High-Performance Motor Drives

Marian P. Kazmierkowski, Leopoldo G. Franquelo, Jose Rodriguez, Marcelo Perez and Jose I. Leon

Abstract-- Power electronic converter-fed high performance AC drives belong to high-tech industry and are one of main factors for energy saving and productivity growth. This paper reviews present state and trends in development of key parts of controlled induction motor drive systems: converter topologies, modulation methods, as well as control and estimation techniques. Following topologies are described: Two- and Multi-level Voltage Source Converters, Current Source Converters and Direct Converters. Among the modulation techniques are briefly presented: conventional bipolar and unipolar PWM, space vector modulation (SVM) with extension for multilevel converters, harmonic control techniques, variable frequency modulation (hysteresis, nearest level, and model predictive techniques). About the control strategies, the generic torque control methods are described in two groups: linear and nonlinear controllers. In linear group are presented field oriented control (FOC), direct torque control (DTC) with voltage SVM, and DTC with flux vector modulation (DTC-FVM). The group of nonlinear control include: classical switching table based hysteresis DTC, direct self control (DSC) and predictive DTC scheme. Also, selected simple flux vector observers/estimators are discussed.

I. INTRODUCTION

The main part of all the produced electric energy is used to feed electric motors and the conversion of electrical power into mechanical power involves motors ranging from below 1 W up to several dozen MW. The contemporary drive systems are expected to meet a variety of requirements among which are:

- Maximum conversion efficiency,
- Wide range steeples adjustment of angular speed, torque, acceleration, angular and linear position,
- Fast error elimination when the control and/or disturbance signals are being changed,
- Maximum utilization of motor power under reduced voltage, current, etc.,
- Reliable, user friendly operation.

The present-day drive control engineering draws heavily on the theory of electric motors, control theory, and industrial electronics, however, it is the last of these disciplines what plays a decisive role in the development of automated electrical drive systems. The two basic divisions of industrial electronics: digital signal processing and power electronics, have opened the way to replace DC brush motors by AC motors in high performance drives.

The roots of power electronics go back to 1901 when P. C. Hewitt invented glass-bulb mercury-arc rectifier [1]. However, the present era of semiconductor power electronics started with the commercially introduced by General Electric (GE) silicon controlled rectifier (SCR), popularly called thyristor, in 1958. Next, the development continued in new semiconductor structures, materials, fabrication, etc. bringing on the market many new devices with higher power ratings and improved characteristics. Today among most common power electronic devices are: power metal oxide semiconductor field effect transistor (MOSFET) and insulated gate bipolar transistors (IGBT), and in very high power range the integrated gate-commutated thyristors (IGCT) **¡Error! No se encuentra el origen de la referencia.**]. Also, integrated intelligent power modules (IPM) are available. Today very promising, opening new era of high voltage, high frequency, and high-temperature technology are semiconductor devices based on wide-band gap silicon carbide (SiC) material.[3]. The new power semiconductor devices have always triggered the development of new converter topologies. At the beginning it started with diode and thyristor line commutated converters and continued with modern forced commutated converters controlled via pulse width modulation (PWM) methods. The list of application of power electronics is too long to list them in all. Therefore, in Figure 1, a classification of the typical applications of power electronic converters from low to high power group is given.

This paper presents an overview of power converter-fed high performance AC drives in three basic parts: power converter topologies, modulation techniques for power converters, control and estimation.

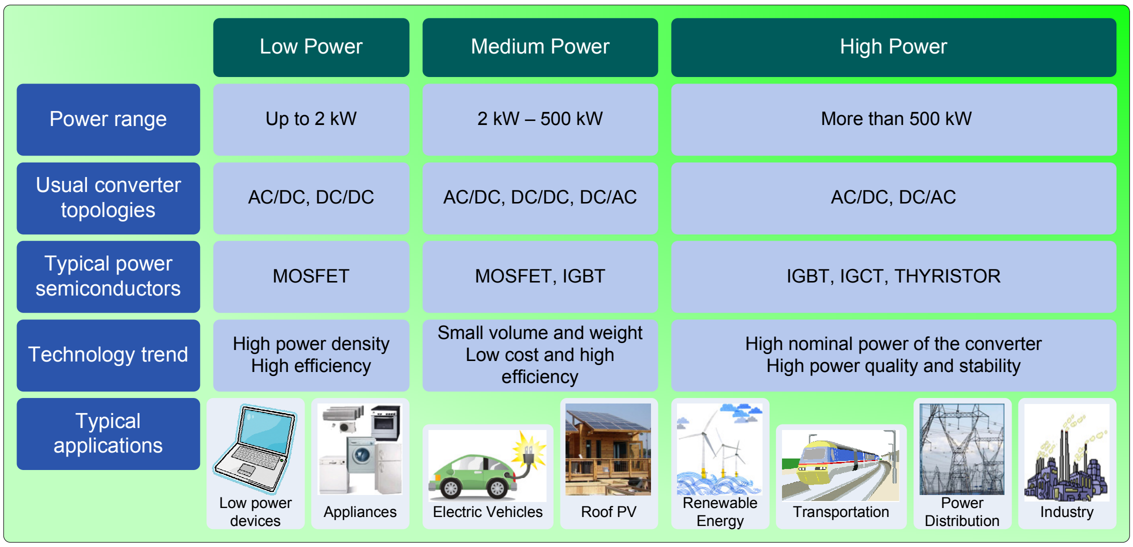

Figure 1. Classification of the power converter applications

II. POWER CONVERTER TOPOLOGIES

IV. 1. Voltage Source Converters

Voltage source converters are considered nowadays a mature technology and have become one of the most common power converter topologies in industry. The basic 2-level voltage source inverter (2L-VSI) is shown in 2a). It is composed by a DC capacitor or voltage source, which gives its name, and an arrangement of two power semiconductors per phase. The gating signals for both power switches are complementary, and the load is connected to the positive or the negative bar and generates only two possible output voltage levels. Using a modulation strategy to generate the gating pulses it is possible to synthesize output voltages with the desired fundamental component. Due to the high harmonic content in the output voltage that need to be filtered to produce nearly sinusoidal currents, these converters are usually applied to very inductive loads such as motors [2], but sometimes, depending on the application, an additional output filter could be required.

The maximum output voltages when a maximum modulation index is used, is defined by the DC link voltage value. In order to efficiently drive high power loads, a large DC link voltage is required, but in practice, this voltage is limited by the blocking voltage of semiconductors and therefore its technology. For instance, some industrial drives use low voltage IGBT (LV-IGBT) to provide up to 690V output voltage. To avoid this voltage limitation, voltage source multilevel converters topologies have been developed during the last decades [3]–[9]. These converters are more complex than the two-level voltage source inverter (2L-VSI) in terms of topology, modulation and control, but the additional complexity can be considered as an opportunity to improve its power quality, reliability, power density, performance and efficiency.

The three-level neutral point clamped (3L-NPC) is shown in Figure 2b) [10]. In this converter the DC link voltage is equally divided using two capacitors and, therefore, the phase output can be connected to the positive bar switching on the two upper switches, to the mid-point using the two central switches or to the negative bar using the two lower switches [11][12]. Each semiconductor needs to block only the half of the total DC link voltage allowing increasing the power rating using the same semiconductor technology than the conventional 2L-VSI. The semiconductors normally used in these converters are the high voltage IGBT (HV-IGBT) [13] and the IGCT [14]. A well documented problem presented in this converter is the unbalance of the capacitors produced by asymmetries in the converter hardware and mainly the operational conditions [15]. Several strategies to deal with this problem have been proposed, mainly focused to modify the modulation strategy [16][17]. Another issue of this converter is the unequal distribution of losses, which causes that the outer switches switching losses differ from the central ones, depending on the operation condition. This problem cannot be easily addressed using the conventional scheme and modified topologies like active neutral point clamped (ANPC) have been proposed [16]. In this converter, the clamping diodes are replaced by controlled switches, then selecting the appropriate combination of switches it is possible to reduce and equally distribute the losses. The NPC converter can be scaled up to achieve more than three levels simply dividing the DC-link in more than two values using several capacitors [18]. Each one of these partial DC-link voltages can be connected to the load using an expanded arrangement of switches and clamping diodes. Along with the increased power rating, the advantages of several output voltage levels are a better power quality, smaller dv/dt and associated EMI [19]. However, when the NPC converter has more than three levels additional problems arise. From the point of view of the power topology, the clamping diodes require to block higher voltage than the main switches, therefore it is necessary to use different technology or use several clamping diodes connected in series. In addition, the unequal use of the power devices of the topology (also present in the 3L-NPC for the inner power semiconductors), becomes critical. Finally, the reliability is also reduced due to the increment of the components count [20]. The mentioned drawbacks limit the use of NPC with more than 3 levels in industrial applications [21].

