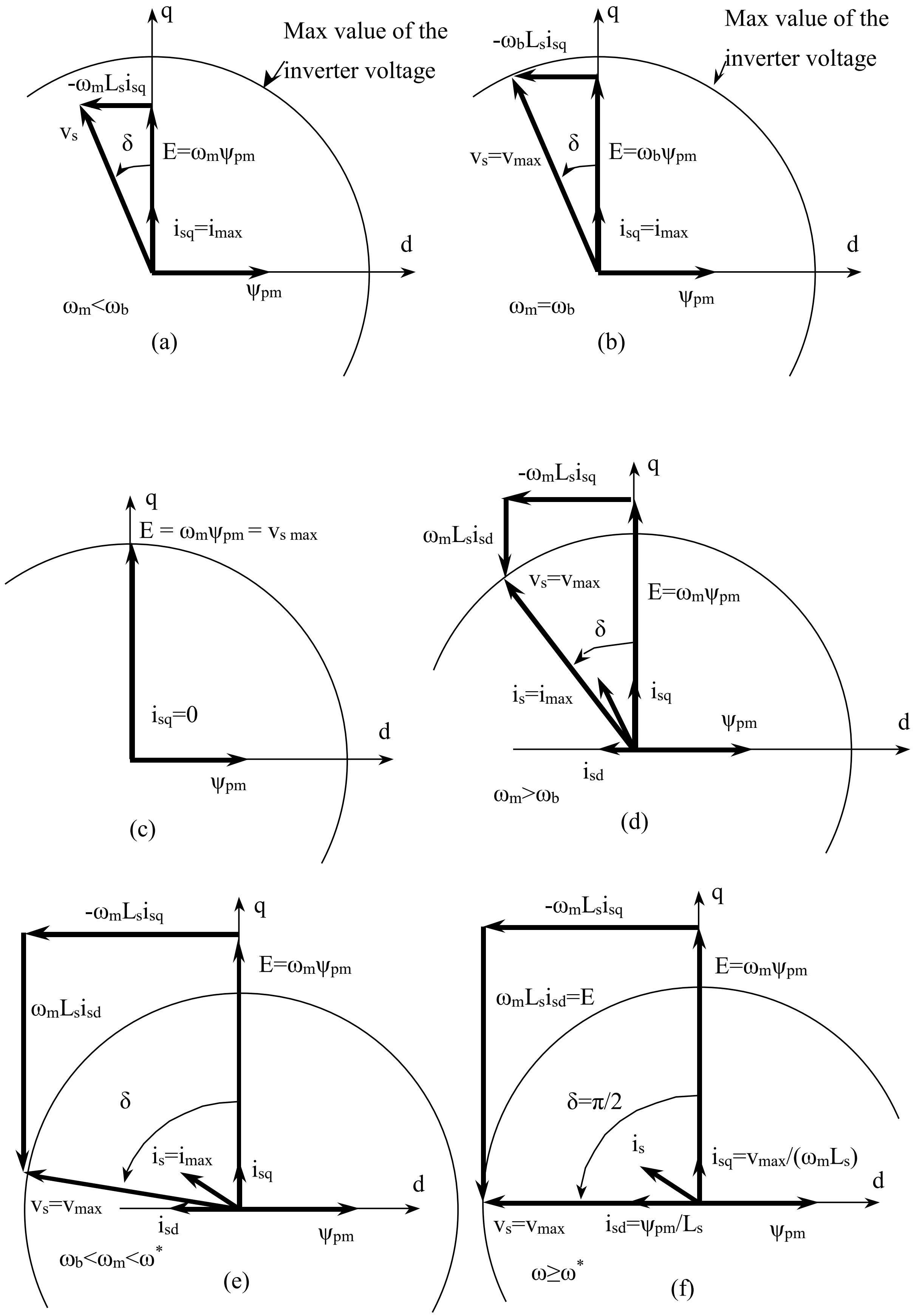


Figure 6-11: Vector diagram as a function of $i_{sd}$ values

You can see (Figure 6-11 (c)) that, in the extreme case of $E=v_{max}$, if $i_{sd} = 0$, $i_{sq}$ should be zero: this condition is characterized by a speed:

$$\omega' = \frac{E'}{\psi_{pm}} = \frac{v_{\max}}{\psi_{pm}}$$

with zero current and zero torque (no-load speed with $i_{sd}=0$).

