9

Electrical Drives and Power Electronics

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Power electronics plays a key role in energy efficiency in electrical systems. Power electronics is the enabling technology for renewable energy sources, such as photovoltaic, wind energy and fuel cell systems, and others, where the injected energy is subjected to strong regulation in order not to perturb the grid.

In distribution, power electronics is becoming more and more present. The inclusion of HVDC systems and STATCOMs are two examples of power electronics based devices that are present in the distribution system.

Also in energy storage systems, e.g. batteries, flywheels and SMES, power electronics is the technology that makes the system run. In addition, power electronics is present in each piece of electrical or electronic device or equipment, whether it is portable or not.

The role of power electronics is to control energy flow, while transforming the energy. Of course, this must act according to electrical specifications, and also with high efficiency. Nowadays, power electronic and control devices can meet these two objectives at the same time.

Electrical drives are one of the major energy-consuming devices used in industry. Transforming electrical energy to mechanical energy is a necessary process in most industrial processes, e.g. manufacturing, pumps, fans and compressors in HVACs.

Motor driven systems use 30% of the electrical energy in Europe and make up 60% of the industrial use of electric energy [1]. These figures show how important electrical drives are in terms of energy savings. A large number of studies and initiatives are currently running with this objective. The Motor Challenge Programme (MCP) is an initiative promoted by the European Commission to help industries to improve energy efficiency in motor driven systems; it has had great success in the last years.

In motor driven applications, energy is the main cost of the system, more than the initial cost and maintenance, so a reduction in energy consumption is a must both for energy and financial savings.

Electrical Energy Efficiency: Technologies and Applications, First Edition. Andreas Sumper and Angelo Baggini.

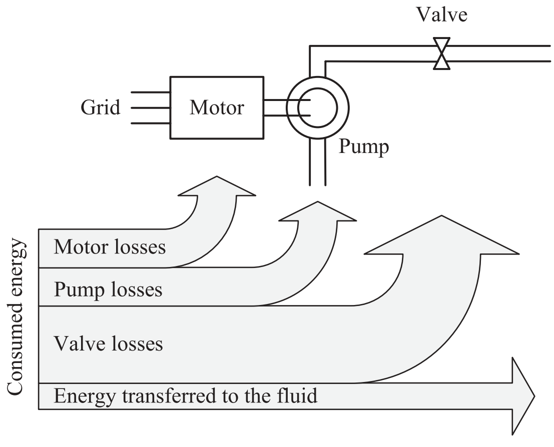

Figure 9.1 Energy losses in a traditional pumping system

Figures 9.1 and 9.2 illustrate an example of energy saving in the electrical drives of a pumping system. Traditional systems (Figure 9.1) use valves to control flow rate. At first, the pump transfers the maximum energy to the fluid using a direct connection of the motor to the grid, and then, to adjust the flow rate, a valve is used to dissipate the excess energy. This is not an efficient system.

In order to increase efficiency, it seems more reasonably to transfer only the necessary amount of energy to the fluid, as shown in Figure 9.2. However, pumps and fans can do this only using a varying rotational speed for the pump or the fan. This can only be done by modifying the rotational speed of the motor. Because the grid has fixed frequency and amplitude, a new interface element must be inserted between the grid and the motor; this provides the needed frequency and voltage amplitude for the motor. This element is the VSD.

A variable speed drive is a power electronics device that allows energy control of the grid to the motor and, consequently, to the fluid. The VSD can adjust the frequency and the voltage

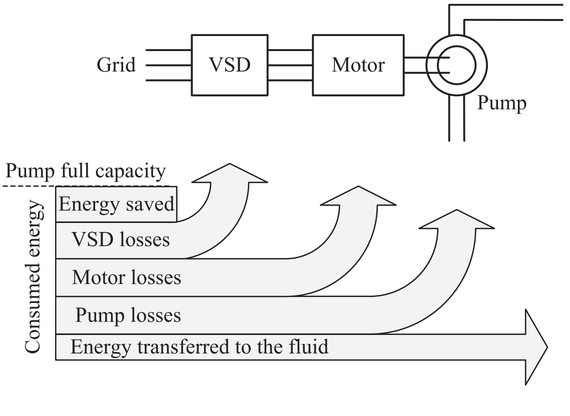

Figure 9.2 Energy savings in a pumping system using a variable speed drive (VSD)

amplitude needed for the system in real-time, transferring only the necessary amount of energy to the fluid.

Available variable speed drives are more than just systems to change the frequency and amplitude of the voltage applied to the motor. They are used as protection devices for the motor; they act as a soft starter, reducing stress on input lines and transformers, and avoiding an oversized design.

In addition, VSDs can act as decentralized control systems because normally they include some automation features, including analogue or digital communications buses.

Adding a new element can increase the cost of the system, and payback analysis must be used in this type of application. The payback is strongly dependent on duty cycle and energy cost, and it can vary from 8 months to 2 years. Even in the worst case of 2 years, this technology can give significant operational cost savings and, of course, a reduction in CO_2 emissions.

The induction motor has been the workhorse for industrial applications for many years. The advantages of induction motors over other types of motors are well known. The first reason is the cost, because of the simple design, simple manufacturing process and the low cost of materials used. Another reason is that it does not need any control devices, and it can be directly connected to the grid, giving the solution as shown in Figure 9.1.

Affordable power electronics and variable speed drives are a ‘novelty’ of the last 20 years. And, in the last 20 years new motor technologies with power electronics have appeared.

Permanent magnet synchronous motors (PMSM) are synchronous motors with magnets in the rotor instead of windings as traditional synchronous motors. The presence of these permanent magnets gives some advantages over induction motors. PMSM are more efficient than induction motors because they do not require magnetizing current [2]. Because of the lack of windings in the rotor, there are no rotor losses, so thermal design of the rotor is much simpler than that of an induction motor. In addition, the rotor can be made with lower inertia, so increasing the ratio of torque to inertia. This is the reason that this motor is the workhorse in such high dynamic applications as servos and robotics.

However, there are some disadvantages. The relative high cost of the permanent magnets, and the need for position sensors and power electronics. The PMSM cannot be directly connected to the grid because of instability at relative low operational speeds, below the nominal speed [3].

Increasing the energy efficiency of the variable speed drives can be achieved in different ways. The first is the control method used. Different control methods can lead to different energy efficiencies with the same performances, so the most efficient control technique must be used. However, high energy efficient control solutions demand a much higher computational cost.

From the converter point of view, there are three main topics that influence system efficiency: first, converter topology, second, modulation technique and, third, semiconductor devices.

The topology of the converter can influence drive efficiency. Some topologies are able to reduce current harmonics in the motor, so reducing losses and torque ripple, and also reducing stress on semiconductor devices, or allowing for lower voltage semiconductors, which normally are more efficient. Multilevel converters and matrix converters are key topologies.

Generating a high-quality sinusoidal voltage to be applied to the motor is the task of the modulator. Different modulation techniques exist and, depending on the modulation technique used, the efficiency of the motor and the converter can be improved.

In power electronic devices, the switching devices are semiconductors that act as on–off switches. Operation in the linear region of the semiconductor is not allowed because of high losses. Even accounting for the lower losses of a semiconductor working at saturation, during conduction and switching there are some losses that must be taking into account. These losses cause the semiconductor to heat to destruction. In this case, as in the motor, the limit is a thermal limit, and a heat removal system must be used. In this sense, semiconductor research focuses on reducing conduction and switching losses, but also on increasing the working temperature of semiconductor devices, reducing the heat removal system energy consumption, and also increasing the drive efficiency.

The most used material in semiconductors is silicon, and different technologies for silicon semiconductor exist. New materials, such as SiC (silicon carbide), GaN (gallium nitride) and even diamond, which have lower losses and a higher working temperature capability, are appearing and these will be the future of semiconductor devices.

Electrical drives usually work in motor mode. However sometimes, and depending on the application, they can work in braking mode. The braking is done electrically: it means that the braking energy is transferred back from the mechanical load to the motor, and converted back into electrical energy.

Variable speed drives must handle this energy. Usually, this energy is dissipated into resistors, but new technologies allow this energy to be stored in supercapacitors or batteries, or fed back into the grid, or even used by another drive that is working in motor mode.

This chapter focuses on all these improvements in drives.

9.1 Control Methods for Induction Motors and PMSM

There are several control methods for induction and PMSM, but only three are considered here because of its importance in industrial applications. The V/f or scalar control is based on steady-state electrical equations of the motor. This method is mainly used in applications that do not require high dynamic performance, such as, HVAC, pumps and fans, because it directly controls the torque developed by the motor.

For applications with high dynamics such as servos and robotics, vector control or DTC methods are more useful. These methods are based on state space motor equations, and they offer torque control. These high performances come at a cost: they need higher computation and sensing effort.

9.1.1 V/f Control

The V/f control is based on the equivalent per phase model of an induction motor, as shown in Figure 9.3, operating at steady state.

In the equivalent circuit of Figure 9.3, the torque can be expressed as:

$$T_{\text{em}} = \frac{P_{\text{m}}}{\omega_{\text{m}}} = 3I_{\text{r}}'\frac{1-s}{s}R_{\text{r}}'. \tag{9.1}$$

The motor equations are

$$u_{\text{s}} = R_{\text{s}}i_{\text{s}} + j\omega_{\text{s}}\lambda_{\text{s}}, \tag{9.2}$$

where u_{s} is the applied voltage, R_{s} is the stator resistance, i_{s} is stator current, ω_{s} is the stator voltage frequency and λ_{s} is the stator flux linkage.

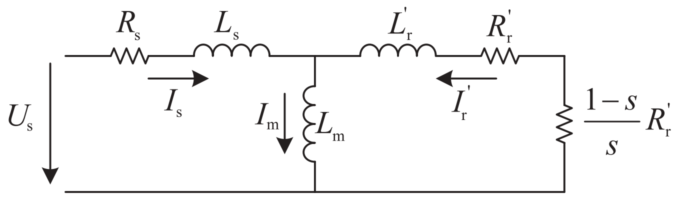

Figure 9.3 Per phase equivalent circuit of an induction motor

From (9.2), the normalized stator magnitude can be calculated as

$$u_s = \sqrt{(R_s i_s)^2 + (\omega_s \lambda_s)^2} \tag{9.3}$$

and, neglecting the stator resistance and for a constant stator flux linkage, equation (9.3) can be expressed as

$$\frac{u_s}{\omega_s} = \lambda_s = ct. \tag{9.4}$$

In the V/f control, this flux linkage λ_s is supposed constant on the whole speed range up to the base speed (ω_{sb}).