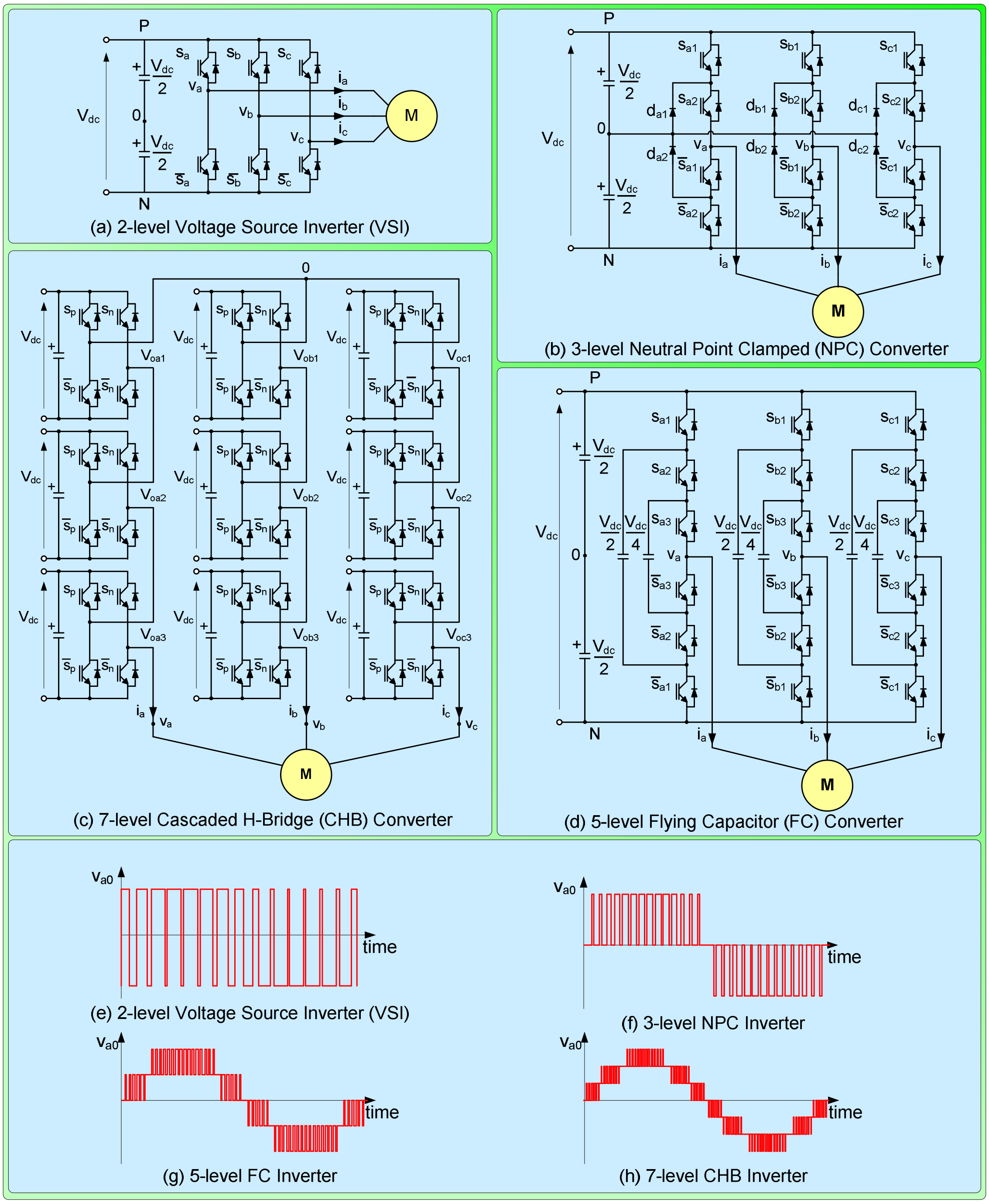

Figure 2. Topologies and phase voltages of the conventional two-level and multilevel voltage source inverters

Converters based on modular power cells as such as the cascaded H-bridge (CHB) and the flying capacitor (FC) converter have been proposed so far to provide higher number of output voltage levels than the 3L-NPC.

The CHB converter topology, shown in Figure 2c), is a highly modular converter based on several single phase inverters, usually called power cells, connected in series to form an phase [22][23]. Each power cell is implemented based on standard low voltage components, which provides an easy and cheap replacement in case of failure [24]. The main advantage of this converter is that using only low voltage components it is possible to drive medium voltage high power loads. Although the switching frequency in each cell is low, the equivalent switching frequency applied to the load is high, which reduces the switching losses, produces low dv/dt and helps to avoid resonances [25]. Additionally, several fault tolerant strategies can be implemented increasing their reliability and overall availability [26].

In the CHB converter, each power cell requires an isolated DC power source which can be obtained for example from photovoltaic (PV) panels or using a multi-pulse transformer with several secondaries diode rectifiers and a DC-link capacitor. The first option provides high scalability and allows the PV energy to be directly connected to the medium voltage grid. The second option has become the standard for drive applications, providing galvanic isolation and high quality input current, cancelling out the low order harmonics produced by the diode rectifiers. The cancellation of current harmonics is produced by a phase shift between the transformer secondaries which leads to a more complex transformer, especially when a high number of cells is used. This input transformer however reduces the scalability of the converter. The DC-link is designed to have a very large capacitance which, on one hand, allows having a good ride-trough capability but on the other hand increasing the size of each cell where at least half of its size are only capacitors and requires additional circuits to safely charge and discharge the DC-link.

As mentioned above, the standard solution to provide the isolated DC sources is using a diode rectifier, which does not provide regenerative operation and cannot be used in applications such as downhill conveyors, or GNL turbines where the load is always regenerating or, at least, great part of its operating time. In these cases a regenerative converter, where the diode based rectifiers are replace by controlled rectifiers, has been proposed [27]. This alternative can control bidirectional active power, which additionally gives precise control of the DC-link voltage, the input reactive power and provides sinusoidal input currents which help to simplify the input transformer configuration. The drawbacks of the regenerative topology are the requirement of inductive input filters and its associated cooling system, the increased losses by the forced commutation of the controlled rectifier and the reduction in the reliability [28][29].

Several topologies based on the cascaded approach have been proposed in the literature. In particular, the asymmetrical CHB converter, built using different values of the dc voltage for each power cell, allows to increase the number of output voltage levels using only a reduced number of power cells [30]. It can produce up to 27 output voltage levels using only three power cells (only nine levels conventional symmetrical CHB converter), but it requires the use of different devices in each power cell and a control system to avoid regeneration in the low voltage cells [31][32].

The FC converter topology is shown in Figure 2d). The output voltage is generated connecting directly the phase output to the positive or negative bars or through the floating capacitors. The number of output voltage levels depends on the number of floating capacitors and the relation among their different DC voltages. For example, the topology shown in Figure 2d) generates 5 output voltage levels. This converter, as in the CHB case, also presents a modular topology, where each cell is composed by a DC capacitor and two complementary switches. However, on the contrary of the CHB case, the addition of extra power cells in the FC converter does not increase the nominal power of the converter but only reduces the dv/dts improving the harmonic content of the output waveforms. As the CHB converter, the modularity reduces the cost of component replacement and maintenance, and also allows to implement fault tolerant strategies. Finally, the FC converter has also its asymmetrical version, and for instance if the floating DC voltage relations of the topology shown in Figure 2d) are changed to 3Vdc/7 and Vdc/7 it is possible to generate 8 different symmetrical output voltage levels [33][34]. As in the asymmetrical CHB case, the increase of levels is achieved at the expense of losing the converter modularity.

The FC converter requires only one DC source to feed all the cells and phases. Therefore the input transformer can be avoided and the number of cells can be arbitrary increased depending on the required output power. Similarly to NPC, this converter requires a control strategy to regulate the voltages in the capacitors.

Along with the classical structures there are several mixed structures, using for example single phase 3L-NPC as one power cell of the CHB [35], or combining and Active NPC with a FC [36]. Recently, the modular multilevel converter (MMC) has been proposed as a completely modular topology [37][38]. This converter can operate with DC or single-phase AC sources. The cells are based on DC-DC boost converter for the DC cell and a half bridge inverter for the AC cell, both cases with a floating DC capacitor. This converter does not require isolated DC sources because each cell has a floating DC capacitor. The number of cell in series in this converter is very large compared to CHB which generating very low dv/dt and a high equivalent switching frequency. This converter always requires a DC voltage control scheme because the floating DC voltages must be kept at the reference value. There are independent controllers implemented to handle the input and output currents and an additional controller to minimize the circulating current which is a current that flows only inside the converter. This circulating current is a particular characteristic of this converter [39].

IV. 2. Current Source Converters

The standard topology of a current source inverter (CSI) is shown in Figure 3a). These converters always require a controlled rectifier in order to provide a constant current flowing through the DC link. In the basic topology a thyristor-based rectifier is usually chosen. A split inductor is used in the DC-link in order to reduce the common mode voltage in the load [40]. The inverter is an arrangement of power semiconductors similar to the VSI, but composed by gate turn-off thyristors (GTO) or IGCT. The output current has a PWM waveform and can not be applied directly to an inductive load such as the motor. An output capacitive filter which eliminates the di/dt and applies a very smooth voltage to the load is mandatory [41]. This converter can achieve medium voltage operation and, moreover, is intrinsically regenerative [42][43]. A back-to-back configuration as shown in Figure 3b) can be used when high quality input current is required. In this case, the input side also requires a second order LC to filter due to the PWM currents presented at the input terminals of the rectifier [44].

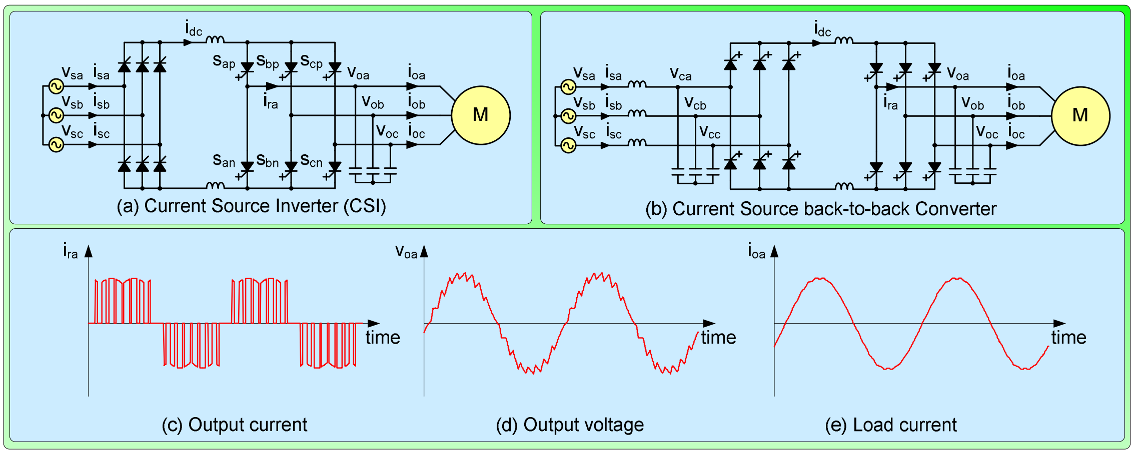

Figure 3. Topologies and output waveforms of the conventional current source inverters

IV. 3. Direct Converters

A third category of converters which do not require energy storage elements is presented in this section. The direct converter transmits the energy directly from the input side to the output side. The main advantage of these converters is the reduction in size but it comes with a complexity in the control scheme.