On the contrary, with $i_{sd} <0$, we can get the situation shown in Figure 6-11 (d) where the machine is still able to develop a torque. In this condition, a kind of field weakening is realized (the component isd is opposite to the flux of the permanent magnets $\psi_{pm}$). The torque decreases as the value of $i_{sq}$ must decrease to respect the limits on the total magnitude of the stator current $i_{sq} = \sqrt{i_{\max}^2 - i_{sd}^2}$ .

Suitably sized motor (using permanent magnets with high coercive force and knee in the third quadrant, with an oversized air gap or internal magnets), you can request a current $i_{sd}$ equal to the ratio between $\psi_{pm}$ and $L_s$ (Figure 6-11 (e)) . The changing point is defined by the value of mechanical speed $\omega^*$. For $\omega>\omega^*$ the voltage $\omega L_s i_{sd}$ is equal to E regardless of mechanical speed. From the Figure 6-11 (f) and looking at the same amplitude of both horizontal axes, the current $i_{sq}$ is equal to $v_{max}/(\omega_m L_s)$ and then the torque is inversely proportional to the mechanical speed.

From this point on, the module of the stator current cannot be maintained at its maximum value but must decrease: $i_s = \sqrt{i_{sd}^2 + i_{sq}^2} = \sqrt{(\psi_{pm}/L_s)^2 + (v_{\max}/\omega_m L_s)^2} < i_{\max}$ .

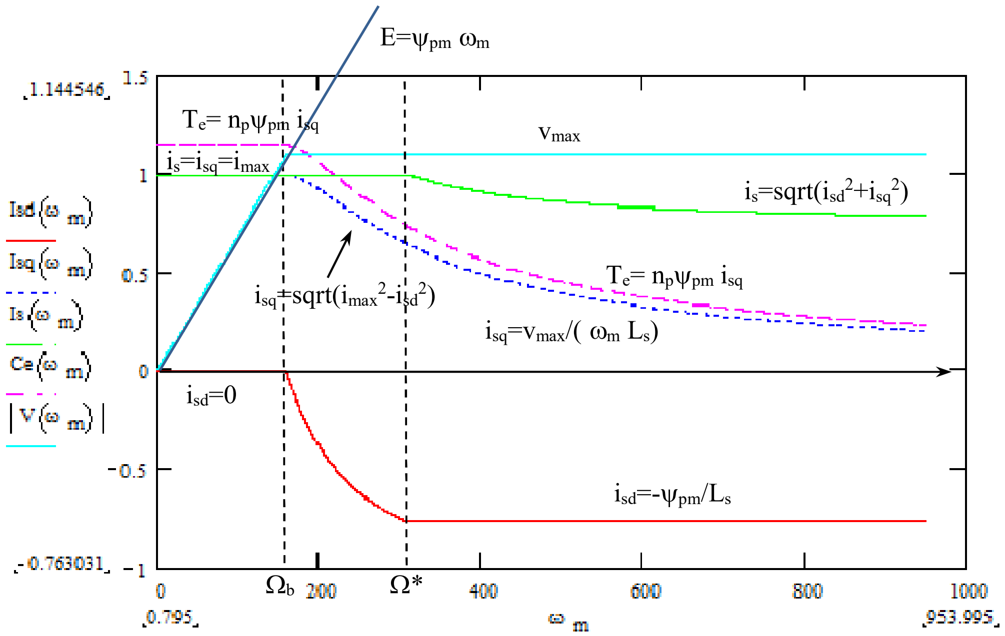


Figure 6-12: AC brushless operating regions

In practice, there are two main areas of application: the most classic of an axis control and the newer type for spindle applications. The first case does not require a speed much higher than the base speed and, therefore, the constant power region is very limited. It is a different story regarding the applications such as spindle. In these cases, the maximum speed can reach considerable values (about 6-8 times the base speed), making these drives a valid alternative to the drives based on induction machines (maximum speed of 5-6 times the base speed). It should be noted that, over the base speed, the emf E continues to grow. At 6 times the base speed E is 6 times the rated E (approximately 6 times the maximum voltage of the inverter). If, in these conditions, the control fails and the current $i_{sd}$ falls to zero, the voltage generated by the machine would destroy static switches, electrolytic capacitors and the winding insulation. It must provide the system with appropriate protections.

