# Power Systems

Ryszard Strzelecki • Grzegorz Benysek
Editors

# Power Electronics in Smart Electrical Energy Networks

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# Preface

The book arises from the conviction that it is necessary to re-think the basic philosophy governing the electricity distribution systems. In the authors' opinion there is a need to exploit fully the potential advantages of renewable energy sources, distributed generation, energy storage and other factors  which should not only be connected but also fully integrated into the system to increase the efficiency, flexibility, safety, reliability and quality of the electricity and networks. Transformation of the current electricity grids into a smart (resilient, interactive *etc.*) network necessitates the development, propagation and demonstration of key cost effective technologies enabling (*e.g.*, innovative interconnection solutions, storage technologies for renewable energy sources, power electronics, communications *etc.*). On the basis of the above, the major aim of this book is to present the features, solutions and applications of the power electronics arrangements likely to be useful in future smart electrical energy networks.

The first part of this book introduces the structure and fundamental problems of the current electricity grids together with the concept of smart electrical energy networks.

Next there is a critical overview of power theories, mainly under non-sinusoidal conditions in single-phase and three-phase systems, in both time and frequency domains. The basic criterion for the choice of the discussed theories is historical development of knowledge in this field and the usefulness of power theory in solving practical problems: reactive power compensation, balancing the supply network load and mitigation of voltage and current distortion. Particular attention is given to the theories defining the current components in the time domain as the basis for present-day interconnection, active compensation and filtering systems. The content of this part is essential for understanding both the principle of operation and the control algorithms of the majority of the currently used power quality improvement and interconnecting systems.

Additionally, in this part an overview of control methods in power systems with the focus on damping of electromechanical oscillations and mitigation of power quality problems is presented. The focus is on power systems with increased levels of uncertainty resulting from deregulation of theelectrical power industry and the presence of non-conventional types of generation (renewable energy sources and distributed generation). The issue of finding the best techno-economical solution for the problems is also briefly mentioned. The focus in thepower quality section is on probabilistic modelling of disturbances and their consequences.

In the next part of the book the main emphasis is on low, medium, and high power conversion issues and the power electronic converters that process power for a variety of applications in smart grids. Following recent trends in power electronics technology, greater stress is placed on modern power electronic converters, such as resonant and multi-level inverters or matrix converters, and these are thoroughly covered. Special features include in-depth discussions of all power conversion types: AC/DC, AC/AC, DC/DC, and DC/AC.

After that, both the relationships and the differences between electrical power quality and electromagnetic compatibility are explained and definitions of these notions are provided. The principles of standardization in both fields are also be discussed. The power quality survey is a useful procedure for identifying and resolving power-related equipment or facility problems. It is an organized, systematic approach to problem solving. If all the steps for a power quality survey are completed, information is obtained that either identifies a solution to a power-related problem or reveals that the problem is not related to the electrical power system.

After that, EMC related problems in smart electrical power systems as well as some EMC regulations are overviewed. Special attention is paid to the origin and the spreading of the conducted EMI over power systems containing power converters. This is true because the diversity of power converters makes difficult the general analysis of the EMI spectra. However, there are some common features which can be derived from typical applications and layouts of the systems with power converters. Specific key aspects of electromagnetic compatibility in power electronics are presented, such as a typical role of power converters and their place in the smart power system, a typical frequency range of generated EMI noises, specific features of the common mode source in three-phase power converter systems and traveling wave phenomena. This part gives a detailed analysis based on the authors' own experimental results in the systems with converters that are common in smart power systems.

The next part of the book introduces high frequency AC power distribution systems as relatively new and promising developments in the field of electric power. Compared with low frequency or DC link power systems, the high frequency system offers many key advantages including system compactness due to small filtering and transforming components, better power quality, freedom from acoustic noise and mechanical resonance. In addition, it is particularly conducive to the distributed and amalgamated structures of future power systems, which are likely to converge with the information superhighways. Also described are the motivations and performances of the earliest high frequency systems used in telecommunications and NASA's Space Station, and to those more recently introduced in the fields of electric vehicles, micro-grids and renewable energies. Additionally there is discussion of the many potential benefits these systems can offer in shaping the future electric power infrastructure, and also the challenges that need to be overcome.

Next addresed are the technical considerations for interconnecting distributed generation equipment with conventional electric utility systems. This discussion arises from the fact that most electric distribution systems are designed, protected, and operated on the premise of being a single source of electric potential on each distribution feeder at any given time. Distributed generation violates this fundamental assumption, and therefore special requirements for connecting to the utility distribution grid are critical to ensure safe and reliable operation. Manufacturers, vendors, and end-users often see distributeed generation interconnection requirements as a huge market barrier, whereas utility engineers consider them to be absolutely necessary. Thus tools to help assess practical interconnection for specific projects and equipment are provided; we also create a clearinghouse for the many ongoing domestic and international efforts to develop uniform standards for interconnection.

After that, the next part of this book is targeted at known electric energy storage systems as well as development of methodologies and tools for assessing the economic value and the strategic aspects of storage systems integrated into electricity grids. Such tools should be ble to evaluate and analyse energy storage solutions in a variety of applications, such as integration of distributed/renewable energy resources, reduction of peak loading, improvement of transmission grid stability and reliability. Additionally, electricity storage is presented as a strategic enabling technology which not only reduces costs and increases the efficient use of grid assets, but is key for accelerating the integration of distributed generation and renewable sources of energy.

The next part of our book deals with grid integration of wind energy systems. The focus of this topic is on the electrical side of wind conversion systems. After a short description of the basics, such as energy conversion, power limitation and speed control ranges, the existing generator types in wind energy conversion system are described. Because of the practical problems arising with wind turbine installations, their grid integration is an interesting field, whereas the characteristics of wind energy conversion itself, the common types of grid coupling and resulting wind park designs are discussed. On the point of common coupling, wind energy generation may produce distortions of the grid, *e.g.*, flicker effects and harmonics. The causes of their generation, superposition and mitigation are described in detail. Existing standards and the requirements of the transmission system operators are also discussed from the point of view of the conversion system.

Because of limited onshore areas for wind energy systems in Europe, powerful wind parks can be installed only at selected places. A solution of this problem is offshore technology which, due to better wind conditions, brings higher energy yields, but also a lot of additional requirements for the installation and operation of the wind turbines. This includes a special generator design necessitated by the salty environment and different possibilities for the wind park structure, which has internal fixed or free adjustable parameters such as frequency, voltage range and transmission type. The external energy transmission to the onshore substation can be realized with different system configurations. Their advantages and disadvantages are explained.

The next part of the book describes grid integration of photovoltaic systems and fuel cell systems. First the cell types and their efficiency and place requirement are explained. The focus lies on grid-connected photovoltaics, mainly their plant design and grid interfacing of systems depending on isolation conditions, and the possible use of different components is a topic of current interest. Power quality becomes an important issue if higher unit powers are installed. Special problems arising from common connection at the low voltage level are discussed. Derived from the existing devices and their assigned problems in the grid, possibilities for future development are presented.

Fuel cells, photovoltaic systems, generate DC voltage and need a power electronic conversion unit for their grid connection. The different types of fuel cells and their typical applications are described. But the focus lies on plant design, grid interfacing and future development. At the moment only a few fuel cell applications exist. The big potential of this technology may lead to large installation numbers within the next five years. Existing standards of this technology are listed to assist the understanding of this technology.

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# 3

# Overview of Power Electronics Converters and Controls

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## 3.1 Power Electronics Background

Power Electronics (PE) is the technology associated with efficient conversion, control and conditioning of electric power by static means from its available input- into the desired electrical output form. Electric energy conversions carried out by

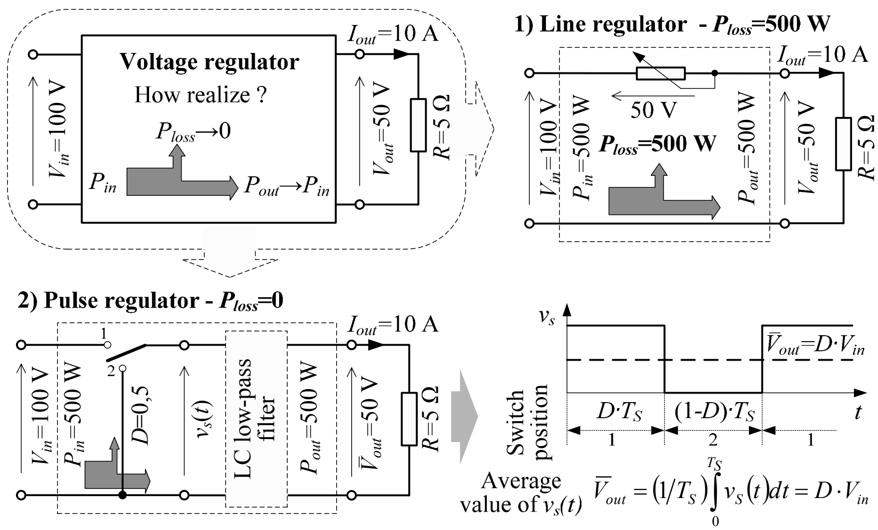



**Figure 3.1.** Efficiency of line and pulse voltage regulators

PE circuits are diverse and apply to power varying from tens of watts up to hundreds of megawatts. This by particular wide range of power is reflected, among others, in the varying overall dimensions of PE arrangements. Some of the arrangements are hand size, while others require spaces especially designed for their utilization.

Our attention should be drawn to the fact that the definition of PE emphasizes high conversion efficiency. That is connected to the fact that the principle of operation of any PE circuit consists in periodical linking of a power source to the electrical energy consumer – load. In such a case, assuming ideal switches (immediate switching, null resistance in the "*on*" state and infinitesimally high resistance in the "*off*" state) as well as the lack of other dissipation elements, power losses equal zero. High efficiency of the PE circuits, in comparison to alternative solutions is demonstrated by the example in Figure 3.1.

Let us assume that we must select only one of the two regulators with input voltage $V_{in}$=100 V and output voltage $V_{out}$=50 V, which are presented in Figure 3.1. Their load is their resistance $R$=5 Ω. In the case of selection of the line regulator, power losses occurring on variable resistor equal $P_{loss}$=500 W. The losses increase along with increasing load (decreasing $R$) as well as increase of the voltage $V_{in}$. Efficiency of the line regulator

$$\eta = P_{out}/P_{in} = \left(P_{in} - P_{loss}\right)/P_{in} = 1 - P_{loss}/P_{in} \tag{3.1}$$

where $P_{out}$ and $P_{in}$ – output and input power, changes same as quotient $V_{out}/V_{in}$. The above provides arguments to the advantage of the pulse regulator – PE converter. In this regulator, the switch takes alternative positions 1 and 2 in the time interval $DT_S$ and $(1\text{-}D)T_S$ pulse repetition period $T_S$, where $D$ – the so-called duty cycle. None of the positions, in the case of ideal switches, causes power losses, $P_{loss}$=0. Thus, efficiency of the pulse regulators, as well as other PE circuits, approaches 100%.

In practice the efficiency PE circuits is somewhat smaller (depending on the type circuit, from 85% up to almost 100%). It relates to the fact that, first, real elements $L$ and $C$ are dissipative. Second, and most of all, none of the actual power

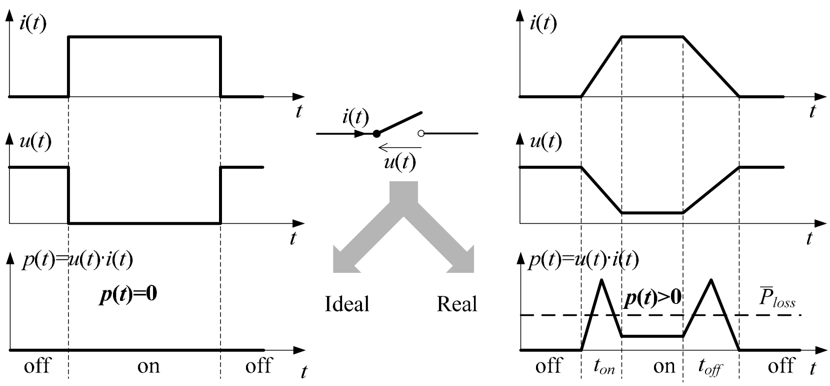



**Figure 3.2.** Power losses in ideal and real switches

switches switch over immediately, and its resistance in the "*on*" state is positive. That causes extra power losses. Figure 3.2 provides illustrative but simplified demonstration of the process in which the power losses arise.

### 3.1.1 Historical Perspective

Historical background of PE backs to the end of the nineteenth century, when in 1896 Karol Pollak, honorary doctor of the Warsaw University of Technology, was issued a German patent (DRP 96564) for an electric aluminum rectifier. Description of the patent also included a scheme of one-phase, full-wave rectifier, known today as Graetz bridge. The description was published in the "*Elektronische Zeitung*" no 25 from 1897, with notation from the editor that at that time professor L. Graetz was working on rectifiers of similar principle of operation. However, the solution of Prof. Graetz was published a year and a half after the patent for Dr. K. Pollak had been issued. Other, successive inventions that are of importance to power electronics and therefore we should be aware of are:

- 1902 – Mercury-arc rectifier (P. Cooper Hewitt);
- 1903 – Phase angle control (PH. Thomas);
- 1904 – Vacuum diode (JA. Fleming);
- 1906 – Triode (L. De Forest);
- 1908 – Iron vessel rectifier (B. Schäfer);
- 1912 – Megamp (E. Alexanderson);
- 1912 – Power rectifier. Sub-synchron cascade one-phase/66 kW, three-phase /300 kW (B. Schäfer);
- 1922 – Cycloconverter (M. Meyer/LA. Hazeltine);
- 1923 – Pooled cathode thyratron (I. Langmuir);
- 1924 – Chopper principle (A. Burnstein);
- 1925 – Parallel inverter commutation (DC. Prince);
- 1925 – Field-effect transistor theoretical development (JE. Lilienfeld);
- 1926 – Hot cathode thyratron (AW. Hull);
- 1928 – Practical grid-controlled mercury-arc rectifier (I. Langmuir, DC. Prince);
- 1929 – Thyratron controlled rectifier (A.W Hull);
- 1931 – Ignitron (J. Slepian);
- 1931 – Cycloconverter for railways (M. Schenkel, I. von Issendorf);
- 1932 – Mercury-arc rectifier for wattles power compensation (M. Schenkel 1932);
- 1932 – First HVDC transmision system (VM. Stör);
- 1934 – Thyratron motor built and tested (E. Alexanderson);
- 1935 – HVDC transmission system 287 kV between Mechanicville and Shenectady, NY, USA;
- 1942 – Frequency changers 20 MW, 25/60 Hz;
- 1947 – Point contact transistor (J. Bardeen, WH. Brattain, WB. Shockley);
- 1951 – Junction transistor (WB. Shockley);
- 1953 – Developed of the germanium power diode 100 A.

The year 1957, which is associated with the development of semiconductor technology, was also adopted as the beginning of modern PE. At that time, Bell Laboratory developed the first p-n-p-n switches – the thyristor or Silicon Controlled Rectifier (SCR). However, the idea of the SCR had been described for the first time by WB. Shockley in 1950. It was referred to as a bipolar transistor with a p-n hook-collector. The operation mechanism of the thyristor was further analyzed in 1952 by JJ. Ebers. In 1956 JL. Moll investigated the switching mechanism of the typical thyristor. Development continued and more was learned about the device such that the first SCR became available in the early 1960s and started gaining a significant level of popularity for power switching. To that day many more innovative and much improved power semiconductor switches were developed [1]. Because of the stimulation of new technical solutions and applications the following arose:

- Triode for Alternating Current (TRIAC) thyristor – developed 1964;
- Bipolar Junction Transistor (BJT) 500 V, 20 A – developed 1970;
- Power Metal-oxide Semiconductor Field-effect Transistor (MOSFET) 100 V, 25 A – developed 1978;
- High power GTO thyristor 2500 V, 1000 A – developed 1981;
- Insulated-gate Bipolar Transistor (IGBT) – developed 1983;
- Intelligent Power Module (IPM) – developed 1990;
- Integrated Gate Commutated Thyristor (IGCT) /emitter turn-off (ETO) thyristor – developed 1997;
- Reverse blocking IGBT (RBIGBT) – developed 2000;
- Matrix converter power module (ECONOMAC) – developed 2001.

However, the present stage of PE development not only results from progress in research on power semiconductors switches [2]. These switches are mostly, and at the same time only, the muscle of PE systems. Also significant are achievements in other related research areas, most of all micro-electronics, control theory and informatics [3]. Without development of these areas we would not be able to equip modern PE arrangements with "*brain and nerves*". All of these areas are interdependent, which is seen in particular on the example of a microprocessor [4]. The application of microprocessors allowed production of practical implementation complex control algorithms, while at the same time stimulating their development. The microprocessor also had significant impact on progress in construction, actual monitoring, diagnostics and remote control of PE systems. Altogether it influenced development of several new technological disciplines [5].

From day to day, changes also occurred in electrical power engineering. The possible place of the PE in the flow of electrical energy from producer to consumer is illustrated in Figure 3.3. Nowadays compensators have become more and more popular on AC transmission and distribution lines as well as feeders, power quality conditioners and power flow controllers [6–11]. It is also unacceptable rationally to apply many renewable energy sources as well as to develop local/distribution generation without PE arrangements [12–15]. The same relates to DC distribution systems and energy storage systems.

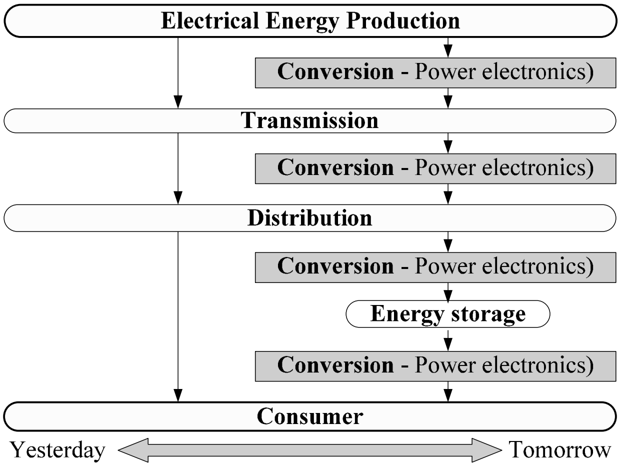



**Figure 3.3.** Power electronics place in the electrical power engineering

As we see, PE as an area of electrical engineering studies continues to develop intensively. Often, such development is described as a "*quiet revolution*". Utilized in all areas of application of electrical energy, modern PE is a research field of interdisciplinary character (Figure 3.4). It is referred to as industrial electronics, and combines multiple diverse technological disciplines [16, 17].

