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FUNDAMENTALS OF POWER ELECTRONICS

This Russian book is recommended by Guidance committee of Ministry of Education of Russian Federation as a textbook for students of specialization of "Power Electronics"

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Zinoviev G.S. Fundamentals of power electronics: Textbook. – Novosibirsk, 2003.
The present textbook pursuant to a principle " three in one " is structured on three levels of depth and accessibility of presentation of a material. Two uppermost levels are intended for "experts" in the power electronics engineering (magistrate and engineering branch of training). The lower level is intended for "no experts" on a power electronics engineering, for whom this course is common technical course (students of electro technical engineering, electro power engineering, radio engineering). A dominating method of the analysis of power quality factors of power electronics converters in the textbook is by an authoring direct computational method of power processes. This method does not require the solution of differential equations.
All chapters of the textbook are supplied with the review questions (tests in electronic version of the textbook) and problems.

ISBN

FOREWORD

Hello, young and unknown tribe
A. S. Pushkin

In power industry there are dc and ac power sources. Overwhelming majority of electrical energy for common mains is generated by 3-phase synchronous generators with standard voltage level (that may differ from country to country) and frequency level (50 Hz – in Russia and in countries of West Europe, 60 Hz – in USA, Canada, and in most of the countries of Central and South America and others). In autonomous power supplies to produce electrical energy asynchronous generators are utilized, and in some cases - special electrical machines with raised frequency (400, 800, 1200 Hz and greater).

Primary dc power sources are generators, accumulators, solar and thermal batteries and MHD-generators.

Because of two types of power sources there are two types of power consumers – ac current ones (single- and multiphase) and dc or rippling current ones.

It is necessary achieve the most effective utilization of electrical energy that is generated with constant parameters, but different consumers require electrical energy with nonstandard parameters: frequency, adjustable voltage, phase number that may differ from source parameters. That is why electrical energy converters are of utmost importance. Nowadays in developed countries almost 40% of electrical energy is converted before utilization. This situation as the baking of bred: from two cereal crops – wheat and rye – a lot of sorts of bread could be baked; among these types customer selects desired ones.

Power electronics is a branch of power industry nowadays, which production is vitally important for all others electrotechnical and electroenergetical branches of industry. Following data is a confirmation of these statements [1-34]:

1. Annual worldwide power consumption is $(8\ldots12)\cdot10^{12}$ Kw·h. Annual costs of production of electrical energy is 400…500 milliards of dollars, at that 72…78 milliards of dollars are spent on real losses of generators, transmitters and loads.
2. Primary consumers of electrical energy nowadays are electromotor drives of different purposes (51%), lighting (19%), heating/cooling (16%), and telecommunications (14%).
3. Today less that 25% of electrical energy is optimally utilized around the world (from losses minimization point of view). But this figure is achieved by utilization of high-performance methods of controlled power conversion from main power to object control energy. At the heart of these methods lies the utilization of high-performance electrical energy converters (power electronics devices).

Below there is enumeration of areas of application of power electronic devices and evaluation of economical result of their application [34]:

1) performance control tools for technology circuits of TES and HES (30…40 % decrease in house energy consumption, decrease in prime cost of 1KW·h of electrical energy on 7…10%).
2) controlled industrial electromotor drives (60…70 milliards of dollars);
3) municipal economy, outdoor lightning by sodium high pressure lamp, indoor lightning by compact luminescent lamp with electronic starting regulators (90…120 milliards of dollars);
4) consumer electronics: vacuum cleaners, refrigerators, lamps, washing machines, induction heaters (only for 10% of refrigerators 1 milliard of dollars over 3 year period);
5) automobile electronics (29 milliards of dollars and fuel economy of 10%);
6) increase in power source efficiency (2…3 milliards of dollars).

Industrial electromotor drive controlled by semiconductor power converter saves up to 40% of electrical energy comparing to uncontrolled one. Nowadays the share of controlled electromotor drives in worldwide technologies not greater than 40%. The utilization of controlled electromotor drive in those areas where it is possible should safe up to 72 milliard oa dollars annually.

Each year for lightning purposes approximately 10 milliards lamps and 500 millions of luminescent lamps are sold. Luminescent lamp with electronic ballast is five times more effective than incandescent lamp, its life time 10 times longer and the economy during the life time of the lamp is 30 dollars; economy potential 119 milliard of dollars. High pressure sodium lamp with electronic starting regulator is 2.5 times more effective than mercury lamps which are used for outdoor lightning. The utilization of sodium lamps for this purpose result is economy of 200 milliards of dollars.

Increase of efficiency of secondary power supplies by utilization of pulse conversion methods has economy potential of 2.5 milliards of dollars.

Utilization of power electronic devices in automobile industry (e.g. in breaking systems, motor control, starter-generator system) only at 10% of fuel economy could save up to 29 milliards of dollars relative to 500 millions of automobiles.

Application of semiconductor converters for performance control of thermal electric station (TES) technology circuits decreases home energy consumption on 30…40% that results in prime cost decrement of 1 KW·h up to 1…10%.

Main consumer of power electronics devices is consumer electronics. High volume of production of household devices defines high reserve of power saving despite small level of power consumption by single device.