The cycloconverter, shown in Figure 4a), has been widely used in high power applications such as grinding mills and belongs to this category [45]. This converter is composed by a thyristor-based dual converter per phase, which can produce a variable DC voltage controlled in order to follow a sinusoidal waveform reference [46]. The input of each converter is fed by a phase shifted transformer where low order harmonics of the input current are cancelled out. The output voltage results in a combination of segments of the input voltages which fundamental component follows a sinusoidal reference as can be seen in Figure 4b). Due to its operation principle, this converter is well suited to drive high-power low-frequency loads [47][48].

The matrix converter, in its direct and indirect versions, also belongs to the direct converters category. The direct matrix converter (DMC) is shown in Figure 4c). The basic operation principle behind this converter is the connection of the phase output to any of the input voltages [49][50]. The converter is composed by nine bidirectional switches which can connect any input phase to any output phase allowing the current flow in both directions. A second order LC input filter is required to improve the input current [51]. The output is directly connected to the inductive load. Not all the possible combinations of switches are allowed, they are restricted to only 27 valid switching states. As mentioned earlier the main advantage of matrix converter is the reduction in size being specially suited for automotive and aircraft applications [52][53].

Several variations on the DMC have been proposed in the literature, the most important is the indirect matrix converter (IMC) which is composed by a bidirectional three-phase rectifier, a virtual DC link and a three phase inverter [54] as shown in Figure 4e). The number of power semiconductors is the same as the DMC, if the bidirectional switch is considered as two unidirectional switches, but the number of possible switching states is different and its analysis is simpler. Using the same configuration of the IMC it is possible to simplify its topology and reduce the components count restricting its operation to positive voltages in the virtual DC-link [55]. This reduced topology is called sparse matrix converter (SMC) and it is shown in Figure 4f).

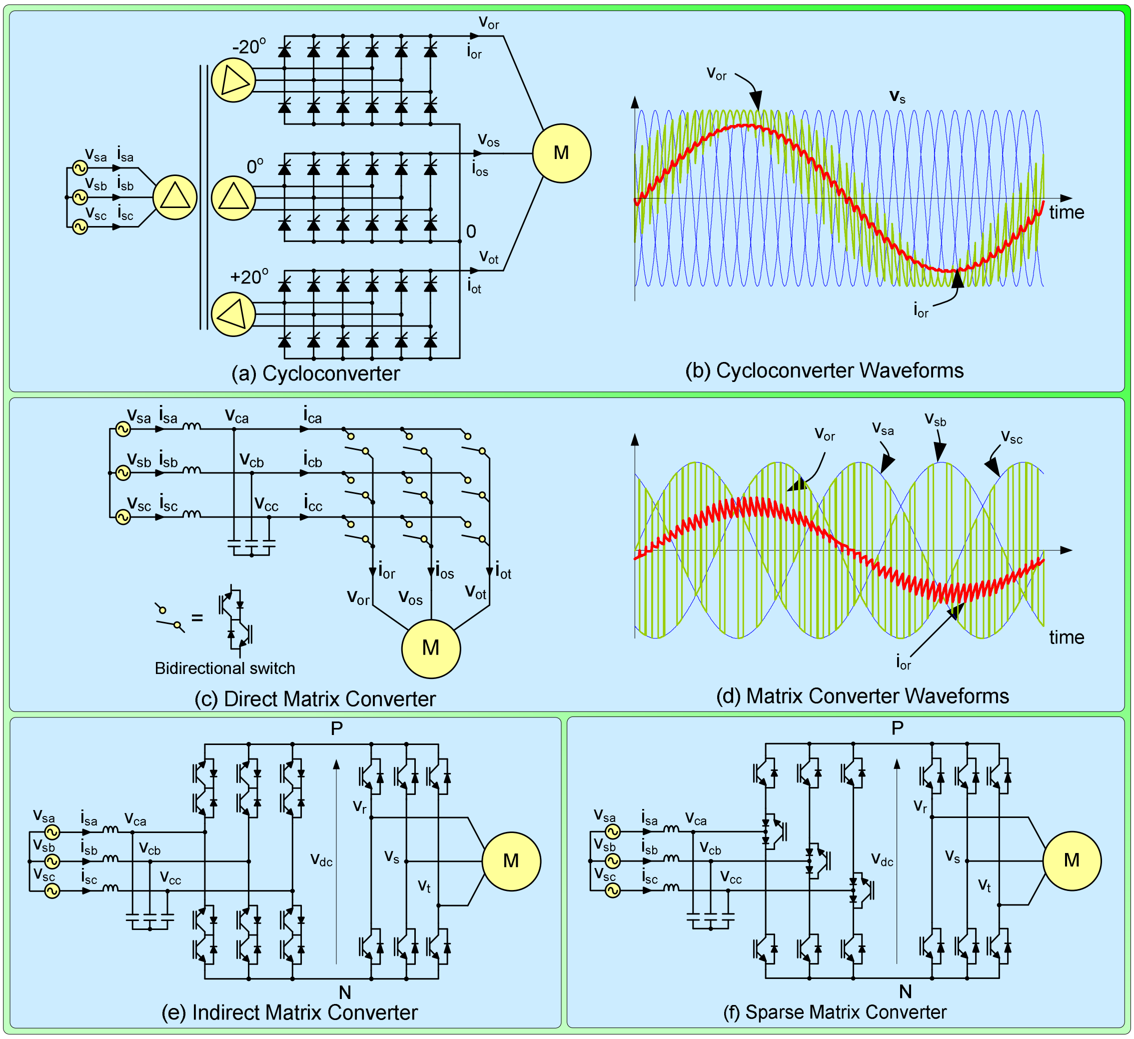

Figure 4. Topologies and output waveforms of the direct converters

III. MODULATION TECHNIQUES FOR POWER CONVERTERS

The vast development of power converters for a large number of applications in the last 50 years has lead to create very important research lines being implemented all over the world. Among these lines, the modulation has been one of the most important topics in the last decades because it is directly related to the efficiency of the overall power system affecting to the economical profit and performances of the final product.

The fundamental objective of a modulation technique is to obtain the best waveforms (voltages and currents) with minimum losses. Other secondary control objectives can be dealt with the proper modulation technique such as common-mode voltage reduction, dc voltage balancing, input current harmonics minimization, low dv/dt’s, among others. To achieve simultaneously all the control targets is impossible, so a trade-off is needed. Each power converter topology and each application has to be in depth studied to determine which modulation technique is the most suitable.

A classification of the modulation techniques for power converters is shown in Figure 5. In this classification, the modulation techniques are divided in four main groups: Pulse-Width Modulation (PWM), Space-Vector Modulation (SVM), harmonic control modulation and other variable switching frequency methods. These modulation techniques were applied to the power converters since several decades and are the most mature modulation techniques implemented in the power converters as commercial products. In the following sections, the proposed classification of modulation techniques will be explained in detail.

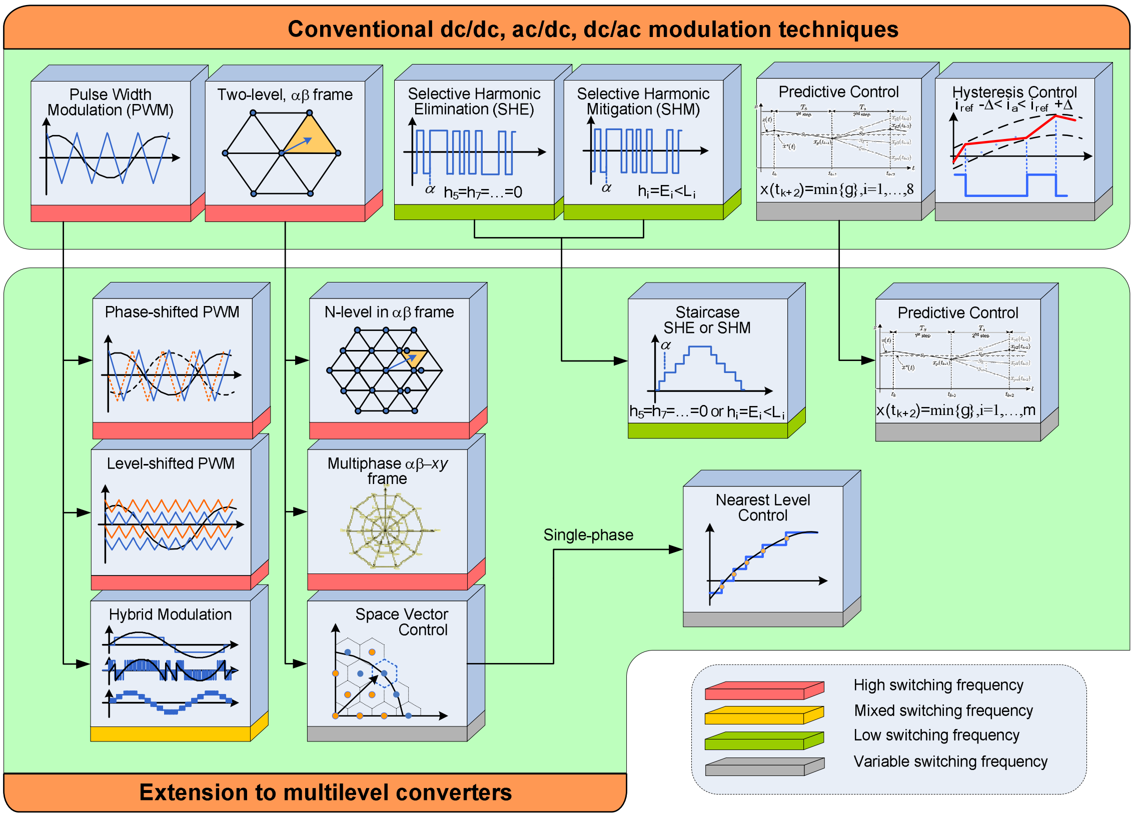

Figure 5. Classification of the most common modulation techniques for power converters