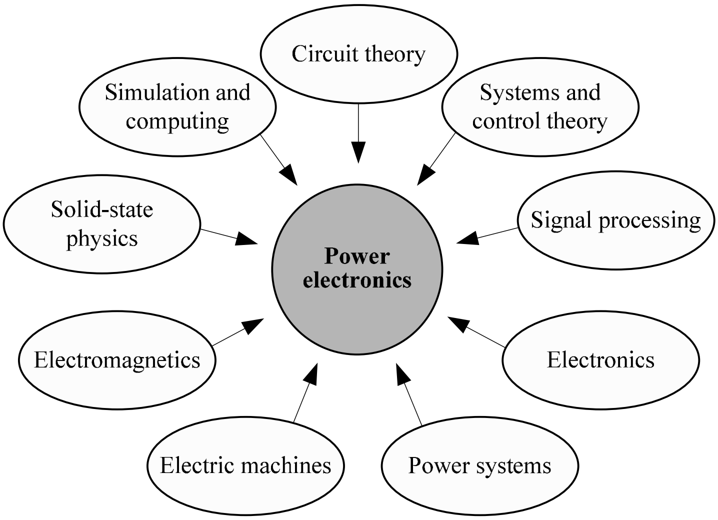



**Figure 3.4.** Interdisciplinary nature of power electronics

### 3.1.2 Generic Power Electronics Arrangements

Conventional PE system usually consists of functional modules, delineated as in Figure 3.5. The PE circuit is the central module, and is constructed with application of semiconductor switches. The second module – internal controller – is responsible for operating the switches according to an assumed operation algorithm and on the basis of physical quantities (most by electric currents and voltages), measured in the PE circuit as well as the output and input PE circuits. Supervisory control of the consumption of electrical energy (*e.g.*, heating), usually assured by external controllers, is nowadays realized together with the internal controllers on the same control board. Some of the applications do not even require additional external controllers.

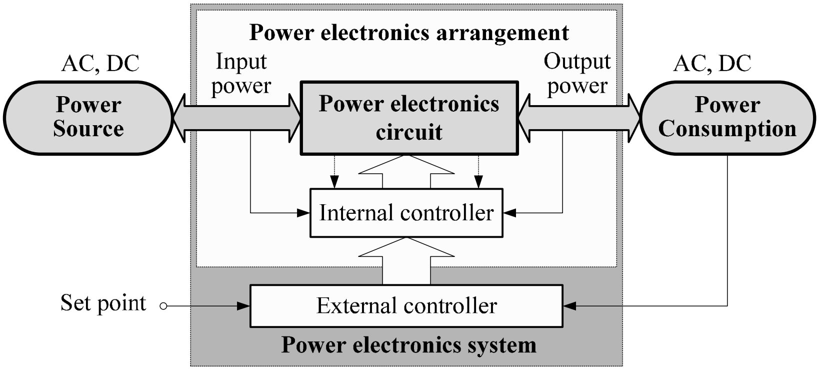



**Figure 3.5.** Block diagram of a power electronics system

Today PE arrangements (Figure 3.6) differ from solutions developed 10–15 years ago by means of realization of particular functions. For example, changes in control layers resulted from development of digital technologies, in particular Digital Signal Processors (DSP), Complex Programmable Logic Devices (CPLD)

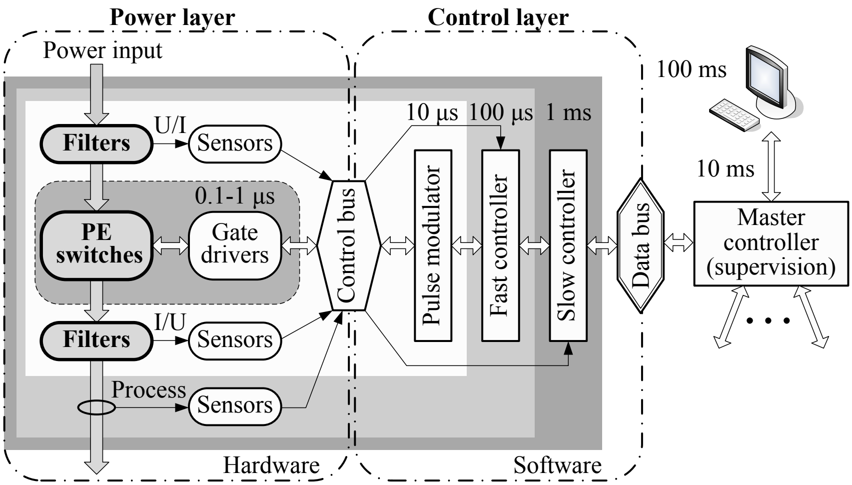



**Figure 3.6.** Block diagram of a modern power electronics arrangements

and Application Specific Integrated Circuit (ASIC). Here software application is dominating. The majority of the functions (*e.g.*, number of safety devices) that used to be realized by discrete components has been moved to the software level. However, discrete components remain at the power layer, which still requires hardware application. Here, we observe integration or packaging of components, for example, in the form of ready modules [18–21]. Examples of such ready blocks are Power Electronics Building Blocks (PEBB), being constructional closed power layers, as well as block EconoMAC [22–24].

In terms of their functionality, power electronics circuits can be divided into two main groups: contactless switches and converters (Figure 3.7). To the first group belong all modern protection and reconfiguration devices, such as static current limiters, static current breakers, and static transfer switches. Application of contactless switches, when compared to contactors, can be characterized by many advantages and especially by very short operating time, high permitted frequency of switch-overs, long lifetime and lack of electric arc while switching them off. This group is less numerous and is not considered further below.

The second, more numerous group, PE circuits, applies to conversion of the form alternating current into direct current and inversely as well as electrical energy parameters (value of voltage/electric current, frequency, number of phases, reactive power *etc.*). The group can be further divided into basic types of PE converters, which is illustrated in Figure 3.7. AC/AC converters can be realized as single-stage (AC regulators, direct frequency converters) or two-stage through DC link. Similarly, DC/DC converters can be realized. However, rectifiers and inverters are characterized by only a single-stage energy conversion.

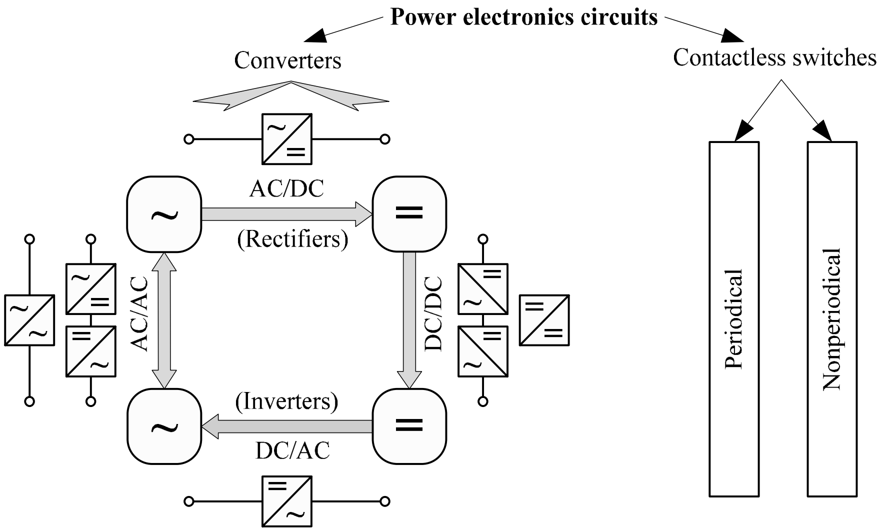



**Figure 3.7.** Basic types of the power electronics circuits and converters

### 3.1.3 Switching and Continuous Models of Converters

PE Converters (PEC), independently of their function, construction details and application, can also be divided into two general classes:

- Direct converters – in which main reactive elements are connected only to input or output terminals of the converter and can be considered as part of the source or the load. Rectifiers and voltage inverters with LC filters are an example of such direct converters;
- Indirect converters – including main reactive elements inside their structure. Because they usually have very few elements, indirect converters are mostly analyzed as connections of direct converters with reactive elements among them.

The division, in most cases is consistent with either single-stage or two-stage realization of the PEC.

Figure 3.8 presents the general model of direct PEC in the form of a switch matrix with "*N*" inputs and "*m*" outputs as well as examples of realization of the switches, determining specific characteristics of PEC. Although inputs as well as outputs are changeable, they must remain of different character, *i.e.,* in case of

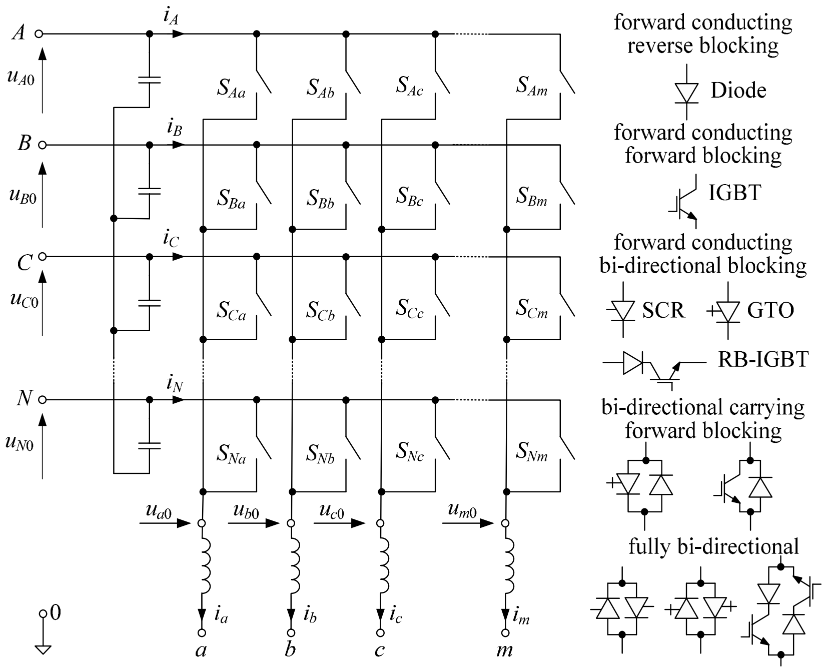



**Figure 3.8.** The general model of the direct PEC and examples of switch realization

voltage input (voltage source or capacitor) the output must be current (current source or reactor), and *vice versa*. The presented model is described by the following equations

$$\begin{bmatrix} u_{a0} \\ u_{b0} \\ u_{c0} \\ \vdots \\ u_{m0} \end{bmatrix} = \underbrace{\begin{bmatrix} S_{Aa} & S_{Ba} & S_{Ca} & \cdots & S_{Na} \\ S_{Ab} & S_{Bb} & S_{Cb} & \cdots & S_{Nb} \\ S_{Ac} & S_{Bc} & S_{Cc} & \cdots & S_{Nc} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ S_{Am} & S_{Bm} & S_{Cm} & \cdots & S_{Nm} \end{bmatrix}}_{\|M\|} \times \begin{bmatrix} u_{A0} \\ u_{B0} \\ u_{C0} \\ \vdots \\ u_{N0} \end{bmatrix} \tag{3.2a}$$

$$\begin{bmatrix} i_A \\ i_B \\ i_C \\ \vdots \\ i_N \end{bmatrix} = \underbrace{\begin{bmatrix} S_{Aa} & S_{Ab} & S_{Ac} & \cdots & S_{Am} \\ S_{Ba} & S_{Bb} & S_{Bc} & \cdots & S_{Bm} \\ S_{Ca} & S_{Cb} & S_{Cc} & \cdots & S_{Cm} \\ \vdots & \vdots & \vdots & \vdots & \vdots \\ S_{Na} & S_{Nb} & S_{Nc} & \cdots & S_{Nm} \end{bmatrix}}_{\|M\|^T} \times \begin{bmatrix} i_a \\ i_b \\ i_c \\ \vdots \\ i_m \end{bmatrix} \tag{3.2b}$$

where $|M|$ – connection matrix; $|M|^T$ – transpose of a matrix $|M|$; $S_{ij}$ – state of the switch $S_{ij}$, where if the switch is "*on*" then $S_{ij}=1$, and if the switch is "*off*" then $S_{ij}=0$, $i=A,B,\ldots, N$ and $j=a,b,\ldots m$.

Switch states $S_{ij}$ can only take values 0 or 1, depending on time. In such a case the natural method to shape output voltage [$u_{a0}$, $u_{b0}$, $u_{c0}$,…, $u_{m0}$] and input currents [$i_A$, $i_B$, $i_C$,…, $i_N$] in direct converters is pulse modulation. The applied modulation algorithm includes practical limitations [25, 26]. In particular, states of all switches, at any given moment, cannot result in short-circuit or overvoltage. For example, in the presented direct converters model (Figure 3.7) with input voltage and output current, the states $S_{ij}$ of all switches must meet the following requirements

$$\sum_{i=A}^{N} S_{ia} = \sum_{i=A}^{N} S_{ib} = \sum_{i=A}^{N} S_{ic} = \cdots = \sum_{i=A}^{N} S_{im} = 1, \qquad \sum_{i=A}^{N}\sum_{j=a}^{m} S_{ji} = m \tag{3.3}$$

The first condition (Equation 3.3) should be understood as the condition where one switch can be connected to one output only – otherwise, input short-circuit occurs. Meeting the second condition (Equation 3.3) ensures the direction of the output current flow – the number of additional switches must always be equal to the number of outputs. On this basis, the direct converters analysis is made together with synthesis of their algorithms. The analysis of general characteristics of direct converters can also be carried out in a simplified way. It is assumed that the relative time to connect a switch $S_{ij}$ is equal to the instantaneous value of the modulating continuous function $d_{ji}(t)$, such thet $0\le d_{ji}(t)\le 1$ and satisfying Equation 3.3, *i.e.*,

$$\sum_{i=A}^{N} d_{ia}(t) = \sum_{i=A}^{N} d_{ib}(t) = \sum_{i=A}^{N} d_{ic}(t) = \cdots = \sum_{i=A}^{N} d_{im}(t) = 1, \qquad \sum_{i=A}^{N}\sum_{j=a}^{m} d_{ji}(t) = m \tag{3.4}$$

In this case, instead of the switch matrix (Figure 3.7) we obtain a continuous model of direct converters, where each switch $S_{ij}$ is exchanged by an ideal transformer with a transformation ratio equaling to the modulating function $d_{ji}(t)$. An example of such a model for 3×3 direct converters with fully bi-directional turn-off switches is presented in Figure 3.9. The same figure shows the fundamental method to determine the state of the matrix switches on the basis of modulating functions. The method is discussed with the example of the simplest Pulse-width Modulation (PWM) and in relation to the output "*a*" of a direct converter.

Additionally, Figure 3.10 presents the basic scheme of a three-phase voltage source inverter supplied by the voltage $U_{DC}$ and the corresponding continuous model. If taking into account this model, the general characteristics of three-phase VSI can be determined from the equations

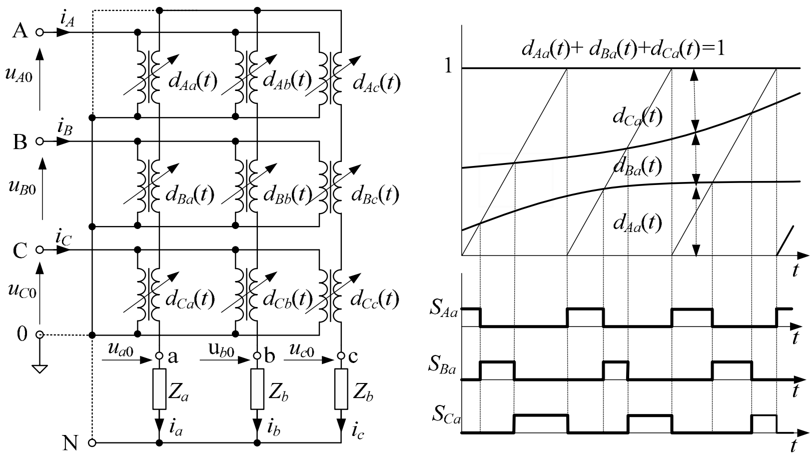
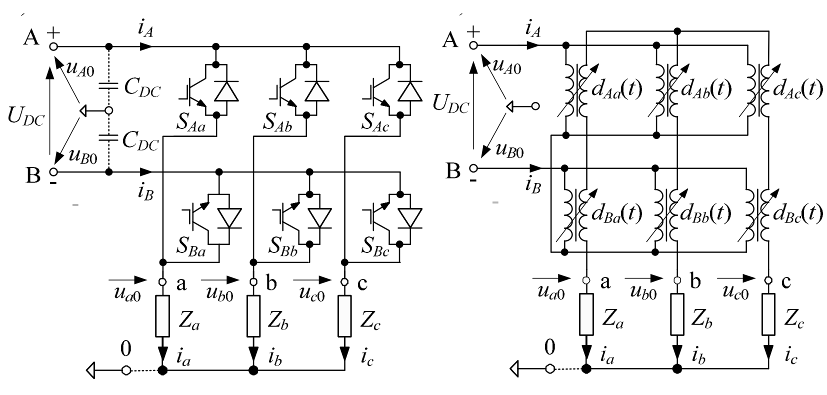



**Figure 3.9.** The continuous model of the 3×3 direct converters




a b

**Figure 3.10. a** Basic scheme of the three-phase VSI. **b** Its continuous model

$$\begin{bmatrix} u_{a0} \\ u_{b0} \\ u_{c0} \end{bmatrix} = \begin{bmatrix} d_{Aa}(t) & d_{Ba}(t) \\ d_{Ab}(t) & d_{Bb}(t) \\ d_{Ac}(t) & d_{Bc}(t) \end{bmatrix} \times \begin{bmatrix} u_{A0} \\ u_{B0} \end{bmatrix} \tag{3.5}$$

$$\begin{bmatrix} i_A \\ i_B \end{bmatrix} = \begin{bmatrix} d_{Aa}(t) & d_{Ab}(t) & d_{Ac}(t) \\ d_{Ba}(t) & d_{Bb}(t) & d_{Bc}(t) \end{bmatrix} \times \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \tag{3.6}$$

where $u_{A0}=-u_{B0}=U_{DC}/2$. Only sinusoidal modulating functions are further considered

$$\begin{bmatrix} d_{Aa}(t) \\ d_{Ab}(t) \\ d_{Ac}(t) \end{bmatrix} = \frac{1}{2} \begin{bmatrix} 1 + A\sin(\omega t + 0) \\ 1 + A\sin(\omega t - 2\pi/3) \\ 1 + A\sin(\omega t - 4\pi/3) \end{bmatrix} \tag{3.7}$$

$$\begin{bmatrix} d_{Ba}(t) \\ d_{Bb}(t) \\ d_{Bc}(t) \end{bmatrix} = \frac{1}{2} \begin{bmatrix} 1 - A\sin(\omega t + 0) \\ 1 - A\sin(\omega t - 2\pi/3) \\ 1 - A\sin(\omega t - 4\pi/3) \end{bmatrix} \tag{3.8}$$

where $\omega=2\pi f$; $f$ – output fundamental frequency; $A$ – modulation factor ($0\leq A\leq 1$).

From Equations 3.5–3.8 it results that, for such modulating functions, the amplitude of the sinusoidal output voltage cannot exceed $U_{DC}/2$. It should be emphasized that the value is not any boundary value. In the case of vector modulation the amplitude of the sinusoidal voltage can be increased by about 15%. However, if output-voltage overmodulation is allowable (it is even advisable), then taking into account the boundary cases, the amplitude of a component of fundamental frequency may even reach the value $2U_{DC}/\pi$ [26, 27].

On the basis of Equations 3.6–3.8 it is easy to show that the input currents of a three-phase VSI with sinusoidal output voltage are

$$i_A = i_{DC} + (1/2)\cdot i_0, \quad i_B = -i_{DC} + (1/2)\cdot i_0, \quad i_0 = i_a + i_b + i_c \tag{3.9}$$

where

$$i_{DC} = (A/2)\cdot[i_a \cdot \sin(\omega t) + i_b \cdot \sin(\omega t - 2\pi/3) + i_c \cdot \sin(\omega t - 4\pi/3)]$$

The above relations indicate the unique characteristic of a three-phase VSI, which is its ability to generate reactive currents, theoretically without application of any input energy storage such as capacitors $C_{DC}$. This characteristic, resulting from the lack of reactive power in DC circuits, is used, for example, in D-STATCOM systems [9, 10].

