1 Principles of vector orientation and vector orientated control structures for systems using three-phase AC machines

From the principles of electrical engineering it is known that the 3-phase quantities of the 3-phase AC machines can be summarized to complex vectors. These vectors can be represented in Cartesian coordinate systems, which are particularly chosen to suitable render the physical relations of the machines. These are the field-orientated coordinate system for the 3-phase AC drive technology or the grid voltage orientated coordinate system for generator systems. The orientation on a certain vector for modelling and design of the feedback control loops is generally called vector orientation.

1.1 Formation of the space vectors and its vector orientated philosophy

The three sinusoidal phase currents i_{su}, i_{sv} and i_{sw} of a neutral point isolated 3-phase AC machine fulfill the following relation:

$$i_{su}(t) + i_{sv}(t) + i_{sw}(t) = 0 \tag{1.1}$$

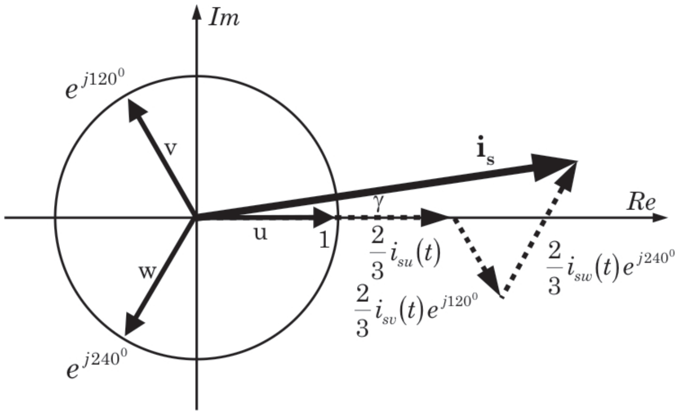

Fig. 1.1 Formation of the stator current vector from the phase currents

These currents can be combined to a vector $\mathbf{i}_s(t)$ circulating with the stator frequency f_s (see fig. 1.1).

$$\mathbf{i}_s = \frac{2}{3}\left[i_{su}(t) + i_{sv}(t)\,e^{j\gamma} + i_{sw}(t)\,e^{j2\gamma}\right] \text{ with } \gamma = 2\pi/3 \qquad (1.2)$$

The three phase currents now represent the projections of the vector $\mathbf{i}_s$ on the accompanying winding axes. Using this idea to combine other 3-phase quantities, complex vectors of stator and rotor voltages $\mathbf{u}_s$, $\mathbf{u}_r$ and stator and rotor flux linkages ψ_s, ψ_r are obtained. All vectors circulate with the angular speed ω_s.

In the next step, a Cartesian coordinate system with *dq* axes, which circulates synchronously with all vectors, will be introduced. In this system, the currents, voltage and flux vectors can be described in two components *d* and *q*.

$$\begin{aligned} \mathbf{u}_s &= u_{sd} + ju_{sq};\, \mathbf{u}_r = u_{rd} + ju_{rq} \\ \mathbf{i}_s &= i_{sd} + ji_{sq};\, \mathbf{i}_r = i_{rd} + ji_{rq} \\ \boldsymbol{\psi}_r &= \psi_{rd} + j\psi_{rq};\, \boldsymbol{\psi}_s = \psi_{sd} + j\psi_{sq} \end{aligned} \qquad (1.3)$$

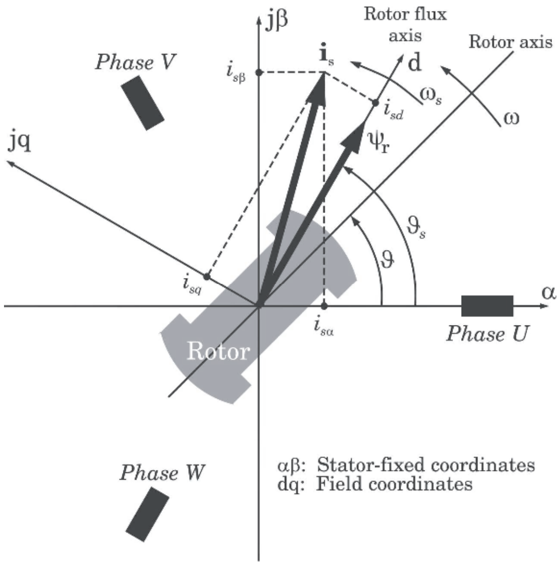

Fig. 1.2 Vector of the stator currents of IM in stator-fixed and field coordinates

Now, typical electrical drive systems shall be looked at more closely. If the real axis *d* of the coordinate system (see fig. 1.2) is identical with the direction of the rotor flux ψ_r (case IM) or of the pole flux ψ_p (case

PMSM), the quadrature component (q component) of the flux disappears and a physically easily comprehensible representation of the relations between torque, flux and current components is obtained. This representation can be immediately expressed in the following formulae.