## 6.6 Control scheme

As seen in the previous section, the controller has to drive the power converter so that it always results $i_{sd} = i_{sd\ ref}$ (depending on the operating region or on a possible voltage controller), while the component $i_{sq}$ must be such that the produced torque is able to maintain the required speed of the machine. The overall scheme of the drive with speed control loop is the following:

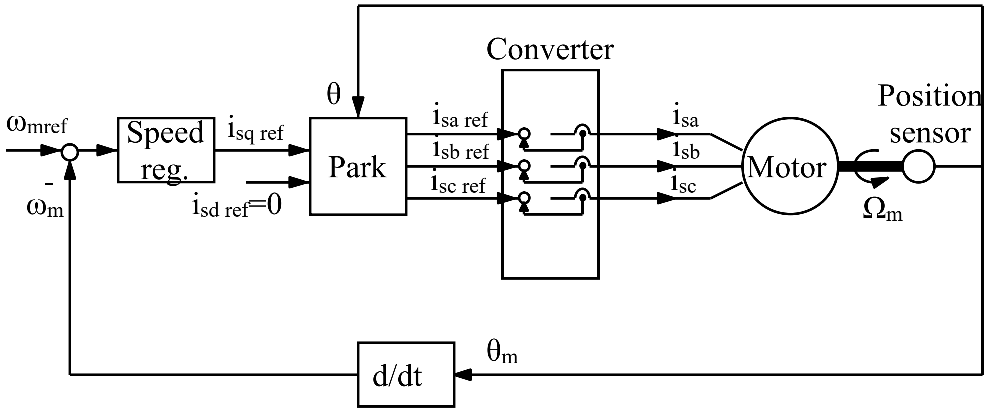


Figure 6-13: Speed control loop (current controlled power converter)

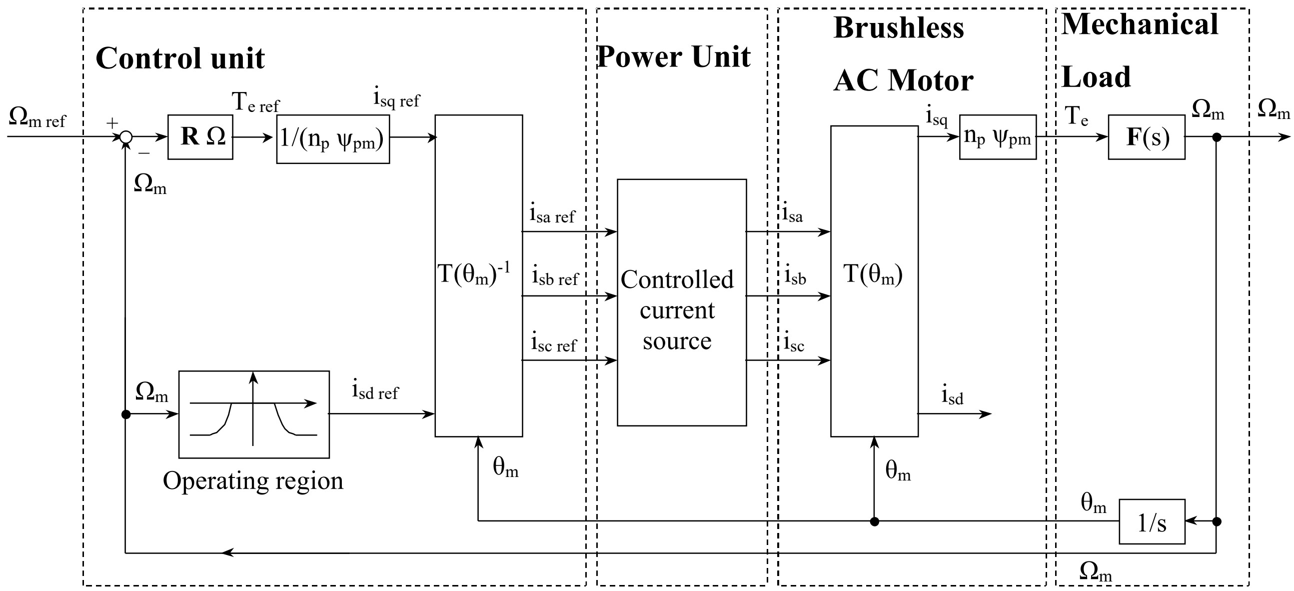


Figure 6-14: Full control scheme (current controller power converter)

In the Figure 6-14, the block $T(\theta_m)^{-1}$ performs the transformation from space phasor variables and phase quantities.