Continuous (average) models, are very supportive when one wants to evaluate usability of PEC in specific application. By avoiding impact of switching process, problems with stiff differential equations are avoided. At the same time, in the case of switching frequency exceeding 5 kHz, simulation error usually does not exceed 5%. Therefore we can successfully focus our attention on functional characteristics of tested application. It is worth noting that application of average models of the PEC usually includes controlled voltage sources and electric current sources, instead of transformers of adjustable ratio of transformation. This results, mainly, from the approach used in the circuits' theory. One example is model three-phase VSI presented in Figure 3.11, corresponding with the model in Figure 3.10b.

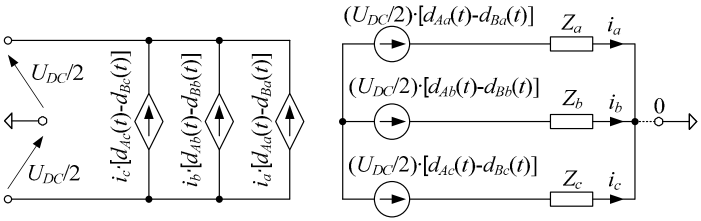



**Figure 3.11.** Continuous model of the three-phase VSI with of controllable sources

## 3.2 High Technology of Converters

The conventional circuit elements applied to PE arrangements can be assigned to one of the classes resistive elements, capacitive elements, magnetic devices, semiconductor devices operated in the linear mode, and semiconductor devices operated in the switched mode (Figure 3.12). At the same time, different classes vary in priorities of application. In the case of controllers one usually avoids applying magnetic devices because of relatively large overall dimensions and integration difficulties. Whereas in PEC, with respect to power losses, semiconductor devices operated in the linear mode are not applied. Moreover, application of resistors should also be limited and replaced with other possibilities. Nowadays, resistors remain in use in cases of dissipative snubbers [28–30] as well as in starting systems PEC, *e.g.*, for initial charging of capacitors in DC circuits. As

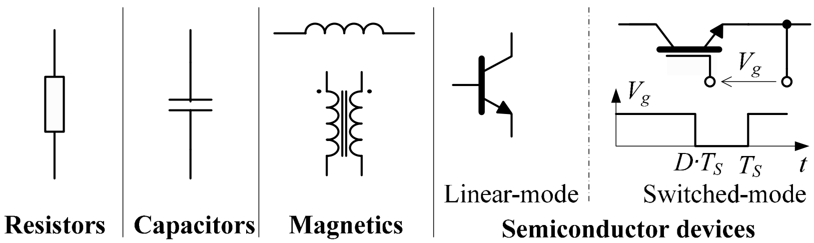



**Figure 3.12.** Different conventional circuit elements of the PE arrangements

supplementation of the above discussion, Figure 3.13 presents usual quantities, expressed in percentage, of the components in weight and volume of the PE arrangements with medium power. As we can see the critical elements are capacitors, semiconductor switches and magnetic devices, and secondary cooling systems and bus work.

### 3.2.1 State-of-the-Art of Power Semiconductors Switches

Recent technology advances in PE have been made by improvements in controllable power semiconductor switches. Figure 3.14 presents probably the most important power semiconductors switches on the market today and their power range [31].

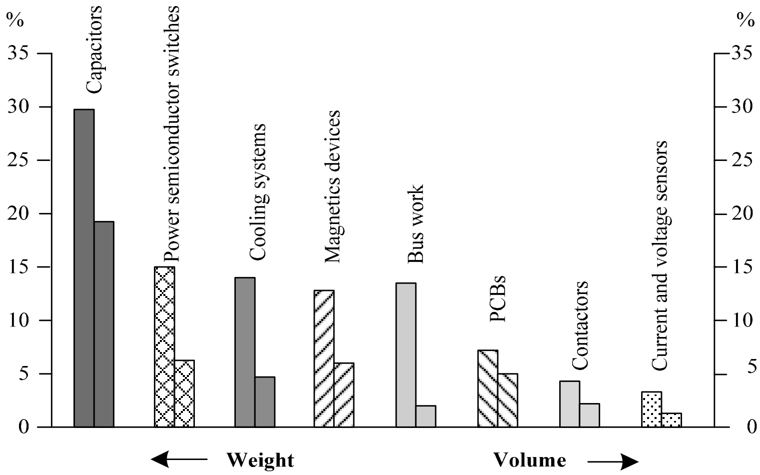



**Figure 3.13**. Typical components in the construction of the PE arrangements

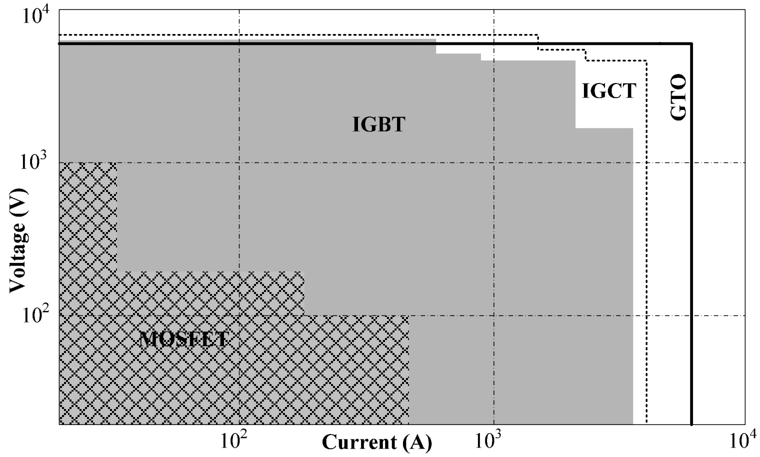



**Figure 3.14**. Power range of commercially available power semiconductors

MOSFETs and IGBTs have replaced BJT almost completely. A remarkable development in MOSFETs took place during the last few years. Today, available MOSFETs achieve maximum switch power up to about 100 kVA.

Conventional GTOs are available with a maximum device voltage of 6 kV in industrial converters. High state current density, high blocking voltages, as well as the possibility to integrate an inverse diode are considered significant advantages of these devices. However, the requirement of bulky and expensive snubber circuits as well as the complex gate drives leads to replacement of GTOs by IGCTs. Just like GTOs, IGCTs are offered only as press-pack devices. A symmetrical IGCT, for example, is offered by Mitsubishi with maximum device voltage of 6.5 kV. It is technically possible today to increase the blocking voltage of IGCTs as well as the inverse diodes to 10 kV. Due to the thyristor latching structure, the GTO offers lower conduction losses than the IGBT of the same voltage class. In order to improve switching performance of classical GTOs, researchers developed Gate-commutated Thyristors (GCTs) with a very short turn-off delay (about 1.5 μs) [32]. Although new asymmetric GCT devices characterized by up to 10 kV with peak controllable currents and 1 kA have been developed, only devices with 6 kV and 6 kA are commercially available.

Also commercially distributed nowadays are IGBTs from 1.2 kV up to 6.5 kV with DC current ratings up to 3 kA [33]. They are optimized to meet the specific requirements of high-power motor drives for industrial applications. Due to the complex and expensive structure of a pres-pack, IGBT are mainly applied to module packages. In IGBT modules, multiple IGBT chips are connected in parallel and bonded to ceramic substrates to provide isolation. Both IGCTs and IGBTs have the potential to lower overall costs of the systems, to increase the number of economically valuable applications as well as to improve the performance of high-power converters, (when compared to GTOs) due to a snubber-less operation at higher switching frequencies.

In the case of insufficient voltage-current parameters of available semiconductor devices in a given application, it is possible to use their parallel and series connections [30, 34, 35]. In a similar manner it is also possible to connect ready converter modules, *e.g.*, PEBB. The possibilities of different connections depend upon achieved uniformity of division of currents and voltages, with the assistance of proper control mechanism. Therefore, parallel connection is usually applied only to MOSFETs or converter modules.

On the other hand, GTOs, GCTs/IGTCs and IGBTs are devices which connect in series relatively well. In such arrangements the most difficult task is to compensate voltages in dynamic states – during switching on and especially during switching off. For example, relatively small differences between switching-off time of the two transistors IGBT equal $\Delta t_{off}$=40 ns can cause differences in dividing voltage of more than 50%. In such a case connecting in series more than *h* devices is purposeless. The same problem can be attenuated in a simper way, mainly by introducing additional Gate Balance Transformers (GTC) [36] into the gate circuits. The solution, which is presented in Figure 3.15, can also be applied to series connections of three or more transistors IGBT. The effectiveness of the proposed solution is illustrated by current and voltage waveforms during switching off (Figure 3.15).

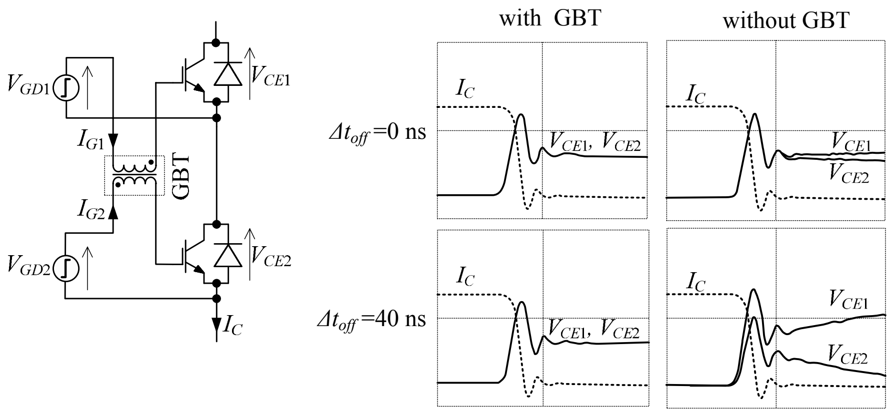



**Figure 3.15**. IGBTs series connection with gate balance transformer

IGBTs, GTOs, GCTs and IGCTs, as well as Emitter Turn-off (ETO) thyristors, are being improved continuously [37, 38]. Research is mainly focused on increase of permissible voltages, currents and switching frequency, and decrease in conduction and switching losses. Hopes are placed in ETO thyristors, which are distinguished by two gate circuits (Figure 3.16). Because of this characteristic, the turn-off time of the ETO thyristor is shorter than GTOs' or IGCTs'.

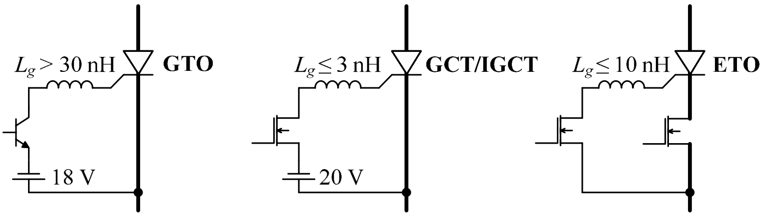



**Figure 3.16**. Gate drivers of conventional GTOs, GCTs/IGCTs, and ETOs

Research also concentrates on semiconductor devices that would be different to siliceous [39], and which would be characterized by higher voltage breakdowns (Figure 3.17) as well as higher permissible work temperatures. Sooner or later, commercial power-electronic devices based on silicon carbide (SiC), of relatively high power and high mean base voltage [40] should be available on the market. However, today the most important and fundamental semiconductor remains silicon.

### 3.2.2 Soft-switching *vs* Hard-switching Techniques

At the beginning of modern PE, that is about 10–15 years after the solid state thyristor had been invented, the elementary arrangements were phase-controlled rectifiers, inverters, and cycloconverters that operate on line or load commutation

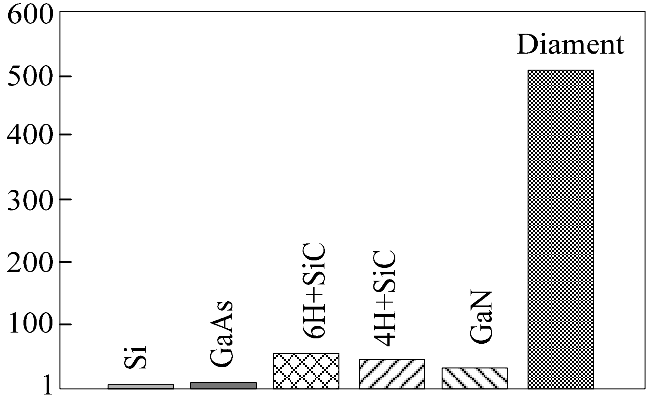



**Figure 3.17**. Relative breakdown voltage of a p-n junction

principle use soft-switching [41–43]. When an incoming thyristor is turned on, the current is gradually transferred from the outgoing to the incoming device, and then the outgoing device turns off by a segment of reverse voltage. Basically, this is soft switching at zero current for both the incoming and outgoing devices. In fact, the classical force-commutated thyristor inverters could also be defined as soft-switched with the help of auxiliary devices and circuit components [44–46]. Their structures, however, were quite developed, and this fact had negative impact on the overall efficiency of the PE circuits. Force-commutated thyristor converters gradually became obsolete due to the emergence of turn-off power semiconductor devices (MOSFET, IGBT, GTO, IGCT *etc.*).

Many modern PEC are used with turn-off power semiconductor devices that apply hard-switching techniques. In this case, during turn-on simultaneous current growth and voltage extinction occur in the switches, whereas in the case of turn-off the exact opposite occurs – simultaneous current extinction and voltage growth. In both situations, in real power switches, significant switching losses occur (Figure 3.2). For that reason, as well as because of other device stresses and EMI problems [47], the typical PEC switching frequency with application of hard-switching technique is limited to a few tens of kilohertz (depending on the type of power and application PEC).

In order to improve operating conditions of the power devices, in particular in switching processes, the circuits forming the switching trajectory are applied (Figure 3.18). The earliest applied device was dissipative passive snubbers, while later active snubbers with energy recovery were introduced [5, 29, 30]. In addition, supportive LC circuits that realize so-called soft-switching were introduced [48]. This concept consists in utilization of resonant tanks in the converters in order to create oscillatory voltage and/or current waveforms. In such a case, Zero Voltage Switching (ZVS) or Zero Current Switching (ZCS) conditions can be created for the power switches.

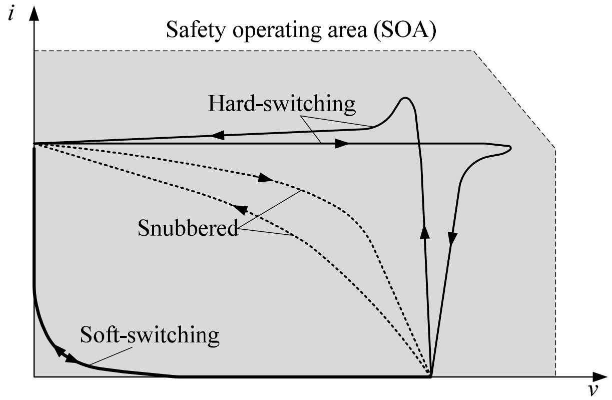



**Figure 3.18**. Typical switching trajectories of power semiconductor devices

Figure 3.19 shows the typical voltage and current waves at hard turn-on and turn-off of a device in a simple buck converter with and without dissipative snubbers. The turn-on snubber $L_1$-$R_1$-$D_1$ allows decreased maximum value of the transistor current $i_T$, decreased stress $di_T/dt$ as well as decreased current component $i_T$ caused by reverse current of the diode $D_0$, limiting turn-on switching losses and transferring them to resistor $R_2$. However turn-off snubber $C_2$-$R_2$-$D_2$ allows one to

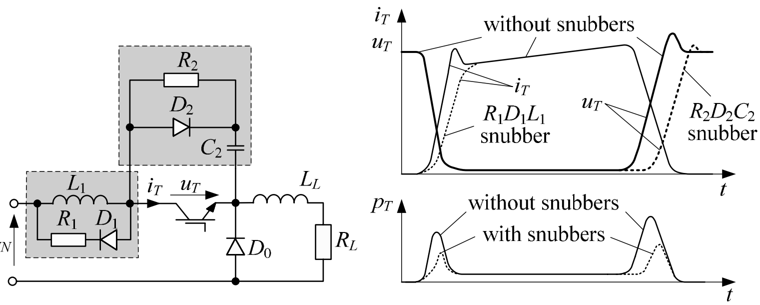



**Figure 3.19**. Switching waveforms of the converter with and without snubbers

transfer turn-off switching losses from the transistor to the resistor $R_2$. Therefore it decreases the maximum voltage in the transistor $u_T$. In such a manner the snubbers produce a more secure switching trajectory of a transistors (Figure 3.18).

Sometimes, in the modern PEC, in particular high power PECs, the number of snubbers is minimized or they are not used at all. This results from the fact that better and better power switches are developed as well as from the pursuit of cost cutting. Obviously semiconductor devices are then more head load and should be over-dimension. Often, however, supportive circuits must be used in order to produce a switching trajectory. In such cases more and more often solutions that allow for soft-switching are utilized [49–53].

Throughout the 1990s, new generations of soft-switched PEC that combine the advantages of conventional hard-switching PWM converters and resonant converters were developed. Unlike typical resonant converters, new soft-switched converters usually utilize resonance in a controlled manner. Resonance is allowed to occur just before and during the turn-on and turn-off processes so as to create ZVS and ZCS conditions. Other than that, they behave just like conventional PWM PEC. With simple modifications, many customized control integrated circuits designed for conventional PEC can be employed for soft-switched converters. Because the switching loss and stress have been reduced, soft-switched PEC can be operated at very high frequencies (reaching even a few megahertz) and allow one to obtain very high packing density (over 10 W/cm$^3$). Soft-switching techniques also provide an effective solution to suppress EMI [54, 55] and have been applied to different PEC converters [28–30].

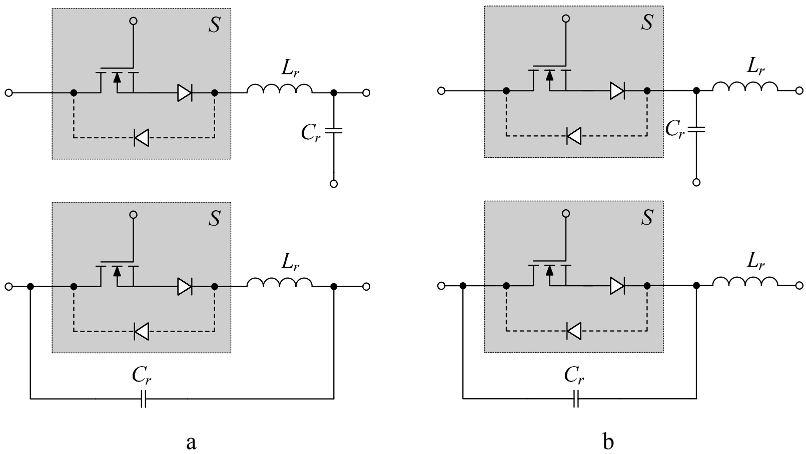



**Figure 3.20**. Types of resonant switches: **a** zero-current; **b** zero-voltage

The fundamental component used in a soft-switching technique is a resonant switch. It is a sub-circuit comprising a semiconductor switch *S* and resonant elements $L_r$ and $C_r$. Uni-directional as well as bi-directional switches are also used as switches *S*. In addition, a type of applied switch *S* determines the operation mode of the resonant switch [28, 56]. The basic two types of resonant switches, including Zero-current (ZC) and Zero-voltage (ZV) resonant switches, are shown in Figure 3.20.