The applied voltage versus the applied frequency is a linear relation, and can be seen in Figure 9.4. Above base or rated speed, the voltage is then maintained constant while the frequency increases because the voltage capability of the converter is limited.

At low speed, the resistive term in (9.3) cannot be neglected and, to maintain a constant flux, some extra voltage must be applied to the motor as shown in Figure 9.4. This extra voltage depends on the current level (load) of the motor, as shown in (9.3).

Figure 9.5 shows a block diagram of the V/f control scheme. The control algorithm computes the voltage amplitude, proportional to the desired frequency. Some resistive drop compensation can be added at this point, but current measurements must be made. The voltage magnitude and frequency are used in the modulator to synthesize the switching signals for the voltage source inverter.

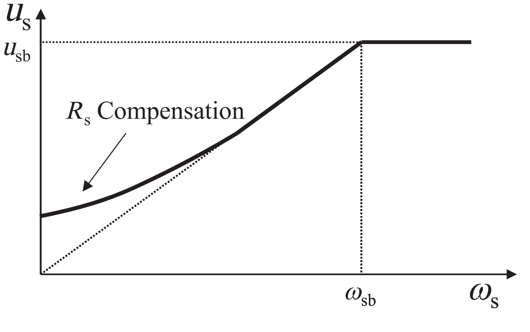

Figure 9.4 Voltage vs. frequency in V/f control method. The resistive drop compensation can also be seen

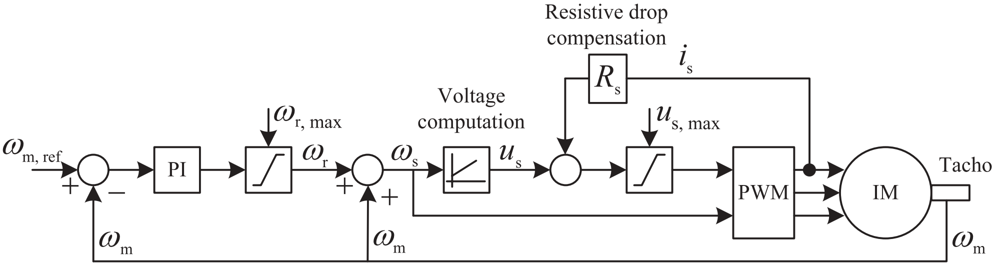

Figure 9.5 Block diagram of the V/f control method

The direct connection of a motor to the grid is a particular case of V/f control. But, by adding a VSD, this particular case can be extended to the whole speed range, adding features as soft starting, protection, and speed regulation.

These features, added to the easy implementation of this control method, have led to the great success of the V/f control method and it is now one of the most used control method for induction motors.

In the case of the induction motor, because it is an asynchronous machine, the mechanical and stator voltage speeds are not the same and depend on load. The slip ratio, *s*, is the relation between mechanical speed and stator voltage speed, and can be defined as

$$s = \frac{\omega_s - \omega_m}{\omega_s} = \frac{\omega_r}{\omega_s}, \tag{9.5}$$

where ω_m is the mechanical speed. Figure 9.6 shows the torque-speed characteristic of an induction motor. To assure the operation in the stable region of this curve, limiting the maximum slip, a PI controller can be added in an external loop as shown in Figure 9.5. This controller assures not only the stable operation and exact desired speed, but can act as current protection for the motor because rotor speed and current (or torque) are related, as shown in equation (9.3) and Figure 9.6.

As shown in Figure 9.5, a voltage limiter must be used because of the limited voltage of the converter, as will be seen in next sections, but this method needs a speed sensor.

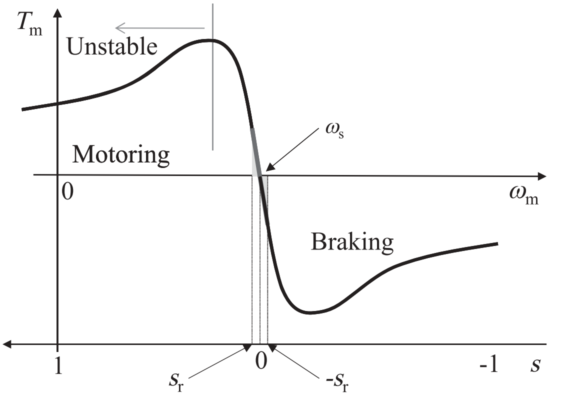

Figure 9.6 Torque vs. speed characteristic of an induction motor under V/f control

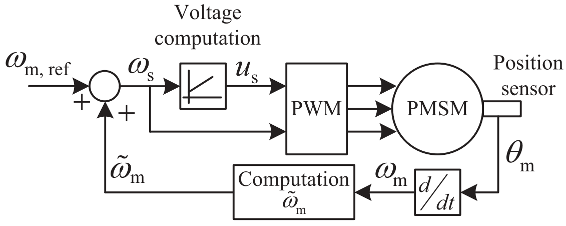

Figure 9.7 Stabilization loop with speed measurement for PMSM

In the case of PMSM, the V/f control strategy cannot be applied directly because of instability at low speeds [4]. The use of PMSM with V/f control requires the use of a squirrel cage in the rotor, combined with permanent magnets [5], or the measurement of the rotor mechanical speed in order to synchronize stator currents and rotor position, as seen in Figure 9.7, but this solution increases the cost and reduces reliability because of the speed sensor.

In [3], a sensorless stable V/f control is proposed. The method, instead of using speed perturbations to stabilize the motor, uses the power perturbations. The power can be computed using motor phase current measurements. This method shows a good response at nominal speed with nominal torque step changes, but shows a bad performance at low speeds.

9.1.1.1 Pumping, Fans and HVAC Applications

Continuous running applications such as pumps, fans and HVAC applications are at the focus of energy saving. Sixty-three percent of energy used in motors is in compressors, ventilation and pumping [1]. The energy cost of these applications is around 75% of the total cost, including investment and maintenance. The Motor Challenge Programme of the European Commission demonstrates that 30% of the energy use in pumping, fans and HVAC applications can be saved using VSD.

Typical installations are designed for the worst-case load conditions (maximum flow), and they only occur 5% of the time. Flow control is normally achieved using an output control valve. However, as can be observed in Figure 9.8, the energy savings using a VSD can be high. As shown in Figure 9.8, pumps and fans are quadratic torque loads with rotating speed.

From (9.1), it can be obtained that the $T_{em} \propto V^2/f^2$, but, as said, the load characteristic is $T_{pump} \propto f^2$. Obviously, at the constant speed operating point, the two torques are equal. Then, the VSD must be programmed to follow a $V \propto f^2$ law, instead of $V \propto f$.

As can be seen in Figure 9.9, in valve control applications, the desired flow rate is obtained by increasing the head of the pump or fan (operating point O_1). This is because the pump or fan characteristics cannot be modified (operation at fixed rotational speed), and the flow rate can only be modified by creating a fictive head (or losses) with the valve. The transferred power to the fluid (*Head* $\times$ *Flow rate*) is the dashed area below point O_1.

By using a VSD, the characteristic of the pump or fan can be moved to the desired flow rate by modifying the rotational speed from n_1 to n_2 (operating point O_2). Now, the transferred power to the fluid is just the power needed, increasing the efficiency of the VSD-Motor-Pump (or fan)-tube system.

Figure 9.9 also shows that as low as the flow rate is, the energy savings are higher.

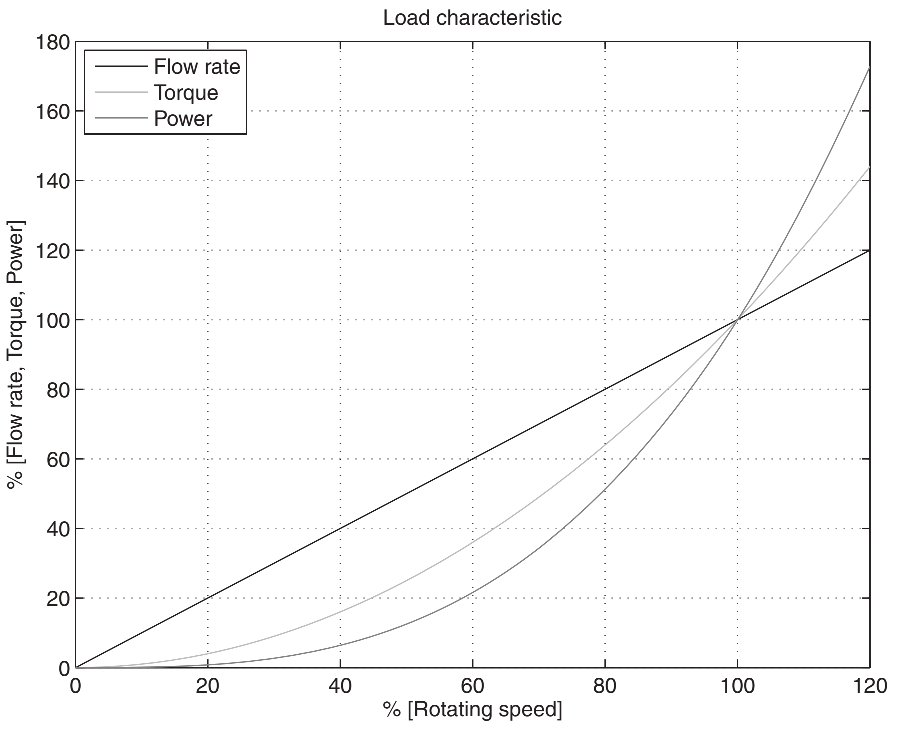

Figure 9.8 Load characteristics of a fan or pumping system. It must be noted that at 60% of rated speed, the torque is 35%, and the power is 20% of rated

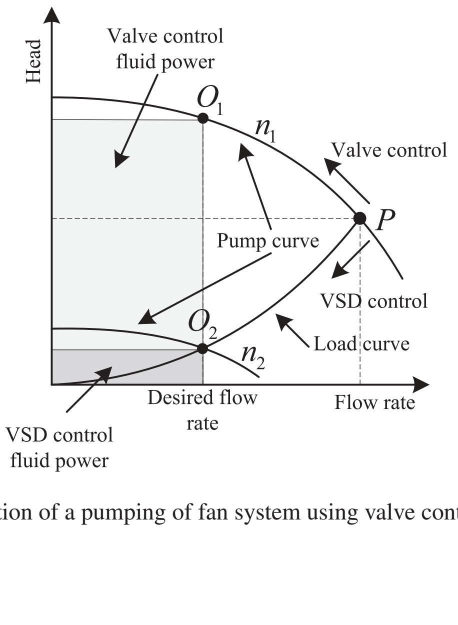

Figure 9.9 Operation of a pumping of fan system using valve control or VSD control

9.1.2 Vector Control

Vector control is based on the instantaneous decomposition of the electrical equations of the motor in order to obtain a relationship between the stator currents and the rotor flux with the torque produced, as in a DC motor. The decomposition is based on the well known Park transformation [6], and the result is the decomposition of the flux-producing current and the torque-producing current, as in a separate commutator motor, where the flux current corresponds to the excitation current and the torque-producing current corresponds to the armature current.