- The induction motor with squirrel-cage rotor:

$$\psi_{rd}(s)=\frac{L_m}{1+sT_r}i_{sd}\ ;\quad m_M=\frac{3}{2}\frac{L_m}{L_r}z_p\,\psi_{rd}\,i_{sq} \tag{1.4}$$

- The permanentmagnet-excited synchronous motor:

$$m_M=\frac{3}{2}z_p\,\psi_p\,i_{sq} \tag{1.5}$$

In the equations (1.4) and (1.5), the following symbols are used:

m_M	Motor torque
z_p	Number of pole pairs
$\psi_{rd}, \psi_p = \psi_p$	Rotor and pole flux (IM, PMSM)
i_{sd}, i_{sq}	Direct and quadrature components of stator current
L_m, L_r	Mutual and rotor inductance
	with $L_r = L_m + L_{\sigma r}$ ($L_{\sigma r}$: rotor leakage inductance)
T_r	Rotor time constant with $T_r = L_r/R_r$ (R_r : rotor resistance)
s	Laplace operator

The equations (1.4), (1.5) show that the component i_{sd} of the stator current can be used as a control quantity for the rotor flux ψ_{rd}. If the rotor flux can be kept constant with the help of i_{sd}, then the cross component i_{sq} plays the role of a control variable for the torque m_M.

The linear relation between torque m_M and quadrature component i_{sq} is easily recognizable for the two machine types. If the rotor flux ψ_{rd} is constant (this is actually the case for the PMSM), i_{sq} represents the motor torque m_M so that the output quantity of the speed controller can be directly used as a set point for the quadrature component i_{sq}^*. For the case of the IM, the rotor flux ψ_{rd} may be regarded as nearly constant because of its slow variability in respect to the inner control loop of the stator current. Or, it can really be kept constant when the control scheme contains an outer flux control loop. This philosophy is justified in the formula (1.4) by the fact that the rotor flux ψ_{rd} can only be influenced by the direct component i_{sd} with a delay in the range of the rotor time constant T_r, which is many times greater than the sampling period of the current control loop. Thus, the set point i_{sd}^* of this field-forming component can be provided by the output quantity of the flux controller. For PMSM the pole flux ψ_p is maintained permanently unlike for the IM. Therefore the PMSM must be controlled such that the direct component i_{sd} has the value zero. Fig. 1.2 illustrates the relations described so far.

If the real axis d of the Cartesian dq coordinate system is chosen identical with one of the three winding axes, e.g. with the axis of winding u (fig. 1.2), it is renamed into $\alpha\beta$ coordinate system. A stator-fixed coordinate system is now obtained. The three-winding system of a 3-phase AC machine is a fixed system by nature. Therefore, a transformation is imaginable from the three-winding system into a two-winding system with α and β windings for the currents $i_{s\alpha}$ and $i_{s\beta}$.

$$\begin{cases} i_{s\alpha} = i_{su} \\ i_{s\beta} = \dfrac{1}{\sqrt{3}}\left(i_{su} + 2 i_{sv}\right) \end{cases} \tag{1.6}$$

In the formula (1.6) the third phase current i_{sw} is not needed because of the (by definition) open neutral-point of the motor.

Figure 1.2 shows two Cartesian coordinate systems with a common origin, of which the system with $\alpha\beta$ coordinates is fixed and the system with dq coordinates circulates with the angular speed $\omega_s = d\vartheta_s/dt$. The current $\mathbf{i}_s$ can be represented in the two coordinate systems as follows.

- In $\alpha\beta$ coordinates: $\mathbf{i}_s^s = i_{s\alpha} + j\, i_{s\beta}$
- In dq coordinates: $\mathbf{i}_s^f = i_{sd} + j\, i_{sq}$

(Indices: s - stator-fixed, f - field synchronous coordinates)

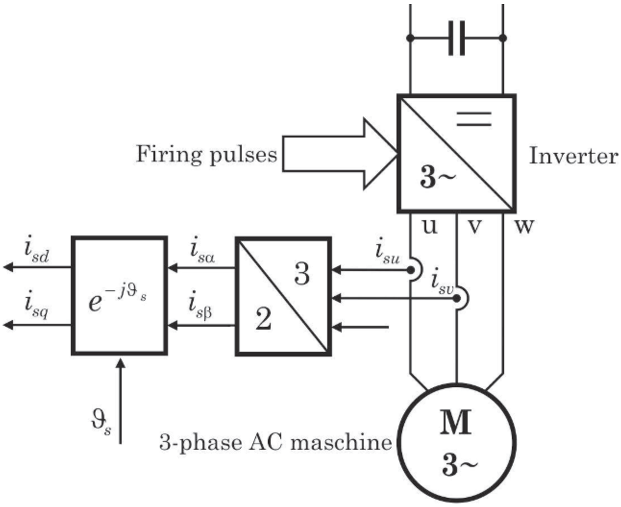

Fig. 1.3 Acquisition of the field synchronous current components

With

$$\begin{cases} i_{sd} = \ \ i_{s\alpha}\cos\vartheta_s + i_{s\beta}\sin\vartheta_s \\ i_{sq} = -i_{s\alpha}\sin\vartheta_s + i_{s\beta}\cos\vartheta_s \end{cases} \tag{1.7}$$