III.1. Pulse width modulation (PWM)

III.2.a. Conventional bipolar and unipolar PWM

Among the most commonly used modulation techniques, PWM is the most successful modulation method because of its high performance, simplicity, fixed switching frequency and easy digital and analog implementation since some decades ago [56]. The PWM technique is based on the comparison of a reference signal with frequency f_1, usually a sinusoidal reference, with a triangular carrier with frequency f_c. If the reference is above or below the triangular carrier, some power semiconductors are switched on and off or vice versa. As the power converter is inherently a non linear system, the obtained output voltage waveform is a switched signal with the possible output voltages of the power converter. The harmonic spectrum of this switched waveform has the fundamental harmonic equal to the fundamental harmonic of the reference signal f_1 but some harmonic distortion appear around the carrier frequency f_c (and their multiple values) due to the switching of the converter. The frequency ratio is defined as $R_f=f_c/f_1$ and has to be high enough in order to avoid the presence of low order harmonic distortion. So, R_f is usually limited to values higher than 10. Ideally, in order to obtain an output waveform with minimum distortion and to avoid the low frequency audible noise, R_f has to be very large. However, if R_f is high the number of commutations of the power semiconductors is large as well leading to inacceptable switching losses from the point of view of efficiency and power dissipation.

The basic PWM can be bipolar or unipolar generating an output waveform with two voltage or three voltage levels respectively using one reference and one carrier (bipolar PWM) or one reference and two opposite carriers (unipolar PWM). For instance, if a conventional two-level three-phase VSI is considered, a bipolar PWM technique can be applied obtaining a phase voltage which is a switched voltage with values $-V_{dc}/2$ and $V_{dc}/2$. On the other hand, if an H-bridge is considered, an output voltage with values $-V_{dc}/2$, 0 and $V_{dc}/2$ can be obtained if a unipolar PWM is applied.

A way to achieve a better utilization of the dc voltage of the converter can be obtained by a third harmonic injection in the reference voltage [57]. In this way, the modulation index (related to the maximum voltage that can be generated by the converter) can be extended up to 15.47%. The third harmonic injection does not represent a problem because it is canceled in the line-to-line voltages. An easier way to carry out this third harmonic injection is to use a min-max sequence injection using the following expression:

$$v_{a_{comp}}^* = v_a^* - \frac{\min\{v_a^*, v_b^*, v_c^*\} + \max\{v_a^*, v_b^*, v_c^*\}}{2}$$

$$v_{b\_comp}^* = v_b^* - \frac{\min\{v_a^*, v_b^*, v_c^*\} + \max\{v_a^*, v_b^*, v_c^*\}}{2}$$

$$v_{c\_comp}^* = v_c^* - \frac{\min\{v_a^*, v_b^*, v_c^*\} + \max\{v_a^*, v_b^*, v_c^*\}}{2}$$

where v_a^*, v_b^* and v_c^* are the reference voltages for phase a, b and c respectively and $v^*_{a\_comp}$, $v^*_{b\_comp}$ and $v^*_{c\_comp}$ are the modified reference voltage for phase a, b and c respectively. The min-max sequence injection has the advantage of avoiding the presence of a PLL because it is not needed to know the angle of the reference voltage (in the conventional third harmonic injection a synchronization is needed).

The PWM technique has been extensively applied to power converters for any application from low power to medium-high power. It has to be noticed that for high power applications, the switching losses have to be taken in consideration and R_f has to be as lowest as possible. This fact leads to create low harmonics distortion because the carrier frequency is inside the bandwidth which has to be kept with maximum quality. Therefore, passive filters usually formed by inductors and capacitors have to be part of the power system when PWM is used for high power applications. The design of these passive filters is extremely important because they are very bulky, heavy and expensive. So, there is a trade-off between switching losses and cost of the filter.

III.2.b. Extension of PWM to multilevel converters

PWM technique is suitable to be used in simple dc/dc, ac/dc and dc/ac converters with low number of power semiconductors per phase. Additional considerations have to be taken into account if a PWM technique is applied to more complex power converter topologies. Among these complex new converter topologies, the multilevel converters are especially attractive for high-power applications due to the increase of nominal power at the expense of a high number of medium-low voltage power semiconductors. This makes that a conventional bipolar or unipolar PWM cannot be directly applied to the topology to determine the switching states of the power devices.

The PWM techniques for multilevel converters are an extension of the conventional PWM techniques increasing the number of triangular carriers to carry out the comparisons with the reference voltage. In this way, as is shown in Figure 5, the most common PWM methods for multilevel converters are the level-shifted and the phase-shifted PWM techniques [58].

Each one of these multicarrier methods is specifically well suited to be applied to a multilevel converter topology. In fact, the level-shifted PWM technique can be easily applied to the neutral-point-clamped converter because the switching of the converter is directly related to the comparisons of the reference voltage with the triangular carriers. On the other hand, the phase-shifted PWM technique is especially well designed to be applied to the flying-capacitor topology. In this case, a bipolar PWM is applied to each power cell, but introducing a phase shift in the triangular carrier of consecutive power cells equal to $\pi/(N-1)$ where N is the number of levels of the converter. In addition, the use of the phase-shifted modulation in a symmetrical flying-capacitor converter leads to achieve a natural balance of the flying capacitor voltages without using sensors. Finally, the phase-shifted PWM technique is also well suited to be applied to the cascaded H-bridge converter. In this case, an unipolar PWM is implemented in each H-bridge with a phase shift in the triangular carriers of consecutive H-bridges equal to $\pi/(2m)$ where m is the number of H-bridges per phase of the converter.

Both modulation techniques can be compared in terms of harmonic distortion and switching losses. It can be said that, assuming the same switching frequency of the output voltage, the level-shifted PWM method achieves a better harmonic performance because it avoids the appearance of side bands around the carrier frequency. However, it is achieved at the expense of non-equally distributed switching losses between the power semiconductors and the consequent issues about asymmetrical power dissipation and reliability reduction. This problem is not present if the phase shifted modulation is applied because it intrinsically imposes a very similar switching pattern to each power cell of the converter. In addition, the effective switching frequency of output voltage is m times the carrier frequency used in the bipolar or unipolar PWM method (where m is the number of power cells of the converter). This makes that the phase shifted modulation is superior to the level shifted method in terms of harmonic distortion as well.

III.3. Space vector modulation (SVM)

III.3.a. Conventional SVM technique

The space vector modulation is an alternative technique to determine the switching gate signals of a power converter. Unlike the PWM method, the SVM technique does not use triangular carriers and comparators to generate the switching of a power converter. The SVM uses a vector representation of the possible phase to neutral voltages than can be generated by the converter [59]. As the converter is a non linear discrete system, it has only a discrete number of switching states which generate a limited number of phase voltages which can be plotted in the $\alpha\beta\gamma$ frame. This representation defines the possible vectors which can be used by the converter (i.e. seven different vectors for the eight possible switching states of the conventional three-phase two-level converter). The γ component of these vectors is zero in power systems where the neutral current is zero and this leads to a two-dimensional representation using the $\alpha\beta$ frame. The reference voltage is represented as well using the same frame and it can be generated as a linear combination of the three nearest switching vectors. It has to be noticed that the conventional SVM technique has been claimed superior to PWM due to its better dc-link voltage utilization but, in fact, the SVM is a vector representation of the PWM method with a third harmonic injection. Despite of this fact, the SVM technique is very interesting because the switching sequence can be changed manually taking into account control targets such as common-mode voltage elimination, current ripple reduction and switching losses minimization, among others.

III.3.b. Extension of SVM for multilevel converters

The extension of the conventional SVM technique for multilevel converters is straight forward because it consists on the inclusion of all the additional switching states in the αβγ frame. As the number of switching states of a multilevel converter is higher than the two-level case, a high number of switching vectors appear. In order to generate the reference vector, again the linear combination of the three closest vectors has to be used [60]. However, it has to be noticed that, in the multilevel case, there are several redundant vectors different of the zero vector. This means that different switching states are represented using the same vector in the αβγ frame. Any of these redundant vectors can be used in the switching sequence without distorting the line-to-line voltage and therefore this concept can be used in order to improve some converter feature such as the dc voltage balancing and the switching losses equalization or minimization.

In the last years, new SVM techniques based on the natural cartesian *abc* coordinates have been proposed [61]. The resulting modulation technique is a three-dimensional SVM technique and is similar in terms of computational cost compared with conventional SVM methods based on the αβγ representation. However, the *abc* frame based SVM techniques can be also applied to converter topologies where the common-mode current is not zero. In addition, it naturally eliminates the common-mode voltage average over a sampling period at the expense of not dealing with the third harmonic injection. This is achieved determining the specific duty cycle of each redundant switching vector of the switching sequence which is the difference with the conventional αβγ frame based SVM technique when is applied to power converters with common-mode current equal to zero. On the other hand, in [62] it is demonstrated that the *abc* frame based SVM techniques are equivalent to a time based level-shifted PWM method applied to each phase of the converter.