Any position control can be obtained by inserting an appropriate control loop outside of the speed control loop.

Figure 6-14 shows a power supply device with a fast current control that allows the motor currents to follow closely the reference values. Taking into account the stator dynamics (most frequent case), the control scheme changes, as shown in Figure 6-15. In this case the power converter has to establish a voltage reference, while the current regulation is now performed by the controller.

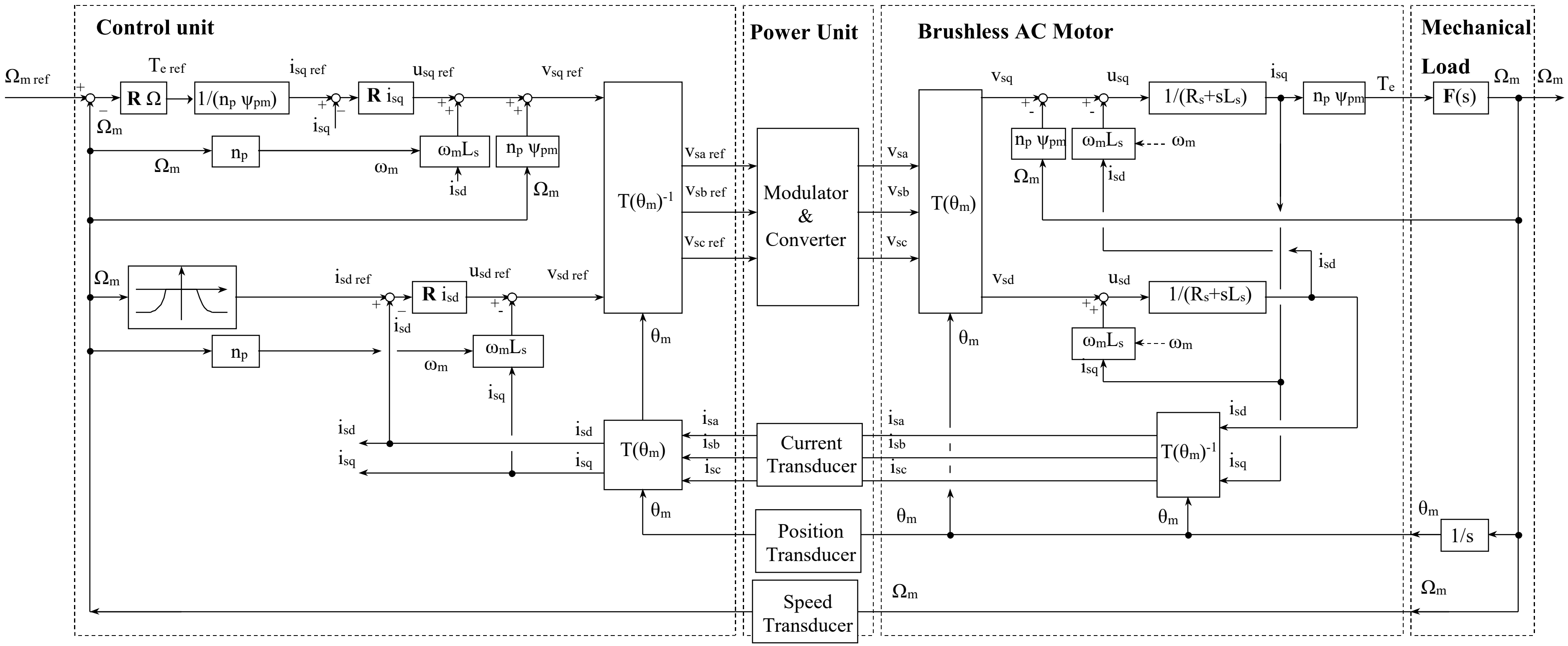


Figure 6-15: Full control scheme (voltage controlled power converter)

The structure of the current controller is based on the differential equations of the machine:

$$v_{sd} = R_s i_{sd} + L_s p i_{sd} - \omega_m L_s i_{sq}$$
$$v_{sq} = R_s i_{sq} + L_s p i_{sq} + \omega_m \psi_{pm} + \omega_m L_s i_{sd}$$