The principles of vector control apply equally to synchronous and asynchronous motors. The main difference is that in the case of PMSM, the rotor flux is produced by permanent magnets, which yield to an increase of efficiency of PMSM in front of an induction or asynchronous motor because there is no need to create rotor flux from stator currents.

From the Park transformation, the transformed equations are:

$$
\begin{aligned}
u_{qs} &= R_s i_{qs} + p\lambda_{qs} + \omega_s \lambda_{ds} \\
u_{ds} &= R_s i_{ds} + p\lambda_{ds} - \omega_s \lambda_{qs} \\
u_{qr} &= 0 = R_r i_{qr} + p\lambda_{qr} + (\omega_s - \omega_r)\,\lambda_{dr} \\
u_{dr} &= 0 = R_r i_{dr} + p\lambda_{dr} - (\omega_s - \omega_r)\,\lambda_{qr} \\
\lambda_{qs} &= L_s i_{qs} + L_m i_{qr} \\
\lambda_{ds} &= L_s i_{ds} + L_m i_{dr} \\
\lambda_{qr} &= L_m i_{qs} + L_r i_{qr} \\
\lambda_{dr} &= L_m i_{ds} + L_r i_{dr} \\
T_{em} &= \frac{3PL_m}{22L_r}\left(\lambda_{dr} i_{qs} - \lambda_{qr} i_{ds}\right),
\end{aligned}
\tag{9.6}
$$

where u_{qs} and u_{ds} are the stator applied voltages, R_s and R_r are the stator and rotor resistance respectively, i_{qs} and i_{ds} are the stator currents, p is the derivative operator, λ_{qs} and λ_{ds} are the stator flux linkages, ω_s is the stator voltages' and currents' frequency, u_{qr} and u_{dr} are the rotor voltages, which are equal to zero because the rotor windings are supposed short-circuited, i_{qr} and i_{dr} are the rotor currents, λ_{qr} and λ_{dr} are the rotor flux linkages, ω_r is the rotor voltages' and currents' frequency, L_s, L_m and L_r are the stator, magnetizing and rotor inductance respectively, T_{em} is the electromagnetic produced torque, and P is the number of poles.

The advantage of vector control is that, with proper selection of the angle in the Park transformation, equation (9.6) can be simplified. Choosing the angle orientated with rotor d axis flux, and then controlling the rotor q axis flux to be zero ($\lambda_{1qr} = 0$), the torque equation becomes

$$T_{em} = \frac{3pL_m}{22L_r}\lambda_{dr} i_{qs}, \tag{9.7}$$

which is similar to the torque equation of a DC motor. Moreover, if at the same time $p\lambda_{dr} = 0$, then λ_{dr} and i_{qs} are orthogonal, producing the maximum possible torque [7].

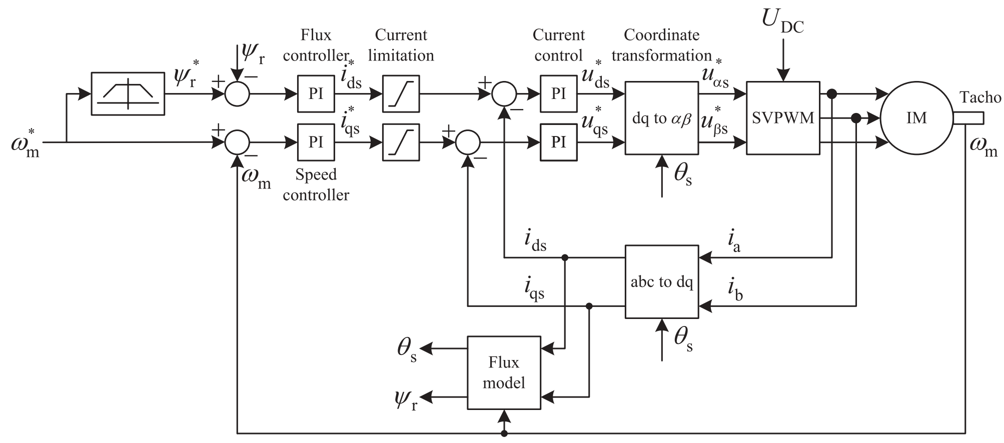

Figure 9.10 Block diagram of the vector control scheme

However, to maintain these conditions, the motor needs an accurate and dedicated control method.

In the case of PMSM, the rotor flux λ_{dr} is created by permanent magnets and has a fixed value. To control the torque it is necessary only to control i_{qs}, but in the case of an induction motor, the flux λ_{dr} must be created from stator currents. From the rotor voltage equation in (9.5), the relation between the rotor flux λ_{dr} and the stator current i_{ds} can be obtained as

$$\lambda_{dr} = i_{ds}\frac{L_m}{\tau_r p + 1}, \tag{9.8}$$

where $\tau_r = {}^{L_r}\!/_{R_r}$ is the rotor time constant. As mentioned in (9.7), the relation is of first order and must be computed in the controller in order to estimate the rotor flux λ_{dr} from the stator current i_{ds}, which can easily be measured.

Figure 9.10 shows that the torque and the flux can be independently controlled. In addition, in this control strategy, current limitation is directly implemented in the control scheme. It allows an inherent protection of the motor and converter. This limitation can be adapted depending on the motor working conditions, such as temperature, allowing the use of the motor in different ranges.

9.1.3 DTC

Direct Torque Control (DTC) was introduced in 1986 by Takahashi and Noguchi [8]. Meanwhile, Depenbrock developed Direct Self-Control (DSC) [9], a control scheme similar to DTC. It was this last author who developed the first VSD commercially available in 1995, based on DTC by ABB.

In DTC, the torque and the stator flux are controlled simultaneously by inverter voltage space-vector selection using a look-up table.

The main advantage of DTC compared with vector control is its simplicity. This simplicity allows DTC to implement a speed sensorless control system.

The main drawbacks of DTC are the high torque ripple and the variable switching frequency caused by the hysteretic controllers.

The DTC is based on the fact that the stator flux vector can be known from the voltage applied to the stator as

$$\Psi_{\mathrm{dqs}} = \int \left(u_{\mathrm{dqs}} - R_{\mathrm{s}} i_{\mathrm{dqs}}\right) \mathrm{d}t. \tag{9.9}$$

On the other hand, the torque produced can be computed knowing if this stator flux vector and stator current vector are known:

$$T_{\mathrm{em}} = -\mathrm{Im}\left(\Psi_{\mathrm{dqs}} i_{\mathrm{dqs}}^{\bullet}\right) = -\left\|\Psi_{\mathrm{dqs}}\right\| \left\|i_{\mathrm{dqs}}\right\| \mathrm{Im}\left(e^{\mathrm{j}(\theta_{\mathrm{s}}-\theta_1)}\right), \tag{9.10}$$

where θ_{s} is the angle of the stator flux vector and θ_1 is the angle of the stator current vector.

The DTC, supposing that the stator flux vector is known, determines which voltage vector must be applied to the motor depending on whether the flux must be increased or not, and also whether the torque must be increased or not.

Figure 9.11 shows, for a given stator flux vector, that each of the eight stator voltage vectors determine a new stator flux vector.

At the instant of time in Figure 9.11, the magnitude of the flux must be reduced in order maintain the stator flux vector within the flux hysteresis band. Then, only stator voltage vectors $\underline{V_3}$ and $\underline{V_4}$ can be used.

The decision to choose $\underline{V_3}$ or $\underline{V_4}$ is taken in the knowledge of whether the torque must be increased or reduced. To make this decision, another hysteresis comparator is used. In the case presented in Figure 9.11, $\underline{V_3}$ increases the torque, and $\underline{V_4}$ decreases the torque. The two zero vectors, $\underline{V_0}$ and $\underline{V_7}$, can be used when the torque needed is zero.

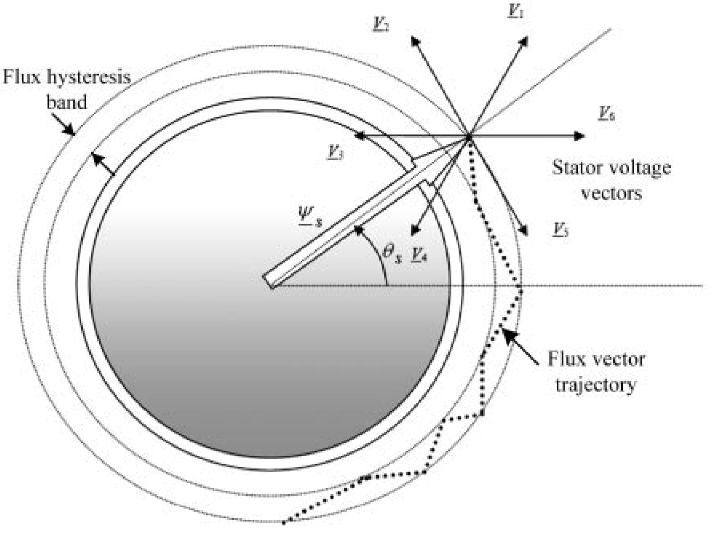

Figure 9.11 Flux trajectory in DTC control

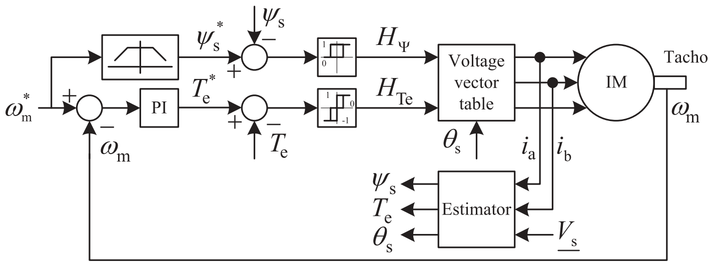

Figure 9.12 Block diagram of the DTC control method

From this methodology, a switching table can be obtained for all cases that determine which voltage vector is to be applied to the stator at each instant in time [8].