Most power-consuming household devices are air conditioners, induction heaters, washing machines, refrigerators and lamps. The utilization of regulated compressor in refrigerators will allow the economy up to 40% of electrical energy and will result in price decrement of whole refrigerator (economy for 200 liter refrigerators is 70 dollars a year.

Washing machine with intellectual power regulator could save 60% of water. Induction heater has efficiency greater than 90% instead of 50% for usual electrical heater. Indoor lightning device with 20W luminescent lamp could be used instead of 100W incandescent lamp.

The values of specific cost of power electronic devices lie within 0.008…2 dollars/watt band. This value is inversely proportional to the device power. Potential Russian market of power electronic devices for next 10 years could be evaluated from 4 to 6 milliards of dollars. In worldwide practice the cost of semiconductor devices is one third of overall power electronics system cost.

As IEEE PELS states (International power electronics society) there is "the Renaissance" in power electronics in the West now; in next years there is a necessity of approximately 100 thousand of new specialists in area of power electronics. So the actuality of supplying study process by contemporary literature is evident.

Overall spreading of different electrical devices equipped with different power electronics devices causes two problems in study process in related area. Firstly there is need of convenient textbooks on power electronics for engineers and electricians for whom this discipline is general only. These specialists should be acquainted with power electronic devices that are presented in devices they deal with. Secondly, there is need of textbooks for specialists oriented on development end research of power electronic devices. Here are the development tasks (engineering type) and research tasks of new devices and their operation modes (scientific "Master's" type). This again proves the necessity of differentiated (three-level) structuring of study material on Power Electronics within single textbook frameworks.

Today there is little of study literature for "nonspecialists" and most of this literature is written 10 years ago [1-6], except last "sectoral" textbook for railway transport universities [7]. This is not enough for such intensively developing area as power electronics. New high effective semiconductor devices (GTO –thyristors, IGBT – transistors, "intellectual" power modules) led to new technical solutions for power electronics devices. New control algorithms for these devices aren't described in study literature yet.

There is very little of study literature for "specialists" on power electronics that corresponds to contemporary requirements in Russia, because textbooks [8-11] and reference books [12-14] printed in USSR are much away of today's problems than books for "nonspecialists".

NSTU published study [15-20] and scientific literature [21] supported by computer courses of laboratory works [22-25]. A series of textbooks and methodical guidance books were written at the department of power electronics [26-33]. Today these books are not enough.

The fundamental of unified approach to study material exposition in this textbook is the fact that depth of the material is defined by the degree of assumption in mathematical model of system under study. By the beginning of explanation from idealized models it is easy to obtain simple analytical expressions for basic systems, which knowledge is necessary for both specialists or no specialist. Then together with complication of mathematical model the more deep theories of model under study are introduced. On such a way of ascension from common simple things to special complicated ones each student reaches his own peak.

Used approach to study the material reckons on three levels of preparation ("three in one"). Material marked with vertical line and with stars in the name of the section is not meant for "nonspecialists"; among all the analytical expressions they should study only those highlighted by bold style. Sections with two stars in its name are for "specialists" of second level of preparation.

The selection of the material for this textbook and its expounding are performed accordingly to the state educational standards for correspondent specializations. If one range the material within single scale then the student on the first level of study has to “know” the material, on the second – “know how” and on the third – “master the material”.

The structure of this book is so: first chapter introduces the conception of power electronics devices analysis; there is summary of methodology, description of its application to valve converters, criteria of quality of electromagnetic processes and devices. There is a set of basic cells and their principal circuits relative to valve converters; consideration of methods of power quality factors calculation with detailed description of direct method; short review of accessible software sets for simulation of power electronic devices; exact description of Parus-Pargraph software that is developed at the department of industrial electronics at NSTU and is a part of laboratory work on “Fundamentals of Power Electronics” course.

Sections 2 and 3 introduce system analysis of rectifier and dependent inverter, i.e. the system of conversion of ac current to dc one and vice versa. Sections 2.1 concretize the procedure of analysis for rectifier, and whole the second chapter is devoted to the analysis of basicc rectifiers composed of ideal elements.

Third chapter is dedicated to the analysis of the processes in generalized naturally commutated valve converter (rectifier and source inverter) considering real circuit parameters, i.e. the chapter aims to obtain general laws for controlled rectification and inverting processes. Overall study of this oldest type of valve converters is a prototype of system analysis of other types of valve converters that have less detailed description within this book.

Chapter 4 presents model example for course projecting of a rectifier. The same method should be used to project other types of converters on the basis of studied circuits. The novelty in this course projecting is that one can verify the results by mathematical simulation of projected converter by ParGraph–Parus software.

Chapter 5 presents the results of development of direct analytical (solving of differential equations is avoided) methods of power electronic devices calculations that are equivalented by mathematical models of any order. Preliminary acquaintance with these models in section 1.5 is enough to understand the material of second, third, and fourth chapters but not enough to expound following material.

Sixth chapter is a brand new one for study literature on the power electronics is dedicated to the electromagnetical compatibility problem of valve converters with the mains and environment.

Chapter 7 presents dc-dc converters for two assumption levels of mathematical models. This type of the converters as a previous one with dc output is a fundamental for building of a converter with ac output if ac current is considered as periodically reversed dc current as a constructive methodology. Chapter 8 tells about dc-ac converters (autonomous inverters), chapter 9 – ac-ac converters without output frequency regulation (ac voltage regulators); chapter 10 – ac current regulators with variable output frequency.