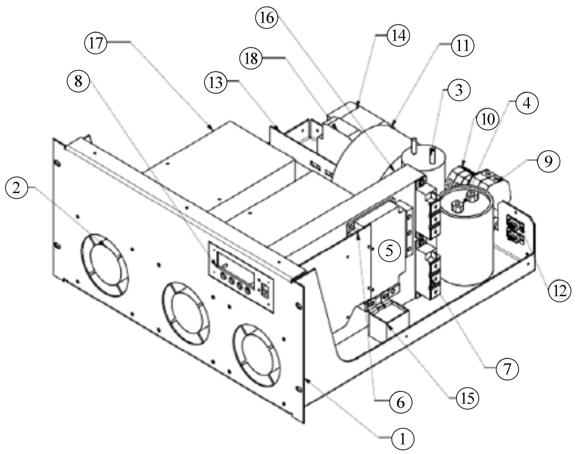
In a ZC resonant switch (Figure 3.20a), an inductor $L_r$ is connected in series with a power switch *S* in order to create ZCS conditions. The objective of this type of switch is to shape the switch current waveform during conduction time in order to create a zero-current condition for the switch to turn off. If a uni-directional switch *S* is applied, the switch current is allowed to resonate in the positive half cycle only, that is, to operate in half-wave mode. If a diode is connected in anti-parallel, the switch current can flow in both directions. In this case, the resonant switch can operate in full-wave mode. At turn-on, the switch current will rise slowly from zero. It will then oscillate because of the resonance between $L_r$ and $C_r$. Finally, the switch can be commutated at the next zero current duration.

In a zero-voltage resonant switch (Figure 3.20b), a capacitor $C_r$ is connected in parallel with the switch *S* in order to create ZVS conditions. The objective of a ZV switch is to use the resonant circuit to shape the switch voltage waveform during the off time in order to create a zero-voltage condition for the switch to turn on. If the switch *S* is uni-directional, the voltage of the capacitor $C_r$ can oscillate freely in both positive and negative half-cycle, and the resonant switch can operate in full-wave mode. However, if the switch *S* is bi-directional (that is when a diode is connected in anti-parallel with the unidirectional switch), the resonant voltage of the capacitor is clamped by the diode to zero during the negative half-cycle. Then the resonant switch will operate in half-wave mode.

This book is not designed to deal with other important aspects of soft-switching in PEC comprehensively. More detailed information about the subject can be found in works referred to in this book.

### 3.2.3 Construction Arrangement and Cooling Systems

Construction arrangments and dimensions of PECs vary. Distribution of the components and execution of the electric connections significantly influence characteristics of the PECs. Considering reliability, particularly important are electromagnetic screening of the control circuits from power circuits. Therefore, nowadays, connections between the circuits are often realized with fiber optic cable. Reliability of all connections is also very important.




1) Chassis
2) Cooling fan
3) Output filter capacitor
4) DC link capacitor
5) IPM module
6) PCB board
7) Diode bridge
8) LCD display
9) Output terminal
10) Input terminals
11) Output filter choke
12) CANON jack
13) Internal power supply
14) Series contactor
15) Snubbers capacitor
16) IPM heatsink
17) DC/DC converter moduls
18) Time relay

**Figure 3.21.** Single module 48 V/ 230 V/ 5 kW of a redundant inverters system

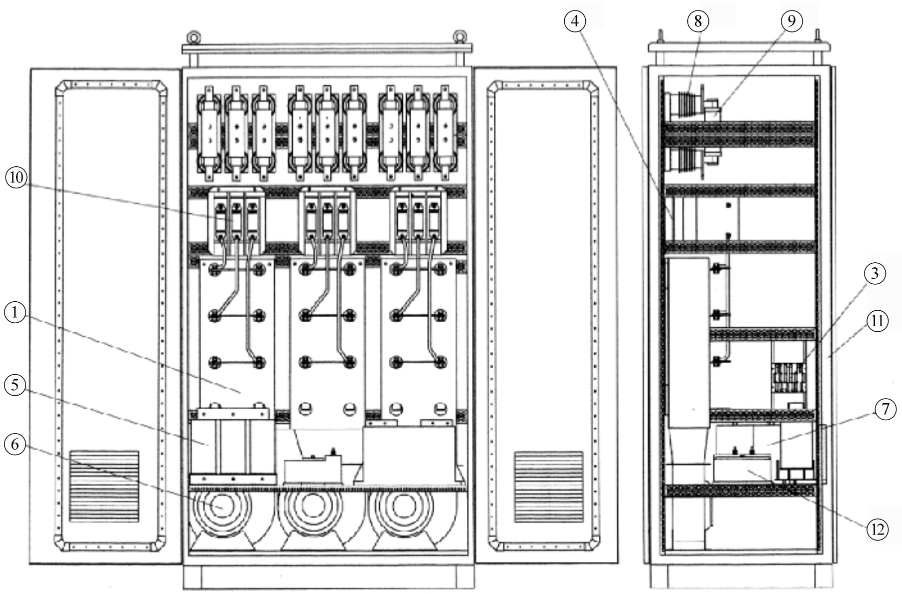
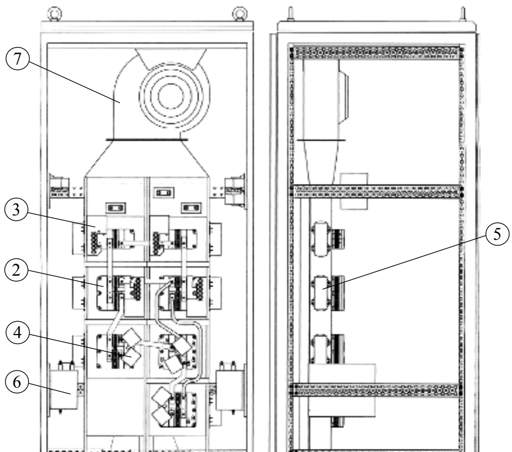


**Figure 3.22**. Design of an 6 kV / 2 MVA 18-pulse diode rectifier: 1) modules of the three-phase diode rectifier bridge with radiators; 2) power supply of the contactor (invisible); 3) resistors of the auxiliary circuits; 4) starting resistors; 5) transformer of the auxiliary circuits; 6) cooling fans; 7) DC link capacitors; 8) fuse base; 9) fuse; 10) vacuum contactor; 11) cubicle; 12) reactor



1) DC link capacitors
2) Transistors or diodes module
3) Gates driver
4) Snubbers capacitor
5) Power supply of a driver
6) Auxiliary fuses circuit
7) Cooling fan

**Figure 3.23.** Design of the prototype of the four-level 6 kV voltage inverter branch

In PECs with power up to about 100 kVA, all components are usually placed in a common box. An example of such a solution could be design of a single modul of a redundant inverters system (Figure 3.21). For high power, particular components or assembly components are placed in a separate cubicle (Figures 3.22 and 3.23).

Relatively often, power semiconductor devices are also offered as complete construction modules – assembly components together with heatsink arrangement. Examples of these modules are shown in Figure 3.24. Similar diode modules were applied, for example, in an 18-pulse rectifier (Figure 3.22). Such a simplifying solution design of the PECs [18, 22, 23, 57] in connection with modern cooling systems [58–60], and modern passive elements [20, 21, 61] can be realized for wide power intervals and for different applications.

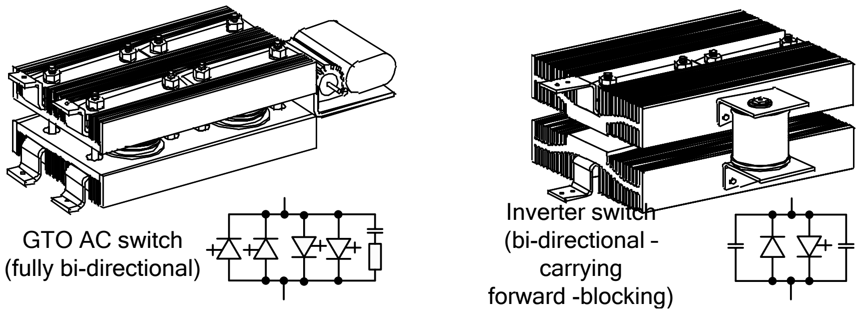



**Figure 3.24**. Example of a construction of high PE modules

To PE arrangements, air-cooling and liquid-cooling systems – direct and indirect – are applied. During the direct cooling, the medium is in direct contact with package semiconductor devices, heatsink or other electric components. However, with indirect cooling, two cooling media are used, one of which transfers heat from the components to a heat exchanger, second to environment. Coolant can consist of natural circulation or forced circulation (fans, pumps). Sometimes mixed coolant and circuits are also applied.

In low power PECs, air-cooling with natural circulation is almost always used. However, in PECs with power from few to hundreds of kilowatts, it is usual to apply forced air-cooling. The selected solutions of cooling systems of cubicles are presented in Figure 3.25. In all examples, cool air is delivered *via* vents in the bottom of the cubicle or in the lower parts of lateral faces. If PECs are to work under conditions of dustiness, then sealing of the cubicle and application of exchangeable dust filters in the vents are required. Sometimes, in order to achieve the desired goal, indirect cooling with a heat exchanger (HE) is used, for example as presented in Figure 3.25f.

In PECs for very high currents, liquid-cooling systems, for which water is a medium [60], are usually used. At certain, oil is used because of its insulating properties. It should be noted that liquid cooling is also used to carrying away heat from other power components of the PECs. The fundamental problem in liquid-cooling systems is leaking of any connections. Moreover, efficiency of liquid-cooling is much higher if a medium is in a boiling state. Then heat transfer takes place not only by convection but also thanks to the phase transition of liquid into steam. Arrangements which rely on this property are referred to as vapor-cooling systems [58].

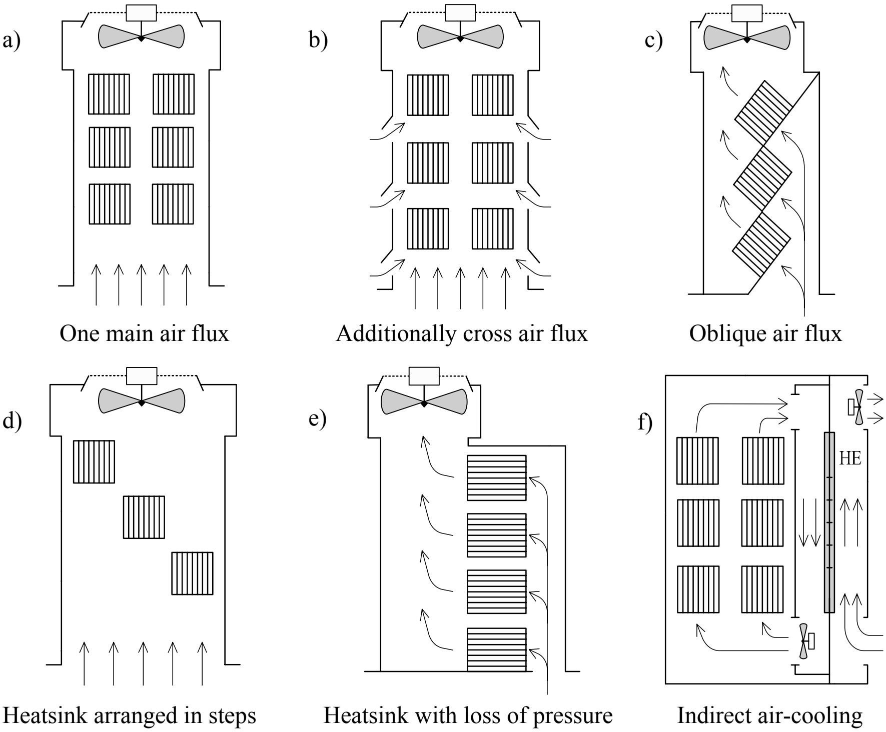



**Figure 3.25**. Air-cooling systems of cubicles with forced circulation

Variations of vapor-cooling systems are heat pipes (Figure 3.26) [59, 62]. Closed pipe is filled with liquid (water, freon, fluorocarbon *etc.*). On the inner wall are layers of a special material with capillary properties, forming a wick structure.

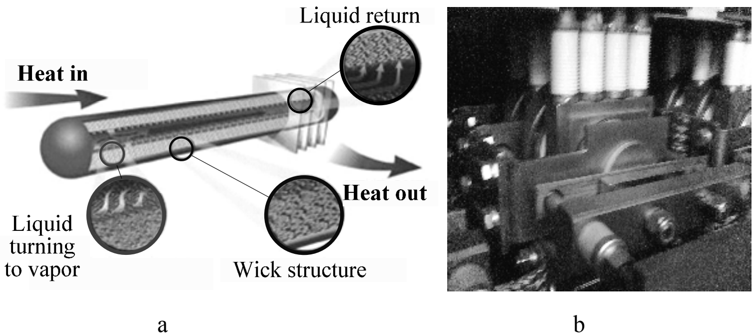



**Figure 3.26**. **a** Principle of the heat pipe. **b** Example of application in PEC

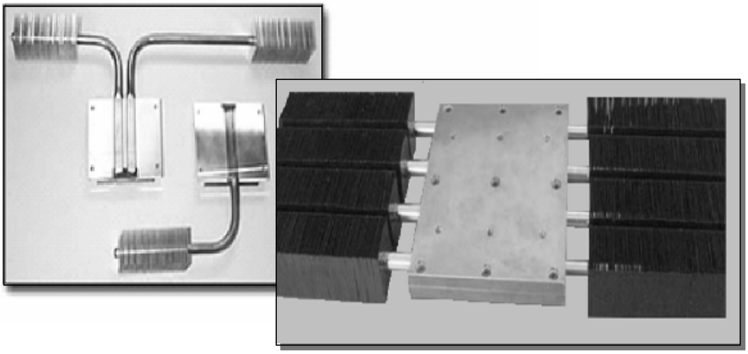


**Figure 3.27**. Example of the heat pipes assisted heat-sink

The steam formed in the heating sphere is transport to the condensation sphere at low temperature under conditions of pressure difference. In this sphere the steam gives up heat and condenses. Due to capillary forces the liquid comes back through the wick structure to the heating sphere and the cooling cycle is repeated. A significant property of heat pipes is also the option of accommodating almost any arrangement. Thanks to that the constructor can apply a cooling component that enables him to transfer heat to the part from which it is easiest to carry away. One example is the heat pipes presented in Figure 3.27.

## 3.3 Multi-level Converters

This section briefly discusses selected basic problems of modern multi-level PECs.

### 3.3.1 Multi-level Converter Concepts

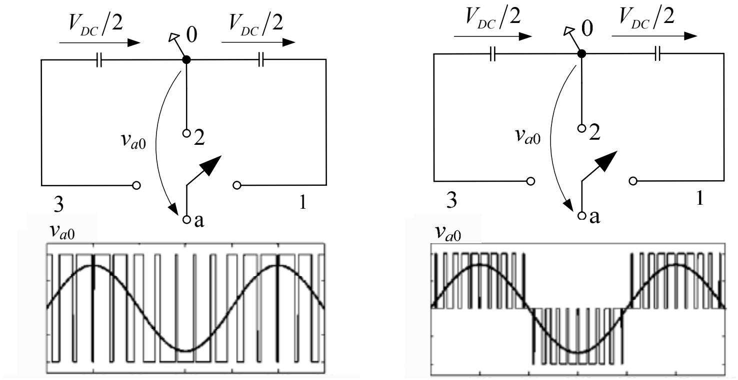
In PECs with PWM of medium/high voltage/power and some specific applications and running conditions, typical solutions (for example three-phase VSI presented in Figure 3.10) are not the most suitable ones. Then too high frequency of the switches in semiconductor devices of high voltage/power (requires a compromise between output-voltage quality and regulation dynamics with application of an output filter), higher voltage stresses, and smaller $dv/dt$ and EMI problems (without any special countermeasures), and sometimes an insufficient value of the peak voltage in semiconductor devices (for a peak voltage of 6 kV, the recommended voltage is about 3.5 kV) would be the main reasons for a second interest in multi-level PECs in the 1980s–1990s, in particular multi-level VSI [63–69]. Many older solutions would then have limited applications [70–75].

As the main advantages of modern multi-level VSI we can count [76]:

- Increased range of output-voltage amplitude changes;
- Greater accuracy in modeling output voltage and current;
- Ability to decrease transformation ratio and even eliminate the output transformer for medium voltage;

- More easy adaptation to low-voltage energy storage;
- Decreased voltage hazard and current elements (dependent on applied typology);
- Decreased level of common-mode disturbances.

The basic differences regarding conventional two-level VSI and the general principle of wave-forming output multi-level voltage are illustrated in Figures 3.28 and 3.29.




a b

**Figure 3.28.** Principles of voltage wave-forming in: **a** two-level VSI; **b** three-level VSI

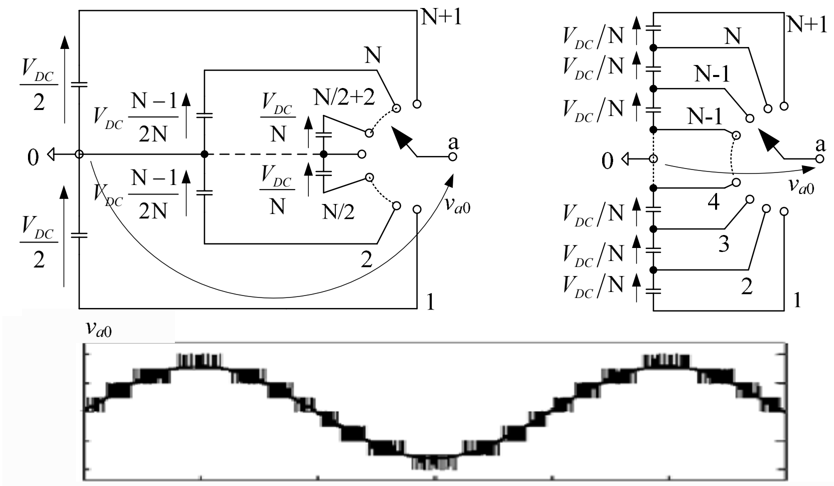



**Figure 3.29.** General principles of voltage wave-forming in N-level VSI

All known topologies of multi-level VSI in the literature, without isolated (galvanic separated) DC voltage sources, can be synthesized on the basic VSI modules and the multi-layer topology presented for the case of three-layers in Figure 3.30.

The manner of their synthesis is to exclude individual switches properly. However the majority of multi-level VSIs obtained in this manner did not find any applications, either because of the complexity of this typology or because of greater losses. In practice, two special cases of multi-layer topology that are realized are multi-level Diode Clamped Inverters (DCI), with which we also include Neutral Point Clamped (NPC) VSIs, and multi-level Flying Capacitor Inverters (FCI) [76]. Furthermore, also applied are multi-level Cascaded H-bridge Inverters (CHBI), which require, in contrast to the first two, isolated DC voltage sources [69, 76, 77]. The main advantage of CHBIs is the possibility of easy development and independent stabilization of the voltage in the DC circuit for each mode H-bridge. However, none of the three selected topologies of multi-level VSI, presented as one branch in Figure 3.31, has so far gained a leading position.