III.4. Harmonic control techniques

III.4.a. Selective harmonic elimination technique

Selective harmonic elimination (SHE) is especially well suited for high-power drives where the switching frequency is restricted by the switching losses. SHE technique is used to achieve very low switching frequency avoiding the presence of low order harmonics. The conventional SHE algorithm is based on the calculation of a reduced number of switching angles of the output voltage pattern to eliminate a given number of harmonics using the Fourier decomposition [63]. The resulting expressions are non-linear and very difficult to solve online. For this reason the switching angles are determined offline using mathematical search algorithms such as genetic, particle swarm, tabu and simulated annealing, among others. The resulting angles are stored in a lookup table for a range of modulation indexes and interpolated to generate the switching pattern.

As the SHE technique is based on the Fourier transform calculation of a period of the voltage waveform, it is difficult to apply this modulation in fast dynamic close loop applications. Several approaches to calculate online the switching angles have been proposed in the literature [64],[65].

III.4.b. Selective Harmonic Mitigation Technique

The grid requirements, imposed by the governments and the electrical providers, usually define maximum limits of the harmonic distortion for each harmonic order up to 50th and a maximum value of the total harmonic distortion (THD). In this way, the elimination of low order harmonic components by the SHE technique does not ensure to comply with harmonic restrictions imposed by the current grid codes. The non eliminated harmonics present when the SHE technique is applied can take any value and mostly are above the limits imposed by the grid codes. This fact leads to design highly cost, bulky and heavy filters in order to smooth the output waveforms keeping the harmonic distortion below the limits and therefore fulfilling the grid code requirements.

To overcome this problem, a modified modulation method, based on the SHE modulation concept, has been proposed where the switching angles are calculated not to eliminate a given number of low order harmonic components but to reduce the distortion below the limits imposed by the grid codes. This modulation technique is called Selective Harmonic Mitigation (SHM) [66]. The SHM method can be used in all the applications where the SHE technique operates obtaining better results in terms of the resulting necessary filter to be used in the power system. The SHM technique is based on the offline calculation of the switching angles defining and cost function to be minimized by a search algorithm. This cost function includes terms for the harmonic distortion components, the THD and the minimum time between consecutive switching angles. If the harmonic distortion components or the THD value are below the limit imposed by the applied grid code, its contribution to the cost function is zero.

III.5.Variable frequency modulation techniques

The fixed switching frequency modulation methods are superior to the variable frequency ones in terms of better harmonic distortion distribution. The fixed frequency methods present a harmonic spectrum where the switching distortion is located around the switching frequency and its multiples. This does not happen when a variable frequency modulation method is used because a spread harmonic spectrum is generated. However, in some specific conditions, variable frequency modulation methods, which are very simple mathematically and conceptually, can be also applied achieving good results. It has to be noticed that the variable frequency modulation techniques are not indeed modulation methods because they do not synthesize a reference voltage as an average of the discrete switching states of the power converter. In this way, the concept of switching sequence is not used in these techniques and the switching of the power devices only occurs when some control conditions are fulfilled. This is why the switching frequency of the output waveforms is not constant.

III.5a. Hysteresis control

Also called bang-bang or on-off control, it consists on the determination of the power devices switching depending on a comparison between the measured control variable x_m (usually a current) with a reference waveform x_{ref} with a hysteresis band $\pm\Delta x$. The aim of the controller is keep the control variable inside the hysteresis band leading to a non-periodic switching of the power devices of the converter topology. The control variable ripple is directly defined by the hysteresis band until the hysteresis band is smaller than the dynamic response of the load. In order to design a hysteresis controller, the effect of the switching of the converter on the control variable has to be known. In this way, the controller design is very simple at the expense of generating a spread harmonic spectrum [67].

III.5b. Nearest level control

The nearest level control determines the switching of the power converter comparing the phase voltage reference with the possible voltages that can be generated by the converter topology. Finally, the switching associated to the nearest voltage level is applied and, in this way, as only one switching state is used during each sampling period, the switching losses are reduced compared with conventional PWM or SVM techniques [68]. This is achieved at the expense of a more distorted output voltage because the reference voltage is not generated in average over a sampling period. The nearest level control can only be applied achieving high performance in power converters where the difference between consecutive voltage levels is not large, which means that is specially well designed to be applied to multilevel converters with a high number of levels (for instance for a cascaded h-bridge converter with more than four power cells per phase).

III.5c. Space vector control

As has been explained above, the nearest level control is a PWM method where only the closest switching state of the switching sequence is finally applied to the converter. The space vector control applies the same concept to the SVM technique. In this way, instead of using a switching sequence formed by the state vectors closest to the reference vector, only the closest one is applied to the power converter [69]. As in the nearest level case, the space vector control reduces the switching losses at the expense of a higher distortion and a spread harmonic spectrum of the output voltages and currents. In the same way, this technique has to be used in power converter topologies where the geometrical distance between the reference vector and the closest vectors is not large. Therefore, as in the nearest level control, it is suitable to be used in multilevel converter with a high number of levels.

III.5.d. Model predictive control

Model predictive control is performed in discrete time, where in each sample time the output current is predicted for each one of the valid switching states using a model of the load and the converter. All the predicted values are evaluated in a cost function and the switching state that minimizes this function is selected to be applied at the next sample time [70]. Similarly to hysteresis, this control has a time-based operation, producing a variable switching frequency and consequently a spread harmonic spectrum. One of the main advantages of this algorithm is its flexibility, because the cost function can be easily modified including simultaneously several primary and secondary control objectives such as a frequency-related term to improve its frequency behavior, the dc voltage balance in a NPC converter, switching frequency and common-mode voltage reduction among others. At the present state of the art, the coefficients (usually called weighting factors) of the different terms which form the cost function are determined by empirical procedure. There is no analytical or numerical solution proposed yet to obtain an optimal solution to the problem for FSC-MPC. In [71], some guidelines are presented to help the weighting factor design process.

IV. CONTROL AND ESTIMATION

IV.a. Introduction

The general block scheme of high performance speed controlled AC motor drives is shown in Figure 6. The core of the scheme are internal flux and torque control loops with the estimator block which can be implemented in different ways, whereas the outer speed control loop is rather unified and generates command values for torque M_c and flux Ψ_c (via Flux program block) controllers. The speed feedback signal can be measured by a mechanical motion (speed/position) sensor Ω_m or delivered from the estimator Ω_m creating possibility of the speed sensorless operation.

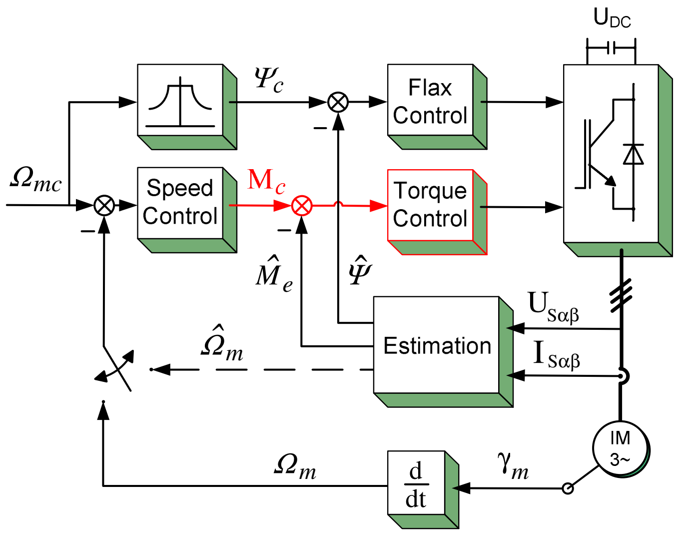

Figure 6. Block diagram of the speed controlled AC motor drive with internal flux and torque loops

IV.2. Control Methods

Several Torque Control (TC) methods have been developed in the last 40 years. Not all of them have found wide industrial applications. Therefore, we present only the most popular strategies used commercially and some of future trends. Existing TC methods can be classified in different ways [72][73]. In this paper, the TC methods are presented in two main groups: *linear* and *nonlinear* controllers. The discussed generic TC methods are presented for the induction motor drives; however, they can be easily expanded for control of permanent magnet synchronous motors (PMSM) with sinusoidal electromotor force (EMF) and currents.

IV.3. Linear Torque Control

The linear torque controllers operate in association of voltage pulse width modulators (PWM). The controllers calculate the required stator voltage vector, averaged over a sampling period. The voltage vector is finally synthesized by a PWM technique which in most cases is the SVM. So, differently from the nonlinear TC schemes where signals are processed on instantaneous values, in a linear TC scheme, the linear (PI) controllers operate on values averaged over the sampling period. Therefore, the sampling frequency can be reduced from about 40 kHz in nonlinear TC, to 2-5 kHz in linear TC schemes. In the linear group, the following TCs are described: Field Oriented Control (FOC), Direct Torque Control with voltage Space Vector Modulation (DTC-SVM) and Direct Torque Control with Flux Space Vector Modulation (DTC-FVM).

1) Field Oriented Control

The Field Oriented Control, proposed in 1970-ties by Hasse [74] and Blaschke [75], is based on an analogy to the mechanically commutated DC brush motor. In this motor, owing to separate exciting and armature winding, flux is controlled by exciting current and torque is controlled independently by adjusting the armature current. So, the flux and torque currents are electrically and magnetically separated. Contrarily, the cage-rotor IM has only a three-phase winding in the stator, and the stator current vector, $\mathbf{I}_s$, is used for both flux and torque control. So, exciting and armature current are coupled (not separated) in the stator current vector and cannot be controlled separately. The decoupling can be achieved by the decomposition of the instantaneous stator current vector, $\mathbf{I}_s$, into two components: flux current, I_{sd}, and torque-producing current, I_{sq}, in the rotor-flux-oriented coordinates (R-FOC) *dq* (see vector diagram in Figure 7). In this way, the control of the IM becomes identical with a separately excited DC brush motor and can be implemented using a current controlled PWM inverter with linear PI controllers and voltage SVM (see block scheme in Figure 7). The core of the FOC scheme are coordinate transformation blocks which allow calculation of field oriented current components I_{sd}, I_{sq} by using αβ/dq transformation, and reference voltage vector components $V_{s\alpha c}$, $V_{s\beta c}$ by using inverse dq/αβ transformation. So, in the FOC torque and flux is controlled indirectly by field oriented current vector components.