Chapter 11 presents quite new study material on power electronics about specific types of the converters – compensators of nonactive components of apparent power, which are used to improve electromagnetic compatibility of any nonlinear power consumers.

General classical principles of valve control of all types of power electronic devices as well as new control tendencies (common vector, intellectual control) are considered in Chapter 12.

Chapter 13 collects general modifications of basic circuits of different types of the converters.

List of references contains utilized and recommended for further study books and articles.

Textbook is equipped with subject index with special power electronics terms (highlighter in italics) and short English-Russian dictionary that expands the same dictionary published by Power Electronics Society.

As a whole the material of this textbook reflects the 30-year author’s experience of teaching of power electronics course at Novosibirsk State Technical University.

The author who is a fan of A.S. Pushkin noted his 200-year jubilee (that concur with the time when this textbook was planed). Accordingly to the phrase of N.V. Gogol that “Pushkin is a extra ordinal phenomenon and may by the only phenomenon of Russian spirit: he is Russian person in its development that may occur 200 years ago” the author used the citations of A.S. Pushkin as the epigraphs to the chapters.

Considering the novelty of the experience of type of building a material (“3 in 1”) author will gratefully consider all the remarks and proposals and most constructive of them will be introduced in next publications. Author expresses his thanks to post-graduates M. Ganin, E. Levin, A. Obuhov and students A. Zimin, I. Proscurin for all their help in calculations, picture arrangements; and special thank to M. Gnatenko for his help in simulation and continuous perfection of ParGraph software and L.A. Laricheva for printing of manuscript.

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Russia, Novosibirsk ,2003 Prof. Zinoviev G. S.

THE FOREWORD TO THE ENGLISH TRANSLATION OF THE HARD TEXTBOOK

The western community of the advanced countries is characterized by occurrence last 10-15 years of significant number of books (textbooks and monographs) on power electronics. Recently we had an opportunity to get acquainted with four textbooks on power electronics in English [1-5] and one textbook in German [6]. As a whole for these books orientation to practical application of devices of power electronics (DPE) is characteristic. Therefore and the statement of a teaching material is directed on studying of designing of typical circuits DPE.

In Russia last 15 years from the moment of the beginning of reorganization ("perestroika") textbooks on power electronics (more in detail see the foreword to the Russian edition) practically were not issued. At the same time there was a reorganization of education with single-level on three-level structure (the bachelor, the engineer - the diploma expert, the master), not provided with the educational literature. Therefore, since the end of 90th years of the last century, we have taken steps on creation of a new complex of textbooks in the field of power electronics by a principle " 3 in 1 ". It is based on the concept of differentiation of a level of complexity and availability of a statement of a material on three levels: initial (the bachelor), intermediate (the engineer) and advanced (master). But as against externally similar technologies " n in 1 ", foreign languages used at studying, here these levels are not combined separately, and as though enclosed each other (Russian nested doll). It allows to not limit interest to studying a discipline at students, and smoothly to pass " borders of the unit " (considered as an obligatory minimum level of knowledge) "upwards up" to a desirable or achievable level. It provides a natural variety necessary for a society in set of the specialists in power electronics.

At initial studying a discipline of power electronics the greater attention is given us to the general problems and laws of transformation of electric energy with the help of power semi-conductor devices. It allows preparing trainees for the profound research if necessary known devices of power electronics, and in a consequence, the most advanced among them, and to synthesis of new devices (a level of the master). To the same purposes wide use of intelligence of the computer in the electronic version of the textbook and for the purposes of animation, research during performance of laboratory works, testing answers also, to the help at structural synthesis by creation of databases and expert systems [7-12]. Work in this direction intensively develops with attraction of students and post-graduated students for its performance. We hope and for external cooperation with the purpose of that this attempt of outline translation of our textbook on the English language is undertaken. The English variant of our educational literature is necessary and for foreign students (Korea, China, Germany, Poland), visitors last years on our faculty.

I express the big gratitude to the former students Anton Shupletsov and Sergey Skrypnik for the big work on "voluntary" amateur (not professional) to translation of my textbook on the English language. We with gratitude shall accept all constructive remarks and offers under the contents of the textbook. Ours e-mail: genstep@mail.ru.

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1.SCIENTIFIC-TECHNICAL AND METHODIC PRINCIPLES OF INVESTIGATION OF POWER ELECTRONICS CONVERTERS

There is no an aim before me
My heart is empty, mind is idle…
Oh, how many wonderful discoveries
A spirit of the enlightenment prepares to us…
A.S.Pushkin

1.1 A methodology of a system approach to analysis of power electronics converters

Problems of studying of power electronic principles, first of all, lean on analysis of basic types of these devices, that is on establishment properties of devices in the function of their parameters. A classic methodology of studying inductive character supposes a motion from private to common, from simple to complicated. But as studying devices are complicated there is a necessity of regulation «good sense» attached to analysis, allowing to make an investigation of different complicated devices given purpose the same type on approach and effective on result. Such approach to investigation, intensively developed in some previous decades, was called the system approach. It is characterized by the next indications:

1) by establishment the boundaries of the investigated system given purpose as the whole, that is by apportionment a system from environment considered as a subsystem;
2) by determination the aims of the system, the criterions of quality of the functioning and the methods of the calculations;
3) by decomposition a system at components or subsystems which are considered as subsystems at the more low stage of hierarchy too, so, as the investigated system is a part of a metasystem;
4) by studying the system in all aspects, demanded by purpose, having a special aim, taking into account all the meaningful bonds both between parts of the system one stage, and between different stages.