*3.3.1.1 Diode Clamped Multi-level VSI*

As a type of VSI configuration, which is important for high-power applications, the diode-clamped inverter provides multiple voltage levels through the connection of the phases to a series bank of capacitors. According to the original invention [64], the concept can be extended to any number of levels by increasing the number of capacitors. Early descriptions of this topology were limited to three levels [70, 71], where two capacitors were connected across the DC bus, resulting in one additional level. The additional level was the neutral point of the DC bus.

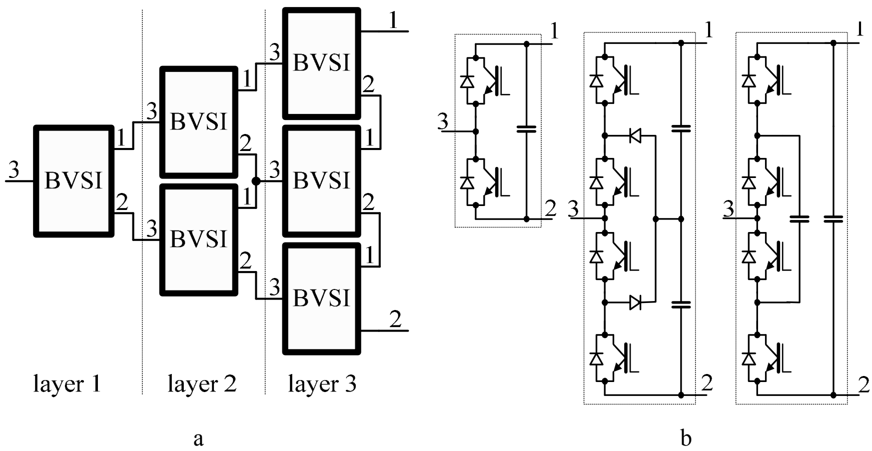



**Figure 3.30. a** Three-layer VSI. **b** Example of the basic-VSI module

For such multi-level PEC, the terminology neutral point clamp converter was introduced [63].

In case $N$+1 number of voltage levels of DCI (Figure 3.31), one phase leg consists of 2$N$ active switches (IGBT, IGCT, GTO) and minimum 2($N$–1) clamping diodes [78]. The total bus voltage $U_{DC}$ is distributed across the capacitors $C_1, ..., C_N$. Hence, if voltage pattern on capacitors is uniform, then output voltage of the DCI can take values $U_{DC}\cdot(n-N)/2$ for $n=0,1,\ldots,N$.

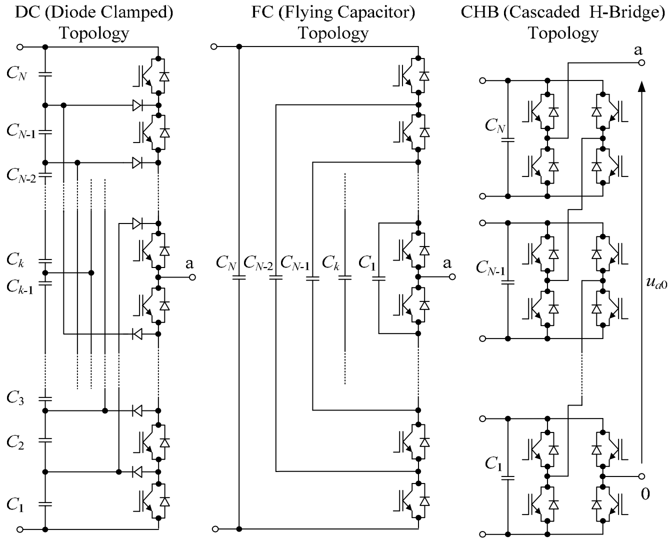



**Figure 3.31.** Generalized typologies of the most frequently applied multi-level VSI

### *3.3.1.2 Flying Capacitor Multi-level VSI*

Another fundamental multi-level topology, the flying capacitor inverter (and other flying capacitor PECs), involves a series connection of capacitor switching cells [66, 79]. This topology, presented in Figure 3.31, reveals several unique and attractive features when compared to the diode-clamped converter. One feature is that added clamping diodes are not needed. Furthermore, the flying capacitor converter has a switching redundancy within the phase, which can be used to balance the flying capacitors so that only one DC source is needed. Traction converters are typical applications of this topology. One phase leg consists of $2N$ active switches and $N$–1 flying capacitors.

### *3.3.1.3 Cascaded H-Bridge VSI*

This class of multi-level PECs is based on a series connection of single-phase VSI bridges (Figure 3.31), and the earliest reference to them appeared in 1975 [70]. The CHBI topology has several advantages that have made it attractive to medium and high power drive applications [77]. Since this topology consists of series power conversion cells, the voltage and power level may be easily scaled. The DC link supply for each H-bridge VSI element must be provided separately. The ability to synthesize quality wave-form of the output voltage with excellent harmonic spectrum is one of its main advantages. Additional, a very important advantage of CHBIs is the possibility to utilize low-cost low-voltage power semiconductors, switches and capacitors [80]. However, drawbacks of this topology are the large number of power devices and of voltages required to supply each cell with a complex, bulky and expensive isolated transformer.

### 3.3.2 Basic Comparison of Multi-level Inverter Topology

Each of the typologies of multi-level inverters presented above differs in the number of semiconductor switches used as well as reactive elements. The cooperative analysis of them could help to decide about appropriate solutions for particular applications.

Based on familiarity with the operation principle we are able to compare topologies of the multi-level voltage inverters according to various criteria. Most often as a criterion we accept a required number of semiconductor and passive components depending on a number $N$ output voltage levels. With these assumptions and one phase of different multi-level VSI, the results obtained are presented in Table 3.1 [81].

**Table 3.1**. Number of the components for one phase of the $N$-level VSIs

| Topology | DCI | FCI | CHBI |
|---|---|---|---|
| Number of active switches | $2(N-1)$ | $2(N-1)$ | $2(N-1)$ |
| Number of clamped diodes | $(N-1)(N-2)$ | 0 | 0 |
| Number of flying capacitors | 0 | $(N-1)(N-2)/2$ | 0 |
| Number of supply capacitors | $(N-1)$ | $(N-1)$ | $(N-1)/2$ |

In Tables 3.2 and 3.3, based on [82, 83, 84, 85], are presented results of the comparative analysis and calculation of realization costs of selected topologies of three-phase multi-level VSIs. This analysis and evaluation by the authors of listed publications was conducted for the following comparable topologies:

- FK-L2: conventional VSI with four IGBTs connected in series;
- NPC-L3: NPC VSI with two IGBTs connected in series;
- DCI-L5: five-level DCI with four capacitors;
- FCI-L5: five-level FCI with neutral point;
- CHBI-L9: nine-level IHBI realized as cascaded connection of the four inverter bridges

with the assumption that

- Value of the constant voltage supplying VSI $U_{DC}$=6.2 kV;
- Inverters should assure line-voltage 4.2 kV;
- IGBTs (3.3 kV, 1200 A) are applied as active switches

with subjective evaluation of degree of complexity of their realization and assuming relative cost per unit for used components: IGBT (generally) – 1 p.u., IGBT for CHBI (lower voltage 1600 V) – 0.5 p.u., power diode with snubbers – 0.5 p.u., diode clamped – 0.3 p.u., capacitor (1.5 kV, 5 mF) – 0.5 p.u., snubber for IGBT – 0.1 p.u.

In typical applications of multi-level inverters in a medium voltage supply network, the number of output voltage levels rarely exceeds four to five. Available commercial turn-off 6.5 kV power switches nowadays allow realization three-phase VSIs of output voltage 6 kV without any problems. Increased number of levels and higher output voltage could be obtained only when cascade VSIs were applied, *e.g.*, CHBIs. Module based construction and isolated power supplies in CHBIs improve their safety in terms of electric shock and ease use. Unfortunately CHBIs have quite large overall dimensions and complicated control and protection.

**Table 3.2**. Number of components for analyzed multi-level VSI (one-phase)

| Topology | FK-L2 | NPC-L3 | DCI-L5 | FCI-L5 | CHBI-L9 |
|---|---|---|---|---|---|
| IGBTs | 8 | 8 | 8 | 8 | 16 |
| Clamped diodes | 0 | 4 | 12 | 0 | 0 |
| Power supply | 1 | 1 | 1 | 1 | 4 |
| Snubbers | 8 | 8 | 8 | 0 | 0 |
| Flying capacitors | 0 | 0 | 0 | 6 | 0 |
| Supply capacitors | 4 | 4 | 4 | 4 | 4 |

**Table 3.3**. Comparison of estimated costs of the analyzed multi-level VSI

| Topology/costs | FK-L2 | NPC-L3 | DCI-L5 | FCI-L5 | CHBI-L9 |
|---|---|---|---|---|---|
| Semiconductor switches | 24 | 24 | 24 | 24 | 24 |
| Supply capacitors | 6 | 6 | 6 | 6 | 12 |
| Extra (diode/capacitors) | --- | 3,6 | 10,8 | 9 | --- |
| Snubbers | 2,4 | 2,4 | 2,4 | --- | --- |

We should note that Table 3.3 does not include the cost of the output filter, which is most expensive in the case of a conventional two-level VSI. The cost of the filter depends to a great extent on harmonic distortion in output voltage. Estimated cost of the filter is equal to the cost of elements of the one-phase VSI, whereas the main component is the cost of the reactor. Also, Table 3.3 does not include realization costs for the controller in this initial charging of capacitors, and the expenses of technological processing of VSI realization and its installation. Therefore the technological-economic evaluation of the presented topologies of multi-level VSIs, without taking into consideration considerable technical problems, cannot be unequivocally final. Each topology (Figure 3.31) and derivative topologies [69, 76–78] have their advantages and disadvantages. Usually the specific application decides which typology should be selected.

### 3.3.3 Space Vector PWM Algorithm of a Multi-level VSI

For the particular states of the switches of the inverter the appropriate voltage space vector can be selected in stationary coordinates $\alpha$-$\beta$ [77, 84, 86]. In the *m*-level VSIs, the area of the space vector is usually divided into six sectors, in which we can distinguish triangular areas among three nearest locations of the space vector. Single sector and equilateral triangles of a side *a* is shown in Figure 3.32.

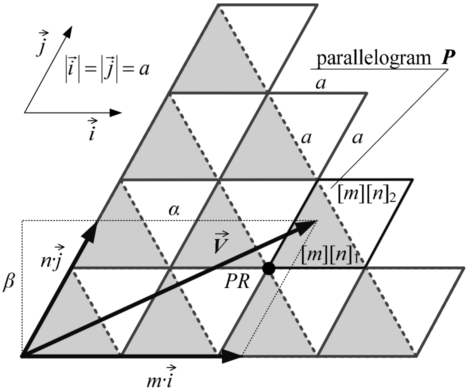



**Figure 3.32**. Single sector for m-level VSI

Reference voltage vector *V* can be presented as a linear combination of the vectors *i* and *j*. In accordance with Figure 3.34, coordinates [$\alpha$, $\beta$] of vector *V* are as follows

$$\vec{V} = [\alpha, \beta] = m \cdot \vec{i} + n \cdot \vec{j} \tag{3.10}$$

where

$$m = \alpha/a - \beta/\left(a\sqrt{3}\right);\ n = 2 \cdot \beta/\left(a\sqrt{3}\right) \tag{3.11}$$

Integers of the *m* and *n* define coordinates of beginning of parallelogram *P* (point *PR* on Figure 3.32), where the reference vector occurs at the time. In order to determine the belonging of the reference vector *V* to one of the two triangles in parallelogram *P*, the value of the following sum *D* needs to be found out

$$D = [m - \mathrm{int}(m)] + [n - \mathrm{int}(n)] \tag{3.12}$$

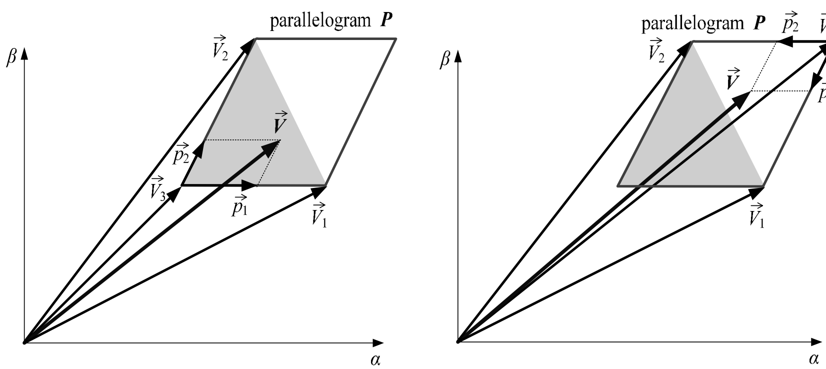



**Figure 3.33**. Positions of the vector *V* in parallelogram *P*

If $D \leq 1$, then the reference vector $V$ occurs in the triangle with index $[m][n]_1$, else the vector $V$ belongs to the triangle with index $[m][n]_2$. Finally, the synthesis of the reference vector $V$ in addition consists of several states of switches, during modulator's work, according to the order defined in the control strategy.

There exist two possible positions of the vector $V$ in the parallelogram $P$ (Figure 3.32). Reference vector $V$ can be projected within the area of each of the two triangles. According to Figure 3.33, the reference voltage vector can be expressed as a vector sum for the position of the vector $V$ as in Figure 3.33 (left)

$$\vec{V} = \underbrace{p_1\left(\vec{V}_1 - \vec{V}_3\right)}_{\vec{p}_1} + \underbrace{p_2\left(\vec{V}_2 - \vec{V}_3\right)}_{\vec{p}_2} + \vec{V}_3 = p_1\vec{V}_1 + p_2\vec{V}_2 + \left(1 - p_1 - p_2\right)\vec{V}_3 \tag{3.13a}$$

or for position of the vector $V$ as in Figure 3.33 (right)

$$\vec{V} = \underbrace{p_1\left(\vec{V}_1 - \vec{V}_4\right)}_{\vec{p}_1} + \underbrace{p_2\left(\vec{V}_2 - \vec{V}_4\right)}_{\vec{p}_2} + \vec{V}_4 = p_1\vec{V}_1 + p_2\vec{V}_2 + \left(1 - p_1 - p_2\right)\vec{V}_4 \tag{3.13b}$$

where $p_1$, $p_2$ are relative lengths (durations) of the active vectors $V_1$ and $V_2$. Duration of zero vectors $V_3$ and $V_4$ result from a difference in carrier period and duration of the active vectors. All space vectors of the four-level VSI are presented in Figure 3.34.

In the selected sector 0 in the Figure 3.34 occur nine numbered regions. Specified position of the space vector is coded as follows – given number defines a point in the linking circuit connected to a load terminal of the particular phase (from left *a*, *b*, *c*). For example, the code "*321*" means that phase *a* was linked to voltage source of value $3\cdot(U_{DC}/3)$, phase *b* was linked to the voltage source of value $2\cdot(U_{DC}/3)$, and phase *c* was linked to the voltage source of value $1\cdot(U_{DC}/3)$. In the case of linear modulation range, value of maximum phase voltage equals $\sqrt{3}\cdot(U_{DC}/3)$, and maximum value of normalized modulation factor $m_a$ in this case equals $3\cdot\sqrt{3}/2$.

Figure 3.35 provides exemplary positions of the space vector of a four-level inverter. When the reference vector $V$ occurs in region 4, relative lengths of the space vectors $V_{331}=V_{220}$ and $V_{321}=V_{210}$ equal

$$p_1 = 2 - n;\ p_2 = 1 - m \tag{3.14}$$

Hence, the vector-duty factors for particular positions of the space vector in modulation period are as shown in Table 3.4.

The discussed algorithm Space Vector PWM (SVPWM) is easy to be implemented in the DSP controller. This algorithm in three-phase VSIs can be easily complemented by a selection procedure of one of few alternative vectors – so-called redundancy vectors. Correct selection of one of the vectors always helps but does not always fully stabilize voltages of the capacitors in multi-level VSIs. In particular, it refers to the typology DCI. For this typology multi-level VSIs

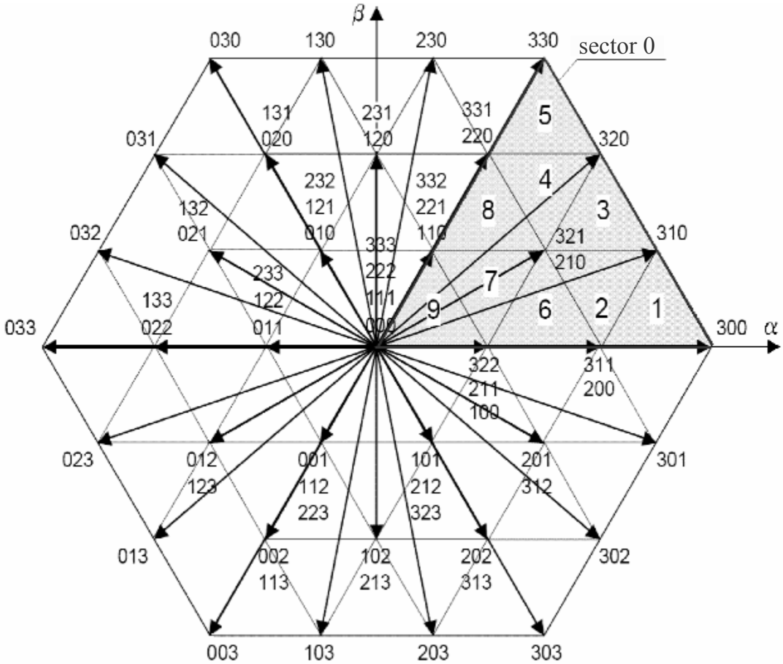



**Figure 3.34.** Space vectors of the four-level VSI

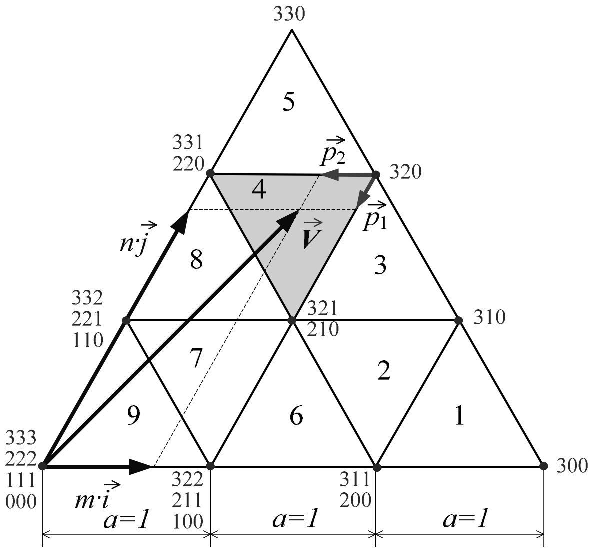



**Figure 3.35.** Example of the position of the normalized vector $V$ in region 4

stabilization of the capacitors' voltages within full admissible range of changes of output voltage is possible only with reactance loads [86–88], that is, for example, in D-STATCOM systems and active power filters. In case of different loads, stabilization of capacitors' voltages is possible only within a limited range of changes of output voltage. The worst case is resistive load. With such load DCIs, above some value of output voltage, it is necessary to change to quasi-three-level and quasi-two-level wave-forming of the output voltage [89]. Active stabilizing circuits could also be applied, or an independent power supply for all supply capacitor DCIs. The final method is most often used in drives. An example is topology of driving frequency converter that is presented in Figure 3.36 [90]. In the controller of these converters the algorithm SVPVM was applied and then completed by selection of appropriate redundancy vectors. This successfully allows a level load of transformer winding in a simple 12-pulse rectifier [91]. Typical the solution are output wave-shapes of the phase voltage and load currents presented in Figures 3.37 and 3.38.