As shown in Figure 9.12, the DTC is a sensorless torque controller technique. The speed closed loop system requires a speed sensor but, as in case of vector control, speed sensorless techniques can be used with DTC.

9.2 Energy Optimal Control Methods

Figure 9.13 shows that in a variable speed drive there are three main sources of losses: the converter, the motor and the mechanical transmission system. Also cabling and connections can be considered in this loss count.

The motor control method can only deal with converter (inverter) and motor losses.

Transmission losses, and losses generated in the grid because of the rectifier, are not related to the motor flux level and current, so the control method cannot help to reduce these losses. Energy optimal control methods are focused on reducing losses in the converter and motor, and operating both in the most optimal point.

As mentioned above, the torque in a motor depends on two terms: the rotor flux and the stator current. The same amount of torque can be produced with different combinations of rotor flux and stator current. Higher rotor flux means higher iron losses, but less copper losses in the stator windings and higher stator current means less iron losses, but higher copper losses. Energy optimal control methods try to find the energy optimal stator current and rotor flux in order to produce the same amount of torque.

In the case of the V/f control scheme, where rotor speed and flux are not decoupled, the main disadvantage in flux reduction is that for a given operating point (torque and speed)

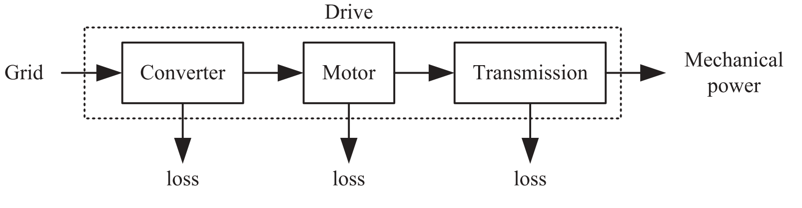

Figure 9.13 Energy process in a drive showing the losses of each element

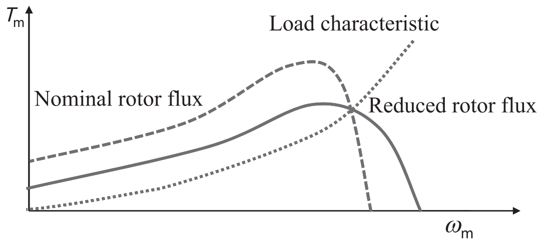

Figure 9.14 Torque-speed characteristics for a given operating point with nominal rotor flux and reduced rotor flux

the stator frequency must be increased, and then the maximum torque is reduced, making the system more sensitive to load perturbations, as shows in Figure 9.14.

In the case of vector control and DTC where the torque and flux are decoupled, the flux can be independently optimized, giving better performance, but the problem is the same (Figure 9.15). What is the optimal flux for a desired torque? Selecting the desired flux level is a balance between efficiency and not reducing the performance of the drive.

Calculating the optimal flux level in terms of efficiency is a complex task that must consider all the losses in the system. Computing all the losses in the VSD is an impossible task in real systems, because some losses are difficult to predict or not all the necessary measurements are available to compute them. Adding extra measurements to estimate the losses better will increase the cost of the system and also will reduce reliability of the control method and VSD.

9.2.1 Converter Losses

Converter losses are due to non-ideal behaviour of switching devices. They can be divided into conduction losses and switching losses.

Conduction losses are due to the resistive behaviour of power devices (transistors and diodes) when they are conducting. These losses depend on circulating current, the modulation index, the power factor of the converter and transistor parameters, which are strongly affected by many parameters such as temperature and voltage. These parameters are difficult to estimate and datasheet values are normally used.

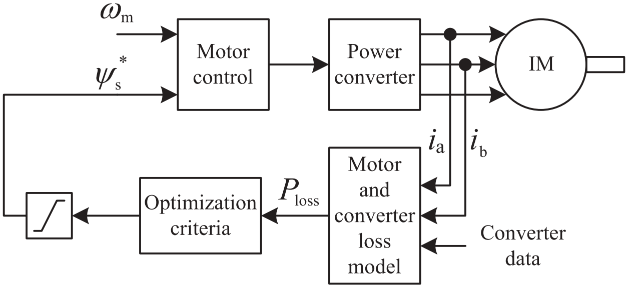

Figure 9.15 Block diagram of the optimization process

Switching losses are due to non-ideal switching behaviour of the power devices. Also, these losses depend on many parameters such as current, voltage temperature and device parameters, and are linearly dependent on frequency.

In order to reduce switching losses, adequate modulation technique must be used. As will be explained later, choosing the adequate modulation technique can reduce the number of commutations, reducing the switching losses, but not reducing the performance of the drive.

In the case of DTC, predicting the converter losses is a difficult task because conduction intervals and switching frequency are not constant.

9.2.2 Motor Losses

Motor losses are divided into copper losses and iron losses. Also, mechanical losses such as friction and windage can easily be taken into account.

Copper losses are due to the Joule effect in the stator and rotor windings (in the case of induction motors). Copper losses depend on the current and winding resistor, which depends on temperature and frequency due to skin effect.

Core losses include eddy currents and hysteresis, and both are due to magnetic flux variation in iron.

9.2.3 Energy Optimal Control Strategies

Many optimal control strategies for energy efficiency are present in the literature [7]. The most important methods are grouped and briefly explained here: simple state control, model-based optimization method, and search control methods. All these methods can be applied to V/f, vector and DTC control methods.

The simplest method is called simple state control. The objective of this method is to maintain a constant power factor operation point. A closed loop system measures cos (φ) and adjusts the flux to maintain a constant power factor.

Model-based methods are based on a motor model. The loss model can be solved analytically or numerically. In both cases, the currents are measured and then the model is solved, providing the optimal operating point.

Solving the model numerically allows for the inclusion of non-linearities, but it requires much time. The analytical solution can directly give the optimal flux.

The search control method requires a precise speed measurement in order to keep the output power of the motor constant. The objective is to modify the flux in order to find the minimum input power but maintain the output power, which gives the optimal operating point.

9.3 Topology of the Variable Speed Drive

A variable speed drive is a complex system where many elements must be present. Commercially available VSD are double conversion systems, even then there exists some converter topology that is able to generate a variable frequency-amplitude voltage from a fixed frequency-amplitude voltage as matrix converters.

Figure 9.16 shows the block diagram of the VSD. When operating in motor mode, the power from the grid is transformed into mechanical power. Normally, the motor can work

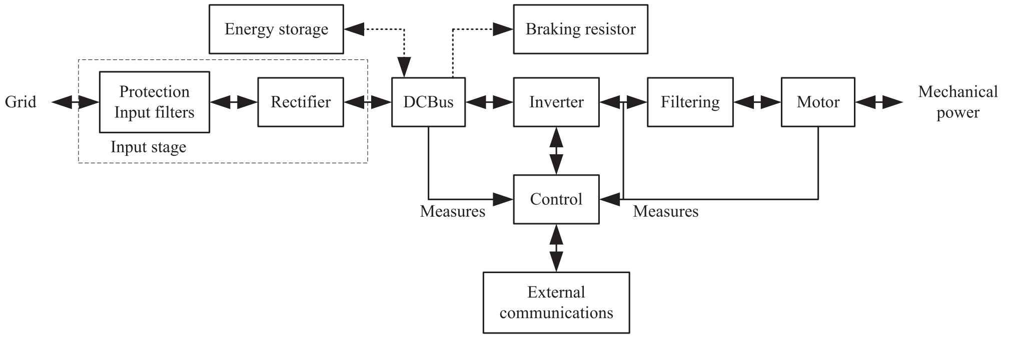

Figure 9.16 Block diagram of a VSD

in regenerative braking, but not the rectifier stage, so this braking energy must be dissipated in a braking resistor, or in more advanced applications, and can be stored in batteries or supercapacitors to be used in the future. Also, a rectifier can be designed to also inject energy to the grid from the motor.

Some measures must exist for control purposes for protection (currents, position/speed, DC bus voltage, grid voltage, etc.). All these measures are used by the control device to generate the appropriate signals to the inverter and other systems to control all the VSD functions. An external communications module is imperative in an actual VSD, not only for VSD configuration and parameterization, but also in a distributed control system, to interact with other control and supervision elements.

9.3.1 Input Stage

As shown in Figure 9.16, the input stage is formed by a protection and filter block and the rectifier. The rectifier is used to convert AC fixed frequency and magnitude voltage from AC grid into the DC voltage need by the inverter.

Figure 9.17 shows the diode bridge rectifier as a cheap and simple solution. It needs no control and line-commutated diodes offer good efficiency.

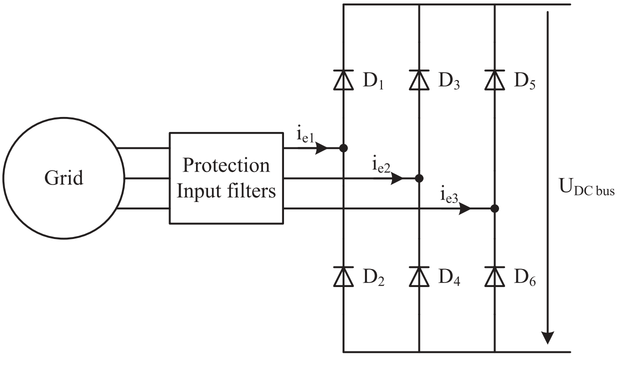

Figure 9.17 Input filter and rectifier on a VSD

The diode bridge rectifier consumes a current with high harmonic content from the grid. Standard IEC61000 [10], limits these harmonics, and some filter must be added in order not to pollute the grid. Also, low-order harmonics can cause losses. Protection is used to protect the VSD against overcurrent, voltage deviations, and voltage spikes coming from the grid. The diode bridge rectifier shown in Figure 9.17 cannot control the DC bus voltage ($\mathrm{U}_{\mathrm{DC\,bus}}$) to a constant value, and U_{DC} depends exclusively on grid voltage. Having a variable value of U_{DC} is disadvantageous for the inverter, and limits the speed range and the performance of the VSD.

But with new regulations on efficiency, [11] and [12], and injected grid harmonics [10], new solutions are on the shelf.

Power factor correctors (PFC) do not pollute the grid with low-order harmonics and consume a unity power factor current, reducing the total current consumed, but they are a more expensive solution, need more components and efficiency can be reduced. The main advantages are that the filter can be reduced, only high frequency harmonics must be filtered, and also that the PFC is able to maintain a constant U_{DC}.