A former classic before approach to investigation was based at what the properties of the whole (system) to a great extend are determined by properties of the components (subsystems) of the whole. System approach is based at another conception: system is not determined simply by a totality of elements and does not come to it, and, inside out, elements are determined by the whole, within the bounds of which it gain the functional purpose; attached to this a system gains new properties in the whole which are absent at elements of system.

Conformably to power electronic devices, studied at the course, four mentioned principles of system approach consist in the next.

At the first, the device of converting electric energy from one sort to another is considered in not itself, and in totality with a power supply at input and with a load (consumer) at output. This triad constitutes a system for investigation. Besides, all kinds of semi-conducting devices of converting electric energy in according to the purpose are exposed.

At the second, the necessary admission of quality criterions of creation and functioning power electronic devices (within the bounds of given course for energy engineers- of energetic quality criterions of devices and of the operating conditions) is determined and existing methods of the calculation are considered.

At the third, the decomposition of power electronic devices is made for the simplification the analysis at the functional and structural stages. In the whole case any converting device must realize a totality of the next functional operations:

-Proper of converting the kind of current;

-Of regulation the parameters of converting energy (of a constant component in the networks of direct current, of the first harmonic in the networks of alternating current);

-Of concordance the stages of voltage of the power supply and the load of a converter;

-Of potential insulation (if it is necessary) of a power supply and a load;

-Of electro-magnetic compatibility of a converter with a power supply and a load.

Two first operations in power electronic devices are realized by the use of semi-conducting control valves, the next two- by the use of a transformer at input, inside or at output of the device, and the last operation- by the use of passive (LC) or active (a control generation of voltage or current of required form) filters.

A structural decomposition of power electronic devices will be executed at two levels. A complicated converting system is divided into a totality of elementary basic cells, characterized by a single-valuation of converting a kind of electric energy, at the upper level (for example, alternating current- direct current). Elementary basic converters are considered as a totality of a transformer, a valve group, filters, a system of control at the lower level.

At the fourth, a principle of a system approach to an investigation of power electronic devices in according to a purpose, having a special aim, of a course will be realized here only in energetic aspect. Attached to this there will be three levels of analysis of electro-magnetic processes in investigated devices in according to three levels of assumption attached to analysis.

* At the first level of analysis all the elements of a converter are ideal (without losses), a maim is a source of infinite power (without losses inside of source, too), a load is ideal, too. A procedure of analysis is elementary.
* At the second level of analysis it is taking account that elements of converting device and maim have a real parameters, a load of converter is ideal. A procedure of analysis is simple and analytic, too.
* At the third level of analysis all the elements of triad: a maim – converter – a load are replaced by models with real parameters of elements. A procedure of analysis becomes more difficult, and not always one can work without means of computer engineering.

Such approach allows to develop a power of the analysis as understanding of the course increases and investigations become deeper, providing enclosures of the results from an analysis of the low levels as private cases into results of the more high level analysis. This fact permits simply to watch for the influence of real parameters of separated system elements to characteristics of the system.

There is built an account of material in according to this procedure of system approach, showed schematically at the table 1.1.1. A sequence of studying the course is determined by maintenance of the cells of the table, scanned from left to right on the lines.

There are considered these operations in the third column of the table conformably to studied objects – power electronic devices in according to the four above-mentioned levels of the system analysis. That demanded preliminary information, on which this operations lean, is presented in the second column of the table. The received results from studying operations at the corresponding stage of the system analysis, mastering of which is controlled by control questions and exercises to each pattern., are presented in the fourth column of the table. Control questions and exercises, marked by asterisks, have a higher difficulty and are intended, as a rule, «for specialists» in power electronics (third and fourth levels of account of the material, in according to a conventional qualification of the training people, brought at the preface).

T A B L E 1.1.1

Characteristic of the stage				
A number of the stage	Demanded knowledge at the stages	Maintenance of the stages of system analysis	Results of the analysis at the stage	A control of the studying
1	Methodology of the system analysis	Formulation of the problem	A common aspect of the system, achieving one`s object, the subsystem	A control questions, exercises
2	Systems of the quality criterions of processes and devices and methods of it calculation	Discomposi-tion of the system, aim and analysis of criterions of the achievement	A tree of the aims, chosen criterions	A control questions, exercises
3	A hierarchy of mathematic models of elements. Procedures of the analysis	A determination of admission of the basic cells of converting systems and the analysis	Properties of the basic cells. Recommendation s about the application	A control questions, exercises
4	Methods of the mathematic-tic modeling of the complicated systems	Composition of properties of the system from proper-ties of the cells	Properties of the whole system. Recommenda-tions about the composition	A control questions, exercises, an example of a calculation

1.2. Energetic quality indexes of converting energy at valve converters

Energetic effectiveness of converting electric energy in power electronic devices is characterized by energetic indexes of electromagnetic elements and device in the whole, determination of which is **an aim of this part.**

1.2.1. Energetic quality indexes of electromagnetic processes

The most important from these indexes are the next.