You notice immediately that, in terms of voltages on the two axes, there are different terms: $R_s i_{sd}+L_s p i_{sd}$ (indicated as $u_{sd}$) and $R_s i_{sq}+L_s p i_{sq}$ (indicated as $u_{sq}$) are the voltages that actually act on their own current; the terms $-\omega_m L_s i_{sq}$ and $\omega_m L_s i_{sd}$ show the existence of a coupling between the two loops; the term $\omega_m \psi_{pm}$ represents an electromotive force, proportional to the mechanical speed (it has full correspondence with the motional term in the dynamical equations of the armature current of a DC machine). So it is a good thing that the outputs of the current regulators are the reference values of the voltages $u_{sd}$ and $u_{sq}$. To obtain the reference value of stator voltages, simply add to $u_{s\,ref}$ the corresponding coupling terms and the motional term $\omega_m \psi_{pm}$.

## 6.7 Current controllers design

With all of the terms of compensation and, if it is reasonable to consider power converter and transducers as pure unity gain, the current regulators can easily be designed on the transfer function: $R_s+sL_s$. ($R_s+sL_d$ e $R_s+sL_q$ in the case of an anisotropic machine).

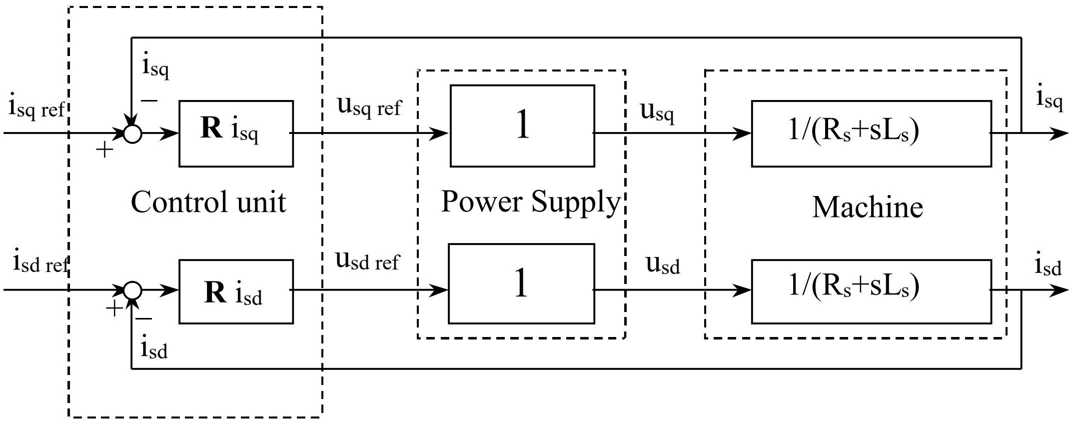


Figure 6-16: scheme for the design of the current controllers

If you do not compensate the coupling terms ($-\omega_m L_s i_{sq}$ and $\omega_m L_s i_{sd}$),, the loop controller should "work" even when the reference of the other loop varies. The motional term ($\omega_m \psi_{pm}$), however, would require full compensation through the integral action of the regulator itself; in the case of a machine starting from a speed different by zero, it would lead to some undesirable effects, such as, for example, a sudden break of the machine in the first instant of the operation.

For a better representation of the power supply, the unity gain could be replaced by a delay whose value is tied to the switching time of the converter switches. If the modulator is based on a PWM technique, for example, the delay between the instant of the reference variation and its implementation varies from 0 to the entire period of switching. On average, we can assume a delay equal to one half of the switching time.

## 6.8 Speed controller design

In order to design the speed regulator it is necessary to know the transfer function of the overall system, described before. The output of the speed controller is the desired value of the electromagnetic torque, which is proportional to the quadrature component of the stator current. The scheme is shown in Figure 6-17.