When the motor is in generating mode, the diode bridge rectifier and PFC are unable to feed back this energy to the grid, and it must be dissipated in the braking resistors or stored in batteries or supercapacitors. The stored energy can be used in the future, increasing the total efficiency of the VSD.

But an active rectifier can be used as an input front-end. The active rectifier allows for a bidirectional power flow, and energy in the generating mode can be back injected to the grid, increasing the efficiency of the system. The topology of the active rectifier is the same as for the inverter, and can be controlled to consume sinusoidal current from the grid and give a constant power factor. The complexity and the cost of the system are high because of the large number of expensive elements and complex control that must run in a microprocessor.

9.3.2 DC Bus

The DC bus is an energy buffer formed mainly of an electrolytic capacitor to supply power transients to the inverter and maintain a constant DC voltage.

When connecting the VSD to the grid, a pre-charge system for DC bus voltage must be used in order to prevent in-rush currents. Frequently, for low-power VSD NTC resistors are used, but for high-power VSD, a resistor and a by-pass switch is desirable.

The bulk energy in the DC bus is stored in electrolytic capacitors. The C rate of this bus depends on the power of the converter. Electrolytic capacitors are a source of losses because they have a high ESR (and hence, high losses) and are not high temperature components (100°C). Care must be taken when designing the DC bus in order not to degrade the life of the bus with high temperature.

The high ESL of electrolytic capacitors can cause high $\mathrm{d}V/\mathrm{d}t$ in transistors, destroying them by over-voltage. To reduce these voltage spikes in transistors, MKP capacitors are connected close to the transistor.

When the motor works in a regenerative mode, the mechanical energy at the rotor shaft is transferred to the DC bus by the inverter, increasing the voltage in the capacitors. If the input stage is a diode bridge rectifier or a PFC, this energy cannot be injected back to the grid and is dissipated in the braking resistor. Figure 9.18 shows the DC bus capacitor, braking chopper and resistor.

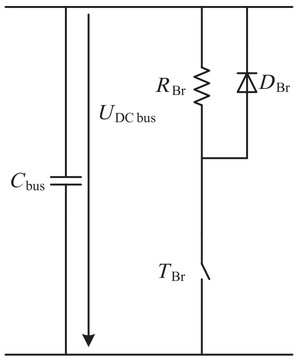

Figure 9.18 DC bus capacitors and braking chopper and resistor

This braking resistor must be sized to dissipate the predicted braking energy. To increase efficiency, this resistor is not always connected to the DC bus. A control system monitors the DC bus voltage. When this voltage reaches some predefined level, the chopper is used to dissipate the excess energy in the braking resistor.

However, this is not an efficient way to use electrical energy. Recently, in some applications where regenerative braking is common, such as in lifts, methods for energy storage with supercapacitors are becoming popular and effective [13]. Using supercapacitors to store the braking energy not only allows for a more efficient way of using energy, but it decreases the power of the rectifying stage. With a storage system, the input rectifier must supply only the average power and system losses, and not the peak power. It can reduce the cost of the system, input filter, protection, and cabling.

9.3.3 The Inverter

In general, when talking about VSD, the inverter is the power electronic element responsible for delivering the appropriated power to the motor [14]. For high voltages (400–800 V), high-power (4 kW–MW), the power electronic converters arena is dominated by the use of power semiconductors, which are mainly used in switching mode applications. This leads to some basic principles and operation modes that apply to all power electronics circuitries.

There are two different types of inverters for VSD [15]. Those that deliver a variable amplitude and frequency voltage, so called Voltage Source Inverters (VSI); and those that deliver a variable amplitude and frequency current, so called Current Source Inverters (CSI). In the medium to low-power drives market, VSI is the most common inverter, due to its simplicity and reliability; in the high-power drives market, both inverters share a part of the market, with a tendency for VSI to increase this share.

The converter output characteristics depend on the converter structure, the type of electronic devices and the method of control. Besides the connection circuit, they impose the proprieties of the output voltage and, consequently, the global impact of the power converter on load supplied.

Silicon-based (Si) static switches are, by far, the prevailing switches in VSD inverters [16]. Insulated gate bipolar transistors (IGBTs) or metal-oxide field-effect transistors (MOSFETs)

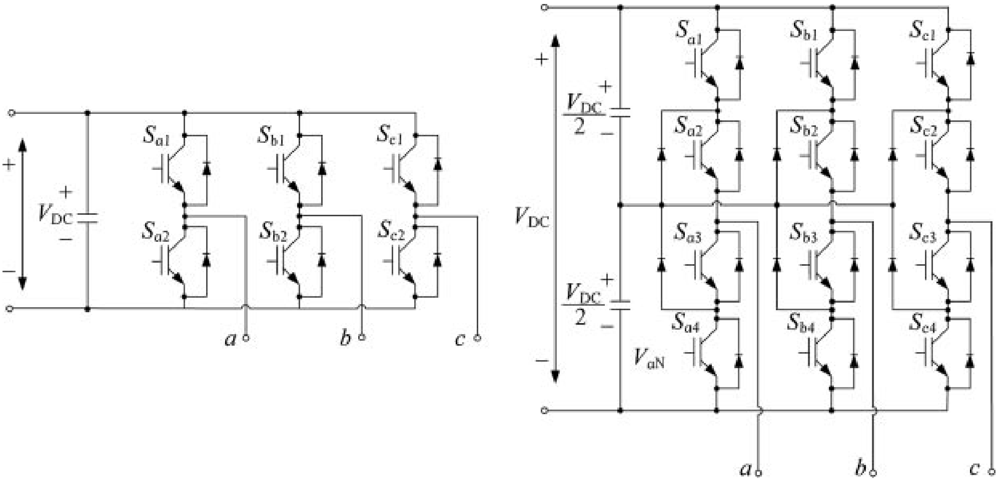

Figure 9.19 Schematics of a three-phase two-level inverter, and a three-phase three-level inverter

are the most usual type of power semiconductor switches; IGBTs in the medium to high-power applications and MOSFETs in the low-power ones. As these switches are the main source of losses in an inverter (conduction and commutation losses), there are great efforts to switch from the Si-based devices to the more energy efficient wide bandgap (WBG) based devices, with silicon carbide (SiC) and gallium nitride (GaN) being at the forefront [17] and [18].

The topology of the power converter in a drive plays an important role in energy efficiency considerations. The most spread inverter topology is the two-level inverter, but multilevel topologies, for high efficiency or high voltage drives, have to be considered. Figure 9.19 shows the three-level and the multilevel NPC topology.

Finally, the way in which switches are turned on and off, i.e. which modulation technique is used, has a great impact on the behaviour of the drive system, affecting losses in the inverter itself, due to commutations, losses in the motor due to harmonic currents, or to a weakened magnetic field, caused by a reduced voltage feeding.

9.4 New Trends on Power Semiconductors

Power electronics uses semiconductors to control energy flow between the source and the load. There are different semiconductor materials, but silicon-based semiconductors are the most used [19]. New materials such as SiC (silicon carbide) and GaN (gallium nitride) are being investigated for use as semiconductors in power electronics.

Semiconductors in power electronics are used only in saturation, i.e. they are only used in an on and off state, but not in the linear operating region. Figure 9.20 shows the switching process of a semiconductor transistor. In the figure there are two main sources of losses: conduction losses and switching losses.

Operation of the converter at high switching frequencies results in a size reduction of the passive components at the expense of increased switching losses. However, Wide Band Gap

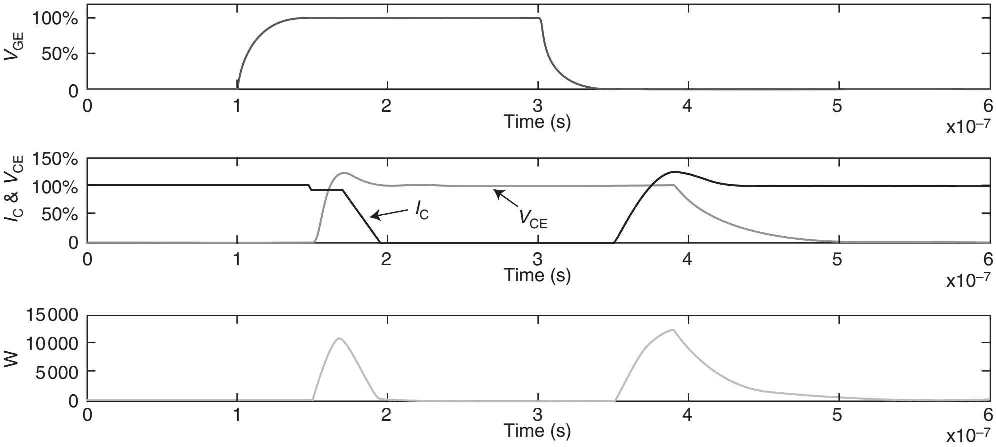

Figure 9.20 Behaviour of an IGBT (turn-on and turn-off)

(WBG, such as SiC and GaN) power devices have the potential to operate at high switching frequencies without the penalty of significant loss because of their fast switching times and their ability to work at high temperatures compared with similar Si devices [20]. Table 9.1 shows different properties of different materials.

9.4.1 Modulation Techniques

The inverter responsible for converting the DC voltage, from the DC bus, to variable amplitude and frequency AC voltages. This conversion from DC to AC can be realized using different techniques, which are usually known as modulation techniques.

In general, all the modulation techniques consist of turning the switches of the inverter on and off at a high frequency in such a way that the average value of the inverter output voltage equals that of the voltage reference. The longer the switch is on compared with the off time, the higher the average output voltage is.

There are several techniques, each with its own advantages and disadvantages, so for each application one should select the one that best fits the demands of the application. Next, a brief description of the main modulation techniques is presented, focusing on the differences between them, in order to make a selection of the modulation technique clear to the end user.