1. *A coefficients of converting of the device on the voltage and current* correspondingly

$$K_{v.c.} = \frac{U_{out.c}}{U_{in.c}}, K_{c.c.} = \frac{I_{out.c}}{I_{in.c}} \tag{1.2.1}$$

They are determined at conditions, corresponding to maximal voltage at output of the converter, which is attached to the absence the regulation, for the useful components of the voltage and current. The useful components, transferring the active power, are, as a rule, first harmonics of the voltage and current in the networks of alternating current, and in the networks of direct current – the average values of the voltage and current.

2. *A coefficient of distortion of the current* (the same for the voltage)

$$v_I = \frac{I_{(1)}}{I} \tag{1.2.2}$$

where $I_{(1)}$- an effective value of the first harmonic of the current, I- an effective value of the current .

3. *A coefficient of harmonics of the current* (a coefficient of non sinusoidal K_{ns}) or *THD (Total Harmonics Distortion)*

$$K_{h.c.} = \frac{I_{h.h.}}{I} \tag{1.2.3}$$

where $I_{h.h}$- an effective value of higher harmonics of the current (different from the first harmonic).

Two this coefficients are connected together correspondingly.

$$v_I = \frac{I_{(1)}}{I} = \frac{I_{(1)}}{\sqrt{I_{(1)}^2 + I_{h.h.}^2}} = \frac{1}{\sqrt{1 + \left(\frac{I_{h.h.}}{I_{(1)}}\right)^2}} = \frac{1}{\sqrt{1 + K_{h.c.}^2}}, \tag{1.2.4}$$

from where

$$K_{h.c.} = \sqrt{\frac{1}{v_I^2} - 1} \tag{1.2.5}$$

4. *A coefficient of displacement of the current* relatively voltage on the first harmonic or *displacement factor*

$$\cos\varphi_{(1)} = \frac{P_{(1)}}{\sqrt{P_{(1)}^2 + Q_{(1)}^2}} \tag{1.2.6}$$

where $P_{(1)}$- the active power in the network, made by the first harmonics of the voltage and current;

$Q_{(1)}$- the reactive power of the displacement in the network, made by the first harmonics of the voltage and current.

5. *A coefficient of the power or power factor*

$$\chi = \frac{P}{S} \qquad (1.2.7)$$

Where P- the active power;

S- the full power.

In the case of the network with sinusoidal voltage and no sinusoidal current

$$\chi = \frac{P_{(1)}}{S} = \frac{EI_{(1)}\cos\varphi_{(1)}}{EI} = \nu_I \cos\varphi_{(1)} \qquad (1.2.8)$$

6. *A coefficient of the useful operation or efficiency*

$$\eta = \frac{P_{out}}{P_{in}} \qquad (1.2.9)$$

In the case of the ideal converter within the bounds of the first level of analysis (the absence of the power losses at elements of the converter) from (1.8) follows the correlation between coefficients of the displacement of the current of input and output networks of the converter.

$$P_{out} = P_{in},\ U_{(1)\,out}\, I_{(1)\,out} \cos\varphi_{(1)out} = U_{(1)\,in}\, I_{(1)\,in} \cos\varphi_{(1)\,in}\ ,$$
$$\cos\varphi_{(1)\,in} = K_{v.c.}K_{c.c}.\cos\varphi_{(1)out}. \qquad (1.2.10)$$

7. *An energetic coefficient of the useful operation or energetic efficiency*

$$\eta_E = \frac{P_{out}}{S_{in}} = \frac{P_{in}}{S_{in}} \cdot \frac{P_{out}}{P_{in}} = \chi\eta. \qquad (1.2.11)$$

8. *A coefficient of pulsations* for the networks of the direct current

$$K_P = \frac{X_{\max}}{X_{ev}} \qquad (1.2.12)$$

where X_{max}-an amplitude of the given (usually, first) harmonic component of the voltage (current),

X_{ev}-an average value of the voltage (current).

An expansion of the traditional system of the quality indexes of processes will be made at part 1.5.3. by an introduction of the *integral coefficients of harmonics.*

In that cases, when by a valve converter there is made an autonomous system of electric supply (a board of a ship, a plane, a ground transport), an admission of the quality indexes of electric energy and the numerical values are determined by corresponding state and trade standards, the same, as the quality of electric energy in electric networks for general use must be corresponding to state standard GOST 13109-87.

It is necessary to know for the calculation of the energetic indexes of processes:

- An effective values of the first harmonics of voltage and current in network and angle of the displacement between them;
- An effective values of the voltage and current;
- An effective values of higher harmonics of the voltage and current;
- An active and reactive power of the network.

It is may to calculate it by one of the three methods: 1) integral, 2) spectral, 3) forward (look part 1.5).

1.2.2. Energetic quality indexes of using of elements of the device and device in the whole

For energetic quality indexes of the using of elements of the converting device it is rationally to accept their relative (to active power of the load) established (standard) powers.