the stator current vector is obtained as:

$$\mathbf{i}_s^f = \left[i_{s\alpha}\cos\vartheta_s + i_{s\beta}\sin\vartheta_s \right] + j\left[i_{s\beta}\cos\vartheta_s - i_{s\alpha}\sin\vartheta_s \right]$$

$$\mathbf{i}_s^f = \left[i_{s\alpha} + j\, i_{s\beta} \right]\left[\cos\vartheta_s - j\sin\vartheta_s \right] = \mathbf{i}_s^s\, e^{-j\vartheta_s}$$

In generalization of that the following general formula results to transform complex vectors between the coordinate systems:

$$\mathbf{v}^s = \mathbf{v}^f\, e^{j\vartheta_s} \quad \text{or} \quad \mathbf{v}^f = \mathbf{v}^s\, e^{-j\vartheta_s} \tag{1.8}$$

v: an arbitrary complex vector

The acquisition of the field synchronous current components, using equations (1.6) and (1.7), is illustrated in figure 1.3.

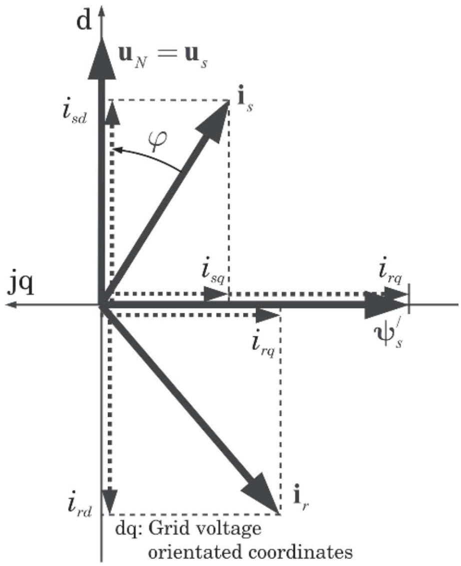

Fig. 1.4 Vectors of the stator and rotor currents of DFIM in grid voltage ($\mathbf{u}_N$) orientated coordinates

In generator systems like wind power plants with the stator connected directly to the grid, the real axis of the grid voltage vector $\mathbf{u}_N$ can be chosen as the *d* axis (see fig. 1.4). Such systems often use doubly-fed induction machines (DFIM) as generators because of several economic advantages. In Cartesian coordinates orientated to the grid voltage vector, the following relations for the DFIM are obtained.

- The doubly-fed induction machine:

$$\sin\varphi = \frac{|\psi_s|/L_m - i_{rq}}{|\mathbf{i}_s|};\ m_G = -\frac{3}{2} z_p \frac{L_m}{L_s} \psi_{sq} i_{rd} \tag{1.9}$$

In equation (1.9), the following symbols are used:

m_G	Generator torque
$\psi_{sq}, \boldsymbol{\psi}_s$	Stator flux
$\mathbf{i}_s$	Vector of stator current
i_{rd}, i_{rq}	Direct and quadrature components of rotor current
L_m, L_s	Mutual and stator inductance with $L_s = L_m + L_{\sigma s}$ ($L_{\sigma s}$: stator leakage inductance)
φ	Angle between vectors of grid voltage and stator current

Because the stator flux ψ_s is determined by the grid voltage and can be viewed as constant, the rotor current component i_{rd} plays the role of a control variable for the generator torque m_G and therefore for the active power P respectively. This fact is illustrated by the second equation in (1.9). The first of both equations (1.9) means that the power factor $\cos\varphi$ or the reactive power Q can be controlled by the control variable i_{rq}.

1.2 Basic structures with field-orientated control for three-phase AC drives

DC machines by their nature allow for a completely decoupled and independent control of the flux-forming field current and the torque-forming armature current. Because of this complete separation, very simple and computing time saving control algorithms were developed, which gave the dc machine preferred use especially in high-performance drive systems within the early years of the computerized feedback control. In contrast to this, the 3-phase AC machine represents a mathematically complicated construct with its multi-phase winding and voltage system, which made it difficult to maintain this important decoupling quality. Thus, the aim of the field orientation can be defined to re-establish the decoupling of the flux and torque forming components of the stator current vector. The field-orientated control scheme is then based on impression the decoupled current components using closed-loop control.

