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Faculty of Electrical Engineering
Institute of Control and Industrial Electronics

Ph.D. Thesis

M. Sc. Mariusz Malinowski

Sensorless Control Strategies for Three - Phase PWM Rectifiers

Thesis supervisor
Prof. Dr Sc. Marian P. Kaźmierkowski

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1. INTRODUCTION

Methods for limitation and elimination of disturbances and harmonic pollution in the power system have been widely investigated. This problem rapidly intensifies with the increasing amount of electronic equipment (computers, radio set, printers, TV sets etc.). This equipment, a nonlinear load, is a source of current harmonics, which produce increase of reactive power and power losses in transmission lines. The harmonics also cause electromagnetic interference and, sometimes, dangerous resonances. They have negative influence on the control and automatic equipment, protection systems, and other electrical loads, resulting in reduced reliability and availability. Moreover, nonlinear loads and non-sinusoidal currents produce non-sinusoidal voltage drops across the network impedance’s, so that non-sinusoidal voltages appears at several points of the mains. It brings out overheating of line, transformers and generators due to the iron losses.

Reduction of harmonic content in line current to a few percent allows avoiding most of the mentioned problems. Restrictions on current and voltage harmonics maintained in many countries through IEEE 519-1992 in the USA and IEC 61000-3-2/IEC 61000-3-4 in Europe standards, are associated with the popular idea of clean power.

Many of harmonic reduction method exist. These technique based on passive components, mixing single and three-phase diode rectifiers, and power electronics techniques as: multipulse rectifiers, active filters and PWM rectifiers (Fig. 1.1). They can be generally divided as:

A) harmonic reduction of already installed non-linear load;

B) harmonic reduction through linear power electronics load installation;
The traditional method of current harmonic reduction involves passive filters LC, parallel-connected to the grid. Filters are usually constructed as series-connected legs of capacitors and chokes. The number of legs depends on number of filtered harmonics ($5^{th}$, $7^{th}$, $11^{th}$, $13^{th}$). The advantages of passive filters are simplicity and low cost [105]. The disadvantages are:

- each installation is designed for a particular application (size and placement of the filters elements, risk of resonance problems),
- high fundamental current resulting in extra power losses,
- filters are heavy and bulky.

In case of diode rectifier, the simpler way to harmonic reduction of current are additional series coils used in the input or output of rectifier (typical 1-5%).

The other technique, based on mixing single and three-phase non-linear loads, gives a reduced THD because the $5^{th}$ and $7^{th}$ harmonic current of a single-phase diode rectifier often are in counter-phase with the $5^{th}$ and $7^{th}$ harmonic current of a three-phase diode rectifier [106].
The other already power electronics techniques is use of multipulse rectifiers. Although easy to implement, possess several disadvantages such as: bulky and heavy transformer, increased voltage drop, and increased harmonic currents at non-symmetrical load or line voltages.

An alternative to the passive filter is use of the active PWM filter (AF), which displays better dynamics and controls the harmonic and fundamental currents. Active filters are mainly divided into two different types: the active shunt filter (current filtering) (Fig. 1.2) and the active series filter (voltage filtering) [7].

The three-phase two-level shunt AF consist of six active switches and its topology is identical to the PWM inverter. AF represents a controlled current source $i_F$ which added to the load current $i_{Load}$ yields sinusoidal line current $i_L$ (Fig. 1.2). AF provide:

- compensation of fundamental reactive components of load current,
- load symetrization (from grid point of view),
- harmonic compensation much better than in passive filters.

In spite of the excellent performance, AFs possess certain disadvantages as complex control, switching losses and EMC problems (switching noise is present in the line current and even in the line voltage). Therefore, for reduction of this effects, inclusion of a small low-pass passive filter between the line and the AF is necessary.
The other interesting reduction technique of current harmonic is a PWM (active) rectifier (Fig. 1.3). Two types of PWM converters, with a voltage source output (Fig. 1.4a) and a current source output (Fig. 1.4b) can be used. First of them called a boost rectifier (increases the voltage) works with fixed DC voltage polarity, and the second, called a buck rectifier (reduces the voltage) operates with fixed DC current flow.

![Fig. 1.4 Two basic topologies of PWM rectifier: a) boost with voltage output b) buck with current output](image)

Among the main features of PWM rectifier are:

- bi-directional power flow,
- nearly sinusoidal input current,
- regulation of input power factor to unity,
- low harmonic distortion of line current (THD below 5%),
- adjustment and stabilization of DC-link voltage (or current),
- reduced capacitor (or inductor) size due to the continues current.

Furthermore, it can be properly operated under line voltage distortion and notching, and line voltage frequency variations.

Similar to the PWM active filter, the PWM rectifier has a complex control structure, the efficiency is lower than the diode rectifier due to extra switching losses. A properly designed low-pass passive filter is needed in front of the PWM rectifier due to EMI concerns.

The last technique is most promising thanks to advances in power semiconductor devices (enhanced speed and performance, and high ratings) and digital signal
processors, which allow fast operation and cost reduction. It offers possibilities for implementation of sophisticated control algorithm.

This thesis is devoted to investigation of different control strategies for boost type of three-phase bridge PWM rectifiers. Appropriate control can provide both the rectifier performance improvements and reduction of passive components. Several control techniques for PWM rectifiers are known [16-23, 30-69]. A well-known method based on indirect active and reactive power control is based on current vector orientation with respect to the line voltage vector (Voltage Oriented Control - VOC) [30-69]. An other less known method based on instantaneous direct active and reactive power control is called Direct Power Control (DPC) [16, 20-23]. Both mentioned strategies do not produce sinusoidal current when the line voltage is distorted. Therefore, the following thesis can be formulated:

“using the control strategy based on virtual flux instead of the line voltage vector orientation provides lower harmonic distortion of line current and leads to line-voltage sensorless operation”.

In order to prove the above thesis, the author used an analytical and simulation based approach, as well as experimental verification on the laboratory setup with a 5kVA IGBT converter.

The thesis consists of six chapters. Chapter 1 is an introduction. Chapter 2 is devoted to presentation of various topologies of rectifiers for ASD’s. The mathematical model and operation description of PWM rectifier are also presented. General features of the sensorless operation focused on AC voltage-sensorless. Voltage and virtual flux estimation are summarized at the end of the chapter. Chapter 3 covers the existing solution of Direct Power Control and presents a new solution based on Virtual Flux estimation [17]. Theoretical principles of both methods are discussed. The steady state and dynamic behavior of VF-DPC are presented, illustrating the operation and performance of the proposed system as compared with a conventional DPC method. Both strategies are also investigated under unbalanced and distorted line voltages. It is shown that the VF-DPC exhibits several advantages, particularly it provides sinusoidal line current when the supply voltage is non-ideal. Test results show excellent
performance of the proposed system. Chapter 4 is focused on the Voltage Oriented and Virtual Flux Oriented Controls. Additionally, development and investigation of novel modulation techniques is described and discussed, with particular presentation of adaptive modulation. It provides a wide range of linearity, reduction of switching losses and good dynamics. Chapter 5 contains comparative study of discussed control methods. Finally Chapter 6 presents summary and general conclusions. The thesis is supplemented by nine Appendices among which are: conventional and instantaneous power theories [A.2], implementation of a space vector modulator [A.3], description of the simulation program [A.4] and the laboratory set-up [A.6].

In the author’s opinion the following parts of the thesis represent his original achievements:

- development of a new line voltage estimator – (Section 2.5),
- elaboration of new Virtual Flux based Direct Power Control for PWM rectifiers – (Section 3.4),
- implementation and investigation of various closed-loop control strategies for PWM rectifiers: Virtual Flux – Based Direct Power Control (VF-DPC), Direct Power Control (DPC), Voltage Oriented Control (VOC), Virtual Flux Oriented Control (VFOC) – (Sections 3.6 and 4.5),
- development of a new Adaptive Space Vector Modulator for three-phase PWM converter, working in polar and cartesian coordinate system (Patent No. P340 113) – (Section 4.4.7),
- development of a simulation algorithm in SABER and control algorithm in C language for investigation of proposed solutions – (Appendix A.4),
- construction and practical verification of the experimental setup based on a mixed RISC/DSP (PowerPC 604/TMS320F240) digital controller – (Appendix A.6).
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List of Symbols

Symbols (general)
x(t), x – instantaneous value
X*, x* – reference
X, x – average value, average (continuous) part
X, x – oscillating part
x – complex vector
x* – conjugate complex vector
|X| – magnitude (length) of function
ΔX, Δx – deviation

Symbols (special)
α – phase angle of reference vector
λ – power factor
ϕ – phase angle of current
ω – angular frequency
ψ – phase angle
ε – control phase angle
cos ϕ – fundamental power factor
f – frequency
i(t), i – instantaneous current
j – imaginary unit
kP, kI – proportional control part, integral control part
p(t), p – instantaneous active power
q(t), q – instantaneous reactive power
t – instantaneous time
v(t), v – instantaneous voltage

ΨL – virtual line flux vector
ΨLa – virtual line flux vector components in the stationary α, β coordinates
ΨLβ – virtual line flux vector components in the stationary α, β coordinates
ΨLd – virtual line flux vector components in the synchronous d, q coordinates
ΨLq – virtual line flux vector components in the synchronous d, q coordinates

uL – line voltage vector
uLa – line voltage vector components in the stationary α, β coordinates
uLβ – line voltage vector components in the stationary α, β coordinates
uLd – line voltage vector components in the synchronous d, q coordinates
uLq – line voltage vector components in the synchronous d, q coordinates

iL – line current vector
iLa – line current vector components in the stationary α, β coordinates
iLβ – line current vector components in the stationary α, β coordinates
List of Symbols

- $i_{ld}$ – line current vector components in the synchronous d, q coordinates
- $i_{lq}$ – line current vector components in the synchronous d, q coordinates

- $u_{s}$, $u_{conv}$ – converter voltage vector
- $u_{sa}$ – converter voltage vector components in the stationary $\alpha$, $\beta$ coordinates
- $u_{sb}$ – converter voltage vector components in the stationary $\alpha$, $\beta$ coordinates
- $u_{sd}$ – converter voltage vector components in the synchronous d, q coordinates
- $u_{sq}$ – converter voltage vector components in the synchronous d, q coordinates

- $u_{dc}$ – DC link voltage
- $i_{dc}$ – DC link current

- $S_a$, $S_b$, $S_c$ – Switching state of the converter

- $C$ – capacitance
- $I$ – root mean square value of current
- $L$ – inductance
- $R$ – resistance
- $S$ – apparent power
- $T$ – time period
- $P$ – active power
- $Q$ – reactive power
- $Z$ – impedance

Subscripts
- ..a, ..b, ..c - phases of three-phase system
- ..d, ..q - direct and quadrature component
- ..+, ..-, ..0 - positive, negative and zero sequence component
- ..$\alpha$, ..$\beta$, ..0 - alpha, beta components and zero sequence component
- ..$h$ – harmonic order of current and voltage, harmonic component
- ..$n$ – harmonic order
- ..max - maximum
- ..min - minimum
- ..L-L - line to line
- ..Load - load
- ..conv - converter
- ..Loss - losses
- ..ref - reference
- ..m - amplitude
- ..rms - root mean square value

Abbreviations

- $AF$ active PWM filter
- $ANN$ artificial neural network
- $ASD$ adjustable speed drives
- $ASVM$ adaptive space vector modulation
- $CB-PWM$ carrier based pulse width modulation
<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>CSI</td>
<td>current source inverter</td>
</tr>
<tr>
<td>DPC</td>
<td>direct power control</td>
</tr>
<tr>
<td>DSP</td>
<td>digital signal processor</td>
</tr>
<tr>
<td>DTC</td>
<td>direct torque control</td>
</tr>
<tr>
<td>EMI</td>
<td>electro-magnetic interference</td>
</tr>
<tr>
<td>FOC</td>
<td>field-oriented control</td>
</tr>
<tr>
<td>IFOC</td>
<td>indirect field-oriented control</td>
</tr>
<tr>
<td>IGBT</td>
<td>insulated gate bipolar transistor</td>
</tr>
<tr>
<td>PCC</td>
<td>point of common coupling</td>
</tr>
<tr>
<td>PFC</td>
<td>power factor correction</td>
</tr>
<tr>
<td>PI</td>
<td>proportional integral (controller)</td>
</tr>
<tr>
<td>PLL</td>
<td>phase locked loop</td>
</tr>
<tr>
<td>PWM</td>
<td>pulse-width modulation</td>
</tr>
<tr>
<td>REC</td>
<td>rectifier</td>
</tr>
<tr>
<td>SVM</td>
<td>space vector modulation</td>
</tr>
<tr>
<td>THD</td>
<td>total harmonic distortion</td>
</tr>
<tr>
<td>UPF</td>
<td>unity power factor</td>
</tr>
<tr>
<td>VF</td>
<td>virtual flux</td>
</tr>
<tr>
<td>VF-DPC</td>
<td>virtual flux based direct power control</td>
</tr>
<tr>
<td>VFOC</td>
<td>virtual flux oriented control</td>
</tr>
<tr>
<td>VOC</td>
<td>voltage oriented control</td>
</tr>
<tr>
<td>VSI</td>
<td>voltage source inverter</td>
</tr>
<tr>
<td>ZSS</td>
<td>zero sequence signal</td>
</tr>
</tbody>
</table>
2. PWM RECTIFIER

2.1. INTRODUCTION

As it has been observed for recent decades, an increasing part of the generated electric energy is converted through rectifiers, before it is used at the final load. In power electronic systems, especially, diode and thyristor rectifiers are commonly applied in the front end of DC-link power converters as an interface with the AC line power (grid) - Fig. 2.1. The rectifiers are nonlinear in nature and, consequently, generate harmonic currents in to the AC line power. The high harmonic content of the line current and the resulting low power factor of the load, causes a number of problems in the power distribution system like:

- voltage distortion and electromagnetic interface (EMI) affecting other users of the power system,
- increasing voltampere ratings of the power system equipment (generators, transformers, transmission lines, etc.).

Therefore, governments and international organizations have introduced new standards (in the USA: IEEE 519 and in Europe: IEC 61000-3)[A8] which limit the harmonic content of the current drown from the power line by the rectifiers. As a consequence a great number of new switch-mode rectifier topologies that comply with the new standards have been developed.

In the area of variable speed AC drives, it is believed that three-phase PWM boost AC/DC converter will replace the diode rectifier. The resulting topology consists of two identical bridge PWM converters (Fig. 2.4). The line-side converter operates as rectifier in forward energy flow, and as inverter in reverse energy flow. In farther discussion assuming the forward energy flow, as the basic mode of operation the line-side converter will be called as PWM rectifier. The AC side voltage of PWM rectifier can be controlled in magnitude and phase so as to obtain sinusoidal line current at unity power factor (UPF). Although such a PWM rectifier/inverter (AC/DC/AC) system is expensive, and the control is complex, the topology is ideal for four-quadrant operation. Additionally, the PWM rectifier provides DC bus voltage stabilization and can also act as active line conditioner (ALC) that compensate harmonics and reactive power at the point of common coupling of the distribution network. However, reducing the cost of the PWM rectifier is vital for the competitiveness compared to other front-end rectifiers. The cost of power switching devices (e.g. IGBT) and digital signal processors (DSP’s) are generally decreasing and further reduction can be obtained by reducing the number of sensors. Sensorless control exhibits advantages such as improved reliability and lower installation costs.
2.2. RECTIFIERS TOPOLOGIES

A voltage source PWM inverter with diode front-end rectifier is one of the most common power configuration used in modern variable speed AC drives (Fig. 2.1). An uncontrolled diode rectifier has the advantage of being simple, robust and low cost. However, it allows only undirectional power flow. Therefore, energy returned from the motor must be dissipated on power resistor controlled by chopper connected across the DC link. The diode input circuit also results in lower power factor and high level of harmonic input currents. A further restriction is that the maximum motor output voltage is always less than the supply voltage.