**Table 3.4**. Vector-duty factors for regions 1–9 in Figure 3.35

| Reg. | Duty factors | | |
|---|---|---|---|
| 1 | $d_{200/311}=3-m-n$ , | $d_{310}=n$ , | $d_{300}=m-2$ |
| 2 | $d_{200/311}=1-n$ , | $d_{210/321}=2-m$ , | $d_{310}=m+n-2$ |
| 3 | $d_{210/321}=3-m-n$ , | $d_{310}=m-1$ ; , | $d_{320}=n-1$ |
| 4 | $d_{210/321}=2-n$ , | $d_{220/331}=1-m$ , | $d_{320}=m+n-2$ |
| 5 | $d_{220/331}=3-m-n$ , | $d_{330}=n-2$ , | $d_{320}=m$ |
| 6 | $d_{100/211/322}=2-m-n$ , | $d_{200/311}=m-1$ , | $d_{200/321}=n$ |
| 7 | $d_{110/221/332}=1-m$ , | $d_{100/211/322}=1-n$ , | $d_{210/321}=m+n-1$ |
| 8 | $d_{110/221/332}=2-m-n$ , | $d_{210/321}=m$ , | $d_{220/331}=n-1$ |
| 9 | $d_{000/111/222/333}=1-m-n$ , | $d_{100/211/322}=m$ , | $d_{110/221/332}=n$ |

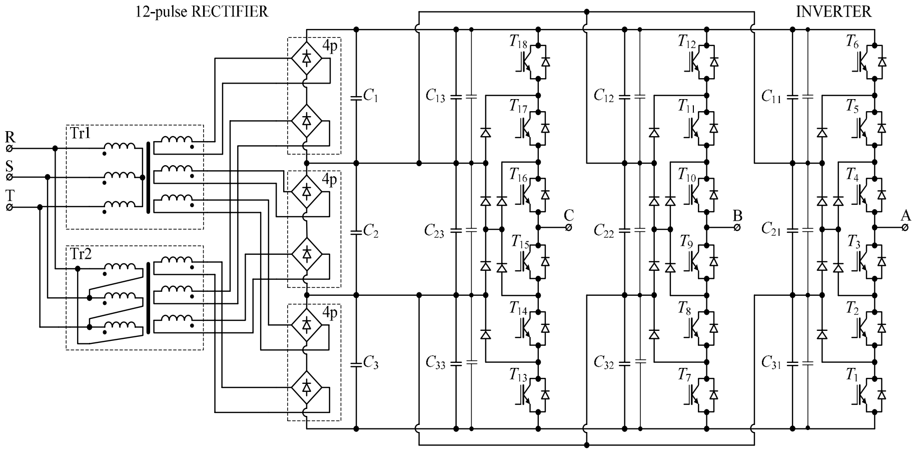
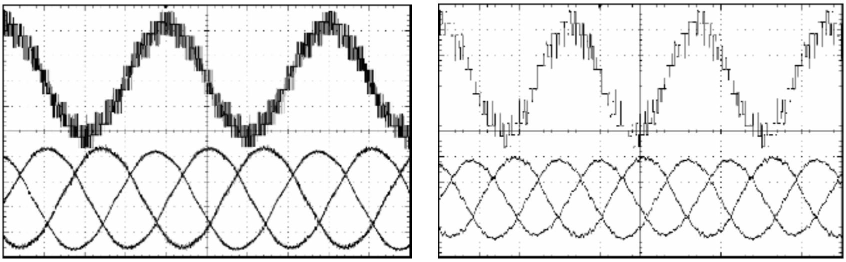
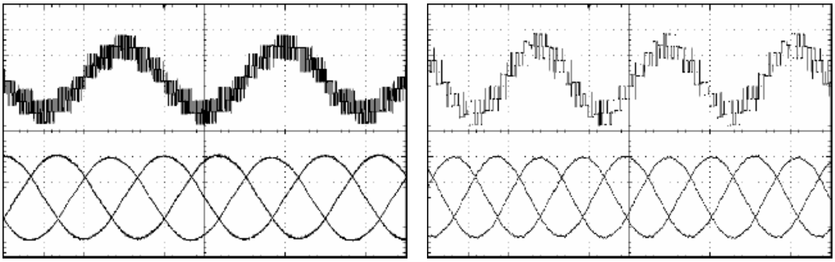



**Figure 3.36.** Frequency converter with four-level DCI and 12-pulse rectifier



a b

**Figure 3.37.** Phase voltage and load currents for modulation factor $m_a$= 2.59 and PWM carriers: **a** $f_c$= 4 kHz; **b** $f_c$= 800 Hz



a b

**Figure 3.38.** Phase voltage and load currents for modulation factor $m_a$= 1.55 and PWM carriers: **a** $f_c$= 4 kHz; **b** $f_c$= 800 Hz

The discussed SVPWM algorithm (slightly modified) is also used in multi-level topology, which is a hybrid connection of typical two-level VSI with additional H-bridge modules in every output [92]. An example of this solution, together with characteristic oscillograms of output line-to-line voltage and phase voltage in individual modes, is presented in Figure 3.39.

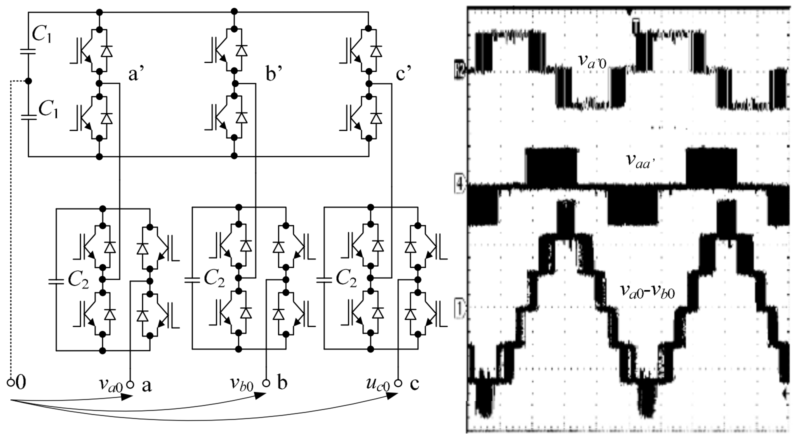



**Figure 3.39.** Connection of a two-level VSI and three H-bridge module

The discussed multi-level inverters obviously do not exhaust all the important solutions that have been tried in recent years, and first of all they concern only voltage systems – VSI. The issue of the multi-level current source inverter, because of its duality when compared to VSI systems, was not dealt with, despite the increasing interest given to these systems and the results achieved [93].

## 3.4 Z-source Converters

Many significant problems that occur in the conventional inverters (Figure 3.40) result from their operating principle. These problems are connected to the following disadvantages:

- In case of voltage source inverters (Figure 3.40a): output voltage $V \leq V_{DC}/1.73$; voltage regulation – only decreasing; problems with short circuits in branches;
- In case of current source inverters (Figure 3.40b): output voltage $Um \geq U_{DC}/1.73$; voltage regulation – only increasing; difficult to apply conventional modules IGBT and open circuits problems.

The issues with short circuits in branches and open circuits are connected with vulnerability of the inverters to damages from EMI distortion.

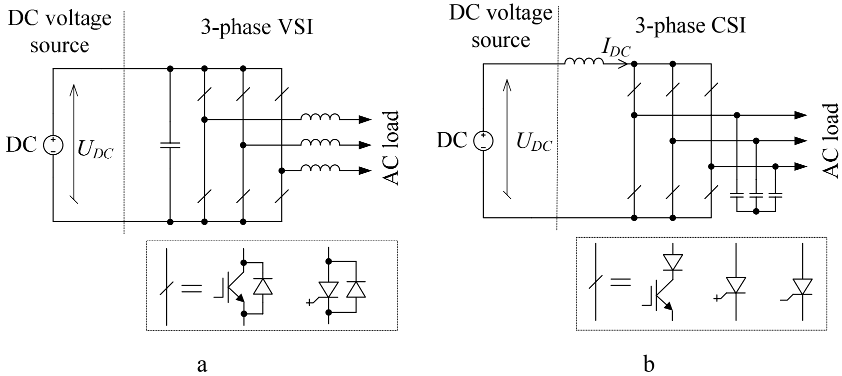



**Figure 3.40.** Conventional inverter systems: **a** VSI; **b** CSI

If the inverters' applications require amplitude to be adjusted outside the limited region, output transformer or additional DC/DC converter (Figure 3.41) can be used. Disadvantages of the solutions with output transformer (Figure 3.41 left) are most of all large overall dimensions, heavy weight and range of regulation limited by transformer voltage ratio. However, if an additional DC/DC converter is applied (Figure 3.41 right), then it results in two-stage conversion of the electrical energy, and therefore we should consider higher costs of the system and increased losses. Moreover, in such a case, one type of inverter cannot be replaced by an other type (*i.e.*, CSI can not be replaced by VSI and *vice versa*) and short circuits or open circuits and transition processes occur. Therefore, the search continues for new solutions in inverter systems with improved adjustment properties. Especially worth attention seems to be the Z-source inverter patented by F.Z. Peng in 2003 [94].

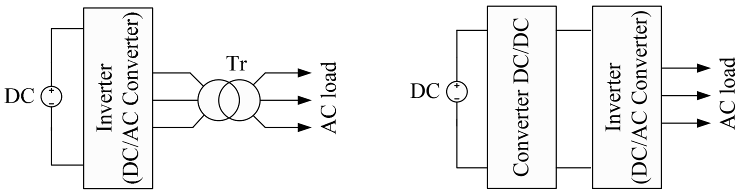



**Figure 3.41.** The inverter systems with increased range of regulation

Figure 3.42 presents basic schemes of the three-phase Z-inverters: voltage (Figure 3.42a) and current (Figure 3.42b) [94, 95]. In contrast to conventional VSI and CSI inverters, on the DC side of the Z-inverter is a *D* diode and a Z-source of "*X*" shape, composed of two capacitors $C_1$ and $C_2$ and two chokes $L_1$ and $L_2$. The *D* diode prevents forbidden reversed current flow (for voltage Z-inverter) or reversed voltage (for current Z-inverter). For this reason, application of the basic Z-inverters are possible only if energy return to the input source is unnecessary. Further, this is forbidden in the case of a fuel cell or photo-voltaic cell. It should be noted that the same diode function can be served by other PE systems as well. The main advantages of the Z-converters are:

- Secures the function of increasing and decreasing of voltage in the one-step energy processing (lower costs and decreased loses);
- Resistant to short circuits on branches and to opening of the circuits that improve resistance to failure switching and EMI distortions;
- Relatively simple start-up (lowered current and voltage surges).

We should acknowledge that two-direction energy flow is only possible due to change of a diode of the source on the switch of the inverter.

Because the operation principle of the voltage and current Z-inverter is similar, all the solutions considered below relate only to the voltage Z-inverter.

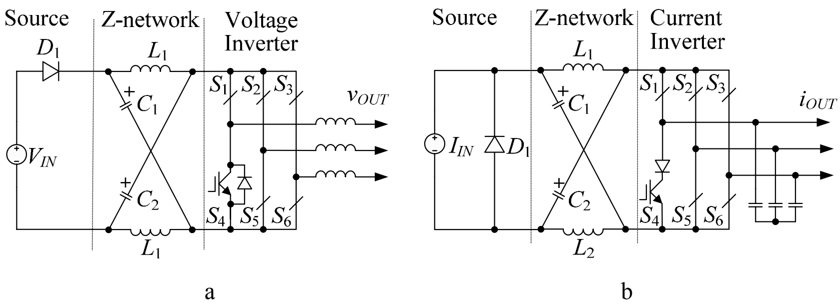



**Figure 3.42.** Basic schemes of the Z-inverter: **a** voltage; **b** current

### 3.4.1. Operation Principle of the Voltage Z-inverter

Conventional three-phase VSI system (Figure 3.40a) can assume eight states: six active states (while exchange of instantaneous power between the load and DC circuit) and two null states (when the load is shorted by transistors). Whereas, three-phase Z-inverter (Figure 3.42a) can assume 9 states, that is one more than in the VSI system — the additional nine state is the third 0 state, occurring when the load is shorted simultaneously by lower and upper groups of transistors. This state is defined as "*shoot-through*" state and may be generated in seven different ways, although of equivalent procedures: independently through every branch (three procedures), simultaneously through two of the branches (three procedures), and simultaneously through all of the three branches (one procedure). The main and unique characteristic of the Z-inverter is that the shoot-through state permits one to raise output voltage above the voltage $V_{IN}$.

Figure 3.43 describes simple equivalent schemes of the Z-inverter examined from the clap site of DC, where a source $v_d$ is modeling inverter $S_1$–$S_6$. In the shoot-through states (Figure 3.43a) a $D$ diode is polarized reversely and does not conduct the inverter bridge input voltage $v_d=0$, and energy stored in capacitors $C$ is transferred to the chokes $L$. In "*non-shoot-through*" states (Figure 3.43b), where every combination of the switches $S_1$–$S_6$ that is allowed in VSI system is also possible, the $D$ diode conducts and the voltage $v_d$ increases stepwise from 0 to its maximum $v_d$*.

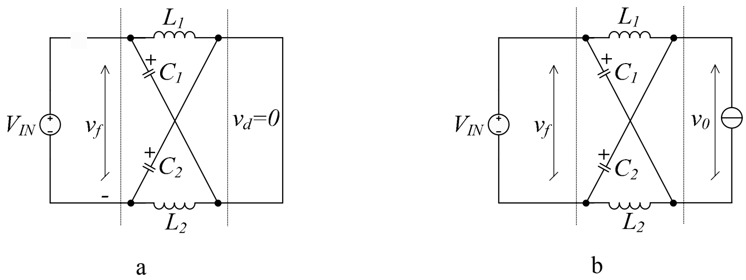



**Figure 3.43.** Equivalent schemes of the Z-source inverter: **a** "*shoot-through*" states; **b** "*non-shoot-through*" states

Since Z-source are symmetric circuits (Figure 3.43), when $C_1$=$C_2$ and $L_1$=$L_2$ and low voltage pulsation $v_{C1}$ and $v_{C2}$ during pulse period $T$,

$$v_{C1} = v_{C2} = V_C \qquad \text{and} \qquad v_{L1} = v_{L2} = v_L \tag{3.15}$$

where $V_C$ is average value of voltage in capacitors, $v_L$ – instantaneous voltage in chokes. Considering Equation 3.15 and equivalent schemes of the Z-inverter (Figure 3.43), voltage $v_d$ is calculated on the basis of following dependencies in "*shoot-through*" states (Figure 3.43a) duration $T_Z$

$$v_L = V_C\,, \qquad v_f = 2\cdot V_C\,, \qquad v_d = 0 \tag{3.16}$$

in "*non-shoot-through*" states (Figure 3.43b) duration $T_N$

$$v_L = V_{IN} - V_C, \qquad v_f = V_{IN}, \qquad v_d = V_C - v_L = 2 \cdot V_C - V_{IN} \tag{3.17}$$

where $v_f$ is Z-source input voltage.

Assuming that in a pulse period $T=T_Z+T_N$, in a steady state the average voltage in chokes $V_L=0$, on the basis Equations 3.16 and 3.17, we should conclude

$$V_L = \frac{1}{T}\left( \int_0^{T_Z} v_L dt + \int_{T_Z}^{T} v_L dt \right) = \frac{T_Z \cdot V_C + T_N \cdot (V_{IN} - V_C)}{T} = 0 \tag{3.18}$$

Hence, average input voltage of the inverter bridge input voltage

$$V_C = V_d = V_{IN} \frac{T_N}{T_N - T_Z} = V_{IN} \frac{1 - D}{1 - 2 \cdot D} \tag{3.19}$$

where $D=T_Z/T$ is "*shoot-through*" duty factor, satisfying a requirement $D<0.5$. Similarly on the basis of Equations 3.17–3.19, the value $v_d^*$ of voltage $v_d$ in "*non-shoot-through*" is determined

$$v_d^* = V_C - v_L = 2 \cdot V_C - V_{IN} = V_{IN} \cdot \frac{1}{1 - 2D} = V_{IN} \cdot B \tag{3.20}$$

where $B=1/(1-2 \cdot D)=T/(T_N-T_Z) \geq 1$ is a peak factor, and the value $v_d^*$ is determined by relative voltage $V_{IN}$.

Further, the value $v_d^*$ determines output voltage amplitude $V_{OUT(max)}$ of the Z-inverter. When applying sinusoidal PWM the amplitude equals

$$V_{OUT(\max)} = M \cdot \frac{v_d^*}{2} = \frac{M}{1 - 2 \cdot D} \cdot \frac{V_{IN}}{2} = K \cdot \frac{V_{IN}}{2} \tag{3.21}$$

where $M$ is modulation index, of maximum value limited by inequity $M \leq 1-D$, related to time $T_Z$ of "*shoot-through*" states. As we conclude, based on Equation 3.21, the Z-inverter output voltage amplitude $V_{OUT(max)}$ can be either lower or higher than in typical VSI system with sinusoidal PWM, *e.g.* $V_{OUT(max)}=M \cdot V_{IN}/2$. This possibility is acknowledged when looking at the 3D diagram of the function

$$K = M / (1 - 2 \cdot D) \tag{3.22}$$

within the acceptable area

$$\Omega \subset \begin{cases} 0 \leq M \leq 1 \\ D < 0{,}5 \cap D \leq 1 - M \end{cases} \tag{3.23}$$

This function is presented in Figure 3.44.

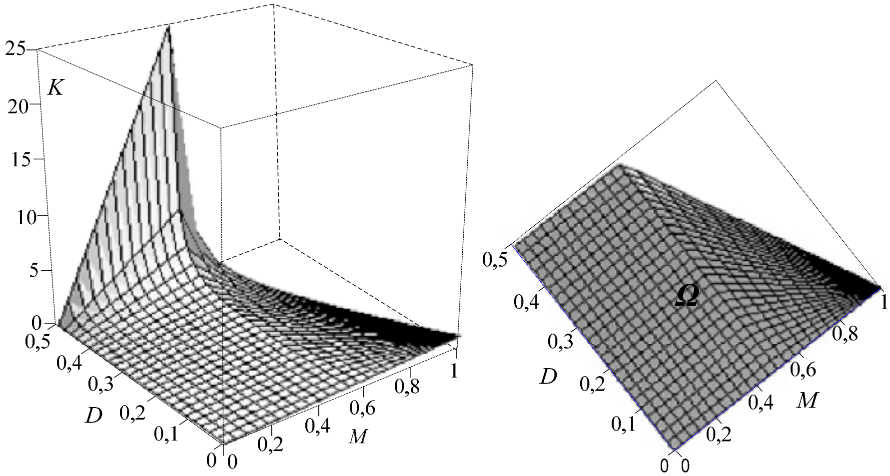



**Figure 3.44.** 3D diagram of the function $K$ (Equation 3.22) within area $\Omega$

The discussed properties of the voltage Z-inverter are confirmed by research presented in a number of publications from the recent past [96–99]. The properties are also illustrated below in selected results from the authors' research. Basic parameters of the system that were assumed in the research are presented in Table 3.5. In order to control switches $V_1$–$V_6$, we applied the algorithm of simple sinusoidal PWM, modified by "shoot-through" states. The essence of this modification is explained in Figure 3.45. Selected results of the research are presented in Figure 3.46.