The established power of the two-winding transformer is calculated as a half of a summa of the products of effective values of a voltage (determines a section of a magneto conductor of the given kind and the number of turns of a winding) and a current (determines a section of the winding wire) for each winding

$$S_{\text{т}}^{*} = \frac{S_{\text{т}}}{P_{\text{н}}} = \frac{U_1 I_1 + U_2 I_2}{2P_{\text{н}}} \tag{1.2.13}$$

The established power of a reactor at a network of an alternating current is calculated, as a power of a transformer, with a coefficient 0,5 because of a presence only one winding

$$S_L^{*} = \frac{S_L}{P_{\text{н}}} = \frac{1}{2}\frac{U_L I_L}{P_{\text{н}}} \tag{1.2.14}$$

The reactor at a network of a direct current is characterized by a stocked already energy attached to a given frequency and a level of pulsation of a current

$$W = LI^2 .$$

The established (a reactive) power of a condenser at a network of a sinusoidal voltage (relatively an active power of a network) is calculated as a product of the effective values of a voltage and a current of a condenser, and attached to a presence of the higher harmonics in a current their amount limits depending on their frequency.

$$Q_C^{*} = \frac{Q_C}{P_{\text{н}}} = \frac{U_C I_C}{P_{\text{н}}} \tag{1.2.15}$$

A condenser at a network of a direct voltage is characterized by a stocked energy CU^2 attached to a given level and a frequency of a pulsation of a voltage (attached to a level of the higher harmonics of a voltage).

$$W_c = CU^2$$

It is possible to use the conventional reduction for a correlation of the energetic indexes of the network elements of an alternating current, expressed in units of the power, with the energetic indexes of the network elements of a direct current, expressed in units of the energy. It is necessary to divide the first indexes to a circular frequency of an alternating voltage w or to multiple the second indexes to this frequency for this procedure.

The established power of the not completely control valves (thyristors) is determined so:

$$S_v = nI_{\text{a}}U_{b\,max} ,$$

where n – the number of valves.

The established power of the completely control valves is determined already not through an average value of an anode current of a valve I_{a} , and through the maximal value:

$$S_v = nI_{a\,max}U_{b\,max} .$$

It is possible to determine the specific weight, overall, cost indexes and the specific indexes of the active power losses at the elements by calculated established powers of the elements and their known constructive execution.

The general

1. *An index of the specific mass of a device* [kg/kVA]

$$M_S = \frac{M}{S} \tag{1.2.16}$$

where M – a mass of a device, kg

S – the established (full) power, kVA.

2. *An index of the specific size of a device* [dm^3/kVA]

$$V_S = \frac{V}{S} \tag{1.2.17}$$

where V – a size of a device, dm^3.

An index of the specific cost of a device [c.u./kVA]

$$C_S = \frac{C}{S} \tag{1.2.18}$$

where C – a cost of a device, c.u.

This may calculate indexes there:

an index of the specific weight of a device [kg/dm^3]

$$M_V = \frac{M}{V} = \frac{M_S}{V_S} \tag{1.2.19}$$

an index of the cost of a mass unit [c.u./kg]

$$C_M = \frac{C}{M} = \frac{C_S}{M_S} \tag{1.2.20}$$

an index of the cost of a size unit [c.u./dm^3]

$$C_V = \frac{C}{V} = \frac{C_S}{V_S} \tag{1.2.21}$$

The additional

An index of the specific losses at a size unit [Wt/dm^3]

$$\Delta P_V = \frac{\Delta P}{V} \tag{1.2.22}$$

An index of the specific losses at a mass unit [Wt/kg]

$$\Delta P_M = \frac{\Delta P}{M} \tag{1.2.23}$$

Indexes of the specific losses at unit of a power (full or reactive) [Wt/kVA] or [Wt/kVAr]:

- for the reactive elements at networks of an alternating current

$$\Delta P_S = \frac{\Delta P}{S} \quad \text{or} \quad \Delta P_Q = \frac{\Delta P}{Q} \tag{1.2.24}$$

- for the reactive elements at networks of a direct current

$$\Delta P_W = \frac{\Delta P}{W} \tag{1.2.25}$$

The specific indexes are connected together by the next evident correlations:

$$\Delta P_M = \frac{\Delta P_S}{M_S} \tag{1.2.26}$$

$$V_S = \frac{M_S}{M_V} \tag{1.2.27}$$

$$\Delta P_V = \Delta P_M M_V = \frac{\Delta P_S M_V}{M_S} \tag{1.2.28}$$

The presence of three equations of a bond between the indexes shows, that only three indexes from six enumerated are independent, and three others may be calculated on the brought equations of a bond.

The numerical values of the specific indexes for the Russian element base of the power electronics (valves, transformers, reactors, condensers) are brought at a textbook [25]. It is possible to make an estimation of the mass-sized and cost indexes of a device else at the stage of a calculation of the electromagnetic parameters of the scheme elements of a converter, knowing values of the specific constructive indexes of the elements. Another way of getting of this indexes – their calculation on the constructive given facts of the ready converting aggregates, brought at reference books [13,36,37].

1.3. AN ELEMENT BASE OF THE ELECTRONIC POWER CONVERTERS

An aim of the given part is an acquaintance with an electric parameters of elements of the power electronics, from which, in according to a principal scheme of the valve converter, there are constructed the concrete power electronic devices.

1.3.1. The power semiconductor devices

All the considered converters, studied at the course « The principles of the power electronics», are realized at the power semiconductor valves: not control (diodes) and control (thyristors, transistors). The control valves are divided into two classes:

1) the valves with not full control;
2) the valves with full control.