Based on the theoretical statements, briefly outlined in chapter 1.1, the classical structure (see fig. 1.5) of a 3-phase AC drive system with field-orientated control shall now be looked at in some more detail. If block 8 remains outside our scope at first, the structure, similar as for the case of a system with DC motor, contains in the outer loop two controllers: one for the flux (block 1) and one for the speed (block 9). The inner loop is formed of two separate current controllers (blocks 2) with PI behaviour for the field-forming component i_{sd} (comparable with the field current of the DC motor) and the torque-forming component i_{sq} (comparable with the armature current of the DC motor). Using the rotor flux ψ_{rd} and the speed ω, the decoupling network (DN: block 3) calculates the stator voltage components u_{sd} and u_{sq} from the output quantities y_d and y_q of the current controllers R_I. If the field angle ϑ_s between the axis d or the rotor flux axis and the stator-fixed reference axis (e.g. the axis of the winding u or the axis α) is known, the components u_{sd}, u_{sq} can be transformed, using block 4, from the field coordinates dq into the stator-fixed coordinates $\alpha\beta$. After transformation and processing the well known vector modulation (VM: block 5), the stator voltage is finally applied on the motor terminals with respect to amplitude and phase. The flux model (FM: block 8) helps to estimate the values of the rotor flux ψ_{rd} and the field angle ϑ_s from the vector of the stator current $\mathbf{i}_s$ and from the speed ω, and will be subject of chapter 4.4.

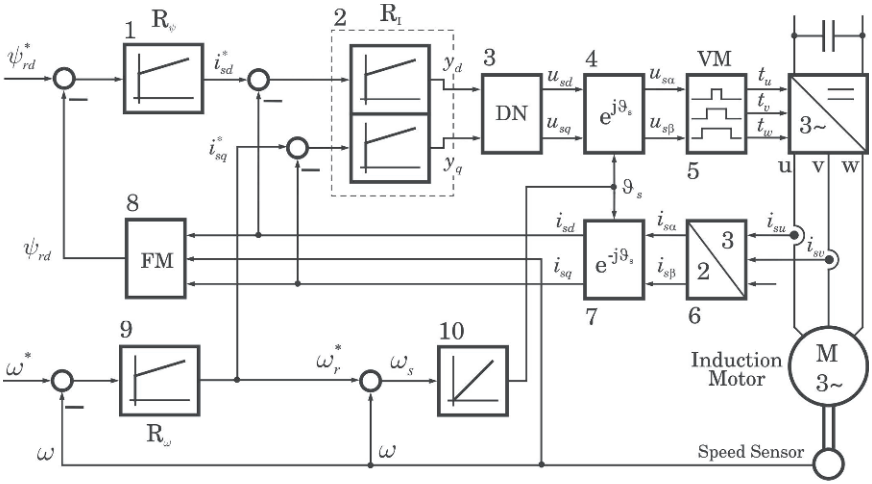

Fig. 1.5 Classical structure of field-orientated control for 3-phase AC drives using IM and voltage source inverter (VSI) with two separate PI current controllers for *d* and *q* axes

If the two components i_{sd}, i_{sq} were completely independent of each other, and therefore completely decoupled, the concept would work perfectly with two separate PI current controllers. But the decoupling network DN represents in this structure only an algebraic relation, which performes just the calculation of the voltage components u_{sd}, u_{sq} from the current-like controller output quantities y_d, y_q. The DN with this stationary approach does not show the wished-for decoupling behavior in the control technical sense. This classical structure therefore worked with good results in steady-state, but with less good results in dynamic operation. This becomes particularly clear if the drive is operated in the field weakening range with strong mutual influence between the axes d and q.