Equations (2.1) and (2.2) can be used to determine the order and magnitude of the harmonic currents drawn by a six-pulse diode rectifier:

\[ h = 6k \pm 1 \quad k = 1, 2, 3, \ldots \]  
\[ \frac{I_h}{I_1} = \frac{1}{h} \]

Harmonic orders as multiples of the fundamental frequency: 5\textsuperscript{th}, 7\textsuperscript{th}, 11\textsuperscript{th}, 13\textsuperscript{th} etc., with a 50 Hz fundamental, corresponds to 250, 350, 550 and 650 Hz, respectively. The magnitude of the harmonics in per unit of the fundamental is the reciprocal of the harmonic order: 20% for the 5\textsuperscript{th}, 14.3\% for the 7\textsuperscript{th}, etc. Eqs. (2.1)-(2.2) are calculated from the Fourier series for ideal square wave current (critical assumption for infinite inductance on the input of the converter). Equations (2.1) is fairly good description of the harmonic orders generally encountered. The magnitude of actual harmonic currents often differs from the relationship described in (2.2). The shape of the AC current depends on the input inductance of converter (Fig. 2.2). The ripple current is equal 1/L times the integral of the DC ripple voltage. With infinite inductance the ripple current is zero and the flap-top wave of Fig. 2.2d results. The full description of harmonic calculation in six-pulse converter can be found in [116].

![Fig. 2.1 Diode rectifier](image)
Besides of six-pulse bridge rectifier a few other rectifier topologies are known [117-118]. Some of them are presented in Fig. 2.3. The topology of Fig. 2.3(a) presents simple solution of boost – type converter with possibility to increase DC output voltage. This is important feature for ASD’s converter giving maximum motor output voltage. The main drawback of this solution is stress on the components, low frequency distortion of the input current. Next topologies (b) and (c) uses a PWM rectifier modules with a very low current rating (20-25% level of rms current comparable with (e) topology). Hence they have a low cost potential provide only possibility of regenerative braking mode (b) or active filtering (c). Fig. 2.3d presents 3-level converter called Vienna rectifier [112]. The main advantage is low switch voltage, but not typical switches are required. Fig. 2.3e presents most popular topology used in ASD, UPS and recently like a PWM rectifier. This universal topology has the advantage of using a low-cost three-phase module with a bi-directional energy flow capability. Among disadvantages are: high per-unit current rating, poor immunity to shoot-through faults, and high switching losses. The features of all topologies are compared in Table 2.1.

### Table 2.1 Features of three-phase rectifiers

<table>
<thead>
<tr>
<th>Topology</th>
<th>Regulation of DC output voltage</th>
<th>Low harmonic distortion of line current</th>
<th>Near sinusoidal current waveforms</th>
<th>Power factor correction</th>
<th>Bi-directional power flow</th>
<th>Remarks</th>
</tr>
</thead>
<tbody>
<tr>
<td>Diode rectifier</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td></td>
</tr>
<tr>
<td>Rec(a)</td>
<td>+</td>
<td>-</td>
<td>-</td>
<td>+</td>
<td>-</td>
<td></td>
</tr>
<tr>
<td>Rec(b)</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>+</td>
<td></td>
</tr>
<tr>
<td>Rec(c)</td>
<td>-</td>
<td>+</td>
<td>+</td>
<td>+</td>
<td>-</td>
<td>UPF</td>
</tr>
<tr>
<td>Rec(d)</td>
<td>+</td>
<td>+</td>
<td>+</td>
<td>+</td>
<td>-</td>
<td>UPF</td>
</tr>
<tr>
<td>Rec(e)</td>
<td>+</td>
<td>+</td>
<td>+</td>
<td>+</td>
<td>+</td>
<td>UPF</td>
</tr>
</tbody>
</table>
Fig. 2.3 Basic topologies of switch-mode three-phase rectifiers
(a) simple boost-type converter  
(b) diode rectifier with PWM regenerative braking rectifier  
(c) diode rectifier with PWM active filtering rectifier  
(d) Vienna rectifier (3-level converter)  
(e) PWM reversible rectifier (2-level converter)

The last topology is most promising therefore was chosen by most global company (SIMENS, ABB and other). In a DC distributed Power System (Fig. 2.5) or AC/DC/AC converter (Fig. 2.4), the AC power is first transformed into DC thanks to three-phase PWM rectifier. It provides UPF and low current harmonic content. The converters connected to the DC-bus provide further desired conversion for the loads, such as adjustable speed drives for induction motors (IM) and permanent magnet synchronous motor (PMSM), DC/DC converter, multidrive operation, etc.
The AC/DC/AC converter (Fig. 2.4) is known in ABB like an ACS611/ACS617 (15 kW - 1,12 MW) complete four-quadrant drive. The line converter is identical to the ACS600 (DTC) motor converter with the exception of the control software [20,121]. Similar solutions possess SIEMENS in Simovert Masterdrive (2,2 kW – 2,3 MW) [127]. Furthermore, AC/DC/AC provide:

- the motor can operate at a higher speed without field weakening (by maintaining the DC-bus voltage above the supply voltage peak),
- decreased theoretically by one-third common mode voltage compared to conventional configuration thanks to the simultaneous control of rectifier - inverter (same switching frequency and synchronized sampling time may avoid common-mode voltage pulse because the different type of zero voltage ($U_o, U_7$) are not applied at the same time) [114],
- the response of the voltage controller can be improved by fed-forward signal from the load what gives possibility to minimize the DC link capacitance while maintaining the DC-link voltage within limits under step load conditions [104, 111].

Other solution used in industry is shown in Fig. 2.5 like a multidrive operation [120]. ABB propose active front-end converter ACA 635 (250 kW - 2,5 MW) and Siemens Simovert Masterdrive in range of power from 7,5 kW up to 1,5 MW.
2.3 OPERATION OF THE PWM RECTIFIER

Fig. 2.6b shows a single-phase representation of the rectifier circuit presented in Fig. 2.6a. $L$ and $R$ represents the line inductor. $u_L$ is the line voltage and $u_S$ is the bridge converter voltage controllable from the DC-side. Magnitude of $u_S$ depends on the modulation index and DC voltage level.

Inductors connected between input of rectifier and lines are integral part of the circuit. It brings current source character of input circuit and provide boost feature of converter. The line current $i_L$ is controlled by the voltage drop across the inductance $L$ interconnecting two voltage sources (line and converter). It means that the inductance voltage $u_l$ equals the difference between the line voltage $u_L$ and the converter voltage $u_S$. When we control phase angle $\varepsilon$ and amplitude of converter voltage $u_S$, we control
indirectly phase and amplitude of line current. In this way average value and sign of DC current is subject to control what is proportional to active power conducted through converter. The reactive power can be controlled independently with shift of fundamental harmonic current \( I_L \) in respect to voltage \( U_L \).

Fig. 2.7 presents general phasor diagram and both rectification and regenerating phasor diagrams when unity power factor is required. The figure shows that the voltage vector \( u_S \) is higher during regeneration (up to 3%) then rectifier mode. It means that these two modes are not symmetrical [67].

Main circuit of bridge converter (Fig. 2.6a) consists of three legs with IGBT transistor or, in case of high power, GTO thyristors. The bridge converter voltage can be represented with eight possible switching states (Fig. 2.8 six-active and two-zero) described by equation:

\[
u_{k+1} = \begin{cases} 
(2/3)u_{dc}e^{jk\pi/3} & \text{for } k = 0...5 \\
0 & \text{for } k = 0...5 
\end{cases}
\]  
(2.3)

Fig. 2.8 Switching states of PWM bridge converter
2.3.1 Mathematical description of the PWM rectifier

The basic relationship between vectors of the PWM rectifier is presented in Fig. 2.9.

![Fig. 2.9 Relationship between vectors in PWM rectifier](image)

**Description of line voltages and currents**

Three phase line voltage and the fundamental line current is:

\[
\begin{align*}
    u_a &= E_m \cos \omega t \\
    u_b &= E_m \cos (\omega t + \frac{2\pi}{3}) \\
    u_c &= E_m \cos (\omega t - \frac{2\pi}{3})
\end{align*}
\]

(2.4a) (2.4b) (2.4c)

\[
\begin{align*}
    i_a &= I_m \cos (\omega t + \varphi) \\
    i_b &= I_m \cos (\omega t + \frac{2\pi}{3} + \varphi) \\
    i_c &= I_m \cos (\omega t - \frac{2\pi}{3} + \varphi)
\end{align*}
\]

(2.5a) (2.5b) (2.5c)

where \(E_m\) (\(I_m\)) and \(\omega\) are amplitude of the phase voltage (current) and angular frequency, respectively, with assumption
we can transform equations (2.4) to \( \alpha - \beta \) system thanks to equations (A.2.22a) and the input voltage in \( \alpha - \beta \) stationary frame are expressed by:

\[
\begin{align*}
    u_{La} &= \frac{3}{2} E_m \cos(\alpha t) \\ u_{L\beta} &= \frac{3}{2} E_m \sin(\alpha t)
\end{align*}
\]  

(2.7)  

(2.8)

and the input voltage in the synchronous \( d-q \) coordinates (Fig. 2.9) are expressed by:

\[
\begin{bmatrix}
    u_{ld} \\
    u_{lq}
\end{bmatrix} = \begin{bmatrix}
    \frac{3}{2} E_m \\
    0
\end{bmatrix} = \begin{bmatrix}
    \sqrt{u_{La}^2 + u_{L\beta}^2} \\
    0
\end{bmatrix}
\]

(2.9)

**Description of input voltage in PWM rectifier**

Line to line input voltages of PWM rectifier can be described with the help of Fig. 2.8 as:

\[
\begin{align*}
    u_{Sab} &= (S_a - S_b) \cdot u_{dc} \\ u_{Sbc} &= (S_b - S_c) \cdot u_{dc} \\ u_{Scb} &= (S_c - S_a) \cdot u_{dc}
\end{align*}
\]  

(2.10a)  

(2.10b)  

(2.10c)

and phase voltages are equal:

\[
\begin{align*}
    u_{Sa} &= f_a \cdot u_{dc} \\ u_{Sb} &= f_b \cdot u_{dc} \\ u_{Sc} &= f_c \cdot u_{dc}
\end{align*}
\]  

(2.11a)  

(2.11b)  

(2.11c)

where:

\[
\begin{align*}
    f_a &= \frac{2S_a - (S_b + S_c)}{3} \\ f_b &= \frac{2S_b - (S_a + S_c)}{3} \\ f_c &= \frac{2S_c - (S_a + S_b)}{3}
\end{align*}
\]  

(2.12a)  

(2.12b)  

(2.12c)

The \( f_a, f_b, f_c \) are assume 0, \( \pm 1/3 \) and \( \pm 2/3 \).

**Description of PWM rectifier**

**Model of three-phase PWM rectifier**

The voltage equations for balanced three-phase system without the neutral connection can be written as (Fig. 2.7b):

\[
u_L = u_I + u_S
\]  

(2.13)
The combination of equations (2.11, 2.12, 2.15, 2.16) can be represented as three-phase block diagram (Fig. 2.10) [34].

![Fig. 2.10 Block diagram of voltage source PWM rectifier in natural three-phase coordinates](image)

**Model of PWM rectifier in stationary coordinates (α-β)**

The voltage equation in the stationary α-β coordinates are obtained by applying (A.2.22a) to (2.15) and (2.16) and are written as:

\[
\begin{bmatrix}
    u_{Lα} \\
    u_{Lβ}
\end{bmatrix} = R \begin{bmatrix}
    i_{Lα} \\
    i_{Lβ}
\end{bmatrix} + L \frac{d}{dt} \begin{bmatrix}
    i_{Lα} \\
    i_{Lβ}
\end{bmatrix} + \begin{bmatrix}
    u_{Sa} \\
    u_{Sb}
\end{bmatrix}
\]  

(2.17)

and

\[
C \frac{du_{dc}}{dt} = (i_{Lα}S_α + i_{Lβ}S_β) - i_{dc}
\]  

(2.18)

where:  
\[S_α = \frac{1}{\sqrt{6}}(2S_a - S_b - S_c)\]  
\[S_β = \frac{1}{\sqrt{2}}(S_b - S_c)\]
A block diagram of $\alpha\beta$ model is presented in Fig. 2.11.

\begin{align*}
\frac{1}{R+sL}i_{\alpha} = i_{\beta} \\
\frac{1}{sC}i_{\beta} = u_{d} \\
\end{align*}

Fig. 2.11 Block diagram of voltage source PWM rectifier in stationary $\alpha$-$\beta$ coordinates

**Model of PWM rectifier in synchronous rotating coordinates (d-q)**

The equations in the synchronous d-q coordinates are obtained with the help of transformation 4.1a:

\begin{align*}
u_{ld} &= R{i}_{ld} + L\frac{di_{ld}}{dt} - \omega L{i}_{lq} + u_{ld} \quad (2.19a) \\
u_{lq} &= R{i}_{lq} + L\frac{di_{lq}}{dt} + \omega L{i}_{ld} + u_{lq} \quad (2.19b) \\
C\frac{du_{dc}}{dt} &= (i_{ld}S_{d} + i_{lq}S_{q}) - i_{dc} \quad (2.20)
\end{align*}

where: $S_{d} = S_{\alpha} \cos \omega \theta + S_{\beta} \sin \omega \theta$; $S_{q} = S_{\beta} \cos \omega \theta - S_{\alpha} \sin \omega \theta$

A block diagram of d-q model is presented in Fig. 2.12.

\begin{align*}
u_{ld} &= R{i}_{ld} + L\frac{di_{ld}}{dt} - \omega L{i}_{lq} + u_{ld} \\
u_{lq} &= R{i}_{lq} + L\frac{di_{lq}}{dt} + \omega L{i}_{ld} + u_{lq} \\
C\frac{du_{dc}}{dt} &= (i_{ld}S_{d} + i_{lq}S_{q}) - i_{dc}
\end{align*}

Fig. 2.12 Block diagram of voltage source PWM rectifier in synchronous d-q coordinates
PWM rectifier

*R* can be practically neglected because voltage drop on resistance is much lower than voltage drop on inductance, what gives simplified equations (2.14), (2.15), (2.17), (2.19).

\[
\frac{du_i}{dt} = L + u_c
\]  
(2.21)

\[
\begin{bmatrix}
  u_a \\
  u_b \\
  u_c
\end{bmatrix} = L \frac{d}{dt} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \begin{bmatrix} u_{Sa} \\ u_{Sb} \\ u_{Sc} \end{bmatrix}
\]  
(2.22)

\[
\begin{bmatrix}
  u_{La} \\
  u_{Lb}
\end{bmatrix} = L \frac{d}{dt} \begin{bmatrix} i_{La} \\ i_{Lb} \end{bmatrix} + \begin{bmatrix} u_{Sa} \\ u_{Sb} \end{bmatrix}
\]  
(2.23)

\[
u_{Ld} = L \frac{di_{Ld}}{dt} - \omega L i_{Lq} + u_{Sd}
\]  
(2.24a)

\[
u_{Lq} = L \frac{di_{Lq}}{dt} + \omega L i_{Ld} + u_{Sq}
\]  
(2.24b)

The active and reactive power supplied from the source is given by [see A.2]

\[
p = \text{Re}\{u \cdot i^*\} = u_a i_a + u_b i_b = u_a i_a + u_b i_b + u_c i_c
\]  
(2.25)

\[
q = \text{Im}\{u \cdot i^*\} = u_b i_a - u_a i_b = \frac{1}{\sqrt{3}}(u_{bc} i_a + u_{ca} i_b + u_{ab} i_c)
\]  
(2.26)

It gives in the synchronous d-q coordinates:

\[
p = (u_{Lq} i_{Lq} + u_{Ld} i_{Ld}) = \frac{3}{2} E_m I_m
\]  
(2.27)

\[
q = (u_{Lq} i_{Ld} - u_{Ld} i_{Lq})
\]  
(2.28)

(if we make assumption of unity power factor, we will obtain following properties

\[
i_{Lq} = 0, u_{Lq} = 0, u_{Ld} = \frac{3}{2} E_m, i_{Ld} = \frac{3}{2} I_m, q = 0\) (see Fig. 2.13)).