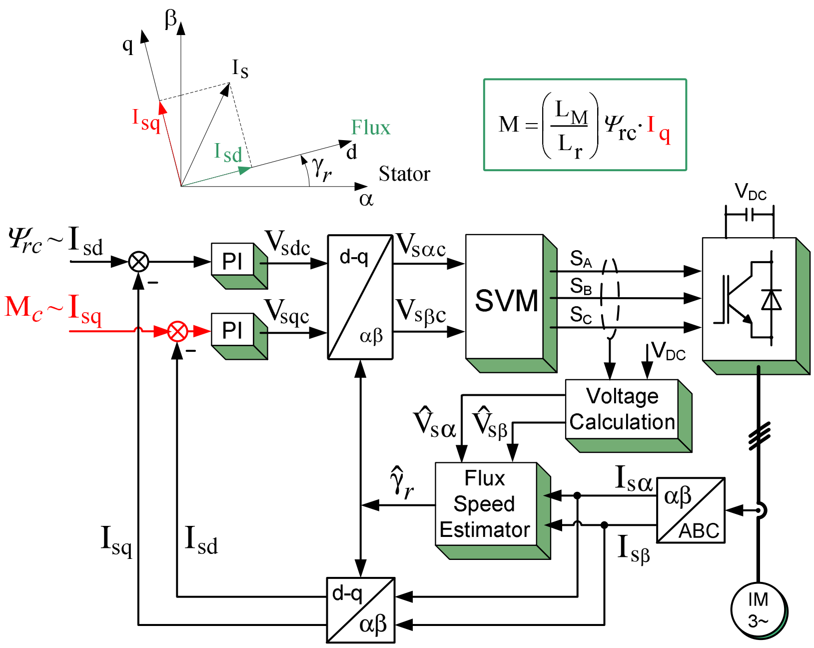

Figure 7. Vector diagram and block diagram of rotor FOC. Torque is controlled indirectly via torque current I_{sq} control loop

2) *Direct Torque Control with Voltage SVM*

A block scheme of DTC-SVM with closed-loop torque and flux control operating in Cartesian stator flux coordinates [76][77] is presented in Figure 8. The output of the PI flux and torque controllers is interpreted as the reference stator voltage components, $V_{\Psi c}$ and V_{Mc}, in S-FOC (*dq*). These DC voltage commands are then transformed into stationary coordinates (αβ), and the commanded values, $V_{s\alpha c}$ and $V_{s\beta c}$, are delivered to the SVM block. Note that this scheme can be seen as a simplified stator flux oriented control (S-FOC) without current control loops [78] or as a classical ST-DTC scheme (see

Figure 10) in which switching table is replaced by a modulator (SVM) and a hysteresis torque and flux controllers are replaced by linear PI [76][77][79]. So, in the DTC-SVM scheme torque and flux are controlled directly in closed loops, and therefore an accurate estimation of motor flux and torque is necessary. Contrarily to classical hysteresis based DTC, the DTC-SVM operates at constant switching frequency. This improves considerably the drive performance in terms of reduced torque and flux pulsations, reliable start-up and low speed operation.

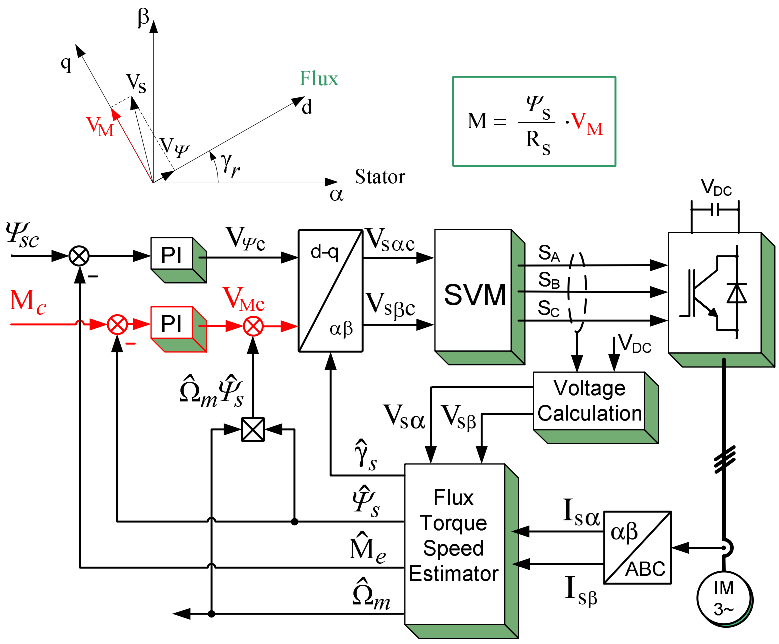

Figure 8. Vector diagram and block scheme of DTC-SVM. Torque is controlled directly via voltage vector component V_M

3) Direct Torque Control with Flux SVM

Further simplification can be achieved in a block scheme of DTC with flux vector modulation - FVM as shown in Figure 9. For torque regulation, a PI controller is applied and its output produces an increment in the torque angle, $\Delta\delta_\Psi$ (see vector diagram in Figure 9) [80][81]. Assuming that the rotor and flux magnitudes are approximately equal, the torque is controlled only by changing the torque angle, δ_Ψ, which corresponds to the increment of the stator flux vector $\Delta\boldsymbol{\Psi}_s$. The commanded stator flux vector is calculated by addition of the estimated flux position γ_s and change of the torque angle $\Delta\delta_\Psi$. Its value is compared with the estimated flux and the stator flux error $\Delta\boldsymbol{\Psi}_s$ is used directly for calculation of VSI switching states in the FVM block [82][83]. Thanks to internal stator flux loop used for calculation of $\Delta\boldsymbol{\Psi}_s$ in flux pulse width modulator, the flux PI controller of Figure 8 is eliminated.

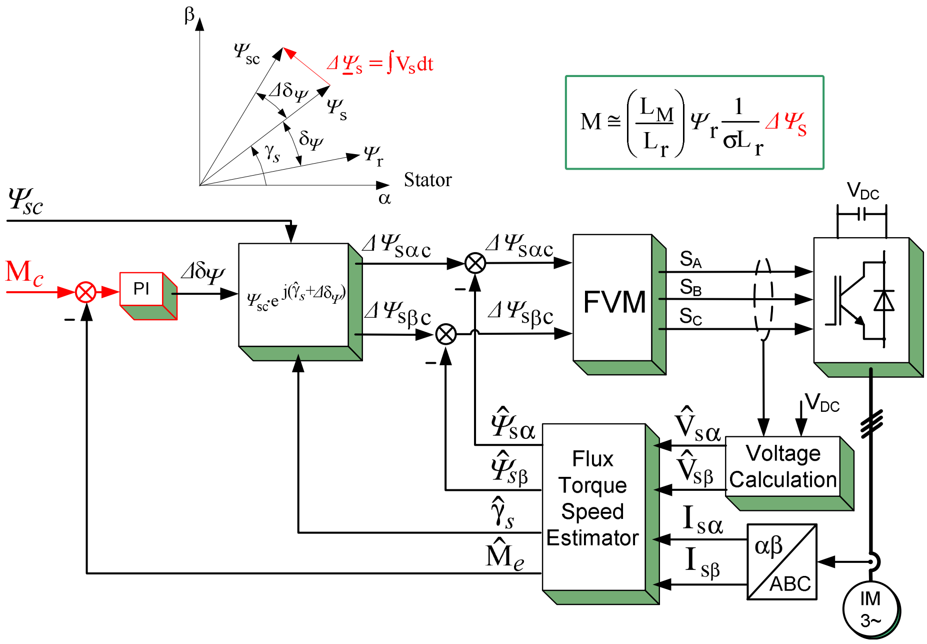

Figure 9. Vector diagram and block scheme of DTC-FVM. Torque is controlled directly via stator flux vector increment $\Delta\boldsymbol{\Psi}_s$ with Flux Vector Modulation (FVM)

IV.4. Nonlinear Torque Control

The presented nonlinear TC group departs from the idea of coordinate transformation and the analogy with DC motor control, which is basis for the FOC. It proposes to replace the decoupling control with the bang-bang control, which meets very well with on-off operation of the inverter semiconductor power devices. Compared to the conventional FOC (Figure 7), the DTC schemes have the following features:

- Simple structure,
- There is no current control loops and current is not regulated directly,
- Coordinate transformation is not required,
- There is no separate voltage pulse width modulator,
- Speed sensor is not required,
- Accurate stator flux vector and torque estimation is required.

This section includes the Switching Table based Direct Torque Control (ST-DTC), Direct Self Control (DSC) and on-line optimized Model Predictive DTC. Also, neural networks (NN's) and fuzzy logic controllers (FLC's) belong to the class of nonlinear control.

1) Classical ST-DTC Scheme

The block diagram of the classical ST-DTC scheme is shown in

Figure 10 [84]. The stator flux magnitude $\varPsi_{sc}$ and the motor torque M_c are the command signals which are compared with the estimated $\hat{\varPsi}_s$ and $\hat{M}_e$ values, respectively. The digitized flux and torque errors generated by the hysteresis controllers $d_{\varPsi}$, d_M and the position sector $N(\gamma_s)$ of the stator flux vector obtained from the angular position $\gamma_s = arctg(\Psi_{s\beta} / \Psi_{s\alpha})$ selects the appropriate voltage vector from the switching selection table. Thus, pulses S_A, S_B, S_C for control the inverter power switches are generated from the vector selection table. The characteristic features of the ST-DTC scheme of

Figure 10 include:

- Sinusoidal stator flux and current waveforms with harmonic content determined by the flux and torque controller hysteresis tolerance bands,
- Excellent torque dynamics (depending on voltage reserve),
- Flux and torque hysteresis bands determine the inverter switching frequency, which varies with the synchronous speed and load changes.