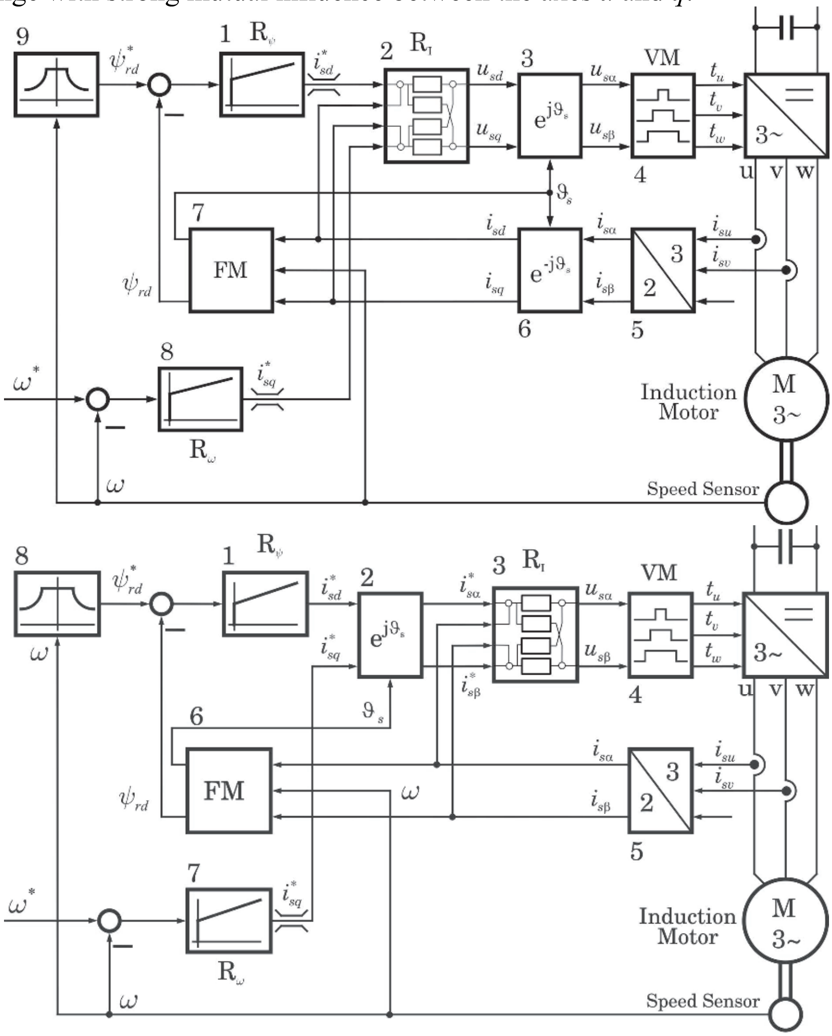

Fig. 1.6 Modern structures with field-orientated control for three-phase AC drives using IM and VSI with current control loop in field coordinates (**top**) and in stator-fixed coordinates (**bottom**)

In contrast to this simple control approach, the 3-phase AC machine, as highlighted above, represents a mathematically complicated structure. The actual internal *dq* current components are dynamically coupled with each other. From the control point of view, the control object „3-phase AC machine“ is an object with multi-inputs and multi-outputs (MIMO process), which can only be mastered by a vectorial MIMO feedback controller (see fig. 1.6). Such a control structure generally comprises of decoupling controllers next to main controllers, which provide the actual decoupling.

Figure 1.6 shows the more modern structures of the field-orientated controlled 3-phase AC drive systems with a vectorial multi-variable current controller $\mathbf{R}_I$. The difference between the two approaches only consists in the location of the coordinate transformation before or after the current controller. In the field-synchronous coordinate system, the controller has to process uniform reference and actual values, whereas in the stator-fixed coordinate system the reference and actual values are sinusoidal.

The set point ψ_{rd}^* for the rotor flux or for the magnetization state of the IM for both approaches is provided depending on the speed. In the reality the magnetization state determines the utilization of the machine and the inverter. Thus, several possibilities for optimization (torque or loss optimal) arise from an adequate specification of the set point ψ_{rd}^*. Further functionality like parameter settings for the functional blocks or tracking of the parameters depending on machine states are not represented explicitly in fig. 1.5 and 1.6.

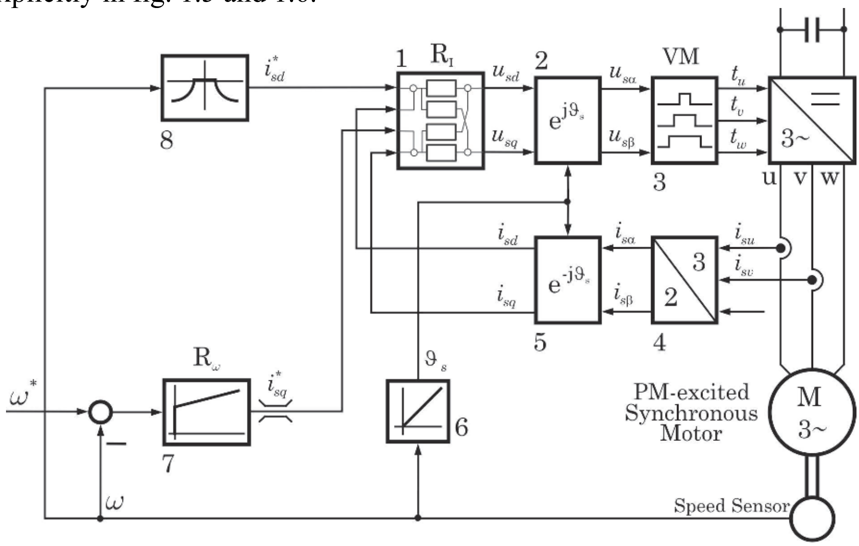

Fig. 1.7 Modern structure with field-orientated control for three-phase AC drives using PMSM and VSI with current control loop in field coordinates

PMSM drive systems with field-orientated control are widely used in practical applications (fig. 1.7). Because of the constant pole flux, the torque in equation (1.5) is directly proportional to the current component i_{sq}. Thus, the stator current does not serve the flux build-up, as in the case of the IM, but only the torque formation and contains only the component i_{sq}. The current vector is located vertically to the vector of the pole flux (fig. 1.8 on the left).