![Fig. 2.13 Power flow in bi-directional AC/DC converter as dependency of i_L direction.](image-url)
2.3.2 Steady-state properties and limitations

For proper operation of PWM rectifier a minimum DC-link voltage is required. Generally it can be determined by the peak of line-to-line supply voltage:

\[ V_{dc\,\text{min}} \geq V_{LN(\text{rms})} \cdot \sqrt{3} \cdot \sqrt{2} = 2.45 \cdot V_{LN(\text{rms})} \]  

(2.29)

It is true definition but not concern all situations. Other publication \[36,37\] defines minimum voltage but do not take into account line current (power) and line inductors. The determination of this voltage is more complicated and is presented in \[59\].

Equations (2.24) can be transformed to vector form in synchronous d-q coordinates defining derivative of current as:

\[ L \frac{dI_{dq}}{dt} = u_{dq} - j\omega L_{dq} - u_{dq}. \]  

(2.30)

Equation (2.30) defines direction and rate of current vector movement. Six active vectors \([U_1-6]\) of input voltage in PWM rectifier rotate clockwise in synchronous d-q coordinates. For vectors \(U_0, U_1, U_2, U_3, U_4, U_5, U_6, U_7\) the current derivatives are denoted respectively as \(U_{pq}, U_{p1}, U_{p2}, U_{p3}, U_{p4}, U_{p5}, U_{p6}, U_{p7}\) (Fig. 2.14).  

![Fig. 2.14 Instantaneous position of vectors](image-url)
The full current control is possible when the current is kept in specified error area (Fig. 2.15). Fig. 2.14 and Fig. 2.15 presents that any vectors can force current vector inside error area when angle created by vectors $U_{p1}$ and $U_{p2}$ is $\xi \leq \pi$. It results from trigonometrical condition that vectors $U_{p1}$, $U_{p2}$, $U_1$ and $U_2$ form an equilateral triangle for $\xi = \pi$ where $u_{Ldq} - j\omega L_{Ldq}$ is an altitude. Therefore, from simple trigonometrical relationship, it is possible to define boundary condition as:

$$|u_{Ldq} - j\omega L_{Ldq}| = \frac{\sqrt{3}}{2} u_{sdq}$$  \hspace{1cm} (2.31)

and after transformation, assumpting that $u_{sdq} = 2/3U_{dc}$, $u_{Ldq} = E_m$, $i_{Ldq} = i_{Ld}$ (for UPF) we get condition for minimal DC-link voltage:

$$u_{dc} \sqrt{3\left[E_m^2 + (\omega L_i L_{sd})^2\right]} \hspace{0.5cm} \text{and} \hspace{0.5cm} \xi > \pi.$$  \hspace{1cm} (2.32)

Above equation shows relation between supply voltage (usually constant), output dc voltage, current (load) and inductance. It also means that sum of vector $u_{Ldq} - j\omega L_{Ldq}$ should not exceed linear region of modulation i.e. circle inscribed in the hexagon (see Section 4.4).

The inductor has to be designed carefully because low inductance will give a high current ripple and will make the design more depending on the line impedance. The high value of inductance will give a low current ripple, but simultaneously reduce the operation range of the rectifier. The voltage drop across the inductance has influence for the line current. This voltage drop is controlled by the input voltage of the PWM rectifier but maximal value is limited by the DC-link voltage. Consequently, a high current (high power) through the inductance requires either a high DC-link voltage or a low inductance (low impedance). Therefore, after transformation of equation (2.32) the maximal inductance can be determinate as:

$$L \left( \sqrt{\frac{u_{dc}^2}{3} - \frac{E_m^2}{\omega i_{Ld}}} \right).$$  \hspace{1cm} (2.33)
2.4 SENSORLESS OPERATION

Normally, the PWM rectifier needs three kinds of sensors:

- DC-voltage sensor (1 sensor)
- AC-line current sensors (2 or 3 sensors)
- AC-line voltage sensors (2 or 3 sensors)

The sensorless methods provide technical and economical advantages to the system as: simplification, isolation between the power circuit and control system, reliability and cost effectiveness. The possibility to reduce the number of the expensive sensors have been studied especially in the field of motor drive application [1], but the rectifier application differ from the inverter operation in the following reasons:

- Zero vector will shorted the line power,
- The line operates at constant frequency 50Hz and synchronization is necessary.

The most used solution for reducing of sensors include:

- AC voltage and current sensorless,
- AC current sensorless,
- AC voltage sensorless.

**AC voltage and current sensorless**

Reductions of current sensors especially for AC drives are well known [1]. The two-phase currents may be estimated based on information of DC link current and reference voltage vector in every PWM period. No fully protection is main practical problem in the system. Particularly for PWM rectifier the zero vectors ($U_0, U_7$) presents no current in DC-link and three line phases are short circuit simultaneously. New improved method presented in [30, 115] is to sample DC-link current few times in one switching period. Basic principle of current reconstruction is shown in Fig. 2.16 together with a voltage vector’s patterns determining the direction of current flow. One active voltage vector takes it to reconstruct one phase current and another voltage vector is used to reconstruct a second phase current using values measured from DC current sensor. A relationship between the applied active vectors and the phase currents measured from DC link sensor is shown in TABLE 2.2, which is based on eight voltage vectors composed of six active vectors and two zero vectors.

![Fig. 2.16 PWM signals and DC link current in sector I](image)

**Table 2.2 Relationship between voltage vectors of converter, DC-link current and line currents.**

<table>
<thead>
<tr>
<th>Voltage Vector</th>
<th>DC link current $i_{dc}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$U_0(000)$</td>
<td>0</td>
</tr>
<tr>
<td>$U_1(100)$</td>
<td>$+i_a$</td>
</tr>
<tr>
<td>$U_2(110)$</td>
<td>$-i_a$</td>
</tr>
<tr>
<td>$U_3(010)$</td>
<td>$+i_b$</td>
</tr>
<tr>
<td>$U_4(011)$</td>
<td>$-i_b$</td>
</tr>
<tr>
<td>$U_5(001)$</td>
<td>$+i_c$</td>
</tr>
<tr>
<td>$U_6(101)$</td>
<td>$-i_c$</td>
</tr>
<tr>
<td>$U_7(111)$</td>
<td>0</td>
</tr>
<tr>
<td>$U_0(000)$</td>
<td>0</td>
</tr>
<tr>
<td>$U_1(101)$</td>
<td>$+i_a$</td>
</tr>
<tr>
<td>$U_2(111)$</td>
<td>$-i_a$</td>
</tr>
</tbody>
</table>
PWM rectifier

The main problem of AC current estimation based on minimum pulse-time for DC-link current sampling. It appears when either of two active vectors is not present, or is applied only for a short time. In such a case, it is impossible to reconstruct phase current. This occur in the case of reference voltage vectors passing one of the six possible active vectors or a low modulation index (Fig. 2.17). The minimum short time to obtain a correct estimation depends on the rapidness of the system, delays, cable length and dead-time [30]. The way to solve the problem is to adjust the PWM-pulses or to allow that no currents information is present in some time period. Therefore improved compensation consists of calculating the error, which are introduced by the PWM pulse adjustment and then compensate this error in the next switching period.

Fig. 2.17. Voltage vector area requiring the adjustment of PWM signals, when a reference voltage passes one of possible six active vectors and in case of low modulation index and overmodulation

The AC voltage and current sensorless methods in spite of cost reduction posses several disadvantages: higher contents of current ripple, problems with discontinuous modulation and overmodulation mode [see Section 4.4], sampling is presented few times per switching state what is not technically convenient, unbalance and start up condition are not reported.

**AC current sensorless**

This very simple solution based on inductor voltage \( u_I \) measurement in two lines. Supply voltage can be estimated with assumption that voltage on inductance is equal to line voltage when the zero-vector occurs in converter (Fig. 2.18)

Fig. 2.18. PWM rectifier circuit when the zero voltage vector is applied.
PWM rectifier

On the basis of the inductor voltage described in equation (2.34)

\[ u_{IR} = L \frac{di_{IR}}{dt} \]  \hspace{1cm} (2.34)

the line current can be calculated as:

\[ i_{LR} = \frac{1}{L} \int u_{IR} dt \]  \hspace{1cm} (2.35)

Thanks to equation (2.35) the observed current will not be affected by derivation noise, but it directly reduces the dynamic of the control. This gains problems with over-current protection

**AC voltage sensorless**

Previous solutions present some over voltage and over current protection troubles. Therefore the DC-voltage and the AC-line current sensors are an important part of the over-voltage and over-current protection, while it is possible to replace the AC-line voltage sensors with a line voltage estimator or virtual flux estimator what is described in next point.

### 2.5 VOLTAGE AND VIRTUAL FLUX ESTIMATION

**Line voltage estimator [44]**

An important requirement for a voltage estimator is to estimate the voltage correct also under unbalanced conditions and pre-existing harmonic voltage distortion. Not only the fundamental component should be estimated correct, but also the harmonic components and the voltage unbalance. It gives a higher total power factor [21]. It is possible to calculate the voltage across the inductance by the current differentiating. The line voltage can then be estimated by adding reference of the rectifier input voltage to the calculated voltage drop across the inductor [52]. However, this approach has the disadvantage that the current is differentiated and noise in the current signal is gained through the differentiatiion. To prevent this a voltage estimator based on the power estimator of [21] can be applied. In [21] the current is sampled and the power is estimated several times in every switching state.

In conventional space vector modulation (SVM) for three-phase voltage source converters, the AC currents are sampled during the zero-vector states because no switching noise is present and a filter in the current feedback for the current control loops can be avoided. Using equation (2.36) and (2.37) the estimated active and reactive power in this special case (zero states) can be expressed as:

\[ p = L \left( \frac{di_a}{dt} i_a + \frac{di_b}{dt} i_b + \frac{di_c}{dt} i_c \right) = 0 \]  \hspace{1cm} (2.36)

\[ q = \frac{3L}{\sqrt{3}} \left( \frac{di_a}{dt} i_c - \frac{di_c}{dt} i_a \right) \]  \hspace{1cm} (2.37)
It should be noted that in this special case it is only possible to estimate the reactive power in the inductor. Since powers are DC-values it is possible to prevent the noise of the differentiated current by use of a simple (digital) low pass filter. This ensures a robust and noise insensitive performance of the voltage estimator.

Based on instantaneous power theory, the estimated voltages across the inductance is:

\[
\begin{bmatrix}
  u_{L\alpha} \\
  u_{L\beta}
\end{bmatrix} = \frac{1}{i_{L\alpha}^2 + i_{L\beta}^2} \begin{bmatrix}
  i_{L\alpha} - i_{L\beta} \\
  i_{L\beta} \\
  i_{L\alpha}
\end{bmatrix} q
\]  

\[(2.38)\]

where:

\(u_{L\alpha}, u_{L\beta}\) are the estimated values of the three-phase voltages across the inductance \(L\), in the fixed \(\alpha-\beta\) coordinates.

The estimated line voltage \(u_{L(eest)}\) can now be found by adding the voltage reference of the PWM rectifier to the estimated inductor voltage [44].

\[u_{L(eest)} = u_s + u_I\]  

\[(2.39)\]

**Virtual flux estimator**

The voltage imposed by the line power in combination with the AC side inductors are assumed to be quantities related to a virtual AC motor as shown in Fig. 2.19.

Thus, \(R\) and \(L\) represent the stator resistance and the stator leakage inductance of the virtual motor and phase-to-phase line voltages: \(U_{ab}, U_{bc}, U_{ca}\) would be induced by a virtual air gap flux. In other words the integration of the voltages leads to a virtual line flux vector \(\Psi_l\), in stationary \(\alpha-\beta\) coordinates (Fig. 2.20).
Similarly to Eq. (2.39) a virtual flux equation can be presented as [65, 102] (Fig. 2.21):

\[
\psi_{L(\text{est})} = \psi_S + \psi_I
\]

(2.40)

Fig. 2.21 Relation between voltage and flux for different power flow direction in PWM rectifier.
Based on the measured DC-link voltage $U_{dc}$ and the converter switch states $S_a$, $S_b$, $S_c$ the rectifier input voltages are estimated as follows

\[ u_{Sa} = \sqrt{\frac{2}{3}} U_{dc} (S_a - \frac{1}{2} (S_b + S_c)) \]  

\[ u_{S\beta} = \frac{1}{\sqrt{2}} U_{dc} (S_b - S_c) \]

(2.41a)  

(2.41b)

Then, the virtual flux $\Psi_L$ components are calculated from the (2.41) in stationary ($\alpha$-$\beta$) coordinates system

\[ \Psi_{L,\alpha\text{ (est)}} = \int (u_{Sa} + L \frac{di_{La}}{dt}) dt \]  

\[ \Psi_{L,\beta\text{ (est)}} = \int (u_{S\beta} + L \frac{di_{L\beta}}{dt}) dt \]

(2.42a)  

(2.42b)

The virtual flux components calculation is shown in Fig. 2.22.