Many modifications of the classical ST-DTC scheme aimed at improving starting, overload conditions, very low speed operation, torque ripple reduction, variable switching frequency functioning, and noise level attenuation have been proposed during the last decade [7].

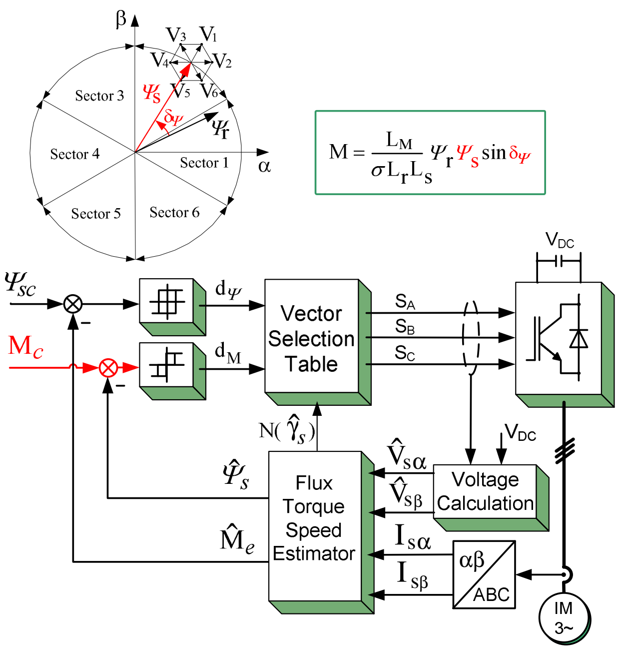

Figure 10. Vector diagram and block scheme of switching table based DTC. Torque is controlled directly via stator flux vector movement by selection of appropriate forward/backward active voltage vector (V_1 or V_6) and stops by selection zero voltage vector V_0. Stator flux vector moves on circular path

2) *Direct Self Control (DSC) Scheme*

The block diagram of the DSC method is shown in

Figure 11 [85]. Based on the command stator flux Ψ_{sc} and the actual phase components $\Psi_{sA}, \Psi_{sB}, \Psi_{sC}$, the flux comparators generate digital variables d_A, d_B, d_C, which corresponds to active voltage states (V_1–V_6). The hysteresis torque controller generates signal d_M, which determines zero states. So, the Stator flux controller imposes the time duration of the active voltage states, which move the stator flux along the commanded trajectory, and torque controller determinates the time duration of the zero voltage states, which keep the motor torque in the defined-by-hysteresis tolerance band.

The characteristic features of the DSC scheme of

Figure 11 are:

- Non-sinusoidal stator flux and current waveforms that, with the exception of the harmonics, are identical for both PWM and six-step operation,
- The stator flux vector moves along a hexagon path also under PWM operation,

- No voltage supply reserve is necessary and the inverter capability is fully utilized,
- The inverter switching frequency is lower than in the ST-DTC scheme of
-
-
- Figure 10,
- Excellent torque dynamics in constant and weakening field regions.

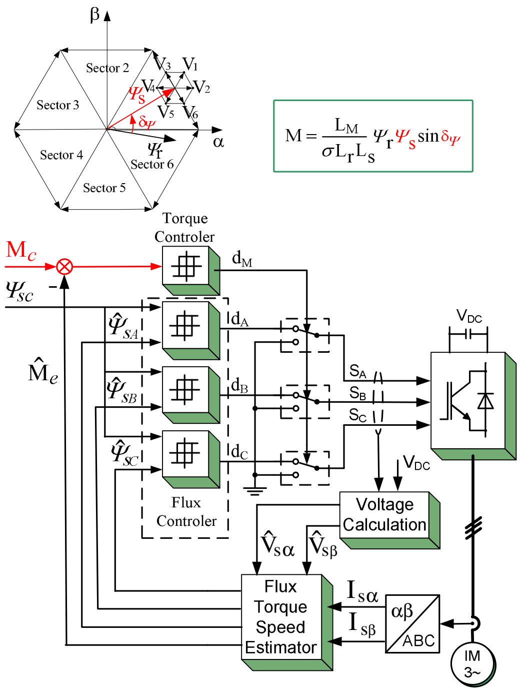

Figure 11. Vector diagram and block scheme of DSC. Torque is controlled directly in similar way as in the ST-DTC scheme, however, stator flux vector moves on hexagonal path because of different sector definition

Note, that the behavior of a DSC scheme can be reproduced by a ST-DTC scheme from Figure 10 for flux hysteresis of 14% wide.

3) *Predictive DTC Scheme*

The current and future trend in control of AC motors is to incorporate more advanced techniques like the model predictive control - MPC [86][87]. This trend is supported by the DSP producers which not only increase the operation speed and computation capacity, but also expand their offer by specialized circuits (for example Texas Instruments C2000 family). Predictive control is a very wide class of controllers which uses the model of the system (called *predictive model*) for the prediction of the future behavior of the controlled variables. This information is used for the controller to calculate the optimal actuation which minimizes a predefined cost function.

The simplified block diagram of the predictive DTC scheme is shown in Figure 12. This type of the MPC is classified as MPC with finite control set FCS (seven switching states are available in two level VSI) and, contrarily to a MPC with continuous control set [86][87], operates without pulse width modulators because the inverter switching state is calculated on-line by minimization a cost function. In most cases the cost function is defined as a weighted sum of torque and flux errors, however other components can be added (for example: reducing unnecessary switching, shaping frequency spectrum, maintaining the voltage balance of the dc-link capacitors in case of multi-level VSI, reducing common voltage, etc.) increasing the flexibility and priority in predefinition of system performances. Among important advantages of the MPC are:

- Intuitive and easy to understand concept,
- Easy inclusion of nonlinearities in the model and simple treatment of constraints,
- Easy to implement,
- Open to include modifications and extensions depending on specific applications.

However, these advantages are offset by problems when implementing predictive DTC:

- Larger amount of on-line calculations, compared to classical controllers,
- The accuracy of the predictive model has a direct influence on the quality of the predictive controller.

As reported in the works by G. Papafotiou et. al. (see in [86][87]) by using a long output horizon predictive DTC, the average switching frequency, and as consequence also inverter switching losses, has been reduced 16,5% with respect to the classical DTC used by ABB while maintaining the same control quality of the motor torque, magnitude of stator flux and 3-level inverter's neutral point voltage. This illustrates a very high improvement potential of the MPC.

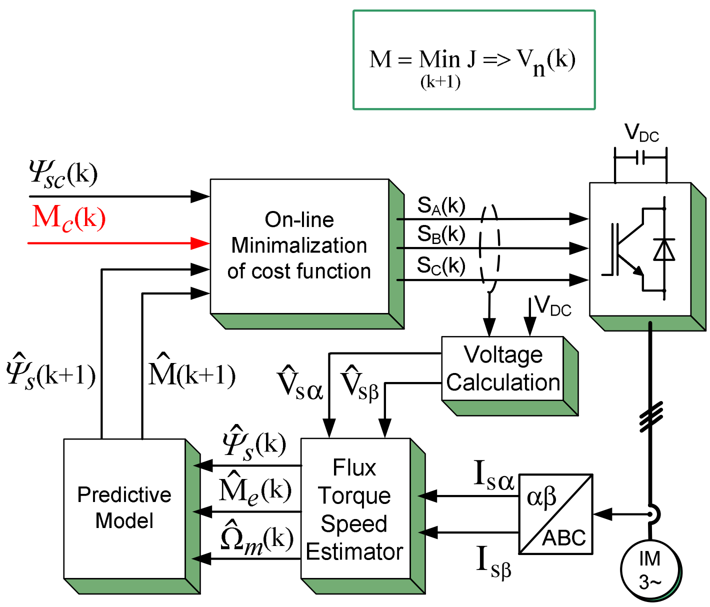

Figure 12. Block diagram of the predictive DTC. Torque is controlled directly by on-line minimization of the cost function in every sampling period

4) Neuro-fuzzy Scheme

In the last decade an interest in development of artificial intelligence based controllers in the area of power electronics and drive control has been observed. Especially, the combination of fuzzy logic and artificial NNs has been proved to be powerful as it offers advantages of both techniques. Several vector control schemes with neuro-fuzzy controller has been proposed and described in the literature [2, 73, 85, 90]. However, at present they represent only some alternative solutions to other existing techniques, and their specific applications areas cannot be clearly defined.

IV.5. Estimation

Implementation of any high performance drive system requires a high accuracy estimation of the actual stator or/and rotor flux vector (magnitude and position) and electromagnetic torque. Once the flux vector is accurately estimated, the torque estimation is performed easily as a cross product of the flux and measured stator current vectors.

IV.5a. Flux vector estimation

To avoid the use of flux sensors or measuring coils in the IM, methods of indirect flux vector generation have been developed, known as *flux models* or *flux estimators*. These are models of motor equations which are excited by appropriate easily measurable quantities, such as stator voltages and/or currents ($\underline{V}_s$, $\underline{I}_s$), angular shaft speed (Ω_m) or position angle (γ_m). There are many types of flux vector models, which usually are classified in terms of the input signals used [89]-[91]. Such models generate the stator or/and rotor flux vector which, in an ideal case, rotates synchronously with the motor magnetic field. Because the motor parameters very often are known only rough, and change during motor operation, therefore an error appears between the motor flux and that estimated in the model. The error depends on: model variants, parameter deviation between motor and model, accuracy of input signal measurement, motor point of operation. To minimize the sensitivity of flux models to motor parameters changes, the model adaptive reference systems (MARS) and the observer technique [76][90][92] can be used. Also, sliding mode approach for robust flux estimation has been proposed [93].