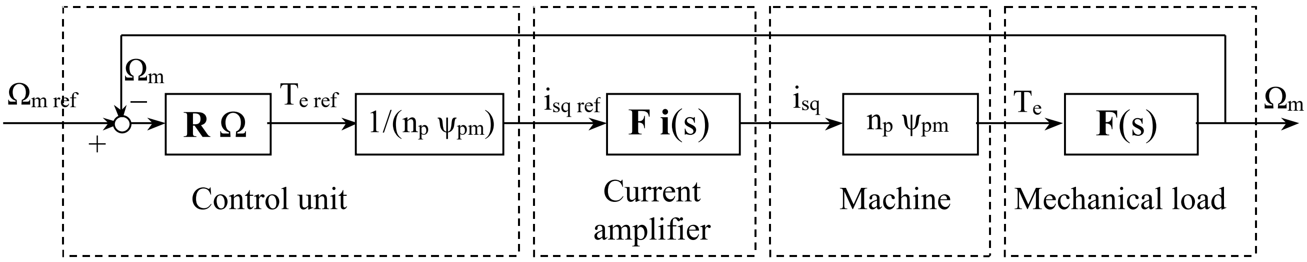


Figure 6-17: scheme for the design of the speed regulator

Again, you can follow two paths. If the bandwidth of the current regulator is very high, compared to that required to the speed control, it can be considered as an ideal current amplifier (unity gain). Otherwise, once designed the current regulator, the transfer function **F i(s)** is known. The speed controller must therefore be designed considering a system characterized by a transfer function equal to the product of **F i(s)** for the transfer function of mechanical load **F(s)**.

## 6.9 Synchronous reluctance machine (SYRM)

Figure 6-18 shows a synchronous reluctance machine with two poles (in Figure 6-20 a four poles machines is presented). It has a typical anisotropic structure of the rotor.