![Fig. 2.22. Block scheme of virtual flux estimator with first order filter.](image-url)
3. VOLTAGE AND VIRTUAL FLUX BASED DIRECT POWER CONTROL

3.1 INTRODUCTION

Control of PWM rectifier can be considered as a dual problem to vector control of an induction motor (Fig. 3.1) [4,110]. Various control strategies have been proposed in recent works on this type PWM converter. Although these control strategies can achieve the same main goals, such as the high power factor and near-sinusoidal current waveforms, their principles differ. Particularly, the Voltage Oriented Control (VOC), which guarantees high dynamics and static performance via an internal current control loops, has become very popular and has constantly been developed and improved [46, 48], [51], [53-54]. Consequently, the final configuration and performance of the VOC system largely depends on the quality of the applied current control strategy [6]. Another control strategy called Direct Power Control (DPC) is based on the instantaneous active and reactive power control loops [21], [22]. In DPC there are no internal current control loops and no PWM modulator block, because the converter switching states are selected by a switching table based on the instantaneous errors between the commanded and estimated values of active and reactive power. Therefore, the key point of the DPC implementation is a correct and fast estimation of the active and reactive line power.

![Diagram 3.1 Relationship between control of PWM line rectifier and PWM inverter – fed IM](image)

The control techniques for PWM rectifier can be generally classified as voltage based and virtual flux based, as shown in Fig. 3.2. The virtual flux based method corresponds to direct analogy of IM control.

![Diagram 3.2 Classification of control methods for PWM rectifier](image)
3.2 BASIC BLOCK DIAGRAM OF DIRECT POWER CONTROL (DPC)

The main idea of DPC proposed in [22] and next developed by [21] is similar to the well-known Direct Torque Control (DTC) for induction motors. Instead of torque and stator flux the instantaneous active ($p$) and reactive ($q$) powers are controlled (Fig. 3.3).

![Block scheme of DPC](image)

The commands of reactive power $q_{ref}$ (set to zero for unity power factor) and active power $p_{ref}$ (delivered from the outer PI-DC voltage controller) are compared with the estimated $q$ and $p$ values (described in section 3.3 and 3.4), in reactive and active power hysteresis controllers, respectively.

The digitized output signal of the reactive power controller is defined as:

\[
\begin{align*}
    dq &= 1 \text{ for } q < q_{ref} - H_q \\
    dq &= 0 \text{ for } q > q_{ref} + H_q,
\end{align*}
\]

(3.1a, 3.1b)

and similarly of the active power controller as

\[
\begin{align*}
    dp &= 1 \text{ for } p < p_{ref} - H_p \\
    dp &= 0 \text{ for } p > p_{ref} + H_p,
\end{align*}
\]

(3.2a, 3.2b)

where: $H_q$ & $H_p$ are the hysteresis bands.

The digitized variables $d_p, d_q$ and the voltage vector position $\gamma_{UL} = \text{arc tg} \left(\frac{u_{L\alpha}}{u_{L\beta}}\right)$ or flux vector position $\gamma_{\psi L} = \text{arc tg} \left(\frac{\psi_{L\alpha}}{\psi_{L\beta}}\right)$ form a digital word, which by accessing the address of the look-up table selects the appropriate voltage vector according to the switching table (described in section 3.5).

The region of the voltage or flux vector position is divided into twelve sectors, as shown in Fig. 3.5 and the sectors can be numerically expressed as:

\[
(n-2)\frac{\pi}{6} \leq \gamma_n < (n-1)\frac{\pi}{6} \quad \text{where } n = 1, 2...12
\]

(3.3)
Note, that the sampling frequency has to be about few times higher than the average switching frequency. This very simple solution allows precisely control of instantaneous active and reactive power and errors are only limited by the hysteresis band. No transformation into rotating coordinates is needed and the equations are easy implemented. This method deals with instantaneous variables, therefore, estimated values contain not only a fundamental but also harmonic components. This feature also improves the total power factor and efficiency [21].

Further improvements regarding VF-DPC operation can be achieved by using sector detection with PLL (Phase-Locked Loop) generator instead of zero crossing voltage detector (Fig. 3.6). This guarantees a very stable and free of disturbances sector detection, even under operation with distorted and unbalanced line voltages (Fig.3.19).
3.3 INSTANTANEOUS POWER ESTIMATION BASED ON THE LINE VOLTAGE

The main idea of voltage based power estimation for DPC was proposed in [21-22]. The instantaneous active and reactive powers are defined by the product of the three phase voltages and currents (2.25-2.26). The instantaneous values of active \( p \) and reactive power \( q \) in AC voltage sensorless system are estimated by Eqs. (3.8) and (3.9). The active power \( p \) is the scalar product of the current and the voltage, whereas the reactive power \( q \) is calculated as a vector product of them. The first part of both equations represents power in the inductance and the second part is the power of the rectifier.

\[
p = L\left(\frac{d}{dt}i_a +\frac{d}{dt}i_b +\frac{d}{dt}i_c\right) + U_{dc}(S_d i_a + S_b i_b + S_c i_c) \quad (3.8)
\]

\[
q = \frac{1}{\sqrt{3}}\left[3L\left(\frac{d}{dt}i_a - \frac{d}{dt}i_b - \frac{d}{dt}i_c\right) - U_{dc}[S_d(i_b - i_c) + S_b(i_c - i_a) + S_c(i_a - i_b)]\right] \quad (3.9)
\]

As can be seen in (3.8) and (3.9), the form of equations have to be changed according to the switching state of the converter, and both equations require the knowledge of the line inductance \( L \). Supply voltage usually is constant, therefore the instantaneous active and reactive powers are proportional to the \( i_{Ld} \) and \( i_{Lq} \).

The AC-line voltage sector is necessary to read the switching table, therefore knowledge of the line voltage is essential. However, once the estimated values of active and reactive power are calculated and the AC-line currents are known, the line voltage can easily be calculated from instantaneous power theory as:

\[
\begin{bmatrix}
    u_{La}
    \\
    u_{L\beta}
\end{bmatrix} = \frac{1}{i_{La}^2+i_{L\beta}^2}\begin{bmatrix}
    i_{La} & \quad i_{L\beta} & \quad p
    \\
    i_{L\beta} & \quad -i_{La} & \quad q
\end{bmatrix}
\]

(3.10)

The instantaneous power and AC voltage estimators are shown in Fig. 3.7.

Fig. 3.7 Instantaneous power estimator based on line voltage.
In spite of the simplicity, this power estimation method has several disadvantages such as:

- high values of the line inductance and sampling frequency are needed (important point for the estimator, because a smooth shape of current is needed).
- power estimation depends on the switching state. Therefore, calculation of the power and voltage should be avoided at the moment of switching, because of high estimation errors.

3.4 Instantaneous Power Estimation Based on the Virtual Flux

The Virtual Flux (VF) based approach has been proposed by Author to improve the VOC [42, 56]. Here it will be applied for instantaneous power estimation, where voltage imposed by the line power in combination with the AC side inductors are assumed to be quantities related to a virtual AC motor as shown in section 2.5.

With the definitions

\[ \Psi_L = \int u_L \, dt \]  
(3.11)

where

\[ u_L = \begin{bmatrix} u_{la} \\ u_{lb} \end{bmatrix} = \sqrt{2/3} \begin{bmatrix} 1/2 \\ \sqrt{3}/2 \end{bmatrix} \begin{bmatrix} u_a \\ u_b \end{bmatrix} \]  
(3.12)

\[ \Psi_L = \begin{bmatrix} \Psi_{la} \\ \Psi_{lb} \end{bmatrix} = \begin{bmatrix} \int u_{la} \, dt \\ \int u_{lb} \, dt \end{bmatrix} \]  
(3.13)

\[ i_L = \begin{bmatrix} i_{la} \\ i_{lb} \end{bmatrix} = \sqrt{2/3} \begin{bmatrix} 3/2 \\ \sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_a \\ i_b \end{bmatrix} \]  
(3.14)

\[ u_s = u_{\text{comp}} = \begin{bmatrix} u_{sa} \\ u_{sb} \end{bmatrix} = \begin{bmatrix} 1 \\ -1/2 \end{bmatrix} \begin{bmatrix} u_{AM} \\ u_{BM} \end{bmatrix} \]  
(3.15)

the voltage equation can be written as

\[ u_L = R_i \frac{d}{dt} (L_i + \Psi_s) \]  
(3.16a)

In practice, \( R \) can be neglected, giving

\[ u_L = L \frac{d}{dt} i_L + \frac{d}{dt} \Psi_s = L \frac{d}{dt} i_L + u_s \]  
(3.16b)

Using complex notation, the instantaneous power can be calculated as follows:

\[ p = \text{Re}(u_L \cdot i_L^*) \]  
(3.17a)

\[ q = \text{Im}(u_L \cdot i_L^*) \]  
(3.17b)
where \( * \) denotes the conjugate line current vector. The line voltage can be expressed by the virtual flux as

\[
 u_L = \frac{d}{dt} \Psi_L = \frac{d}{dt}(\Psi_L e^{j\alpha}) = \frac{d\Psi_L}{dt} e^{j\alpha} + j\omega \Psi_L e^{j\alpha} = \frac{d\Psi_L}{dt} e^{j\alpha} + j\omega \Psi_L \tag{3.18}
\]

where \( \Psi_L \) denotes the space vector and \( \Psi_L \) its amplitude. For the virtual flux oriented \( d-q \) coordinates (Fig. 2.20), \( \Psi_L = \Psi_{Ld} \), and the instantaneous active power can be calculated from (3.17a) and (3.18) as

\[
p = \frac{d\Psi_L}{dt} i_{Ld} + \omega \Psi_{Ld} i_{Lq} \tag{3.19}
\]

For sinusoidal and balanced line voltages, equation (3.19) is reduced to

\[
\frac{d\Psi_{Ld}}{dt} = 0 \tag{3.20}
\]

\[
p = \omega \Psi_{Ld} i_{Lq} \tag{3.21}
\]

which means that only the current components orthogonal to the flux \( \Psi_L \) vector, produce the instantaneous active power. Similarly, the instantaneous reactive power can be calculated as:

\[
q = -\frac{d\Psi_{Ld}}{dt} i_{Lq} + \omega \Psi_{Ld} i_{Ld} \tag{3.22}
\]

and with (3.20) it is reduced to:

\[
q = \omega \Psi_{Ld} i_{Ld} \tag{3.23}
\]

However, to avoid coordinate transformation into \( d-q \) coordinates, the power estimator for the DPC system should use stator-oriented quantities, in \( \alpha-\beta \) coordinates (Fig.2.20).

Using (3.17) and (3.18)

\[
u_L = \frac{d\Psi_L}{dt} \bigg|_\alpha + j\frac{d\Psi_L}{dt} \bigg|_\beta + j\omega \left( \Psi_{L\alpha} + j\Psi_{L\beta} \right) \tag{3.24}
\]

\[
u_L = \left\{ \frac{d\Psi_L}{dt} \bigg|_\alpha + j\frac{d\Psi_L}{dt} \bigg|_\beta + j\omega \left( \Psi_{L\alpha} + j\Psi_{L\beta} \right) \right\} \left( i_{L\alpha} - j i_{L\beta} \right) \tag{3.25}
\]

That gives

\[
p = \left\{ \frac{d\Psi_L}{dt} \bigg|_{i_{L\alpha}} + \frac{d\Psi_L}{dt} \bigg|_{i_{L\beta}} \right\} i_{L\alpha} + \omega \left( \Psi_{Ld} i_{Ld} - \Psi_{Lq} i_{Lq} \right) \tag{3.26a}
\]

and

\[
q = \left\{ \frac{d\Psi_L}{dt} \bigg|_{i_{L\beta}} + \frac{d\Psi_L}{dt} \bigg|_{i_{L\alpha}} \right\} i_{L\alpha} + \omega \left( \Psi_{Ld} i_{Ld} + \Psi_{Lq} i_{Lq} \right) \tag{3.26b}
\]
For sinusoidal and balanced line voltage the derivatives of the flux amplitudes are zero. The instantaneous active and reactive powers can be computed as [17-19]

\[
p = \omega \cdot (\Psi_{La} i_{La} - \Psi_{L\beta} i_{L\beta}) \quad (3.27a)
\]

\[
q = \omega \cdot (\Psi_{La} i_{La} + \Psi_{L\beta} i_{L\beta}) . \quad (3.27b)
\]

The measured line currents \(i_a, i_b\) and the estimated virtual flux components \(\Psi_{La}, \Psi_{L\beta}\) are delivered to the instantaneous power estimator block (\(PE\)) as depicted in Fig. 3.8.

**3.5 SWITCHING TABLE**

It can be seen in Fig. 3.9, that the instantaneous active and reactive power depends on position of converter voltage vector. It has indirect influence on inductance voltage as well as phase and amplitude of line current. Therefore, different pattern of switching table can be applied to direct control (DTC, DPC). It influence control condition as: instantaneous power and current ripple, switching frequency and dynamic performance. Some works, propose different switching tables for DTC but we cannot find too much reference for DPC. For drives exist more switching table techniques because of wide range of output frequency and dynamic demands [24-29]. For PWM rectifier we have constant line frequency and only instantaneous power varies. Fig. 3.9 presents four different situations, which illustrate the variations of instantaneous power. Point \(M\) presents reference values of active and reactive power.
The selection of vector is made so that the error between $q$ and $q_{ref}$ should be within the limits (Eqs. (3.1),(3.2)). It depends not only on the error of the amplitude but also the direction of $q$ as shown in Fig. 3.10.
Some behaviour of DPC are not satisfactory. For instance when the instantaneous reactive power vector is close to one of sector boundary, two of four possible active vectors are wrong. These wrong vectors can only change the instantaneous active power without correction of the reactive power error. This is easy visible on a current. A few methods to improve the DPC behaviour in the sector bonders is well known. One of them is to add more sectors or hysteresis levels. Therefore, switching table are generally constructed with difference in:

- number of sectors,
- dynamic performance,
- two and three level hysteresis controllers.

**Number of sectors**

Usually the vectors plane is divided for 6 (3. 28) or 12 (3. 29) sectors (Fig. 3.11). It has influence for switching table construction (Table 3.1).

![Voltage plane with 6 sectors](image-a)

![Voltage plane with 12 sectors](image-b)

Fig. 3.11 Voltage plane with a) 6 sectors  b) 12 sectors

\[
(2n-3)\frac{\pi}{6} \leq \gamma_n < (2n-1)\frac{\pi}{6} \quad n = 1, 2, ..., 6 \tag{3.28}
\]

\[
(n-2)\frac{\pi}{6} \leq \gamma_n < (n-1)\frac{\pi}{6} \quad n = 1, 2, ..., 12 \tag{3.29}
\]

<table>
<thead>
<tr>
<th>(d_x)</th>
<th>(d_y)</th>
<th>Sector A</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0</td>
<td>(U_A)</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>(U_B)</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>(U_0)</td>
</tr>
</tbody>
</table>

\(U_A=U_{1}(100),U_{2}(110),U_{3}(010),U_{4}(011),U_{5}(001),U_{6}(101)\)
\(U_B=U_{6}(101),U_{1}(100),U_{2}(110),U_{3}(010),U_{4}(011),U_{5}(001)\)
\(U_0=U_0(000),U_7(111)\)

<table>
<thead>
<tr>
<th>(d_x)</th>
<th>(d_y)</th>
<th>Sector A</th>
<th>Sector B</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0</td>
<td>(U_A)</td>
<td>(U_B)</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>(U_B)</td>
<td>(U_A)</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>(U_0)</td>
<td>(U_0)</td>
</tr>
</tbody>
</table>

\(U_A=U_{1}(100),U_{2}(110),U_{3}(010),U_{4}(011),U_{5}(001),U_{6}(101)\)
\(U_B=U_{6}(101),U_{1}(100),U_{2}(110),U_{3}(010),U_{4}(011),U_{5}(001)\)
\(U_0=U_0(000),U_7(111)\)

When region of the voltage vector position is divided into twelve sectors, the area between adjoining vectors contain two sectors. Sector \(A\) is located closer to \(U_A\) and sector \(B\) closer to \(U_B\).
Hysteresis controllers

The wide of the instantaneous active and reactive hysteresis band have a relevant effect on the converter performance. In particular, the harmonic current distortion, the average converter switching frequency, the power pulsation and the losses are strongly affected by the hysteresis wide. The controllers proposed by [21] in classical DPC are two level comparators for instantaneous active and reactive power (Fig 3.12a). Three level comparators can provide further improvements. Possible combinations of hysteresis controllers for active and reactive power are presented in Fig. 3.12.