Table 9.1 Different properties of Si, SiC and GaN materials [21]

Property	Si	4H-SiC	GaN
Bandgap, Eg (eV)	1.1	3.26	3.45
Dielectric constant, ε_r	11.9	10.1	9
Breakdown field (kV/cm)	300	2200	2000
Thermal conductivity (W/cm K)	1.5	4.9	1.3

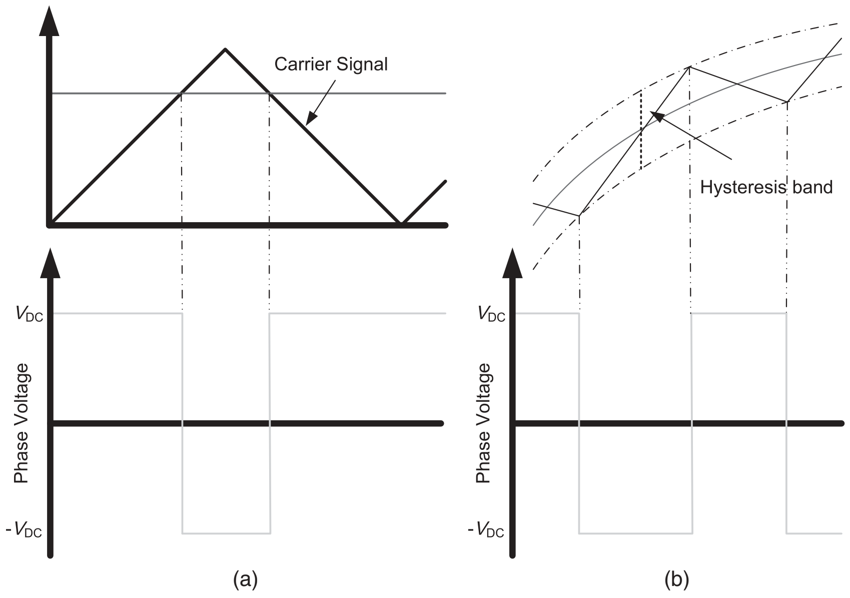

Figure 9.21 Schematics of: (a) a carrier-based modulation and (b) hysteresis modulation

The different modulation techniques can be classified into two large groups, according to the constancy of the switching frequency:

1. The first group, presenting a constant switching frequency, is usually known as Carrier-Based PWM methods (CBPWM). Figure 9.21 shows an example of a CBPWM where the main signals and characteristics of these modulation techniques arise. In these modulation techniques, the current spectrum presents as harmonic at the switching frequency and at its multiples [22], making it easy to filter such harmonic components. The higher the switching frequency, the higher the order of the current harmonics. Considering that in electrical motors, current harmonics implies torque vibration, in applications where mechanical resonances are of great importance, this is the kind of modulation that is preferred, thanks to its ability to move torque frequencies away from the resonances ones.
2. The second group, presenting a variable switching frequency, is usually known as hysteresis modulations (when they are implemented in a hardware manner), or dead-beat control (when they are implemented in a software manner). Compared with the previous group, they deliver a superior performance in terms of the bandwidth of the close loop, but presenting a worst current spectrum.

Apart from the aforementioned difference, modulation techniques differ in other important aspects, the next being the most important as discussed below.

• *Modulation Index (MI)*: in some sense, the modulation index is a figure of merit of the use made by the modulation technique of the DC bus voltage. That is, one inverter controlled by two different modulation techniques, each one with its own MI, can obtain different output voltages from the same DC bus. This is the modulation method with the greater MI, the one that obtain the maximum output voltage. From the energy efficiency side, this has a great impact on the efficiency of an induction motor operated at a fixed speed and close to its nominal speed. In these situations, the maximum output voltage of the inverter tends to be below the nominal voltage of the motor, feeding the motor with a reduced voltage, which finally produces additional losses.

- *Commutation losses*: each time that a switch of an inverter is turned on or off, a certain amount of energy is lost. Some modulation techniques make better use of these commutations (optimize the commutation sequence, obtaining the same voltage distortion) than others with higher commutations, implying reduced commutation losses and thus higher efficiencies.

The selection of a modulation technique for a particular application is often a balance between these competing advantages and disadvantages. Next, the principal modulation techniques are introduced, ordered in a chronological order, and with a brief introduction to each one.

9.4.2 Review of Different Modulation Methods

9.4.2.1 The Square-Wave Modulation

The square-wave modulation was the first method used and a simpler one (better known today as the *six-steps* modulation). Of all the various modulation techniques, this is the one that obtains the maximum fundamental output voltage, but giving, at the same time, the maximum current harmonic distortion.

Owing to this high current THD (Total Harmonic Distortion), this method is currently rarely used, and its use is restricted to applications for controlling the speed of induction motors in the areas of maximum speed and maximum torque, where the inverter has to deliver the maximum possible voltage. In this case, current harmonics are minimized thanks to the effective impedance of the phase motor, which increases linearly with the voltage frequency, and it reaches its maximum at the rated speed.

Nevertheless it is important to calculate the maximum output voltage of the *six-step* modulation, because the modulation index of a modulation method, is defined as the ratio between the maximum voltage of this modulation method and the maximum voltage of the *six-step* method [23]. As is obvious, the modulation index of the *six-step* modulation, by definition, is one.

The *six-step* modulation consists of turning on and off each leg of the inverter at the frequency of the reference output voltage. In Figure 9.22, the three leg voltages in respect to the virtual mid-point of the inverter are shown. Considering a balanced load (as is usually the case in a three-phase induction motor), it shows the voltage between the neutral point of the load and the virtual mid-point of the inverter [22].

The line-to-line voltages are shown in Figure 9.23 along with the line-to-neutral voltage. The presence of the *quasi-square* waveform can be observed in the line-to-line voltages, which gives the name to this kind of modulation; and in the line-to-neutral waveform, the presence of six 'steps' can be observed, which gives the name to this other kind of modulation. From Figure 9.23, it is possible to calculate the line-to-line voltage using the Fourier series (Table 9.3).

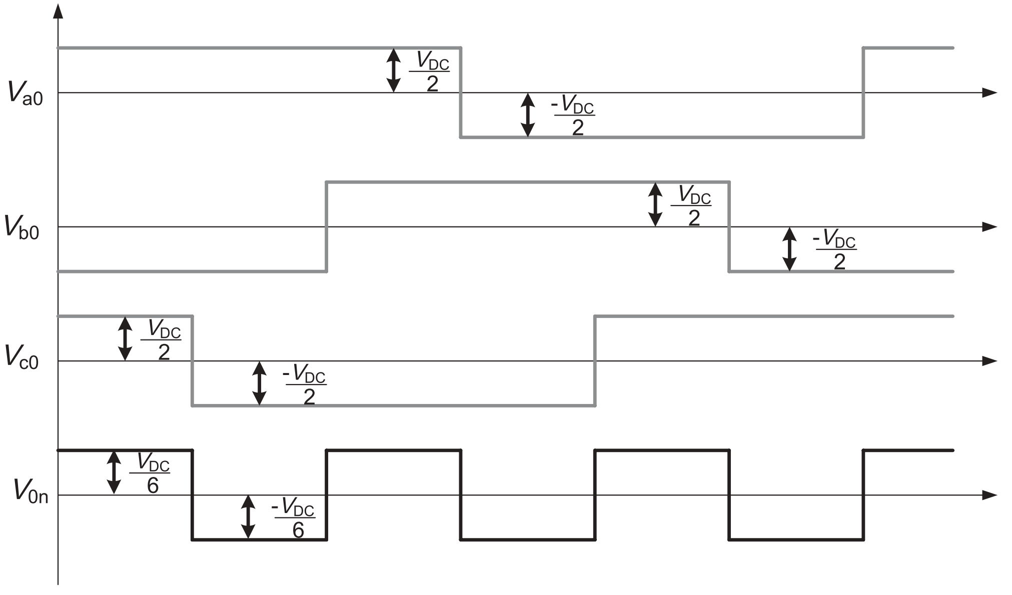

Figure 9.22 Phase to virtual mid-point of the DC bus voltages in the six-step modulation and neutral to virtual mid-point of the inverter voltage

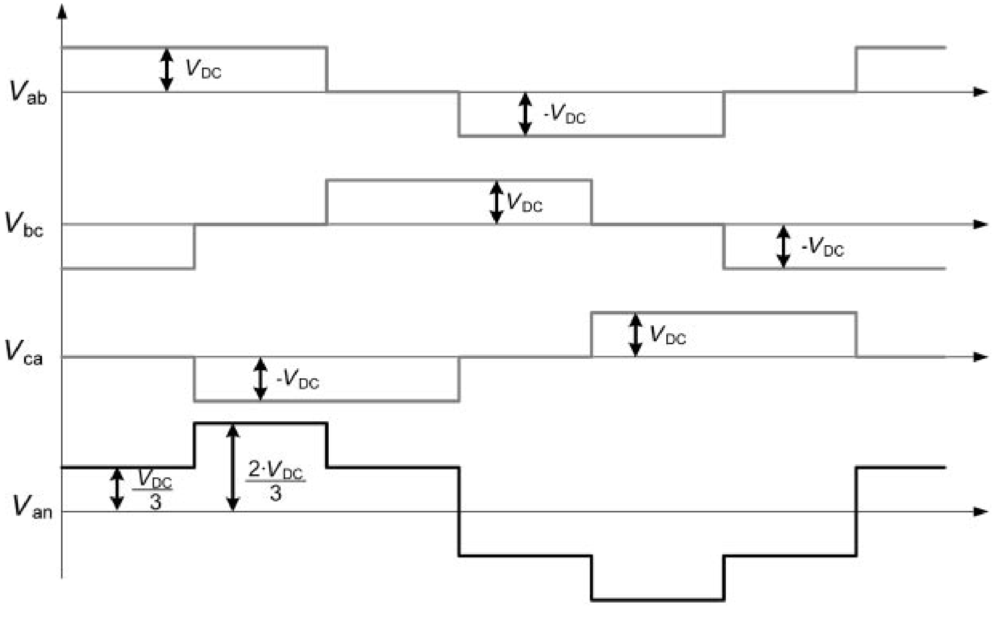

Figure 9.23 Line-to-line voltage and line-to-neutral voltage in a six-step modulation

9.4.2.2 Sinusoidal Carrier-Based PWM (Figure 9.24)

The sinusoidal carrier-base Pulse Width Modulation (SPWM) is one the most widely used modulation technique [24]. It was first introduced by Schönung in 1964, and it consists in the comparison of a carrier signal (usually a triangular one) with a modulated signal (in this case a sinusoidal control signal). The upper switch of an inverter leg is turned on each time the modulated signal is greater than that of the carrier, and on the other hand, when the modulated signal is below that of the carrier, it is the lower switch that is turned on, and the upper switch is turned off [15].

Although this modulation technique has been presented in its single-phase case, it is easy extended to its three-phase implementation. It is worth noting that this kind of modulation technique is easily implemented in a digital manner, thanks to what are known as PWM outputs, which are implemented in micro-controllers or digital signal processors (DSP) that are intended for digital motor control.