**Table 3.5**. Parameters of researched voltage Z-inverter (Figure 3.42a)

| | | | |
|---|---|---|---|
| Supply DC | | $U_{IN}$ | 150 V |
| Z-source | Chokes | $L_1, L_2$ | 0.2 mH |
| | Capacitors | $C_1, C_2$ | 0.2 mF |
| Output filter | Chokes | $L_f$ | 100 μH |
| | Capacitors | $C_f$ | 50 μF |
| Load (resistance) | | $R_0$ | 60 Ω |
| PWM frequency carrier | | $1/T$ | 10 kHz |

Further research on the Z-inverter confirmed the theoretical analysis. Little error arose from assumed values of loads $R_0$ and parameters $L=L_1=L_2$ and $C=C_1=C_2$ of Z-source. In addition, a harmonic distortion coefficient in input voltage $v_{OUT}$, that was calculated every time, has never exceeded 3–4%. The research also showed that transitions in the Z-inverter, resulting from changes of load and factors $M$ and $D$ (in open control system), are relatively fast. Furthermore, they inspired many researchers to elaborate and investigate other Z-source inverters, including NPC inverters and four-wire inverters.

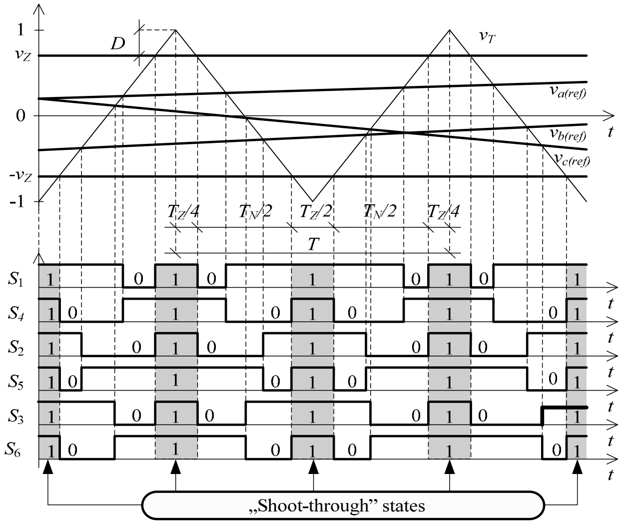



**Figure 3.45.** Control algorithms implementing of the Z-inverter

### 3.4.2. Three-level and Four-wire Inverters with Z-source

In the three-level Z-NPC inverter [100, 101], presented in Figure 3.47, two

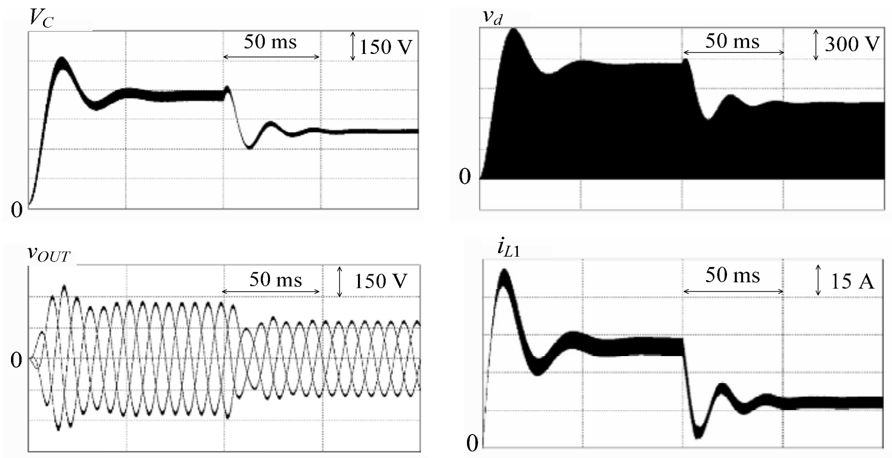



**Figure 3.46.** Selected currents and voltages waveform before and after change of coefficient $D = 0.43 \rightarrow 0.35$ for the time $t = 10$ ms ($M = 0.48$)

Z-sources with input voltage $V_{IN1}$ and $V_{IN2}$ without common point were applied. That allows joint as well as separated voltage $u_{d1}$ and $u_{d2}$ control. The possibility is explained by equivalent diagrams of the Z-NPC inverter in "*shoot-through*" states, shown in Figure 3.48.

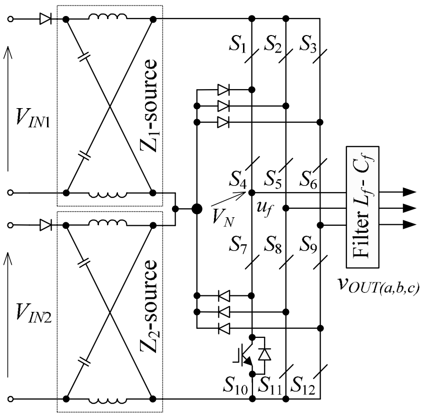



**Figure 3.47.** Three-level Z-NPC inverter with two Z-sources

In the "*shoot-through*" state of upper branches (Figure 3.48a), the switches $S_1$–$S_6$ & $S_7$–$S_9$ are turn-on, whereas "*shoot-through*" states of lower branches (Figure 3.48b) switches $S_7$–$S_{12}$ and $S_4$–$S_6$. These two states, together with the duration $T_{Z1}$ and $T_{Z2}$ in pulse period $T$ (Figure 3.49) cause an average voltage increase to $V_C$ and maximal $v_d^*$ up to a value on output of the upper $Z_1$-source

$$V_{C1} = V_{IN1} \cdot \frac{1 - D_1}{1 - 2 \cdot D_1}, \qquad v_{d1}^* = V_{IN1} \cdot \frac{1}{1 - 2 \cdot D_1} \tag{3.24}$$

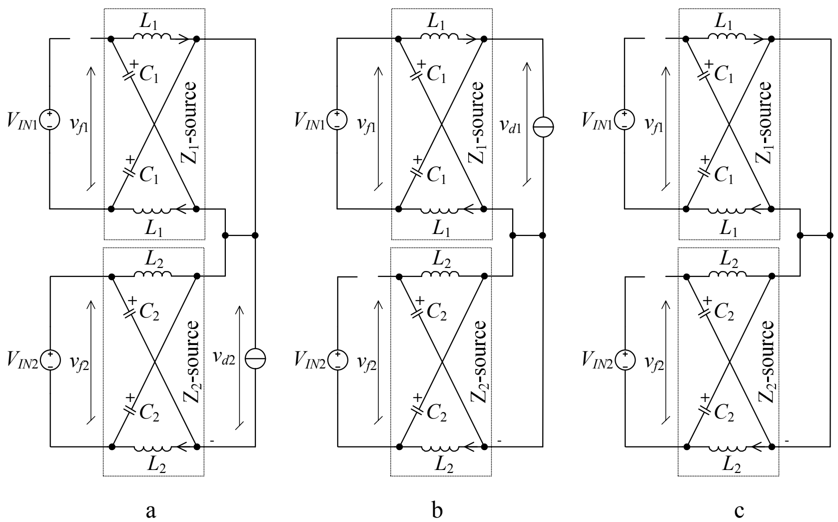



**Figure 3.48.** Equivalent schemes of the Z-NPC inverter (Figure 3.47) in states of "*shoot-through*": **a** upper branches; **b** lower branches; **c** full

on output o lower $Z_2$-source

$$V_{C2} = V_{IN2} \cdot \frac{1-D_2}{1-2\cdot D_2}, \qquad v_{d2}^{*} = V_{IN2} \cdot \frac{1}{1-2\cdot D_2} \tag{3.25}$$

where $D_1=T_{Z1}/T$ and $D_2=T_{Z2}/T$ are "*shoot-through*" coefficients of upper and lower branches.

Physical process occurring and following deduced Equations 3.24 and 3.25, are analogous to those in a basic system of Z-inverter (Figure 3.42) and Equations 3.19 and 3.20. It remains unchanged by total "*shoot-through*" state (Figure 3.48c), occurring when short circuits of upper and bottom branched happen simultaneously at the time $T_{Z2}$–$\Delta T_Z$ (Figure 3.49).

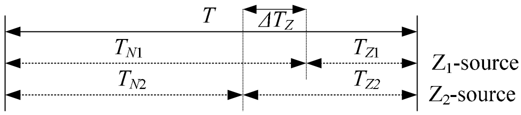



**Figure 3.49.** Exemplary schedule of branches "*shoot-through*" in T period

When considering the Z-NPC inverter presented in Figure 3.47, which is supplied by a source of different voltage $V_{IN1} \neq V_{IN2}$ and controlled on the basis of sinusoidal PWM, and taking into account Equations 3.24 and 3.25, input voltage peak-to-peak value can be determined on the basis of the following dependence

$$V_{OUT(p/p)} = M_1 \cdot \frac{V_{IN1}}{1-2\cdot D_1} + M_2 \cdot \frac{V_{IN2}}{1-2\cdot D_2} \tag{3.26}$$

where $M_1$ and $M_2$ are modulation indexes for positive and negative halves of the output voltage. Hence, if the following condition is true

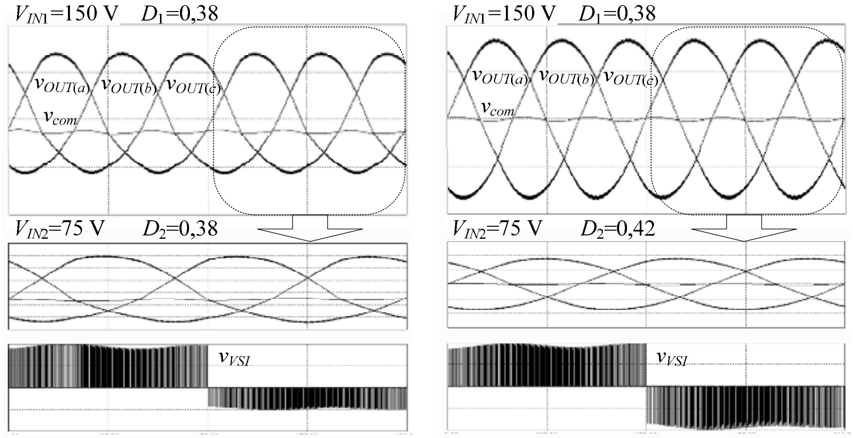
$$M_1 \cdot \frac{V_{IN1}}{1-2\cdot D_1} = M_2 \cdot \frac{V_{IN2}}{1-2\cdot D_2} \tag{3.27}$$

then the voltage $v_{com}$ between reference potential $V_N$ (Figure 3.47) and star point in symmetrical load three-phase load (*e.g.*, DC-offset) is as follows

$$v_{com} = V_N - \left(v_{OUT(a)} + v_{OUT(b)} + v_{OUT(c)}\right) = 0 \tag{3.28}$$

If Equation 3.27 is not satisfied, then additional distortion in the output voltage occurs that is mainly related to even harmonics.

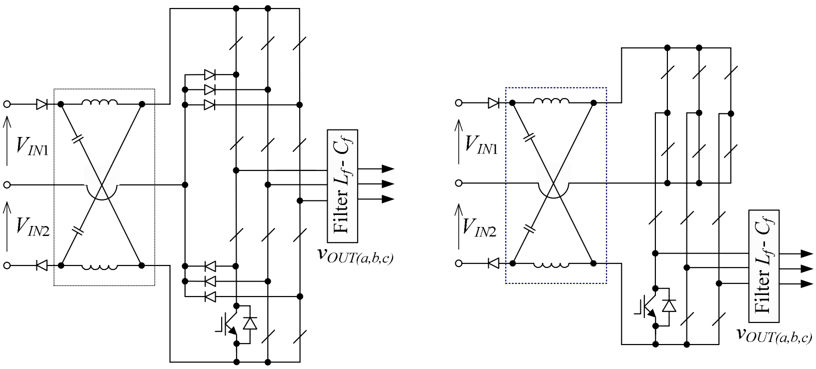
Taking into account Equation 3.27, elimination of output voltage distortion and DC-offset, is possible through: (a) selection of different modulation index's $M_1$ and $M_2$ for the positive and negative half; (b) selection of different "*soot-through*" coefficient $D_1$ and $D_2$ for upper and bottom branches (Figure 3.50). In the second case maximal voltages in switches $S_1$–$S_6$ and $S_7$–$S_{12}$ (Figure 3.47) are equal. Obviously it opens up the possibility to join both procedures.




a b

**Figure 3.50**. The output phase voltages $v_{OUT(a,b,c)}$, DC-offset voltage $v_{com}$ and VSI output voltage $v_{VSI}$ (phase a): **a** in cases of differentiated $V_{IN1} \neq U_{IN2}$ and equal $D_1$=$D_2$; **b** after correction of coefficient $D_2$ on the basis of Equation 3.27

In [102], an alternative for Z-NPC inverter presented in Figure 3.47 topology of the NPC inverter using a single Z-source was proposed. Single Z-source was also tested in application for the DC-link Cascaded Inverter (DCLC). Both solutions that are presented in Figure 3.51 must unfortunately be supplied by two input voltage sources. This disadvantage is not of concern in the Z-NPC inverter, of which the fundamental topology is presented in Figure 3.52 [103]. This inverter can be realized when one of the Z-sources with high frequency transformer of transformation ratio 1:1 is applied. It needs to be supplied from only one source.




a b

**Figure 3.51**. Inverters with a single Z-source: **a** NPC; **b** DCLC

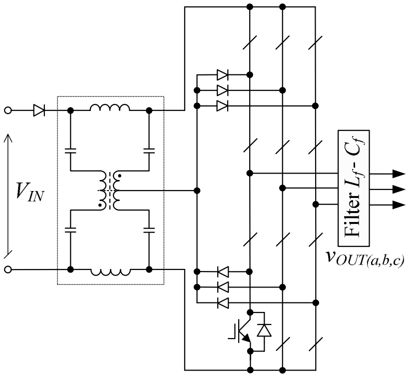



**Figure 3.52.** Z-NPC inverter supplied only from one source $V_{IN}$

Presented typologies of Z-inverters for obvious reasons do not consider all detailed solutions, such as other Z-source converters [104–111]. Many are also

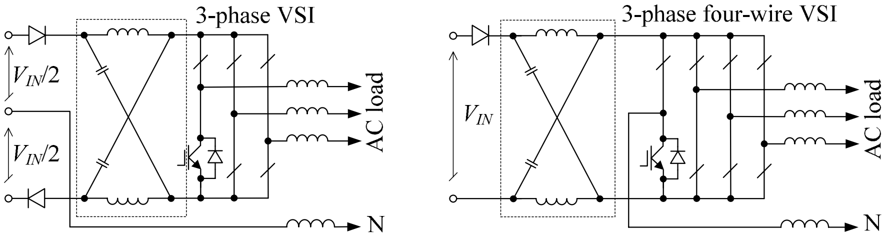



**Figure 3.53.** Simple topology of the voltage Z-inverters for four-wire systems

applications of these inverters, such as those for distributed generation [112–115]. Dedicated literature, however, has discussed only three-wire systems so far. There are practically no publications about application of Z-inverters in four-wire systems including significant research results. Under these circumstances it seems necessary to conduct further research of even such simple solutions as those presented in Figure 3.53.

## 3.5 Summary

The following text should only be considered as a comment. Also, in this manner, we would like to explain ourselves for disregarding many significant issues in the chapter. We are aware that relatively important issues discussed in the field of power electronics are not included here.

For example, we did not discuss diode and thyristor converters, including significant and large groups of conventional PECs with different applications, starting with generating systems, power transmission systems and local DC supply networks to improve the power quality for various technological applications. These PECs, however, have already been discussed by the major stream literature for many years, for example [41–43], or the newest trends focusing on issues of improvement of power quality factors [116, 117]. Valuable publications such as [118, 119] continue to be released. Similar circumstances apply with regard to DC/DC converters [120–122], various AC/AC matrix converters [123–129] or even systems of power factor correctors, which include Vienna converters [130–133] and resonant converters [134–137]. It seems difficult to present carefully yet comprehensively all the most significant issues and the most utilized solutions. Well written and valuable guides to the above-mentioned topics are available in books, in particular [1, 5, 11, 26, 28–30, 77].

The authors also hope that the content of the chapter, as well as referred literature, inspires the reader and initiates individual thinking about the issues discussed as well as the possibilities of applications of power electronics converters in smart energy networks.

## References

[1] Benda V, Gowar J, Grant DA, (1999) Power semiconductor devices – theory and applications. John Wiley, New York

[2] Baliga BJ, (2001) The future of power semiconductor device technology. IEEE Proceedings, vol.89, no.6:822–832

[3] Stemmler H, (2000) State of the art and future trends in high power electronics. IPEC Conference:4–13

[4] Van Wyk JD, (2000) Power electronics at the dawn of a new century – past achievements and future expectations, Proceedings CES/IEEE IPEMC Conference:9–20

[5] Bose B, (2006) Power electronics and motor drives: advances and trends. Academic Press, New York

[6] Strzelecki R, Supronowicz H, (2000) Power factor correction in AC supply systems and improving methods. (in Polish), Warsaw University of Technology Publishing House, Warsaw

[7] Hingorani N, Gyugyi L, (2000) Understanding FACTS: concepts and technology of flexible AC transmission systems. IEEE Press, New York

[8] Acha E, Agelidis VG, Anaya-Lara O, Miller TJE, (2002) Power electronic control in electrical systems. Newnes, Oxford

[9] Ghosh A, Ledwich G, (2002) Power quality enhancement using custom power devices. Kluwer Academic Publishers, Boston

[10] Moreno-Münoz A, (2007) Power quality: mitigation technologies in a distributed environment. Springer-Verlag, London

[11] Akagi H, Watanabe EH, Aredes M, (2007) Instantaneous power theory and applications to power conditioning. John Wiley-IEEE Press, New York

[12] Simões MG, Farret FA, (2004) Renewable energy systems: design and analysis with induction generators. CRC Press, Boca Raton

[13] Green TC, Arámburo H, (2005) Future technologies for a sustainable electricity system: the role of power electronics in future power systems. Cambridge University Press, Cambridge

[14] Jamasab T, Nuttall WJ, Michael G, Pollitt MG, (2006) Future electricity technologies and systems. Cambridge University Press, Cambridge

[15] Blaabjerg F, Chen Z, (2006) Power electronics for modern wind turbines (synthesis lectures on power electronics). Morgan & Claypool Publishers

[16] Maloney TJ, (2003) Modern industrial electronics. Prentice Hall, Upper Saddle River

[17] Irwin JD, (1997) The industrial electronics handbook. CRC Press LCL, Boca Raton

[18] Lee FC, VanWyk JD, Boroyevich D, Lu GQ, Liang Z, Barbosa P, (2002) Technology trends toward a system-in-a-module in power electronics. IEEE Circuits Syst. Mag., vol.2, no.4:4–22

[19] Van Wyk, JD, Lee FC, Boroyevich D, Zhenxian L, Kaiwei Y, (2003) A future approach to integration in power electronics systems. IECON Conference, vol.1: 1008–1019

[20] Chen R, Canales F, Yang B, Barbosa P, Van Wyk JD, Lee FC, (2003) Integration of electromagnetic passive components in DPS front-end DC/DC converter – a comparative study of different integration steps. APEC Conference, vol.2:1137–1142

[21] Rengang Ch, Canales F, Bo Y, Van Wyk, JD, (2005) Volumetric optimal design of passive integrated power electronics module (IPEM) for distributed power system (DPS) front-end DC/DC converter. IEEE Transactions on Industry Applications, vol.41, no.1:9–17