The two level hysteresis controllers for instantaneous reactive power can be described as

- If $\Delta q > H_q$ then $d_q = 1$
- If $-H_q \leq \Delta q \leq H_q$ and $d\Delta q/dt > 0$ then $d_q = 0$
- If $-H_q \leq \Delta q \leq H_q$ and $d\Delta q/dt < 0$ then $d_q = 1$
- If $\Delta q < -H_q$ then $d_q = 0$.

The three level hysteresis controllers for the instantaneous active power can be described as a sum of two level hysteresis

- If $\Delta p > H_p$ then $d_p = 1$
- If $0 \leq \Delta p \leq H_p$ and $d\Delta p/dt > 0$ then $d_p = 0$
- If $0 \leq \Delta p \leq H_p$ and $d\Delta p/dt < 0$ then $d_p = 1$
- If $-H_q \leq \Delta p \leq 0$ and $d\Delta p/dt > 0$ then $d_p = -1$
- If $-H_q \leq \Delta p \leq 0$ and $d\Delta p/dt < 0$ then $d_p = 0$
- If $\Delta p < -H_p$ then $d_p = -1$.

Dynamic performance

Combinations of each converter voltage space vector used for instantaneous active and reactive power variation are summarized in Table 3.3. Situation is presented for vector located in the $k$-th sector ($k = 1, 2, 3, 4, 5, 6$) of the $\alpha, \beta$ plane as shown in Fig. 3.13 [24]. In the table, a single arrow means a small variation, whereas two arrows mean a large variation. As it appears from the table, an increment of reactive power ($\uparrow$) is obtained by applying the space vector $U_k, U_{k+1}$ and $U_{k+2}$. Conversely, a decrement of reactive power ($\downarrow$) is obtained by applying vector $U_{k-2}, U_{k-1}$, or $U_{k+3}$. Active power increase when $U_{k+2}, U_{k+3}, U_{k+1}, U_{k-2}$ or $U_0, U_7$ are applied and active power decrease when $U_k, U_{k+1}$ are applied.
Table 3.3 Instantaneous active and reactive variations due to the applied voltage vectors

<table>
<thead>
<tr>
<th></th>
<th>$U_{K-2}$</th>
<th>$U_{K-1}$</th>
<th>$U_K$</th>
<th>$U_{K+1}$</th>
<th>$U_{K+2}$</th>
<th>$U_{K+3}$</th>
<th>$U_0$</th>
<th>$U_7$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$q$</td>
<td>↓↓</td>
<td>↓</td>
<td>↑↑</td>
<td>↑</td>
<td>↑</td>
<td>↓</td>
<td>↑</td>
<td>↑↓</td>
</tr>
<tr>
<td>$p$</td>
<td>↑</td>
<td>↓</td>
<td>↓</td>
<td>↑</td>
<td>↑↑</td>
<td>↑↑</td>
<td>↑</td>
<td>↑</td>
</tr>
</tbody>
</table>

General features of switching table and hysteresis controllers

- The switching frequency depends on the hysteresis wide of active and reactive power comparators.
- By using three-level comparators, the zero vectors are naturally and systematically selected. Thus, the number of switching is considerably smaller than in the system with two-level hysteresis comparators.
- Zero vectors decrease switching frequency but it provides short-circuit for the line to line voltage.
- Zero vectors $U_0(000)$ and $U_7(111)$ should be appropriate chosen.
- For DPC only the neighbour vectors should be selected what decrease dynamics but provide low current and power ripples (low THD).
- Switching table with PLL (Phase-Locked Loop) sector detection guarantees a very stable and free of disturbances operation, even under distorted and unbalanced line voltages.
- 12 sectors provide more accurate voltage vector selection.
3.6 SIMULATION AND EXPERIMENTAL RESULTS

To study the operation of the *VF-DPC* system under different line conditions and to carry out a comparative investigation, the PWM rectifier with the whole control scheme has been simulated using the *SABER* software [A.4]. The main electrical parameters of the power circuit and control data are given in the Table A.4.1. The simulation study has been performed with two main objectives:

- explaining and presenting the steady state operation of the proposed by Author *VF-DPC* with a purely sinusoidal and distorted unbalanced supply line voltage, as well as performance comparison with the conventional scheme where the instantaneous power is estimated based on calculated voltage (not virtual flux) signals [21];
- presenting the dynamic performance of power control.

The simulated waveforms for the proposed by Author *VF-DPC* and for the *DPC* reported in [21] are shown in Fig. 3.14. These results were obtained for purely sinusoidal supply line voltage. Similarly Fig. 3.15 shows on oscillogram for distorted (5% of 5-th harmonic) and unbalanced (4,5%) line voltages (see A.1). Fig. 3.15 and Fig. 3.16 show that *VF-DPC* provides sinusoidal and balanced line currents even at distorted and unbalanced supply voltage. This is thanks to fact that voltage was replaced by virtual flux.

The dynamic behaviour under a step change of the load is presented in Fig. 3.21. Note, that in spite of the lower sampling frequency (50 kHz), the *VF* based power estimator gives much less noisy instantaneous active and reactive power signals (Fig. 3.21b) in comparison to the conventional *DPC* system with 80 kHz sampling frequency (Fig. 3.21a). This is thanks to the natural low-pass filter behaviour of the integrators used in (2.42) (because *k*-th harmonics are reduced by a factor 1/*k* and the ripple caused by high frequency power transistor switching is effectively damped). Consequently, the derivation of the line current, which is necessary in conventional *DPC* for sensorless voltage estimation, is in the *VF-DPC* eliminated. However, the dynamic behaviour of both control systems, are identical (see Fig. 3.21). The excellent dynamic properties of the *VF-DPC* system at distorted and unbalanced supply voltage are shown in Fig. 3.22.

Experimental results were realized on laboratory setup presented in A.6. The main electrical parameters of the power circuit and control data are given in the Table A.6.2. The experimental results are measured for significantly distorted line voltage what is presented in Fig. 3.17. Steady state operation for *DPC* and *VF-DPC* are shown in Fig. 3.18 - 3.20. The shape of the current for conventional DPC is strongly distorted because two undesirable conditions are applied:

- sampling time was 20μs (should be about 10μs [21]),
- the line voltage was not purely sinusoidal.

*VF-DPC* in comparison with the conventional solution at the same condition provides sinusoidal current (Fig. 3.19-3.20) with low total harmonic distortion. The dynamic behaviour under a step change of the load for *VF-DPC* are shown in Fig. 3.23-3.24.
**STEADY STATE BEHAVIOUR**

➤ **RESULTS UNDER PURELY SINUSOIDAL LINE VOLTAGE (SIMULATION)**

Fig. 3.14 Simulated basic signal waveforms and line current harmonic spectrum under purely sinusoidal line voltage: a) conventional DPC presented in [21], b) proposed VF-DPC. From the top: line voltage, estimated line voltage (left) and estimated virtual flux (right), line currents, instantaneous active and reactive power, harmonic spectrum of the line current.

DPC THD = 5.6%, VF-DPC THD = 5.2%.

➤ **RESULTS UNDER NON SINUSOIDAL LINE VOLTAGE (SIMULATION)**

Fig. 3.15. Simulated waveforms and line current harmonic spectrum under pre-distorted (5% of 5th harmonic) and unbalanced (4.5%) line voltage for conventional DPC and VF-DPC. From the top: line voltage, estimated line voltage(left) and virtual flux (right), line currents, harmonic spectrum of the line current.
Direct Power Control (DPC)

Fig. 3.16. Simulated basic signal waveforms in the VF-DPC under pre-distorted (5% of 5th harmonic) and unbalanced (4.5%) line voltage.
From the top: line voltages, line currents. THD = 5.6%

**RESULTS UNDER NON SINUSOIDAL LINE VOLTAGE (EXPERIMENT)**

Fig. 3.17. Line voltage with harmonic spectrum
(u_L – line voltage, u_diff - distortion from purely sinusoidal supply line voltage).
Fig. 3.18. Experimental waveforms with distorted line voltage for conventional DPC. From the top: line voltage, line currents (5A/div) and estimated virtual flux.

Fig. 3.19. Experimental waveforms with distorted line voltage for VF- DPC. From the top: line voltage, line currents (5A/div) and estimated virtual flux.
Fig. 3.20. Experimental waveforms with distorted line voltage for VF-DPC. From the top: line voltage, line currents (5A/div), instantaneous active (2 kW/div) and reactive power (2 kVAr/div), harmonic spectrum of line current ($THD = 5.6\%$) [17].
DYNAMIC BEHAVIOUR

➤ RESULTS UNDER PURELY SINUSOIDAL LINE VOLTAGE (*SIMULATION*)

![Graphs of DPC and VF-DPC](image)

Fig. 3.21. Transient of the step change of the load:
(a) conventional *DPC* presented in [21], (b) proposed *VF-DPC*.
From the top: line voltage, line currents, instantaneous active and reactive power.

➤ RESULTS UNDER NON SINUSOIDAL LINE VOLTAGE (*SIMULATIONS*)

![Graphs of DPC and VF-DPC](image)

Fig. 3.22. Transient to the step change of the load in the *VF-DPC*:
(a) load increasing (b) load decreasing.
From the top: line voltages, line currents, instantaneous active and reactive power.
RESULTS UNDER NON SINUSOIDAL LINE VOLTAGE (EXPERIMENT)

Fig. 3.23. Transient of the step change of the load in the improved VF-DPC: load increasing. From the top: line voltages, line currents (5A/div), instantaneous active (2 kW/div) and reactive power (2 kVAr/div).

Fig. 3.24 Transient of the step change of the load in the improved VF-DPC: start-up of converter. From the top: line voltages, line currents (5A/div), instantaneous active (2 kW/div) and reactive power (2 kVAr/div).
3.8 SUMMARY

The presented DPC system constitutes a viable alternative to the VOC system [see Chapter 4] of PWM line rectifiers. However, conventional solution shown by [21] possess several disadvantages:

- the estimated values are changed every time according to the switching state of the converter, therefore, it is important to have high sampling frequency. (good performance is obtained at 80kHz sampling frequency, it means that result precisely depends on sampling time),
- the switching frequency is not constant, therefore, a high value of inductance is needed (about 10%). (this is an important point for the line voltage estimation because a smooth shape of current is needed),
- the wide range of the variable switching frequency can be problem, when designing the necessary LC input filter,
- calculation of power and voltage should be avoided at the moment of switching because it gives high errors of the estimated values.

Based on duality with a PWM inverter-fed induction motor, a new method of instantaneous active and reactive power calculation has been proposed. This method uses the estimated Virtual Flux (VF) vector instead of the line voltage vector. Consequently, voltage sensorless line power estimation is much less noisy thanks to the natural low-pass behaviour of the integrator used in the calculation algorithm. Also, differentiation of the line current is avoided in this scheme. So, the presented VF-DPC of PWM rectifier has the following features and advantages:

- no line voltage sensors are required,
- simple and noise robust power estimation algorithm, easy to implement in a DSP,
- lower sampling frequency (as conventional DPC [21]),
- sinusoidal line currents (low THD),
- no separate PWM voltage modulation block,
- no current regulation loops,
- coordinate transformation and PI controllers are not required,
- high dynamic, decoupled active and reactive power control,
- power and voltage estimation gives possibility to obtain instantaneous variables with all harmonic components, what have influence for improvement of total power factor and efficiency [21].

The typical disadvantages are:

- variable switching frequency,
- fast microprocessor and A/D converters, are required.

As shown in the Chapter 3, thanks to duality phenomena, an experience with the high performance decoupled PWM inverter-fed induction motor control can be used to improve properties of the PWM rectifier control.
4. VOLTAGE AND VIRTUAL FLUX ORIENTED CONTROL (VOC, VFOC)

4.1 INTRODUCTION

Similarly as in FOC of an induction motor [4], the Voltage Oriented Control (VOC) and Virtual Flux Oriented Control (VFOC) for line side PWM rectifier is based on coordinate transformation between stationary $\alpha-\beta$ and synchronous rotating $d-q$ reference system. Both strategies guarantees fast transient response and high static performance via an internal current control loops. Consequently, the final configuration and performance of system largely depends on the quality of applied current control strategy [6]. The easiest solution is hysteresis current control that provides a fast dynamic response, good accuracy, no DC offset and high robustness. However the major problem of hysteresis control is that its average switching frequency varies with the load current, which makes the switching pattern uneven and random, thus, resulting in additional stress on switching devices and difficulties of LC input filter design. Therefore, several strategies are reported in literature to improve performance of current control [2], [38-40], [68-69]. Among presented regulators the widely used scheme for high performance current control is the $d-q$ synchronous controller, where the currents being regulated are DC quantities what eliminates steady state error.

4.2 BLOCK DIAGRAM OF THE VOLTAGE ORIENTED CONTROL (VOC)

The conventional control system uses closed-loop current control in rotating reference frame, the Voltage Oriented Control (VOC) scheme is shown in Fig. 4.1.