The maximum output voltage of the PWM modulation is presented in Table 9.3. Along with the maximum output voltage and the maximum modulation index, Table 9.3 shows the ratio between the maximum output voltage and the three-phase input voltage, assuming that the inverter is preceded by a three-phase diode rectifier, and that the DC bus gets charged with the peak voltage of the input three-phase voltage (which would be the idealized value when the DC link capacitors tend to infinity). This value of 0.86 means that, if a three-phase induction motor that was specified to work directly connected to the mains, was fed by an inverter controlled by the SPMW, this motor could never be working at its rated speed or

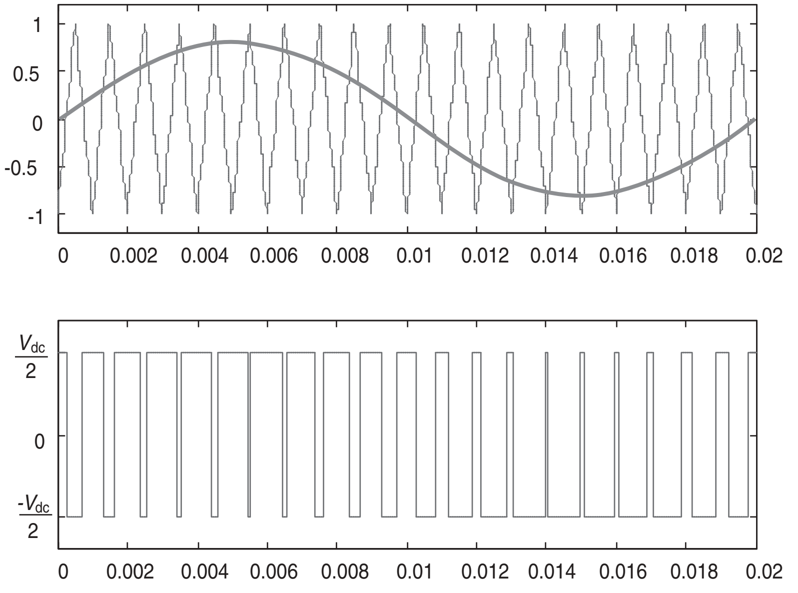

Figure 9.24 Sinusoidal carrier-based PWM. In the upper figure the carrier signal) and the modulated signal are depicted. In this case, the modulated signal consists of a 50 Hz sinusoidal. In the bottom figure, the voltage of the phase to mid-point of the inverter is shown

otherwise it would work in an undervoltage condition, which always ends with an increment in the losses of the motor, and a drop in its efficiency.

9.4.2.3 Third Harmonic Injection PWM (THIPWM) (Figure 9.25)

In 1997, Buja and Indri presented the THIPWM to overcome this inconvenience of the SPWM method [25]. This modulation technique consists in adding an adequate third harmonic waveform to each of the reference three-phase voltages, so allowing an increase in the fundamental maximum output voltage by a factor of $\sqrt{3}/2$, which is nearly 15.47% higher than the SPWM.

The THIPWM has been object of an intensive research, and different authors have proposed different *appropriate* third harmonic waveform. Some have proposed a maximum peak value of 1/6 and others peak values for the sinusoidal modulating signal [24], while others have proposed discontinuous zero-sequence components, not always sinusoidal [26].

9.4.2.4 Space-Vector PWM (SVPWM)

With the generalization of the Park strategy to control the torque and the speed of the induction motor, the carrier-based modulating techniques were no longer suitable for control of the inverter. Instead, a complex voltage vector for control of the inverter arose, the space-vector PWM.

This method, in a naturally manner, includes in its algorithm the discretization in eight unique vectors, which are the different states that can impose the three-phase inverter [27]. Usually, the

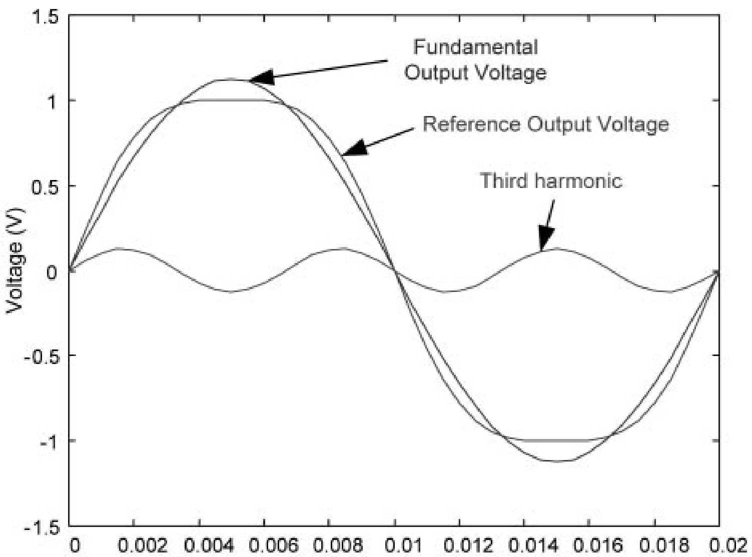

Figure 9.25 THIPWM: The fundamental output voltage, reference output voltage and third harmonic waveform

voltages resulting from these eight states are expressed in Park's variables. The highly coupled nature of the inverter loads, such as induction and synchronous machines, has led to the use of *artificial* variables rather than actual (phase) variables [22]. The mathematical expression which allows one to obtain from the actual variables the Clarke variables is as follows:

$$\begin{pmatrix} v_0 \\ v_\alpha \\ v_\beta \end{pmatrix} = \sqrt{\frac{2}{3}} \begin{pmatrix} {}^1\!/\!\sqrt{2} & {}^1\!/\!\sqrt{2} & {}^1\!/\!\sqrt{2} \\ 1 & -{}^1\!/_2 & -{}^1\!/_2 \\ 0 & {}^{\sqrt{3}}\!/_2 & -{}^{\sqrt{3}}\!/_2 \end{pmatrix} \begin{pmatrix} v_r \\ v_s \\ v_t \end{pmatrix}. \tag{9.11}$$

In Figure 9.26 a brief explanation of the Clarke transformation is presented. Representing in Cartesian coordinate system the phase variables (v_r, v_s, v_t), of a symmetrical balanced three-phase voltage, all the points are located in the homopolar plane. That is, at any instant $v_r + v_s + v_t = 0$. Now, one can define a new coordinate orthogonal frame, with two of the axes located in the homopolar plane ($\alpha - \beta$) and the third axis perpendicular to this same plane (0).

The SVPWM modulation consists of synthesizing a reference voltage, expressed in Clarke variables $\underline{V}^\bullet = (V_\alpha, V_\beta)$, by the combination of any of the eight possible states of the inverters, as depicted in Figure 9.27 and 9.28. Table 9.2 summarized the eight possible vectors for a two-level inverter.

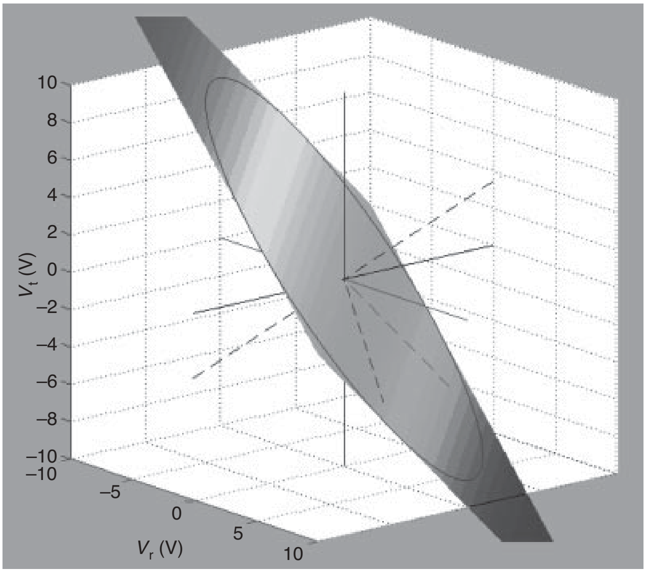

Figure 9.26 Interpretation of the Park's transformation. The Cartesian coordinate system, formed by the three-phase vectors, and the homopolar plane (that is, all the points of the space where $v_r + v_s + v_t = 0$), where the $\alpha - \beta$ axes are located

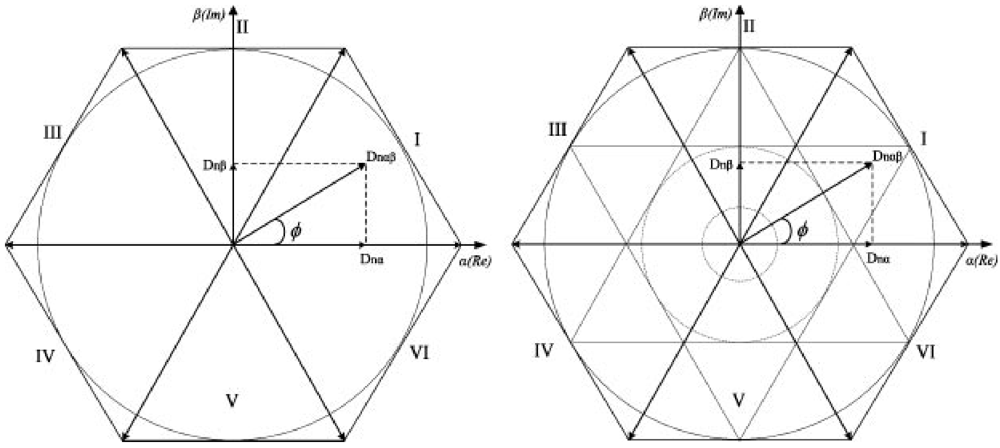

Figure 9.27 A two-level space-vector PWM and a three-level SVPWM

9.4.2.5 Multilevel Inverter Topologies

Recently, multilevel power converters have received a great deal of attention in numerous high-power medium-voltage industrial applications [28–33]. A multilevel converter uses a series of power semiconductor switches to perform the power conversion by synthesizing the AC output terminal voltage from several DC voltage levels and, as a result, staircase waveforms can be generated. Compared with standard two-level converters, multilevel converters offer great advantages such as lower harmonic distortion, lower voltage stress on loads, a lower common-mode voltage, and less electromagnetic interference. By reducing filtering requirements, they not only improve the efficiency of converters, but also increase the load power and, hence, the load efficiency by improving the load voltage with a lower harmonic content.