An example of early developed [94] very simple flux vector observer based on voltage model with compensation signal calculated from the commanded $\underline{\Psi}_{Rc}$ and estimated $\underline{\Psi}_R^{\wedge}$ rotor flux vector is presented in Figure 13. It operates without the measured speed signal, therefore can be used in a speed sensorless drives. At very low speed the flux vector is generated by the reference value $\underline{\Psi}_{Rc}$ because the stator voltage signal is almost zero. For higher speed level (determinated by selection of the compensation gain $\underline{K}_{a,b}$) the voltages signal becomes dominant and the flux vector is calculated by the voltage model. This observer allows starting up and breaking the drive system, however, control torque operation at zero speed is not possible.

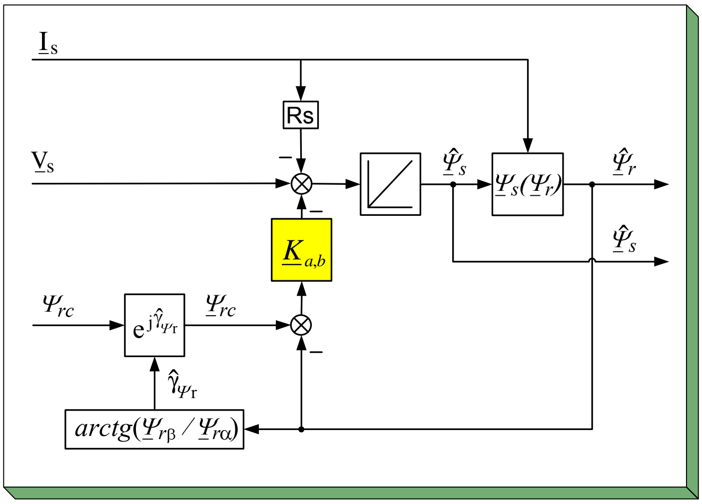

Figure 13. Simple flux vector observer based on voltage model with compensation signal calculated from the commanded $\underline{\Psi}_{Rc}$ and estimated $\underline{\Psi}_R^{\wedge}$ rotor flux vector [21]

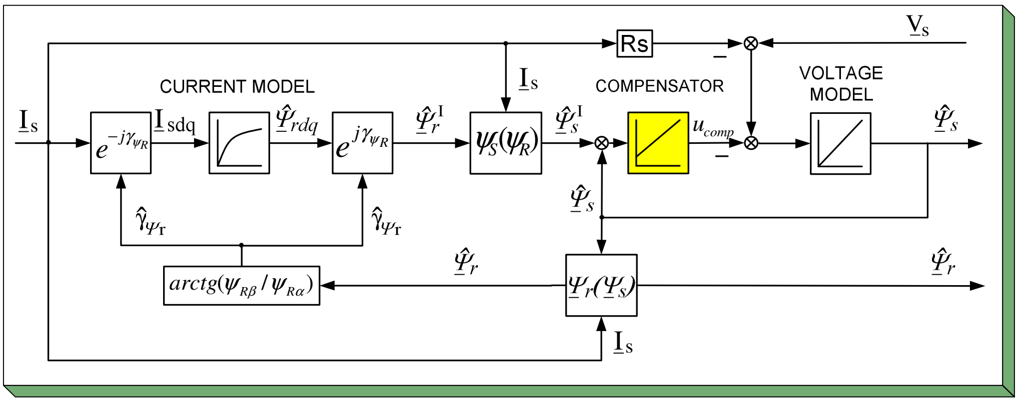

Figure 14. Speed sensorless flux vector observer based on “voltage model” compensated by “current model” [76]

Another flux vector observer also operating without measured speed signal is shown in Figure 14. The stator flux vector is calculated by the “voltage model” and compensated by the “current model” operating in the rotor flux oriented coordinates *dq*. At higher speed the voltage model guarantees better accuracy because the stator resistance is small while, at low speeds, the current model is better and operates even at zero frequency. The transition between the two models is defined by design of the PI controller which generates compensation voltage u_{comp}.

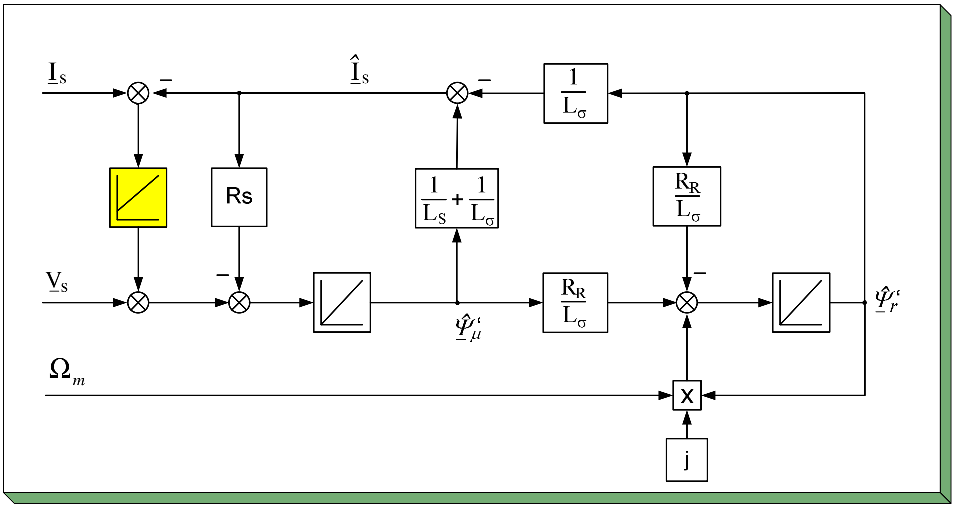

Figure 15. Flux vector observer based on measured $\underline{V}_s$, $\underline{I}_s$, Ω_m with compensation signal calculated from measured $\underline{I}_s$ and estimated $\underline{I}_s^{\wedge}$ stator current vectors [85][95]

Finally, in Figure 15, an flux vector observer based on measured stator voltage $\underline{V}_s$, current $\underline{I}_s$, and mechanical speed Ω_m with compensation signal calculated from measured $\underline{I}_s$ and estimated $\underline{I}_s^{\wedge}$ stator current vectors is shown. In this observer the motor model is fed with current vector estimated from the stator and rotor flux. The measured stator current is delivered only to PI controller for calculation the compensation voltage u_{comp}. Therefore, the model can be easily used for testing and investigation. With mechanical motion sensor it operates well at zero speed [85][95].

IV.5b. Speed estimation

There is a strong trend to avoid mechanical motion (speed/position) sensors because it reduces cost and improves reliability and functionality of the drive system. Various methods of the rotor speed estimation has been developed for induction motors, however the detailed analyze is beyond scope of this paper. A good review of IM speed sensorless control schemes is presented in [85] and [89].

IV.6. Closing remarks and overview

The overview of basic parameters and application areas of main discussed control strategies are presented in the Table 1. It can be concluded that the group of high performance control achieves similar parameters and area of applications. In FOC and DTC-SVM schemes the control action is usually synchronized with PWM-generation and executed with sampling time equal to the switching time ranging from 50μs to 500μs. The typical torque rise time is about 4-6 sampling and is limited by the VSI switching frequency. For comparison also parameters of scalar constant V/Hz control are given.

Table 1: Overview of main control strategies in low and medium power drives

Control Strategy	Speed Control Range	Static Speed Accuracy	Torque Rise Time	Starting Torque	Cost	Typical Applications
Scalar Control (Constant V/Hz)	1:10 (Open Loop)	5%-10%	Not Available	Low	Very Low	Low performance: Pumps, Fans, Compressors, HVAC, etc.
Field Oriented Control (FOC)	>1:200 (Closed Loop)	0%	<1-2ms	High	High	High performance: Crane, Lifts, Transportation, etc.
Direct Torque Control with Voltage/Flux Space Vector Modulation (DTC-SVM)	>1:200 (Closed Loop)	0%	<1-2ms	High	High	High performance: Crane, Lifts, Transportation, etc
Swtiching Table based Direct Torque Control (ST-DTC)	>1:200 (Closed Loop)	0%	<1ms	High	High	High performance: Crane, Lifts, Transportation, etc.
Direct Self Control (DSC)	>1:200 (Closed Loop)	0%	<1-2ms	High	High	High performance: Traction, Fast Field Weakening.
Servo Drives	1:10.000 (Closed Loop)	0%	<1ms	High	High	High performance, Speed Rise Time 10 ms: Robots, Manipulators, Ind. Automation, etc.

V. SUMMARY

Power electronic controlled high performance AC drives belong to High-tech industry and is one of main factors for energy saving and productivity growth. This paper reviews present state and trends in development of key parts of controlled drive systems: converter topologies, pulse width modulation (PWM) methods, as well as control and estimation techniques.

Following topologies are described: Two- and Multi-level Voltage Source Converters, Current Source Converters and Direct Converters.

Among PWM techniques are briefly presented: Conventional bipolar and unipolar PWM, Space Vector Modulation (SVM) with extension for multilevel converters, Harmonic control techniques, Variable frequency modulation (hysteresis, nearest level, and model predictive technique).

Finally, the generic Torque Control methods are described in two groups: Linear and nonlinear controllers. In linear group are presented Field Oriented Control (FOC), Direct Torque Control (DTC) with voltage SVM, and DTC with Flux Vector Modulation (DTC-FVM). The group of nonlinear control include: classical Switching Table based hysteresis DTC, Direct Self Control (DSC), and future trends solutions based on predictive DTC and neuro-fuzzy schemes. Also, selected simple flux vector observers/estimators are discussed.

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