![Fig. 4.1 Block scheme of AC voltage sensorless VOC](image)

A characteristic feature for this current controller is processing of signals in two coordinate systems. The first is stationary $\alpha-\beta$ and the second is synchronously rotating $d-q$ coordinate system. Three phase measured values are converted to equivalent two-phase system $\alpha-\beta$ and then are transformed to rotating coordinate system in a block $\alpha-\beta/d-q$: 
\[
\begin{bmatrix}
    k_d \\
    k_q
\end{bmatrix} =
\begin{bmatrix}
    \cos \gamma_{UL} & \sin \gamma_{UL} \\
    -\sin \gamma_{UL} & \cos \gamma_{UL}
\end{bmatrix}
\begin{bmatrix}
    k_\alpha \\
    k_\beta
\end{bmatrix} \quad (4.1a)
\]

Thanks to this type of transformation the control values are DC signals. An inverse transformation \(d-q/\alpha-\beta\) is achieved on the output of control system and it gives a result the rectifier reference signals in stationary coordinate:

\[
\begin{bmatrix}
    k_\alpha \\
    k_\beta
\end{bmatrix} =
\begin{bmatrix}
    \cos \gamma_{UL} & -\sin \gamma_{UL} \\
    \sin \gamma_{UL} & \cos \gamma_{UL}
\end{bmatrix}
\begin{bmatrix}
    k_d \\
    k_q
\end{bmatrix} \quad (4.1b)
\]

For both coordinate transformation the angle of the voltage vector \(\gamma_{UL}\) is defined as:

\[
\sin \gamma_{UL} = \frac{u_{L\beta}}{\sqrt{(u_{L\alpha})^2 + (u_{L\beta})^2}} \quad (4.2a)
\]

\[
\cos \gamma_{UL} = \frac{u_{L\alpha}}{\sqrt{(u_{L\alpha})^2 + (u_{L\beta})^2}}. \quad (4.2b)
\]

In voltage oriented \(d-q\) coordinates, the AC line current vector \(i_L\) is split into two rectangular components \(i_L = [i_{Ld}, i_{Lq}]\) (Fig. 4.2). The component \(i_{Lq}\) determinates reactive power, whereas \(i_{Ld}\) decides about active power flow. Thus the reactive and the active power can be controlled independently. The UPF condition is met when the line current vector, \(i_L\), is aligned with the line voltage vector, \(u_L\). (Fig. 2.7b) By placing the d-axis of the rotating coordinates on the line voltage vector a simplified dynamic model can be obtained.

![Fig. 4.2: Vector diagram of VOC. Coordinate transformation of line current, line voltage and rectifier input voltage from stationary \(\alpha-\beta\) coordinates to rotating \(d-q\) coordinates.](image)

The voltage equations in the \(d-q\) synchronous reference frame in accordance with equations 2.19 are as follows:

\[
u_{Ld} = R \cdot i_{Ld} + L \frac{di_{Ld}}{dt} + u_{Sd} - \omega \cdot L \cdot i_{Lq} \quad (4.3)
\]

\[
u_{Lq} = R \cdot i_{Lq} + L \frac{di_{Lq}}{dt} + u_{Sq} + \omega \cdot L \cdot i_{Ld} \quad (4.4)
\]

Regarding to Fig. 4.1, the \(q\)-axis current is set to zero in all condition for unity power factor control while the reference current \(i_{Ld}\) is set by the DC-link voltage controller and
Voltage and Virtual Flux Oriented Control

controls the active power flow between the supply and the DC-link. For $R \approx 0$ equations (4.3), (4.4) can be reduced to:

$$u_{Ld} = L \frac{di_{Ld}}{dt} + u_{Sd} - \omega \cdot L \cdot i_{Lq}$$  \hspace{1cm} (4.5)$$

$$0 = L \frac{di_{Lq}}{dt} + u_{Sq} + \omega \cdot L \cdot i_{Ld}$$  \hspace{1cm} (4.6)$$

Assuming that the q-axis current is well regulated to zero, the following equations hold true

$$u_{Ld} = L \frac{di_{Ld}}{dt} + u_{Sd}$$  \hspace{1cm} (4.7)$$

$$0 = u_{Sq} + \omega \cdot L \cdot i_{Ld}$$  \hspace{1cm} (4.8)$$

As current controller, the PI-type can be used. However, the PI current controller has no satisfactory tracing performance, especially, for the coupled system described by Eqs. (4.5), (4.6). Therefore for high performance application with accuracy current tracking at dynamic state the decoupled controller diagram for the PWM rectifier should be applied what is shown in Fig. 4.3 [49]:

$$u_{Sd} = \omega L i_{Lq} + u_{Ld} + \Delta u_d$$  \hspace{1cm} (4.9)$$

$$u_{Sq} = -\omega L i_{Ld} + \Delta u_q$$  \hspace{1cm} (4.10)$$

where $\Delta$ is the output signals of the current controllers

$$\Delta u_d = k_p (i_d^* - i_d) + k_i \int (i_d^* - i_d) dt$$  \hspace{1cm} (4.11)$$

$$\Delta u_q = k_p (i_q^* - i_q) + k_i \int (i_q^* - i_q) dt$$  \hspace{1cm} (4.12)$$

The output signals from PI controllers after $dq/\alpha\beta$ transformation (Eq. (4.1b)) are used for switching signals generation by a Space Vector Modulator [see Section 4.4].

![Fig. 4.3 Decoupled current control of PWM rectifier](image-url)
4.3 BLOCK DIAGRAM OF THE VIRTUAL FLUX ORIENTED CONTROL (VFOC)

The concept of Virtual Flux (VF) can also be applied to improve VOC scheme, because disturbances superimposed onto the line voltage influence directly the coordinate transformation in control system (4.2). Sometimes this is solved only by phase-locked loops (PLL’s) only, but the quality of the controlled system depends on how effectively the PLL’s have been designed [31]. Therefore, it is easier to replace angle of the line voltage vector $\gamma_{UL}$ by angle of VF vector $\gamma_{\Psi L}$, because $\gamma_{\Psi L}$ is less sensitive than $\gamma_{UL}$ to disturbances in the line voltage, thanks to the natural low-pass behavior of the integrators in (2.42) (because $n$th harmonics are reduced by a factor $1/k$ and the ripple related to the high frequency transistor is strongly damped). For this reason, it is not necessary to implement PLL’s to achieve robustness in the flux-oriented scheme, since $\Psi L$ rotates much more smoothly than $u L$. The angular displacement of virtual flux vector $\Psi L$ in $\alpha$-$\beta$ coordinate is defined as:

$$\sin \gamma_{\Psi L} = \Psi_{L\beta} / \sqrt{\left(\Psi_{L\alpha}\right)^2 + \left(\Psi_{L\beta}\right)^2}$$  \hspace{1cm} (4.13a)

$$\cos \gamma_{\Psi L} = \Psi_{L\alpha} / \sqrt{\left(\Psi_{L\alpha}\right)^2 + \left(\Psi_{L\beta}\right)^2}$$  \hspace{1cm} (4.13b)

The Virtual Flux Oriented Control (VFOC) scheme is shown in Fig. 4.4.

![Fig. 4.4 Block scheme of VFOC](image)

The vector of virtual flux lags the voltage vector by $90^\circ$ (Fig. 4.5). Therefore, for the UPF condition, the d-component of the current vector, $i_d$, should be zero.
In the virtual flux oriented coordinates voltage equations are transformed into

\[ u_{Lq} = L \frac{di_{Lq}}{dt} + u_{S_q} + \omega \cdot L \cdot i_{Ld} \]  \hspace{1cm} (4.17)

\[ 0 = L \frac{di_{Ld}}{dt} + u_{S_d} - \omega \cdot L \cdot i_{Lq} \] \hspace{1cm} (4.18)

for \( i_{Ld} = 0 \) equations (4.17) and (4.18) can be described as:

\[ u_{Lq} = L \frac{di_{Lq}}{dt} + u_{S_q} \] \hspace{1cm} (4.19)

\[ 0 = u_{S_d} - \omega \cdot L \cdot i_{Lq} \] \hspace{1cm} (4.20)
4.4 PULSE WIDTH MODULATION (PWM)

4.4.1 Introduction
Application and power converter topologies are still expanding thanks to improvements in semiconductor technology, which offer higher voltage and current rating as well as better switching characteristics. On the other hand, the main advantages of modern power electronic converters such as: high efficiency, low weight and small dimensions, fast operation and high power densities are being achieved through the use of the so-called switch mode operation, in which power semiconductor devices are controlled in ON/OFF fashion. This leads to different types of Pulse Width Modulation (PWM), which is a basic energy processing technique applied in power converter systems. In modern converters, PWM is a high-speed process ranging – depending on the rated power – from a few kHz (motor control) up to several MHz (resonant converters for power supply). Therefore, an on-line optimisation procedure is hard to be implemented especially for three or multi-phase converters. Development of PWM methods is, however, still in progress [70-101].

Fig. 4.7 presents a three-phase voltage source PWM converter, which is the most popular power conversion circuit used in industry. This topology can work in two modes:

- **inverter** - when energy, of adjusted amplitude and frequency, is converted from DC side to AC side. This mode is used in variable speed drives and AC power supply including uninterruptible power supply (UPS),
- **rectifier** - when energy of mains (50 Hz or 60Hz) is converted from AC side to DC side. This mode has application in power supply with Unity Power Factor (UPF).

![Fig. 4.7. Three-phase voltage source PWM converter](image)

**Basic requirements and definitions**
Performance significantly depends on control methods and type of modulation. Therefore the PWM converter, should perform some general demands like:

- *wide range of linear operation*, [3, 72, 74, 78, 81, 85, 89],
- *minimal number of (frequency) switching* to keep low switching losses in power components, [5, 72, 74, 80, 87, 93],
- *low content of higher harmonics in voltage and current*, because they produce additional losses and noise in load [5, 77],
- *elimination of low frequency harmonics* (in case of motors it generates torque pulsation)
- *operation in overmodulation region including square wave* [75, 79, 85, 89, 96].
Additionally, investigations are lead with the purpose of:

- simplification because modulator is one of the most time-consuming part of control algorithm and reducing of computations intensity at the same performance is the main point for industry (it gives possibility to use simple and inexpensive microprocessors) [76, 95, 101],
- reduction of common mode voltage [90],
- good dynamics [28, 93],
- reduction of acoustic noise (random modulation)[70].

Basic definition and parameters, which characterize PWM methods, are summarized in Tab.4.1.

<table>
<thead>
<tr>
<th>lp.</th>
<th>Name of parameter</th>
<th>Symbol</th>
<th>Definition</th>
<th>Remarques</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>Modulation index</td>
<td>M</td>
<td>( M = \frac{U_{1m}}{U_{(six-step)} = \frac{U_{1m}}{2/\pi U_d}} )</td>
<td>Two definition of modulation index are used. For sinusoidal modulation ( 0 \leq M \leq 0.785 ) or ( 0 \leq m \leq 1 )</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>( m = \frac{U_m}{U_{mit}} )</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>Max. linear range</td>
<td>( M_{max} )</td>
<td>0 ... 0.907</td>
<td>Depends on shape of modulation signal</td>
</tr>
<tr>
<td></td>
<td></td>
<td>( m_{max} )</td>
<td>0 ... 1.154</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>Overmodulation</td>
<td></td>
<td>( M &gt; M_{max} ) ( m &gt; m_{max} )</td>
<td>Nonlinear range used for increase of output voltage</td>
</tr>
<tr>
<td>4</td>
<td>Frequency modulation ratio</td>
<td>( m_f )</td>
<td>( m_f = \frac{f_s}{f_1} )</td>
<td>For ( m_f &gt; 21 ) asynchronous modulation is used</td>
</tr>
<tr>
<td>5</td>
<td>Switching frequency (number)</td>
<td>( f_r )</td>
<td>( f_r = \frac{1}{T_s} )</td>
<td>Constant</td>
</tr>
<tr>
<td>6</td>
<td>Total Harmonic Distortion</td>
<td>( THD )</td>
<td>( THD = 100 % \frac{I_h}{I_{rms}} )</td>
<td>Used for voltage and current</td>
</tr>
<tr>
<td>7</td>
<td>Current distortion factor</td>
<td>( d )</td>
<td>( h_{(rms)} / h_{(six-step)(rms)} )</td>
<td>Independent of load parameters</td>
</tr>
<tr>
<td>8</td>
<td>Polarity consistency rule</td>
<td>( PCR )</td>
<td></td>
<td>Avoids ± 1 DC voltage transition</td>
</tr>
</tbody>
</table>

### 4.4.2. Carrier Based PWM

**Sinusoidal PWM**

Sinusoidal modulation is based on triangular carrier signal. By comparison of common carrier signal with three reference sinusoidal signals \( \bar{U}_a, \bar{U}_b, \bar{U}_c \) (moved in phase of \( 2/3\pi \)) the logical signals, which define switching instants of power transistor (Fig. 4.8) are generated. Operation with constant carrier signal concentrate voltage harmonics around switching frequency and multiple of switching frequency. Narrow range of linearity is a limitation for CB-SPWM modulator because modulation index reaches \( M_{max} = \pi/4 = 0.785 \) \( (m = 1) \) only, e.g. amplitude of reference signal and carrier are equal. Overmodulation region occurs above \( M_{max} \) and PWM converter, which is treated like a power amplifier, operates at nonlinear part of characteristic (see Fig. 4.21).
**Voltage and Virtual Flux Oriented Control**

**CB-PWM with Zero Sequence Signal (ZSS)**

If neutral point on AC side of power converter $N$ is not connected with DC side midpoint $0$ (Fig. 4.7), phase currents depend only on the voltage difference between phases. Therefore, it is possible to insert an additional Zero Sequence Signal (ZSS) of 3-th harmonic frequency, which does not produce phase voltage distortion $U_{aN}$, $U_{bN}$, $U_{cN}$ and without affecting load average currents (Fig. 4.10). However, the current ripple and other modulator parameters (e.g. extending of linear region to $M_{\text{max}} = \pi / 2\sqrt{3} = 0.907$, reduction of the average switching frequency, current harmonics) are changed by the ZSS. Added ZSS occurs between $N$ and $0$ points and is visible like a $U_{N0}$ voltage and can be observed in $U_{aN}$, $U_{bN}$, $U_{cN}$ voltages (Fig. 4.10).

**Fig. 4.8.** a) Block scheme of carrier based sinusoidal modulation (CB-SPWM)  
(b) Basic waveforms

**Fig. 4.9.** Block scheme of modulator based on additional Zero Sequence Signal (ZSS).

**Fig. 4.10.** Different waveforms of additional ZSS, corresponding to different PWM methods. It can be divided in two groups: continuous and discontinuous modulation (DPWM) [92]. The most known of continuous modulation is method with sinusoidal ZSS with 1/4 amplitude, it corresponds to minimum of output current harmonics, and with 1/6 amplitude it corresponds to maximal linear range [86]. Triangular shape of ZSS with 1/4 peak corresponds to conventional (analogue) space vector modulation with symmetrical placement of zero vectors in sampling time [83]
Discontinuous modulation is formed by unmodulated 60° segments (converter power switches do not switch) shifted from 0 to π/3 (different shift \( \Psi \) gives different type of modulation Fig. 4.11). It finally gives lower (average 33%) switching losses. Detailed description of different kind of modulation based on ZSS can be found in [80].

![Diagrams](image)

Fig. 4.10. Variants of PWM modulation methods in dependence on shape of ZSS.

![Diagram](image)

Fig. 4.11. Generation of ZSS for DPWM method.