Multilevel converters have been basically developed to increase a nominal power in the converter. The higher number of voltage levels in these topologies results in higher quality

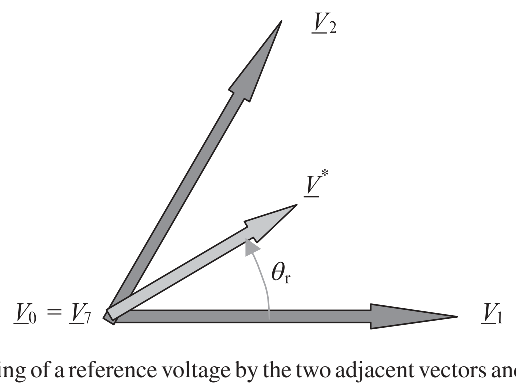

Figure 9.28 Synthesizing of a reference voltage by the two adjacent vectors and the two homopolar ones

Table 9.2 The eight different states of a three-phase inverter, and the vectors resulting from these states. In the fifth column it expressed in terms of the phase voltages, and in the last column in the $\alpha-\beta-0$ frame

	S_1	S_2	S_3	$\underline{V}_{\text{rst}} = (V_r, V_s, V_t)$	$\underline{V}_{\text{park}} = (V_\alpha, V_\beta, V_0)$
$\underline{V}_0$	−1	−1	−1	$\left(\frac{-V_{DC}}{2}, \frac{-V_{DC}}{2}, \frac{-V_{DC}}{2}\right)$	$\left(0, 0, \frac{-\sqrt{3}V_{DC}}{2}\right)$
$\underline{V}_1$	1	−1	−1	$\left(\frac{V_{DC}}{2}, \frac{-V_{DC}}{2}, \frac{-V_{DC}}{2}\right)$	$\left(\frac{\sqrt{2}V_{DC}}{\sqrt{3}}, 0, \frac{-\sqrt{3}V_{DC}}{6}\right)$
$\underline{V}_2$	1	1	−1	$\left(\frac{V_{DC}}{2}, \frac{V_{DC}}{2}, \frac{-V_{DC}}{2}\right)$	$\left(\frac{V_{DC}}{\sqrt{6}}, \frac{V_{DC}}{\sqrt{2}}, \frac{\sqrt{3}V_{DC}}{6}\right)$
$\underline{V}_3$	−1	1	−1	$\left(\frac{-V_{DC}}{2}, \frac{V_{DC}}{2}, \frac{-V_{DC}}{2}\right)$	$\left(\frac{-V_{DC}}{\sqrt{6}}, \frac{V_{DC}}{\sqrt{2}}, \frac{-\sqrt{3}V_{DC}}{6}\right)$
$\underline{V}_4$	−1	1	1	$\left(\frac{-V_{DC}}{2}, \frac{V_{DC}}{2}, \frac{V_{DC}}{2}\right)$	$\left(\frac{-\sqrt{2}V_{DC}}{\sqrt{3}}, 0, \frac{\sqrt{3}V_{DC}}{6}\right)$
$\underline{V}_5$	−1	−1	1	$\left(\frac{-V_{DC}}{2}, \frac{-V_{DC}}{2}, \frac{V_{DC}}{2}\right)$	$\left(\frac{-V_{DC}}{\sqrt{6}}, \frac{-V_{DC}}{\sqrt{2}}, \frac{-\sqrt{3}V_{DC}}{6}\right)$
$\underline{V}_6$	1	−1	1	$\left(\frac{V_{DC}}{2}, \frac{-V_{DC}}{2}, \frac{V_{DC}}{2}\right)$	$\left(\frac{V_{DC}}{\sqrt{6}}, \frac{-V_{DC}}{\sqrt{2}}, \frac{-\sqrt{3}V_{DC}}{6}\right)$
$\underline{V}_7$	1	1	1	$\left(\frac{V_{DC}}{2}, \frac{V_{DC}}{2}, \frac{V_{DC}}{2}\right)$	$\left(0, 0, \frac{\sqrt{3}V_{DC}}{2}\right)$

Table 9.3 Summary of the four modulation techniques. For each technique, the following results are presented: the maximum output voltage (with respect to the voltage of the DC bus), the maximum modulation index and, finally, the ratio between the output voltage and the three-phase input voltage, assuming that the inverter is preceded by a three-phase diode rectifier, and that the DC bus is charged with the peak voltage of the input three-phase voltage

Modulation technique	Maximum output voltage	$\frac{V_{\text{eff}}\text{out}}{V_{\text{eff}}\text{in}}$	Modulation index
Square-wave	$\frac{\sqrt{6}}{\pi} \cdot V_{DC} = 0.779 \cdot V_{DC}$	1.1	1
PWM	$\frac{\sqrt{3}}{2 \cdot \sqrt{2}} \cdot V_{DC} = 0.612 \cdot V_{DC}$	0.86	0.785
PWM + Third harmonic injection (THIPWM)	$\frac{1}{\sqrt{2}} \cdot V_{DC} = 0.707 \cdot V_{DC}$	1	0.907
Space-vector PWM	$\frac{1}{\sqrt{2}} \cdot V_{DC} = 0.707 \cdot V_{DC}$	1	0.907

output voltages. The concept of multilevel converters was introduced in 1975, and the term ‘multilevel’ first meant ‘three-level’ [34] but now refers to converters with more than a two-level output voltage. Multilevel topologies have been developed by increasing the number of semiconductor switches or the number of power converter modules (i.e., multiple converter modules). The trend toward a greater number of voltage levels is necessary because of the benefits of higher voltage ratings with a very low total harmonic distortion. By increasing the number of voltage levels, the converter’s fundamental output voltage can be produced with a lower harmonic content, and it will significantly improve the quality of the output voltage and eventually approach a desired sinusoidal waveform.

The conventional two-level converter can produce high-quality outputs for low-power applications by using a high switching frequency. However, for medium- and high-power applications, the maximum switching frequency is limited by the switching devices due to the high switching losses. In this case, multilevel converters can be used to lower the switching frequency, and a high-quality output waveform can be produced. The superior features of multilevel converters over two-level converters can be briefly summarized as follows [28, 35, 36].

- They can generate output voltages with very low THD. Multilevel output PWM voltage can reduce the inverter switches blocking voltage and the $\mathrm{d}v/\mathrm{d}t$ stress on the load such as a motor. The lower voltage stress on a load can reduce the number of Electro-Magnetic Compatibility (EMC) problems.
- They can produce a lower common-mode voltage. Thus, the lifetime of a motor connected to a multilevel motor drive can be increased due to the reduced stress on the bearings of the motor.
- By generating a staircase voltage waveform, they can produce lower converter input current distortion. The lower current ripple can reduce the size of a capacitor filter in a DC link.
- They are capable of operating at both a fundamental switching frequency and a high switching frequency PWM. In high-power applications, a lower switching frequency can reduce the switching loss, resulting in an efficiency increase.
- However, the trade-off for such an increased performance in multilevel converters is that they require a greater number of power switching devices. The number of semiconductor switches together with their related gate drive circuits can increase the overall system cost and the control complexity. In addition, several DC voltage source are required, which are usually provided by capacitors. Balancing the voltages of these capacitors according to an operating point is a difficult challenge.
- Despite these drawbacks, multilevel converters have turned out to be a very good alternative for high-power applications, since the cost of the control for these cases is a small portion of the whole system cost. Furthermore, as prices of power semiconductors and DSPs continue to decrease, the use of multilevel topologies is expected to extend to low-power application (those of less than 10 kW) as well. Fast power devices (CMOS transistors), which can operate at very high switching frequencies, can be used for low voltages. Therefore, the values of the reactive components will undergo significant reduction. Furthermore, new power devices are expected to appear in the near future, and these may also extend the application of multilevel topologies.

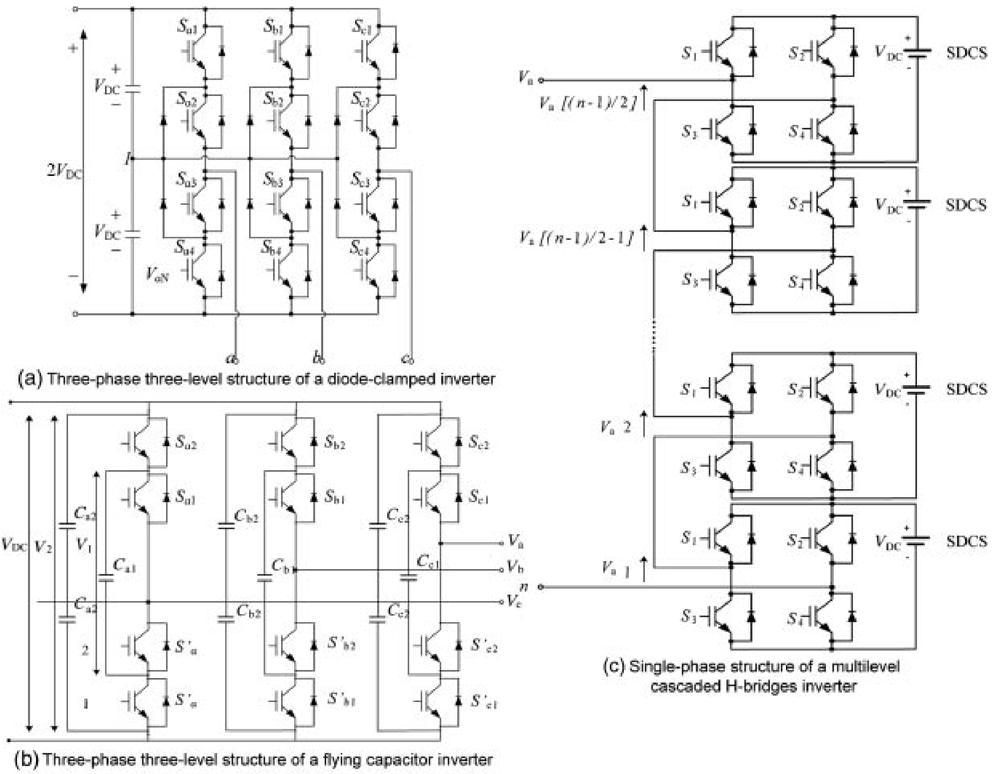

Figure 9.29 Three different structures of multilevel converters

Multilevel topologies are classified into three categories (Figure 9.29):

1. Diode clamped (neutral-clamped) converter
2. Flying capacitors (capacitor clamped) converter
3. Cascaded H-bridges converter with separate DC sources.

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