1.3.1.1. The valves with not full control

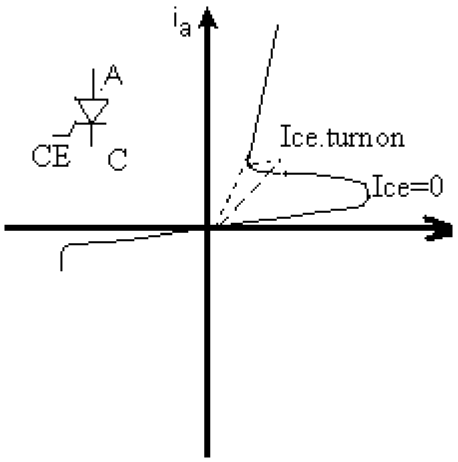
The valves with not full control are characterized by that junction from the state «turn off» to a state «turn on» may be a by momentary influence of a lacking power signal on the network of a control attached to a condition of a presence the forward voltage at the valve, that is a voltage of such polarity, attached to which the valve can conduct a current through itself. A junction of the valve from a state « turn on» to a state «turn off», that is turn off of the valve and stopping of a conducting of a forward current through it, may be only if to change a polarity of a voltage at a valve (a reverse voltage) on a power network, and not in a result of the influence on the network of control. So, *not full control* means that you can turn on a valve by an influence on the network of control, but you can't turn off a valve by an influence on the network of control, and it is demanded to change a polarity of a voltage at a valve to a reverse.

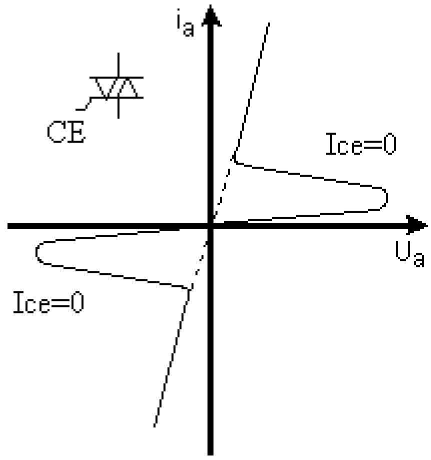
The valves with a full control are characterized by that turn on and turn off of the valves may be by an influence of a lacking power signals on the networks of a control attached to a presence a forward voltage at a valve.

The main representatives of the not completely control valves are *thyristors*-the four-layer p-n-p-n-semiconductor devices with an anode A (an extreme p-oblast), a cathode C (an extreme n-oblast) and a control electrode CE (an inside oblast) and Bi-directional Triode Thyristors- five-layer p-n-p-n-p-semiconductor devices, which it is may to represent as a combination of the two cross-parallel turned four-layer (thyristors) p-n-p-n- structures. There are represented a schematic designation of a thyristor and the voltage-ampere characteristic at the figure 1.3.1. There are showed a schematic designation of a Bi-directional Triode Thyristor (a symmetrical thyristor, TRIAK) and the voltage-ampere characteristic at the figure 1.3.1

figure 1.3.1 figure 1.3.2

The general parameters of the thyristors, which determine the possibilities of their using in different concrete schemas of the converters, are the next:





- *an average value of the anode current of a thyristor* I_a, on which it is marked by the factory-manufacturer issue from a level of a permissible losses of an active power (an apportionment of a heat) ay the valve attached to a conducting a forward current. An experimental current of a valves attached to their production has a type of half-wave of a sinusoid in each period of a network voltage (50 Hz). Attached to this *a coefficient of an amplitude* of a such current $C_a=\pi$ (the relation of an amplitude of a current to an average value), *a coefficient of a form* $C_f = 0{,}5\pi$ (the relation of an effective value of a current to an average). Thyristors are produced for the average current from 1 A to some thousands of amperes;
- *a latching current* I_{lat}, a minimal value of a forward current of a thyristor in case of the absence of control, when a thyristor is still staying to be conducting. Attached to the lowing of an anode current lower then this value a thyristor passes into the state «turn off»;
- *a maximally permissible forward and reverse voltages* U_{max} at the valve, which it must conduct without a breakdown. It is marked depending on the class of a valve on the voltage (there are valves from 1 to 50 classes), multiplying of which at 100 determines the maximally permissible voltage;
- *a reverse-recovery time of a control properties of a thyristor* t_r, which determines as a minimally necessary duration of the application to a valve the reverse voltage (attached to it`s turn off) after conducting a forward current while it restores the turn off properties and it is may to apply a maximal forward voltage again to it. A modern thyristors have a restoration times about from ten microseconds (for high frequency thyristors) till two hundreds microseconds (for low frequency thyristors);
- *a reverse-recovery charge a thyristor* Q_r, a full charge (accumulated in a valve attached to a conducting of a forward current), which flows out from a valve attached to a junction from a state of a conducting of a forward current to a state of the appearing a reverse voltage at a valve;
- *an amplitude of a reverse current of a valve* $I_{b\ max}$, conditioned by a discharge of the restoration charge Q_R from a valve at a moment of the slump till zero of the forward current of a valve (attached to turn off) with a determined speed *di/dt*:

$$I_{b\max} = \sqrt{2Q_R\left(-\frac{di}{dt}\right)}; \qquad (1.3.1)$$

- *a limit speed of an increase of a forward voltage at a valve*, attached to exceeding of which it is possible a turn on of a thyristor at a forward direction even attached to an absence of a control from the appearing a signal-noise in a network of the control electrode, «percolating» through a parasitic capacity between it and the anode of a thyristor. Usually this speed is limited from a hundred till a thousand volts in a microsecond for the different types of thyristors;