[22] Ghizoni D, Burgos R; Francis G, Ma X, Guo J, Solero L, Wang F, Boroyevich D, Cartes DA, (2005) Design and evaluation of a 33kW PEBB module for distributed power electronics conversion systems. IEEE PESC Conference:530–536

[23] Wang F, Rosado S, Boroyevich D, (2003) Open modular power electronics building blocks for utility power system controller applications. IEEE PESC Conference, vol.4:1792–1797

[24] Hornkamp M, Loddenkötter M, Münzer M, Simon O, Bruckmann M, (2001) EconoMAC the first all-in-one IGBT module for matrix converter. PCIM Conference:19–21

[25] Holtz J, (1994) Pulse-width modulation for electronic power conversion. IEEE Proceedings, vol.82, no.8:1194–1214

[26] Holmes DG, Lipo TA, (2003) Pulse width modulation for power converters: principles and practice. John Wiley & Sons, Hoboken

[27] Kaźmierkowski MP, Krishnan R, Blaabjerg F, (2002) Control in power converters. Selected problems. Academic Press, San Diego

[28] Skvarina TL, (2002) The power electronics handbook. CRC Press, Boca Raton

[29] Erickson RW, Maksimović D, (2001) Fundamentals of power electronics. Kluwer Academic Publisher, Norwell

[30] Rashid MH, (2006) Power electronics handbook: devices, circuits and applications. Academic Press, San Diego

[31] Tschirley S, (2007) Automatisierte messtechnische charakterisierung von 10kV integrierten gate-kommutierten thyristoren (IGCTs), Technische Universität Berlin
[32] De Doncker RW, (2002) Medium-voltage power electronic technologies for future decentralized power systems, IEEE Power Conversion Conference:927–932
[33] Bernet S, (2006) State of the art and developments of medium voltage converters – an overview. Przeglad Elektrotechniczny (Electrical Review), vol.82, no.5:1–10
[34] Bock B, (2005) Switching IGBTs parallel connection or with enlarged commutation inductance. Ph.D. Thesis, Ruhr-University, Bochum
[35] Thalheim J, (2003) Control strategies for balancing of series and parallel connected IGBT/diode modules. Ph.D. Thesis, Swiss Federal Institute of Technology, Zurich
[36] Abe Y, Matsubara K, Mochida T, Sasagawa K, Matsuse K, (2005) A novel method for loss reduction in high-voltage inverters. IEEE IAS Conference:1849–1854
[37] Wang H, Huang AQ, Chen B, Wang F, (2005) Development of a scalable power semiconductor switch (SPSS). IEEE no.0-7803-8975-1/05:347–353
[38] Zhang B, Huang AQ, Liu Y, Atcitty S, (2002) Performance of the new generation emitter turn-off (ETO) thyristor. IEEE IAS Conference:559–563
[39] Ozpineci B, Tolbert LM, (2003) Comparison of wide-bandgap semiconductors for power electronics applications**.** Oak Ridge National Laboratory Report for the U.S. Department Energy, Tennessee
[40] Sugawara Y, Takayama D, Asano K, Agarwal A, Ryu S, Palmour J, Ogata S, (2004) 12.7kV ultra high voltage SiC commutated gate turn-off thyristor: SICGT. International Symposium on Power Semiconductor Devices & ICs:365–368
[41] Takeuchi T, (1968) Theory of SCR circuit and application to motor control. Tokyo Electrical Engineering College Press
[42] Shepherd W, (1976) Thyristor control in AC circuits. Bradford University Press, Bradford
[43] Pelly B, (1971) Thyristor phase-controlled converters and cycloconverters. John Wiley & Sons Inc., New York
[44] Larsen AB, Murray JE, (1964) Optimization study of high power static inverters and converters. NASA Report CR-42, Thompson Ramo Wooldridge Inc., Cleveland
[45] Bedford BD, Hoft RG, (1965) Principles of inverter circuits. John Wiley & Sons Inc., New York
[46] Glazenko TA, Goncarenko RE, (1967) Semiconductors frequency converters in electric drives. (in Russian), Energia, Leningrad
[47] Bose BK, (2007) Need a switch? IEEE Industrial Electronics Magazine, vol.1, no.4:30–39
[48] Divan DM, (1986) The resonant DC link converter – a new concept in static power conversion. IEEE IAS Conference:758–766
[49] Azzopardi S, Vinassa JM, Woirgard E, Zardini C, Briat O, (2004) A systematic hard- and soft-switching performances evaluation of 1200V punchthrough IGBT structures, IEEE Transactions on Power Electronics, vol.19, no.1:231–241
[50] Hua G, Lee FC, (1993) An overall view of soft-switching techniques for PWM converters. European Power Electronics Journal, vol.3, no.1:39–50

[51] Cavalcanti MC, da Silva ER, Jacobina CB, Boroyevich D, Dong W, (2003) Comparative evaluation of losses in soft and hard-switched inverters. IEEE IAS Conference, vol.3:1912–1917
[52] Bloh J, De Doncker RW, (1999) Characterizing medium-voltage high-power devices under conventional and soft-switching conditions. IEEE IAS Conference, vol.1:373–378
[53] Citko T, Tunia H, Winiarski B, (2001) Resonant circuits in power electronics. (in Polish), Białystok University of Technology Publishing House, Białystok
[54] Chung HSH, Hui SYR, Tse KK, (1998) Reduction of power converter EMI emission using soft-switching technique. IEEE Transactions on Electromagnetic Compatibility, vol.40, no.3:282–287
[55] Schmidt S, Kempski A, (2004) Analysis of inverters operation with hard and soft switching in context of electromagnetic compatibility (EMC). Electrotechnical Rewiev, vol.80, no.6:620–3
[56] Liu KH, Oruganti R, Lee FC, (1985) Resonant switches – topologies and characteristics. IEEE PESC Conference:62–65
[57] Ericsen T, Hingorani N, Khersonsky Y, (2006) Power electronics and future marine electrical systems. IEEE Transactions on Industrial Applications, vol.42, no.1:155–163
[58] Ponnapan R, Leland J, Beam J, Fronista G, Weimer J, (1995) Effective cooling of MCT and IGBT using Venturi flow. Intersociety Energy Conversion Engineering Conference, paper AP-97
[59] Cao Y, Beam J, Donovan B, (1996) Air-cooling system for metal oxide semiconductor controlled thyristors employing miniature heat pipes. Journal of Thermophysics and Heat Transfer, vol.10:484–489
[60] Lee TY, (2000) Design optimization of an integrated liquid-cooled IGBT power module using CFD technique. IEEE Transactions on Components and Packaging Technologies, vol.23, no.1:50–60
[61] Goldman A, (2001) Magnetic components for power electronics. Kluwer Academic Publisher, Norwell
[62] Peterson GP, (1985) Heat pipe thermal control of electronic components**.** AIAA Thermophysics Conference:937–944
[63] Nabae A, Takahashi I, Akagi H, (1981) A new neutral point clamped PWM inverter. IEEE Transactions on Industry Applications, vol.17, no.5:518–523
[64] Andrejczak JM, Lescure M, (1987) High voltage converter promising technological developments. EPE Conference:1159–1162
[65] Salutari R, Sciutto G, (1992) A new multilevel PWM method: a theoretical analysis, IEEE Transactions on Power Electronics, vol.7, no.3:497–505
[66] Meynard TA, Foch H, (1992) Multilevel conversion: high voltage choppers and voltage source inverters. IEEE PESC Conference:397–403
[67] Fracchia M, Ghiara T, Marchesoni M, Mazzucchelli M, (1992) Optimized modulation techniques for the generalized N-level converter. IEEE PESC, vol.2:1205–1213
[68] Lai JS, Peng FZ, (1996) Multilevel converters – a new breed of power converters. IEEE Transactions on Industry Applications, vol.32, no.3:509–517

[69] Fazel SS, (2007) Investigation and comparison of multi-level converters for medium voltage applications. Ph.D. Thesis, Berlin Technical University

[70] Baker RH, (1975) Electric power converter. US Patent 3867643

[71] Monin VS, Vojtovic VS, (1977). Transistors inverter. USSR Patent 575751

[72] Kobzev AV, (1979) Multi-zone impulse modulation. (in Russian), Novosibirsk, Russia, Nauka

[73] Monin VS, (1986) Stabilize semiconductor converters. Energoatomizdat, Moscow

[74] Baker RH, (1980) High-voltage converter circuit. US Patent 4203151

[75] Tonkal VE, Mielniciuk LP, Novosielcev AV, Dychenko JI, (1981) Power electronics converters with modulation and high frequency link. (in Russian), Kiev, Ukraine, Naukova Dumka

[76] Rodriguez J, Lai JS, Peng FZ, (2002) Multilevel inverters: a survey of topologies, controls, and applications. IEEE Transactions on Industrial Electronics, vol.49, no.4:724–738

[77] Wu B, (2006) High power converters and AC drives. IEEE Press, John Wiley & Sons Inc., New Jersey

[78] Hartman M, (2006) Multilevel voltage inverters. (in Polish), Gdynia Maritime Publishing House, Gdynia

[79] Meynard TA, Foch H, Forest F, Turpin C, Richardeau F, (2002) Multi-cell converters: derived topologies. IEEE Transactions on Industrial Electronics, vol.49, no.5:978–987

[80] Manjrekar P, Steimer P, Lipo T, (2000) Hybrid multilevel power conversion system: a competitive solution for high-power applications. IEEE Transactions on Industry Application, vol.36, no.3:834–841

[81] Sirisukpraset S, (1999) Optimized harmonic stepped-waveform for multilevel inverter. M.Sc. Thesis, Virginia Polytechnic Institute and State University, Blacksburg

[82] Shakweh Y, Lewis EA, (1999) Assessment of medium voltage PWM VSI topologies for multi-megawatt variable speed drive applications. IEEE PESC Conference:965–970

[83] Shakweh Y, Lewis EA, (1999) The universal medium voltage adjustable speed drive. EPE Conference, CD-ROM:1–8

[84] Walker GR, (1999) Modulation and control of multilevel converters. Ph.D. Thesis, University of Queensland

[85] Shakweh Y, (2001) MV inverter stack topologies. Power Engineering Journal:139–149

[86] Gupta AK, Khambadkone AM, (2006) A space vector PWM scheme for multilevel inverters based on two-level space vector PWM. IEEE Transactions on Industrial Electronics, vol.53, no.5:1631–1639

[87] Pou J, (2002) Modulation and control of three-phase PWM multilevel converters. Ph.D. Thesis, Technical University of Catalonia, Barcelona

[88] Pou J, Pindado R, Boroyewich D, (2005) Voltage-balance limits in four-level diode-clamped converters with passive front ends. IEEE Transactions on Industrial Electronics, vol.52, no.1:190–196

[89] Adam GP, Finney SJ, Williams BW, (2007) Quasi two-level operation of a five-level inverter. Electrotechnical Review, vol.10:120–125

[90] Zymmer K, Zakrzewski Z, Strzelecki R, Szczepankowski P, (2008) 6kV four-level diode clamped inverter. Design and control. Electrotechnical Review, vol.84, no.4:4–9

[91] Lopatkin NN, Usachev EP, Zinoviev GS, Weiss H, (2006) Three-level rectifier fed four-level inverter for electric drivers. EPE-PEMC Conference:775–780

[92] Strzelecki R, Jarnut M, Kot E, Kempski A, Benysek G, (2003) Multilevel voltage source power quality conditioner. IEEE PESC Conference:1043–1048

[93] Xu Z, (2003) Advanced semiconductor device and topology for high power current source converter. Ph.D. Dissertation, Virginia Polytechnic Institute

[94] Peng FZ, (2003) Impedance source power converter. USA Patent no.2003107522, International Publication Number WO 03/107522 A1

[95] Peng FZ, (2003) Z-source inverter. IEEE Transactions on Industry Applications, vol.39, no.2:504–510

[96] Shen M, Wang J, Joseph A, Peng FZ, Tolbert LM, Adam DJ, (2004) Maximum constant boost control of the Z-source inverter. IAS Annual Meeting:142–147

[97] Miaosen S, Peng FZ, (2005) Operation modes and characteristics of the Z-source inverter with small inductance. IAS Annual Meeting, vol.2:1253–1260

[98] Loh PC, Vilathgamuwa DM, Gajanayake CJ, Lim YR, Teo CW, (2007) Transient modeling and analysis of pulse-width modulated Z-Source inverter. IEEE Transactions on Power Electronics, vol.22, no.2:498–507

[99] Rabi BJ, Arumugam R, (2005) Harmonics study and comparison of Z-source inverter with traditional inverters. American Journal of Applied Sciences, vol.2, no.10:1418–1426

[100] Loh PC, Blaabjerg F, Feng SY, Soon KN, (2006) Pulse-width modulated Z-source neutral-point-clamped inverter. Annual IEEE APEC Conference:940–947

[101] Strzelecki R, Wojciechowski D, Adamowicz M, Wilk A, Mosoń I, (2006) Three-level NPC Z-inverter. (in Polish), Electrotechnical Review, no.10:54–60

[102] Loh PC, Lim SW, Gao F, Blaabjerg F, (2007) Three-level Z-Source inverters using a single LC impedance network. IEEE Transactions on Power Electronics, vol.22, no.2:706–711

[103] Strzelecki R, Adamowicz M, Wojciechowski D, (2007) Buck-boost inverters with symmetrical passive four-terminal networks. Compatibility in Power Electronics Conference, CD-ROM:1–9

[104] Loh PC, Gao F, Blaabjerg F, (2006) Topological and modulation design of three-level Z-source inverters, IPEMC Conference, CD-ROM:1–5

[105] Strzelecki R, Wojciechowski D, Adamowicz M, (2006) Multilevel, multiphase inverter supplying by many sources, especially different voltage and non-connection sources. Polish Patent P379977

[106] Strzelecki R, Wojciechowski D, Adamowicz M, (2006) Principle of symmetrization of the output voltage of the multilevel inverter supplying by many different voltage sources, especially with four-terminal impedance networks. Polish Patent P379978

[107] Gao F, Loh PC, Blaabjerg F, Teodorescu R, Vilathgamuwa DM, (2007) Five-level Z-source neutral point-clamped inverter. IEEE PEDS Conference:1054–1061

[108] Xinping D, Zhaoming Q, Yeyuane X, Zhengyu L, (2005) Three phase Z-source rectifier. IEEE PESC Conference:494–500
[109] Xinping D, Zhaoming Q, Yeyuan X, Peng FZ, (2006) A novel ZVS Z-source rectifier. IEEE APEC Conference, CD-ROM:1–6
[110] Fang FX, (2006) Three-phase Z-source AC/AC converter. PEMC Conference:621–624
[111] Chaohua TYZ, Shaojun X, (2007) Single-phase four switches Z-source AC/AC converter. IEEE APEC Conference:621–625
[112] Gajanayake CJ, Vilathgamuwa DM, Loh PC, (2006) Modeling and design of multi-loop closed loop controller for Z-source inverter for distributed generation. IEEE PESC Conference, CD-ROM:1–7
[113] Gajanayake CJ, Vilathgamuwa DM, Loh PC, (2007) Development of a comprehensive model and a multiloop controller for Z-source inverter DGs systems**.** IEEE Transactions on Industrial Electronics, vol.54, no.4:2352–2359
[114] Gajanayake CJ, Vilathgamuwa DM, Loh PC, Blaabjerg F, (2007) Z-source inverter based power quality compensator with enhanced ride-through capability. IEEE IAS Conference:955–962
[115] Po Xu P, Zhang X, Zhang CW, Cao RX, Chang L, (2006) Study of Z-source inverter for grid-connected PV systems. IEEE PESC Conference, CD-ROM:1–5
[116] Pejović P, (2007) Three-phase diode rectifiers with low harmonics. Current injection methods. Springer Science & Business Media, LLC, New York
[117] Paice DP, (1999) Power electronics converter harmonics: multipulse methods for clean power. IEEE Press, Wiley & Sons Inc., New York
[118] Choi S, Enjeti PN, Pitel IJ, (1996) Polyphase transformer arrangements with reduced kVA capacities for harmonic current reduction in rectifier type utility interface. IEEE Transactions on Power Electronics, vol.11, no.5:680–689
[119] Singh B, Garg V, Bhuvaneswari G, (2007) A novel T-connected autotransformer-based 18-pulse AC/DC converter for harmonic mitigation in adjustable-speed induction-motor drives. IEEE Transactions on Industrial Electronics, vol.54, no.5:2500–2511
[120] Luo FL, Ye H, (2004) Advances DC/DC converters. CRC Press, New York
[121] Mazumder SK, (2001) Nonlinear analysis and control of standalone, parallel DC/DC and parallel multi-phase PWM converters. Ph.D. Thesis, Virginia Polytechnic Institute and State University, Blacksburg
[122] Bo Y, (2003) Topology investigation for front end DC/DC power conversion for distributed power system. Ph.D. Thesis, Virginia Polytechnic Institute and State University, Blacksburg
[123] Gyugyi L, Pelly BR, (1976) Static frequency changers. John Wiley, New York
[124] Wheeler P, Rodriguez J, Clare JC, (2002) Matrix converter: a technology revived. IEEE Transactions on Industrial Electronics, vol.49, no.2:276–288
[125] Fedyczak Z, Strzelecki R, (1997) Power electronics agreement for AC power control. (in Polish), Adam Marszalek Publishing House, Toruń
[126] Ratanapanachote S, (2004) Applications of an electronics transformer in a power distribution system, Ph.D. Thesis, Texas A&M University

[127] Petry CA, Fagundes JC, Barbi I, (2005) Direct AC/AC converters using switching modules. COBEP Conference:94–99

[128] Herrero LC, de Pablo S, Martin F, Ruiz JM, Gonzalez JM, (2007) Comparative analysis of the techniques of current commutation in matrix converters. IEEE ISIE Symposium:521–526

[129] Casadei D, Serra G, Tani A, Zarri L, (2006) A review on matrix converters. Electrotechnical Review, vol.82, no.2:15–25

[130] Bartosa PM, (2002) Three-phase power factor correction circuits for low-cost distributed power systems. Ph.D. Thesis, Virginia Polytechnic Institute and State University, Blacksburg

[131] Cichowlas M, (2004) PWM rectifier with active filtering. Ph.D. Thesis, Warsaw University of Technology

[132] Kwak S, Toliyat HA, (2006) Current-source-rectifier topologies for sinusoidal supply current: theoretical studies and analyses. IEEE Transactions on Industrial Electronics, vol.53, no.3:984–987

[133] Kolar JW, Drofenik U, Zach FC, (1999) VIENNA rectifier II – a novel single-stage high-frequency isolated three-phase PWM rectifier system. IEEE Transactions on Industrial Electronics, vol.46, no.4:674–691

[134] Kazimierczuk MK, Czarkowski D, (1995) Resonant power converters. John Wiley & Sons Inc., New York

[135] Vlatkovic V, Borojevic D, Lee FC, (1994) Soft-transition three-phase PWM conversion technology. IEEE PESC Conference:79–84

[136] Pfisterer HJ, Spath H, (2000) Switching behavior of an auxiliary resonant commutated pole (ARCP) converter. IEEE International Power Electronics Congress CIEP:359–364

[137] Lakshminarasamma N, Swaminathan B, Ramanarayanan V, (2004) A unified model for the ZVS DC/DC converters with active clamp. IEEE PESC, vol.3:2441–2447