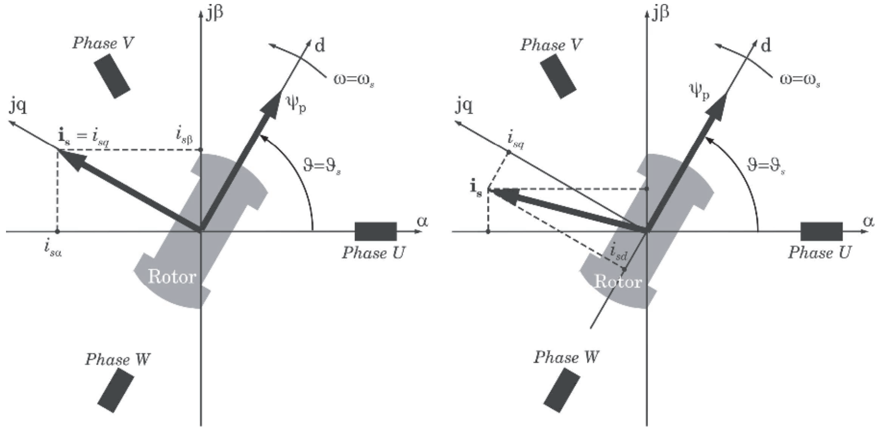

Fig. 1.8 Stator current vector $\mathbf{i}_s$ of the PMSM in the basic speed range (**left**) and in the field-weakening area (**right**)

Using a similar control structure as in the case of the IM, the direct component i_{sd} has the value zero (fig. 1.8 on the left). A superimposed flux controller is not necessary. But a different situation will arise, if the synchronous drive shall be operated in the field-weakening area as well (fig. 1.8 on the right). To achieve this, a negative current will be fed into the *d* axis depending on the speed (fig. 1.7, block 8). This is primarily possible because the modern magnets are nearly impossible to be demagnetized thanks to state-of-the-art materials. Like for the IM, possibilities for the optimal utilization of the PMSM and the inverter similarly arise by appropriate specification of i_{sd}. The flux angle ϑ_s will be obtained either by the direct measuring – e.g. with a resolver – or by the integration of the measured speed incorporating exact knowledge of the rotor initial position.

1.3 Basic structures of grid voltage orientated control for DFIM generators

One of the main control objectives stated above was the decoupled control of active and reactive current components. This suggests to choose the stator voltage oriented reference frame for the further control design. Let us consider some of the consequences of this choice for other variables of interest.

The stator of the machine is connected to the constant-voltage constant-frequency grid system. Since the stator frequency is always identical to the grid frequency, the voltage drop across the stator resistance can be neglected compared to the voltage drop across the mutual and leakage inductances L_m and $L_{\sigma s}$. Starting point is the stator voltage equation

$$\mathbf{u}_s = R_s \mathbf{i}_s + \frac{d\boldsymbol{\psi}_s}{dt} \quad \Rightarrow \quad \mathbf{u}_s \approx \frac{d\boldsymbol{\psi}_s}{dt} \text{ or } \mathbf{u}_s \approx j\omega_s \boldsymbol{\psi}_s \tag{1.10}$$

with the stator and rotor flux linkages

$$\begin{cases} \boldsymbol{\psi}_s = \mathbf{i}_s L_s + \mathbf{i}_r L_m \\ \boldsymbol{\psi}_r = \mathbf{i}_s L_m + \mathbf{i}_r L_r \end{cases} \tag{1.11}$$

Since the stator flux is kept constant by the constant grid voltage (equ. (1.10)) the component i_{rd} in equation (1.9) may be considered as torque producing current.