### 4.4.3. Space Vector Modulation (SVM)

**Basics of SVM**

The SVM strategy, based on space vector representation (Fig. 4.12a) becomes very popular due to its simplicity [97]. A three-phase two-level converter provides eight possible switching states, made up of six active and two zero switching states. Active vectors divide plane for six sectors, where a reference vector \( U^* \) is obtained by switching on (for proper time) two adjacent vectors. It can be seen that vector \( U^* \) (Fig. 4.12a) is possible to implement by the different switch on/off sequence of \( U_1 \) and \( U_2 \), and that zero vectors decrease modulation index. Allowable length of \( U^* \) vector, for each of \( \alpha \) angle, is equal \( U^*_{\text{max}} = \frac{U_{dc}}{\sqrt{3}} \). Higher values of output voltage (reach six-step mode) up to maximal modulation index \( (M = 1) \), can be obtained by an additional
non-linear overmodulation algorithm (see Section 4.4.5).
Contrary to CB-PWM, in the SVM there is no separate modulators for each phase.
Reference vector $U^*$ is sampled with fixed clock frequency $2f_s = 1/T_s$, and next $U'(T_s)$ is used to solve equations which describe times $t_1$, $t_2$, $t_0$, and $t_7$ (Fig. 4.12b). Microprocessor implementation is described with the help of simple trigonometrical relationship for first sector (4.21a and 4.21b), and, recalculated for the next sectors (n).

$$t_1 = \frac{2\sqrt{3}}{\pi} MT_s \sin(\frac{\pi}{3} - \alpha)$$  \hspace{1cm} (4.21a)

$$t_2 = \frac{2\sqrt{3}}{\pi} MT_s \sin \alpha$$  \hspace{1cm} (4.21b)

After $t_1$ and $t_2$ calculation, the residual sampling time is reserved for zero vectors $U_0$, $U_7$ with condition $t_1 + t_2 \leq T_s$. The equations (4.21a), (4.21b) are identical for all variants of SVM. The only difference is in different placement of zero vectors $U_0(000)$ and $U_7(111)$. It gives different equations defining $t_0$ and $t_7$ for each of method, but total duration time of zero vectors must fulfil conditions:

$$t_{0,7} = T_s - t_1 - t_2 = t_0 + t_7$$  \hspace{1cm} (4.22)

The neutral voltage between $N$ and $0$ points is equal: (see Tab. 4.2) [91]

$$U_{N\alpha} = \frac{1}{T_s} (-\frac{U_{dc}}{2} - t_0 + \frac{U_{dc}}{6} t_1 + \frac{U_{dc}}{6} t_2 + \frac{U_{dc}}{2} t_7) = \frac{1}{T_s} \left( -t_0 + \frac{t_1}{3} + \frac{t_2}{3} + t_7 \right)$$  \hspace{1cm} (4.23)

Table 4.2. Voltages between $a$, $b$, $c$ and $N$, $0$ for eight converter switching state

<table>
<thead>
<tr>
<th>$U_{a\alpha}$</th>
<th>$U_{b\alpha}$</th>
<th>$U_{c\alpha}$</th>
<th>$U_{aN}$</th>
<th>$U_{bN}$</th>
<th>$U_{cN}$</th>
<th>$U_{SN}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$U_0$</td>
<td>$-U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>$U_1$</td>
<td>$U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>$2U_{dc}/3$</td>
<td>$-U_{dc}/3$</td>
<td>$-U_{dc}/3$</td>
</tr>
<tr>
<td>$U_2$</td>
<td>$U_{dc}/2$</td>
<td>$U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>$U_{dc}/3$</td>
<td>$U_{dc}/3$</td>
<td>$-2U_{dc}/3$</td>
</tr>
<tr>
<td>$U_3$</td>
<td>$-U_{dc}/2$</td>
<td>$U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>$-U_{dc}/3$</td>
<td>$2U_{dc}/3$</td>
<td>$-U_{dc}/3$</td>
</tr>
<tr>
<td>$U_4$</td>
<td>$-U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>$U_{dc}/2$</td>
<td>$-2U_{dc}/3$</td>
<td>$U_{dc}/3$</td>
<td>$U_{dc}/3$</td>
</tr>
<tr>
<td>$U_5$</td>
<td>$-U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>$U_{dc}/3$</td>
<td>$-U_{dc}/3$</td>
<td>$2U_{dc}/3$</td>
</tr>
<tr>
<td>$U_6$</td>
<td>$U_{dc}/2$</td>
<td>$-U_{dc}/2$</td>
<td>$U_{dc}/2$</td>
<td>$-2U_{dc}/3$</td>
<td>$U_{dc}/3$</td>
<td>$U_{dc}/3$</td>
</tr>
<tr>
<td>$U_7$</td>
<td>$U_{dc}/2$</td>
<td>$U_{dc}/2$</td>
<td>$U_{dc}/2$</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>
Three-phase SVM with symmetrical placement of zero vectors (SVPWM)

The most popular SVM method is modulation with symmetrical zero states (SVPWM):

\[ t_0 = t_7 = \frac{(T_s - t_1 - t_2)}{2} \] (4.24)

Figure 4.13a shows gate pulses for (SVPWM) and correlation between duty time \( T_{on} \), \( T_{off} \) and duration of vectors \( t_1 \), \( t_2 \), \( t_0 \), \( t_7 \). For the first sector commutation delay can be computed as:

\[ T_{on} = t_0 / 2 \]
\[ T_{off} = t_0 / 2 + t_1 \]
\[ T_{con} = t_0 / 2 + t_1 + t_2 \]

For conventional SVPWM times \( t_1 \), \( t_2 \), \( t_0 \) are computed for one sector only. Commutation delay for other sectors can be calculated with the help of matrix:

\[
\begin{bmatrix}
T_{aon} \\
T_{bof} \\
T_{cfof}
\end{bmatrix}
= \begin{bmatrix}
0.5T_0 & t_1 & t_2
\end{bmatrix}
\]

Two-phase SVM

This type of modulation proposed in [98] was developed in [72,74,88] and is called discontinuous pulse width modulation (DPWM) for CB technique with an additional Zero Sequence Signal (ZSS) in [80]. The idea bases on assumption that only two phases are switched (one phase is clamped by 60° to lower or upper DC bus). It gives only one zero state per sampling time (Fig. 4.13b). Two-phase SVM provides 33% reduction of effective switching frequency. However, switching losses also strongly depend on a load power factor angle (see Chapter 4.4.6). It is very important criterion, which allows farther reduction of switching losses up to 50% [80].

Fig. 4.14a shows several different kind of two-phase SVM. It can be seen that sectors are adequately moved on \( 0^0 \), \( 30^0 \), \( 60^0 \), \( 90^0 \), and denoted as \( PWM(0) \), \( PWM(1) \), \( PWM(2) \), \( PWM(3) \) respectively (\( t_0 = 0 \) means that one phase is clamped to one, while \( t_7 = 0 \) means that phase is clamped to zero). Fig. 4.14b presents phase voltage \( U_{aN} \), pole voltage \( U_{a0} \).
...and voltage between neutral points $U_{N0}$ for these modulations. Zero states description for $PWM(1)$ can be written as:

$$t_0 = 0 \Rightarrow t_7 = T_s - t_1 - t_2 \text{ when } 0 \leq \alpha < \pi/6$$

$$t_7 = 0 \Rightarrow t_0 = T_s - t_1 - t_2 \text{ when } \pi/6 \leq \alpha < \pi/3$$

(4.27)

Fig. 4.14 a) Placement of zero vectors in two-phase SVM. Succession: $PWM(0) = 0^\circ$, $PWM(1) = 30^\circ$, $PWM(2) = 60^\circ$ and $PWM(3) = 90^\circ$ b) Phase voltage $U_{aN}$, pole voltage $U_{40}$ and voltage between neutral points $U_{N0}$ for each of modulation

**Variants of Space Vector Modulation**

From equations (4.21)-(4.23) and knowledge of $U_{N0}$ (Fig. 4.14b), it is possible to calculate duration of zero vectors $t_0$, $t_7$. An evaluation and properties of different modulation method shows Table 4.3.

<table>
<thead>
<tr>
<th>Vector modulation methods</th>
<th>Calculation of $t_0$ and $t_7$</th>
<th>Remarques</th>
</tr>
</thead>
</table>
| Vector modulation with $U_{N0} = 0$ | $t_0 = \frac{T_s}{2}(1 - \frac{4}{\pi} M \cos \alpha)$  
$t_7 = \frac{T_s}{2} - t_1 - t_2$ | • Equivalent of classical CB-SPWM (no difference between $U_{aN}$ and $U_{40}$ voltages)  
• Linear region $M_{max} = 0.785$ |
| Vector modulation with 3-th harmonic | $t_0 = \frac{T_s}{2}(1 - \frac{4}{\pi} M (\cos \alpha - \frac{1}{6} \cos 3\alpha))$  
$t_7 = \frac{T_s}{2} - t_1 - t_2$ | • Low current distortions  
• More complicated calculation of zero vectors  
• Extended linear region: $M = 0.907$ |
| Three-phase SVM with symmetrical zero states (SVPWM) | $t_0 = t_7 = (T_s - t_1 - t_2)/2$ | • Most often used in microprocessor technique for the sake of simple zero vector calculation (symmetrical in sampling time $2T_s$)  
• Current harmonic content almost identical like in previous method |
| Two-phase SVM | $t_0 = 0 \Rightarrow t_7 = T_s - t_1 - t_2$  
when $0 \leq \alpha < \pi/6$  
$t_7 = 0 \Rightarrow t_0 = T_s - t_1 - t_2$  
when $\pi/6 \leq \alpha < \pi/3$  
(for $PWM(1)$) | • Equivalent of DPWM methods in CB-PWM technique  
• 33% switching frequency and switching losses reduction  
• Higher current harmonic content at low modulation index  
• Only one zero state per sampling time, simple calculation (Fig. 4.14) |
The space vector modulation techniques with one zero state in sampling time may be additionally changed for the sake of different harmonic content what is presented in Tab.4.4 and Fig.4.15 [73].

Tab. 4.4 Different zero vector placement in $PWM(0)$

<table>
<thead>
<tr>
<th>sector</th>
<th>$PWM(0)$</th>
<th>Different $PWM(0)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$U_{0}-U_{1}-U_{2}-U_{1}-U_{0}$</td>
<td>$U_{1}-U_{2}-U_{0}-U_{2}-U_{1}$</td>
</tr>
<tr>
<td>2</td>
<td>$U_{1}-U_{2}-U_{1}-U_{2}-U_{1}$</td>
<td>$U_{2}-U_{1}-U_{2}-U_{1}-U_{2}$</td>
</tr>
<tr>
<td>3</td>
<td>$U_{2}-U_{1}-U_{2}-U_{1}-U_{0}$</td>
<td>$U_{1}-U_{2}-U_{1}-U_{2}-U_{0}$</td>
</tr>
<tr>
<td>4</td>
<td>$U_{1}-U_{2}-U_{1}-U_{2}-U_{1}$</td>
<td>$U_{2}-U_{1}-U_{2}-U_{1}-U_{2}$</td>
</tr>
<tr>
<td>5</td>
<td>$U_{0}-U_{1}-U_{2}-U_{1}-U_{0}$</td>
<td>$U_{1}-U_{2}-U_{1}-U_{2}-U_{0}$</td>
</tr>
<tr>
<td>6</td>
<td>$U_{1}-U_{2}-U_{1}-U_{2}-U_{1}$</td>
<td>$U_{1}-U_{2}-U_{1}-U_{2}-U_{0}$</td>
</tr>
</tbody>
</table>

a)

\[ T_s \]

\[
\begin{array}{cccccc}
\nu_s & \nu_s & \nu_s & \nu_s & \nu_s & \nu_s \\
0 & 1 & 1 & 1 & 1 & 0 \\
0 & 0 & 1 & 1 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 \\
\end{array}
\]

b)

\[ T_s \]

\[
\begin{array}{cccccc}
\nu_s & \nu_s & \nu_s & \nu_s & \nu_s & \nu_s \\
1 & 1 & 0 & 0 & 1 & 1 \\
1 & 0 & 0 & 0 & 0 & 1 \\
0 & 0 & 0 & 0 & 0 & 0 \\
\end{array}
\]

Fig. 4.15 Different $PWM(0)$ methods presented above
a) vectors placement b) voltage harmonic content.

4.4.4 Carrier Based PWM Versus Space Vector PWM

Comparison of $CB-PWM$ methods with additional ZSS to $SVM$ is shown on Fig. 4.16. Upper part shows pulse generation through comparison of reference signal $U_a$, $U_b$, $U_c$ with triangular carrier signal. Lower part of figure shows gate pulses generation in SVM (obtained by calculation of duration time of active vectors $U_1$, $U_2$ and zero vectors $U_0$, $U_7$). It is visible that both methods generate identical gate pulses. Also it can be observed from Fig. 4.14 and Fig. 4.16 that the degree of freedom represented in selection of ZSS waveform in $CB-PWM$, corresponds to different placement of zero vectors $U_0(000)$ and $U_7(111)$ in sampling time $T_s = 1/2f_e$ of the $SVM$. Therefore, there is no exist difference between $CB-PWM$ and $SVM$ ($CB-DPWM1 = PWM(1)-SVM$). The difference is only in the treatment of the three-phase quantities: CB-PWM operates in terms of three natural components, whereas $SVM$ uses artificial (mathematically transformed) vector representation.
4.4.5 Overmodulation

Modulation is a basic techniques in power electronics, therefore for full description of this topic is necessary to presents also overmodulation. This part of modulation is not so important for PWM rectifier in the sake of higher harmonic contents in current but it is possible to find some application with similar mode [119].

Many approaches have been reported in the literature to increase the range of the PWM voltage source inverter [75,79,85,89]. Some of them are proposed as extensions of the Sinusoidal PWM (SPWM), and others as extensions of the Space Vector PWM (SVPWM). In CB-PWM by increasing the reference voltage beyond the amplitude of the triangular carrier signal, some switching cycles are skipped and the voltage of each phase remains clamped to one of the dc bus. This range shows a high non-linearity between reference and output voltage amplitude and requires infinite amplitude of reference in order to reach a six-step output voltage.

In SVM allowable length of reference vector $U^*$ which provide linear modulation is equal $U^*_{max} = U_{dc} / \sqrt{3}$ (circle inscribed in hexagon $M = 0.906$) (Fig. 4.17). To obtain higher values of output voltage (reach six-step mode) up to maximal modulation index $M = 1$, an additional non-linear overmodulation algorithm has to be apply. This is because minimal pulse width becomes shorter than critical (mainly dependent on power switches characteristic – usually few µs) or even negative. Zero vectors are never used in this type of modulation.

![Fig. 4.16. Comparison of CB-PWM with SVM a) SVPWM b) DPWM](attachment:fig4_16.png)

From the top: CB-PWM with pulses, short segment of reference signal at high carrier frequency (reference signals are straight lines), formation of pulses in SVM.
Algorithm Based on Two Modes of Operation
Two overmodulation regions are considered (Fig. 4.18). In region I the magnitude of reference voltage is modified in order to keep space vector within the hexagon. It defines the maximum amplitude that can be reached for each angle. This mode extends the range of the modulation index up to 0.95. Mode II starts from $M = 0.95$ and reaches six step mode $M = 1$. Mode II defines both the magnitude and the angle of the reference voltage. To implement both modes a lookup table or neural network [96] based approach can be applied.