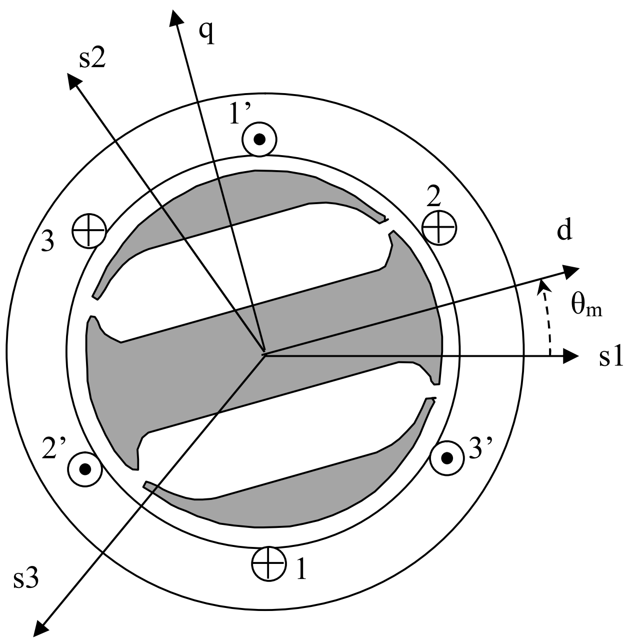


Figure 6-18: synchronous reluctance machine (two poles) $L_d >> L_q$

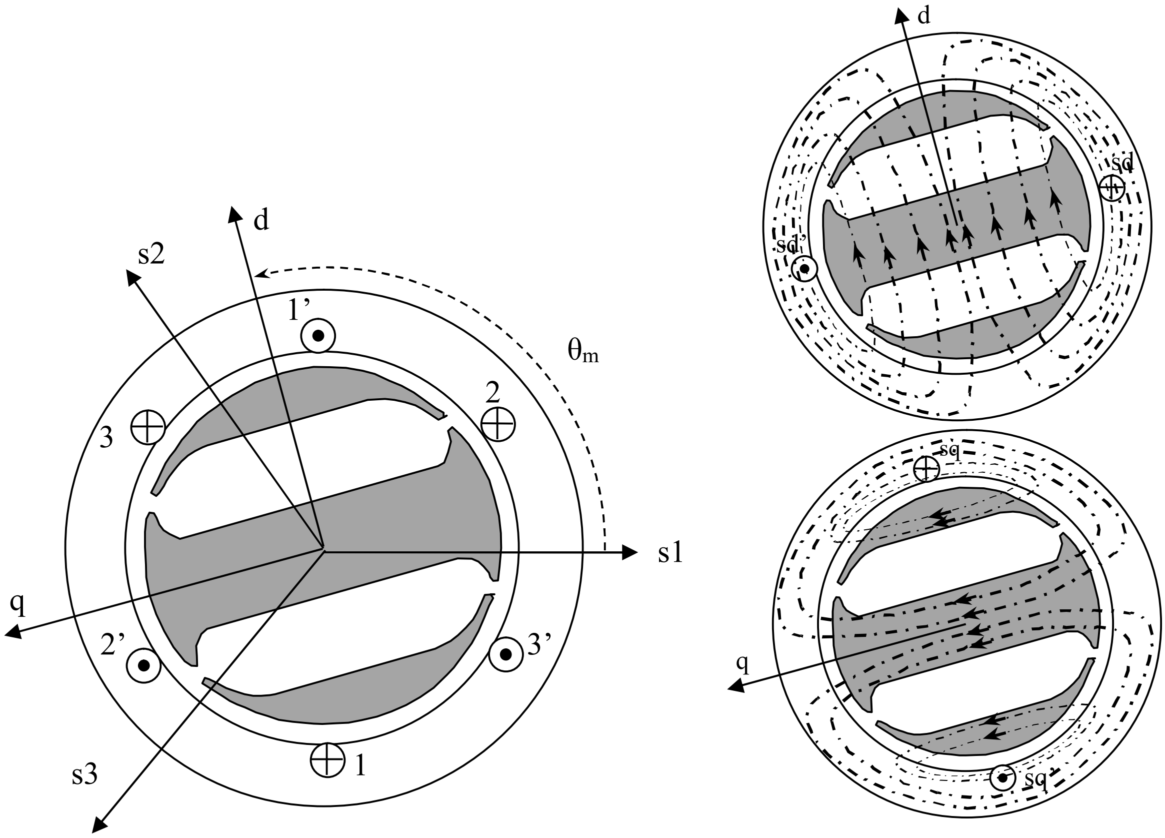


Figure 6-19: synchronous reluctance machine (two poles) with a different choice of the axes: $L_q$>>$L_d$; the flux lines due to the current $i_{sd}$ has to pass through large air gaps.

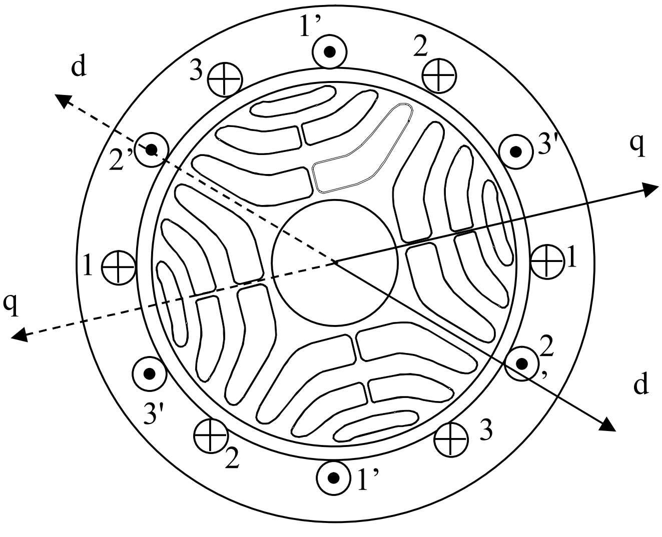


Figure 6-20: Example of a synchronous reluctance machine (four poles)

The stator is equal to the AC brushless stator but the rotor has no permanent magnets and it is designed in order to have a very high difference between the two inductances $L_d$ and $L_q$; the higher depends on the choice of the reference frame: in Figure 6-18 $L_d$>>$L_q$ while in Figure 6-19 $L_q$>>$L_d$. Henceforth the reference as in Figure 6-18 will be adopted.

The torque is only given by its reluctance term:

$$T_e = n_p \left[ (L_d - L_q) i_{sd} i_{sq} \right]$$

So it is necessary to feed the machine with both the $i_{sd}$ and $i_{sq}$ currents.

The electrical equations become:

$$v_{sd} = R_s i_{sd} + L_d p i_{sd} - \dot{\theta}_m L_q i_{sq}$$
$$v_{sq} = R_s i_{sq} + L_q p i_{sq} + \dot{\theta}_m L_d i_{sd}$$

There are different control strategies. One among the other comes from the fact that the time constant of the electric circuit on the "d" axis is higher than on the "q" axis ($L_d >> L_q$ implies $\tau_d >> \tau_q$).

So a constant value of $i_{sd}$ (equal to the rated value $i_{sdn}$) is performed, while $i_{sq}$ is used to control the torque. Above the base speed (due to the power converter), it is necessary to decrease $i_{sd}$.

The control scheme is very similar to the scheme of an AC brushless, shown in Figure 6-15.