In the grid voltage orientated reference frame the fundamental power factor, or displacement factor $\cos\varphi$ respectively, with φ being the phase angle between voltage vector $\mathbf{u}_s$ and current vector $\mathbf{i}_s$, is defined according to figure 1.4 as follows:

$$\cos\varphi = \frac{i_{sd}}{|\mathbf{i}_s|} = \frac{i_{sd}}{\sqrt{i_{sd}^2 + i_{sq}^2}} \tag{1.12}$$

However, it must be considered that according to equation (1.11) for near-constant stator flux any change in $\mathbf{i}_r$ immediately causes a change in $\mathbf{i}_s$ and consequently in $\cos\varphi$. To show this in more detail the stator flux in equation (1.11) can be rewritten in the grid voltage oriented system to:

$$\begin{cases} \psi_{sd}^{/} = \frac{L_s}{L_m} i_{sd} + i_{rd} \approx 0 \\ \psi_{sq}^{/} = \frac{L_s}{L_m} i_{sq} + i_{rq} \approx |\boldsymbol{\psi}_s^{/}| \end{cases} \quad \text{with} \quad \boldsymbol{\psi}_s^{/} = \boldsymbol{\psi}_s / L_m \tag{1.13}$$

For $L_s/L_m \approx 1$ equation (1.13) may be simplified to:

$$\begin{cases} i_{sd} + i_{rd} \approx 0 \\ i_{sq} + i_{rq} \approx \left|\boldsymbol{\psi}_s^{/}\right| = \psi_{sq}^{/} \end{cases} \tag{1.14}$$

The phasor diagram in figure 1.4 illustrates the context of (1.14). With the torque producing current i_{rd} determined by the torque controller according to (1.13) the stator current i_{sd} is pre-determined as well. To compensate the influence on $\cos\varphi$ according to equation (1.12) an appropriate modification of i_{sq} is necessary. The relation between the stator phase angle φ and i_{sq} is defined by:

$$\sin\varphi = \frac{i_{sq}}{\left|\mathbf{i}_s\right|} = \frac{i_{sq}}{\sqrt{i_{sd}^2 + i_{sq}^2}} \tag{1.15}$$

Equation (1.15) expresses a quasi-linear relation between $\sin\varphi$ and i_{sq}, for small phase angles directly between φ and i_{sq} because of $\sin\varphi \approx \varphi$ in this area. This implies to implement a $\sin\varphi$ control rather than the $\cos\varphi$ control considered initially. Due to the fixed relation between i_{sq} and i_{rq} expressed in the second equation of (1.14) the rotor current component i_{rq} is supposed to serve as $\sin\varphi$- or $\cos\varphi$-producing current component. Another advantage of the $\sin\varphi$ control is the simple distinction of inductive and capacitive reactive power by the sign of $\sin\varphi$.

The DFIM control system consists of two parts: Generator-side control and grid-side control. The generator-side control is responsible for the adjustment of the generator reference values: regenerative torque m_G and power factor $\cos\varphi$. For these values suitable control variables must be found. It was worked out in the previous section, that in the grid voltage reference system the rotor current component i_{rd} may be considered as torque producing quantity, refer to equation (1.9). Therefore, if the generator-side control is working with a current controller to inject the desired current into the rotor winding, the reference value for i_{rd} may be determined by an outer torque control loop.

With this context in mind the generator-side control structure may be assembled now like depicted in figure 1.9. Assuming a fast and accurate rotor current vector control this control structure enables a very good decoupling between torque and power factor in both steady state and dynamic operation. With a fast inner current control loop, torque and power factor might be impressed almost delay-free; the controlled systems for both values have proportional behaviour.

However, in practical implementation measurement noise and current harmonics might cause instability due to the strong correlation of the signals in both control loops. Feedback smoothing low-pass filters are necessary to avoid such effects (fig. 1.9). These feedback filters then form the actual process model and the control dynamics has to be slowed down.

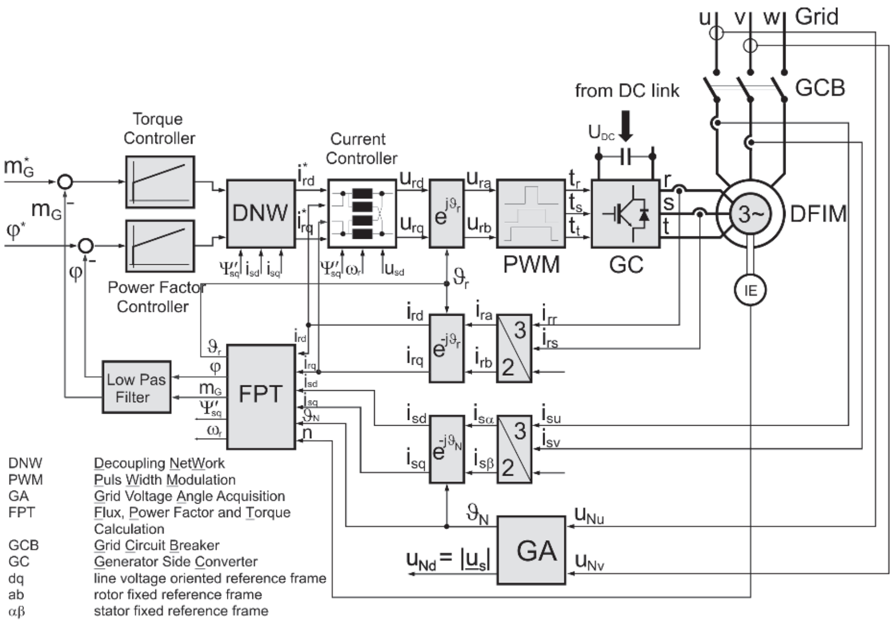

Fig. 1.9 Modern structure with grid voltage orientated control for generator systems using DFIM and VSI with current control loop in grid voltage coordinates