Overmodulation mode I: distorted continuous reference signal
In this range, the magnitude of the reference vector is changed while the angle is transmitted without any changes ($\alpha_r = \alpha$). However, when the original reference trajectory passes outside the hexagon, the time average equation gives an unrealistic on duration for the zero vectors. Therefore, to compensate reduced fundamental voltage, i.e. to track with the reference voltage $U^*$, a modified reference voltage trajectory $\tilde{U}$ is selected (Fig. 4.19a). The reduced fundamental components in region where reference trajectory surpass hexagon is compensated by a higher value in corner (equal areas in one sector - see Fig. 4.19a) [85].
Voltage and Virtual Flux Oriented Control

Overmodulation: (a) mode I (0.907 < \( M \) < 0.952), (b) mode II (0.952 < \( M \) < 1)

\( U^* \) - reference trajectory (dashed line), \( U \) – modified reference trajectory (solid line)

The on time durations for region where modified reference trajectory is moved along hexagon are calculated as:

\[
t_1 = T_s \frac{\sqrt{3} \cos \alpha - \sin \alpha}{\sqrt{3} \cos \alpha + \sin \alpha} \quad (4.28 \text{ a})
\]

\[
t_2 = T_s - t_1 \quad (4.28 \text{ b})
\]

\[
t_0 = 0 \quad (4.28 \text{ c})
\]

Overmodulation mode II: distorted discontinuous reference signal.

The operation in this region is illustrated in Fig. 4.19b. The trajectory changes gradually from a continuous hexagon to the six-step operation. To achieve control in overmodulation mode II, both the reference magnitude and reference angle (from \( \alpha \) to \( \alpha_p \)) are changed:

\[
\alpha_p = \begin{cases} 
0 & 0 \leq \alpha \leq \alpha_h \\
\frac{\alpha - \alpha_h}{\pi/6 - \alpha_h} \cdot \frac{\pi}{6} & \alpha_h \leq \alpha \leq \pi/3 - \alpha_h \\
\frac{\pi/3 - \alpha_h}{\pi/3} - \alpha & \pi/3 - \alpha_h \leq \alpha \leq \pi/3
\end{cases} \quad (4.29)
\]

The modified vector is held at a vertex of the hexagon for holding angle \( \alpha_h \) over particular time and then partly tracking the hexagon sides in every sector for the rest of the switching period. The holding angle \( \alpha_h \) controls the time interval when active switching state remains at the vertices, which uniquely controls the fundamental voltage. It is a nonlinear function of the modulation index, which can be piecewise linearized as [89]:

\[
\begin{align*}
\alpha_h &= 6.4 \cdot M - 6.09 & (0.95 \leq M \leq 0.98) \\
\alpha_h &= 11.75 \cdot M - 11.34 & (0.98 \leq M \leq 0.9975) \\
\alpha_h &= 48.96 \cdot M - 48.43 & (0.9975 \leq M \leq 1.0)
\end{align*}
\]

The six-step mode is characterized by selection of the switching vector, which is closest to the reference vector for one-sixth of the fundamental period. In this way the modulator generates the maximum possible converter voltage. For a given switching
frequency, the current distortion increases with the modulation index. The distortion factor strongly increases when the reference waveform becomes discontinuous in the mode II.

**Algorithm Based on Single Mode of Operation**

In a simple technique proposed in [75], the desired voltage angle is held constant when the reference voltage vector is located outside of hexagon. The value, at which the command angle is held, is determined by the intersection of the circle (respond with modulation index) with the hexagon (Fig. 4.20). The angle at which the command is held (hold angles) depends on the desired modulation index \( M \) and can be found from Eq. (4.31) (max circular trajectory is related to the maximum possible fundamental output voltage \( 2/\pi U_{dc} \) not to \( 2/3 U_{dc} \) – see Fig. 4.17):

\[
\alpha_1 = \arcsin \left( \frac{\sqrt{3}}{2M'} \right) \quad (4.31a)
\]

\[
M' = \left( \frac{2\sqrt{3}-3}{2\sqrt{3}-\pi} \right) M + \left( \frac{3-\pi}{2\sqrt{3}-\pi} \right) \quad (4.31b)
\]

\[
\alpha_2 = \frac{\pi}{3} - \alpha_1 \quad (4.31c)
\]

Fig. 4.20 Overmodulation: single mode of operation

\( U^* \) - reference trajectory (dashed line), \( U \) – modified reference trajectory (solid line)

For a desired angle between \( \theta \) and \( \alpha_1 \), the commanded angle tracks its value. When the desired angle increases over \( \alpha_1 \), the commanded angle stays at \( \alpha_1 \) until the desired angle becomes \( \pi/6 \). After that, the commanded angle jumps to value of \( \alpha_2 = \pi/3 - \alpha_1 \). The commanded value of \( \alpha \) is kept constant at \( \alpha_2 \) for any desired angle between \( \pi/6 \) and \( \alpha_2 \). For a desired angle between \( \alpha_2 \) and \( \pi/3 \), the commanded angle tracks the value of desired angle, as in Fig. 4.20. The advantage of linearity and easy implementation is obtained on the cost of higher harmonic distortion.
4.4.6 Performance criteria

Several performance criteria are considered for selection of suitable modulation method [3]. Some of them are defined in the Table 4.1. Below further important criteria as: range of linear operation, current distortion factor and switching losses are discussed.

Range of linear operation

The linear range of the control characteristic for sinusoidal \( CB-PWM \) ends at \( M = \pi/4 = 0.785 \) \((m = 1)\) of modulation index (Fig. 4.21) i.e. to equal of reference and carrier peak. The \( SVM \) or \( CB-PWM \) with ZSS injection provide extension of linear range up to \( M_{\text{max}} = \pi / 2\sqrt{3} = 0.907 \) \((m_{\text{max}} = 1.15)\). The region above \( M = 0.907 \) is the non-linear overmodulation range.

![Fig. 4.21 Control characteristic of PWM converter](image)

Switching losses

Power losses of the \( PWM \) converter can be generally divided into: conduction and switching losses (see in [87]). Conduction losses are practically the same for different \( PWM \) techniques and they are lower than switching losses. For the switching losses calculation, the linear dependency of a switching energy loss on the switched current is assumed. This also was proved by the measurement results [87]. Therefore, for high switching frequency, the total average value of the transistor switching power losses can be for the continuous \( PWM \) expressed as:

\[
P_{s(c)} = \frac{1}{2\pi} \int_{-\frac{\pi}{2}+\varphi}^{\pi/2+\varphi} k_{TD} i \cdot f_s d\alpha = \frac{k_{TD} I_s}{\pi} f_s \tag{4.32}
\]

where: \( k_{TD} = k_T + k_D \) - proportional relation of the switching energy loss per pulse period to the switched current for the transistor and the diode.
In the case of discontinuous PWM the following properties hold from the symmetry of the pole voltage:

\[
P_{sl}(-\phi) = P_{sl}(\phi) \\
P_{sl}(\phi) = P_{sl}(\pi - \phi) \quad \text{where} \quad 0 < \phi < \pi. \quad (4.33)
\]

Therefore, it is sufficient to consider the range of from 0 to \(\pi/2\) for the DPWM as follows [87]:

\[
P_{PWM}(1) \Rightarrow P_{sl}(\phi) = \begin{cases} 
  P_{sl(c)} \cdot \left(1 - \frac{1}{2} \cos \phi\right) & \text{for } 0 < \phi < \pi/3 \\
  \frac{\sqrt{3}}{2} \sin \phi & \text{for } \pi/3 < \phi < \pi/2
\end{cases} \quad (4.34)
\]

\[
P_{PWM}(0) \Rightarrow P_{sl}(\phi) = P_{sl(PWM(1))} \cdot (\phi - \frac{\pi}{6}) \quad (4.35)
\]

\[
P_{PWM}(2) \Rightarrow P_{sl}(\phi) = P_{sl(PWM(1))} \cdot (\phi + \frac{\pi}{6}) \quad (4.36)
\]

\[
P_{PWM}(3) \Rightarrow P_{sl}(\phi) = \begin{cases} 
  P_{sl(c)} \cdot \left(1 - \frac{\sqrt{3} - 1}{2} \cos \phi\right) & \text{for } 0 < \phi < \pi/6 \\
  \frac{\sin \phi + \cos \phi}{2} & \text{for } \pi/6 < \phi < \pi/3 \\
  P_{sl(c)} \cdot \left(1 - \frac{\sqrt{3} - 1}{2} \sin \phi\right) & \text{for } \pi/3 < \phi < \pi/2
\end{cases}
\]

Switching losses depends on type of discontinuous modulation and power factor angle what is shown in Fig. 4.22 (comparison to continuous modulation). Since the switching losses increase with the magnitude of the phase current (approximately linearly), selecting a suitable modulation can significantly improve performance of the converter. Switching losses are average reduced about 33%. In favour conditions, when modulation is clamped in phase conducting max. current, switching losses decrease up to 50%.

![Fig. 4.22. Switching losses (P_{sl}(\phi)/P_{sl(c)}) versus power factor angle](image-url)
Distortion and Harmonic Copper Loss Factor

The current waveform quality of the PWM converter is determined by harmonics of switching frequency what have influence for copper losses and the instantaneous power ripple. Harmonics are changed according to the selected switching sequence. Detailed description is presented in [84, 98]. The rms harmonic current defined as:

\[ I_{h(rms)} = \frac{1}{T} \int_0^T [i(t) - i_L(t)]^2 dt, \]  

depends on type of PWM and AC side impedance. To eliminate influence of AC side impedance parameters, the distortion factor is commonly used (see Table 4.1):

\[ d = \frac{I_{h(rms)}}{I_{h(six-step)(rms)}} \]  

(4.39)

For six-step operation the distortion factor is \( d = 1 \). It should be noted that harmonic copper losses in the AC-side are proportional to \( d^2 \). Therefore, \( d^2 \) can be considered as a loss factor. Values of loss factor can be compute for different modulation methods [3,87]. It depends on switching frequency, modulation index \( M \), and shape of the ZSS (Fig. 4.23):

- for continuous modulation:

\[
\begin{align*}
SPWM & \quad d = \frac{4M}{\sqrt{6\pi k_{fas}}} \sqrt{1 - \frac{32M}{\sqrt{3} \pi^2} + \frac{3M^2}{\pi}} \quad M \in \left[0, \frac{\pi}{4}\right] \\
SVPWM & \quad d = \frac{4M}{\sqrt{6\pi k_{fas}}} \sqrt{1 - \frac{32M}{\sqrt{3} \pi^2} \frac{9M^2}{2\pi} \left(1 + \frac{3\sqrt{3}}{4\pi}\right)} \quad M \in \left[0, \frac{\pi}{2\sqrt{3}}\right]
\end{align*}
\]  

(4.40)

(4.41)

- for discontinuous modulation (DPWM):

\[
\begin{align*}
DPWM1 & \quad d = \frac{4M}{\sqrt{6\pi k_{fas}}} \sqrt{4 - \frac{4M}{\sqrt{3} \pi^2} \left(8 + 15\sqrt{3}\right) + \frac{9M^2}{2\pi} \left(2 + \frac{\sqrt{3}}{2\pi}\right)} \quad M \in \left[0, \frac{\pi}{2\sqrt{3}}\right] \\
DPWM0(2) & \quad d = \frac{4M}{\sqrt{6\pi k_{fas}}} \sqrt{4 - \frac{140M}{\sqrt{3} \pi^2} + \frac{9M^2}{2\pi} \left(2 + \frac{3\sqrt{3}}{4\pi}\right)} \quad M \in \left[0, \frac{\pi}{2\sqrt{3}}\right] \\
DPWM3 & \quad d = \frac{4M}{\sqrt{6\pi k_{fas}}} \sqrt{4 - \frac{4M}{\sqrt{3} \pi^2} \left(62 - 15\sqrt{3}\right) + \frac{9M^2}{2\pi} \left(2 + \frac{\sqrt{3}}{\pi}\right)} \quad M \in \left[0, \frac{\pi}{2\sqrt{3}}\right] 
\end{align*}
\]  

(4.42)

(4.43)

(4.44)

where \( k_{fas} \) is defined as a ratio of carrier frequency (sampling time) to base of carrier frequency. All continuous PWM have the advantage over DPWM methods for the sake of small distortion factor in the low range of modulation. When the modulation index increases and the PWM performance rapidly decreases, the SVPWM maintain at lowest distortion factor. The harmonic content for SVPWM and DPWM at the same carrier frequency is similar at high modulation index only (Fig. 4.23). However, we should remember that DPWM possess lower switching losses. Therefore, the carrier frequency can be increased by factor 3/2 for 33% reduction of switching losses, or 2 times increased for 50% reduction of switching losses. It provides to lower current distortion for DPWM in comparison to SVPWM.
4.4.7 Adaptive Space Vector Modulation (ASVM)

The concept of adaptive space vector modulation (ASVM) proposed by Author [93, Patent No. P340113] provides:

- full control range including overmodulation and six-step operation,
- theoretically, up to 50% reduction of switching losses at 33% reduction of average switching frequency,
- high dynamics.

The above features are achieved by use of four different modes of SVM with an instantaneous tracking of the AC current peak and an optimal switching table for fast response to step changes of the load. Four PWM operation modes are distributed in the range of modulation index ($M$) as follows (Fig. 4.24a):

- **A**: $0 < M < 0.5$ – conventional SVM with symmetrical zero switching states,
- **B**: $0.5 < M < 0.908$ – discontinuous SVM with one zero state per sampling time (two-phase or flap top PWM),
- **C**: $0.908 < M < 0.95$ – overmodulation mode I, (see Section 4.4.6)
- **D**: $0.95 < M < 1$ – overmodulation mode II.

The combination of regions **A** with **B** without current tracing, suggested in [72,80] is known as hybrid PWM. In the region **B** of discontinuous PWM, for maximal reduction of switching losses, the peak of the current should be located in the centre of “flat” parts. Therefore, it is necessary to observe the peak current position. Components $i_{Lα}, i_{Lβ}$ of the measured current are transformed into polar coordinates and compared with voltage reference angle (Eq. (4.45)). It gives possibility to identify power factor angle $ϕ$, which decide about placement of clamped region. Thus, the ring from Fig. 24b will be adequately moved ($ϕ$). For each of sector:
if \( \alpha < \varphi + \kappa \) \( \Rightarrow t_0 = 0 \) 
if \( \alpha > \varphi + \kappa \) \( \Rightarrow t_7 = 0 \) 

(4.45)

where: \( \alpha \) - reference voltage angle, \( \varphi \) – power factor angle, \( \kappa \) - for successive sectors \( \pi/6, \pi/2, 5\pi/6, 7\pi/6, 3\pi/2, 11\pi/6 \)

This provides tracking of the power factor angle in full range of \( \varphi \) (from \(-\pi\) to \(\pi\)), what guarantees maximal reduction of switching losses (Fig. 4.25)

![Diagram](image)

Fig. 4.24. Adaptive modulator
a) effect of modulation index b) effect of power factor angle

![Graph](image)

Fig. 4.25. Switching losses versus power factor angle for conventional SVPWM and ASVM

The dynamic state is identified after step change of load what results that switching table is used. After returning to steady state the ASVM operates like a conventional SVM. The full algorithm of adaptive modulator is presented in Fig. 4.26. Fig. 4.27 shows an example of implementation in a current regulator. Adaptive modulation with simplified switching time calculation is described in A.3.