The DFIM is often used in wind power plants thanks to the fundamentally smaller power demand for the power electronic components compared to systems with IM or SM. The demand for improved short-circuit capabilities (ride-through of the wind turbine during grid faults) seems to be invincible for DFIM, because the stator of the generator is directly connected to the grid. Practical solutions require additional power electronics equipment and interrupt the normal system function. Thanks to the power electronic control equipment between the stator and the grid, this problem does not exist for IM or SM systems.

Figure 1.10 presents a nonlinear control structure, which results from the idea of the exact linearization and contains a direct decoupling between active and reactive power. However, the most important advantage of this concept consists of the improvement of the system performance at grid faults, which allows to maintain system operation up to higher fault levels.

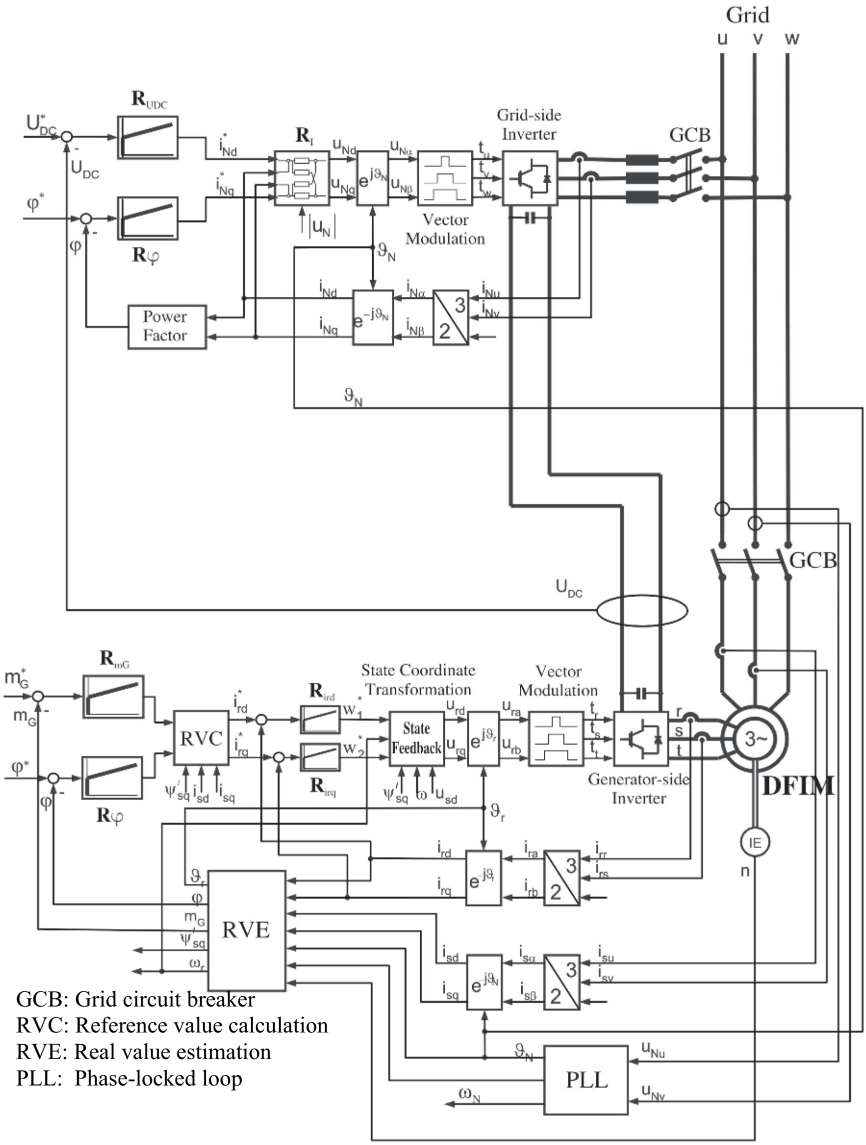

Fig. 1.10 Complete structure of wind power plant with grid voltage orientated control using a nonlinear control loop in grid voltage coordinates

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