- *a limit speed of an increasing of a forward current of a thyristor attached to it's turn on*, connected with a not uniform distribution of a current on a square of a p-n junction of a thyristor, what can bring to a local damage (to a burn through) of a p-n junction. Usually this amount is limited by a manufacturer at the level from some ten till some hundreds of amperes in a microsecond;
- *a limit frequency of impulses of a forward current of a valve*, till which a valve can work without lowing a permissible average value of an anode current. This amount is equal 400 Hz for a low frequency thyristors and diodes, and for a high frequency thyristors – till 10…20 kHz;
- *a time of the turn on* t_{on} and *a time of the turn off* t_{off} of a semiconductor valve characterize a time of a junction of a valve from a state «turn off» to a state «turn on» and from a state « turn on» to a state «turn off» correspondingly;
- *a parameters of a control signal in a network of a control electrode of a thyristor*, which provide it`s reliable turn on: *a control voltage* U_{ce} (some of volts), *a control current* I_{ce} (a parts of ampere), *a speed of increasing of a control current* dI_{ce}/dt (1-2 A/mcsec), *a minimal duration of the control impulse* (20-100 mcsec). Attached to this the power of the control signal is lower in thousands of times then a power, switched by a thyristor at the anode network;
- *a voltage of a cut-off of the straighten voltage-ampere characteristic of a valve in the forward direction* ΔU_0 and the *dynamic resistance* R_{dyn}. There are showed a real not linear and a partly-linear model (simplificated) voltage-ampere characteristics of a valve at a forward direction at a figure 1.3.3.

A value of a voltage of a cut-off for a siliceous valves is equal about 1 V, a value of a dynamic resistance is a reverse proportional to a nominal average value of an anode current of a valve I_a and is a variable amount from a parts of Om for the lacking power thyristors till the thousand parts of Om for the power thyristors, having an order $1/I_a$ [Om]. figure 1.3.3

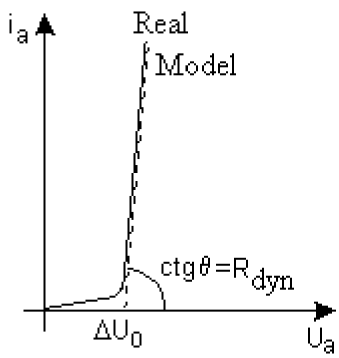


Figure 1.3.3

This parameters determine the losses of an active power at a valve attached to a conducting of a forward current, what brings the heating of the semiconductor structure; figure 1.3.3

- *a heat resistance of a valve* characterizes the ability to lead a heat from the place of it apportionment, that is from p-n junction, and determines as a relation of the temperature overfull between two mediums ΔTon the unit of the dispersed at a valve power ΔP_v [degr/Wt]. Three heat resistances of a valve are meaningful first of all: p-n junction – a body of a valve R_{jb}, p-n junction- a cooler R_{jc}, p-n junction – environment R_{je}. A different heat resistances, through which is determined the limit power of the losses at a valve (a limit average value of an anode current of a valve), issue from a maximal permissible temperature of the p-n junction (for a siliceous diodes – 150 °C, for a siliceous thyristors – 110…120°C) are corresponding to the different ways of the cooling of a valve;
- *a protective index* $\int i^2 dt$ is a value of the time integral from a square of the percussive forward current, appearing attached to a crash, attached to an increasing of which the valve is destroyed. In according to this index, if to increase a value of a crash forward current through the valve, so the duration of it must decrease.

1.3.1.2. The valves with a full control

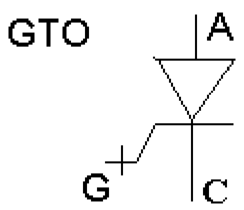
A valves with a full control are characterized by that it is possible to turn on and to turn off it attached to a presence of a forward voltage at it by an influence just on the network of a control.

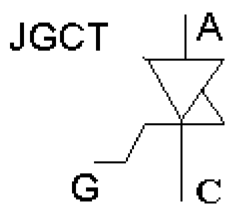
The general representatives of the valves with full control are *Gate Turn Off (gate-off) GTO* thyristors and a power transistors (*a bipolar, field-effect transistors* and *a combined,* called as *a bipolar transistors with an insulated gate,* designated as IGBT – Isolated Gate Bipolar Transistor).

1.3.1.2.1. A Gate Turn Off thyristors

A Gate Turn Off (gate-off) thyristors are differ from an usual (one-operating) thyristors by that it is possible to turn off it by a supply of a short, but a power current impulse of a reverse polarity, to a network of a control electrode of a thyristor. A big amount of this current impulse is determined by that a coefficient of amplification on a current attached to a turn off a thyristor is not high, usually it is not more then 4-5. Because it is important not an average value of a forward current for a gate-off thyristor, and the maximal (momentary) value, because of which gate-off thyristors are marked. A reached limit parameters of a gate-off thyristors abroad: on a forward current – till 2,5 kA, on a voltage – till 4 kV, on a frequency of a switching – till 1 kHz, on an amplification coefficient on a current of a turn off – till 3-5. A conventional designation of a GTO-thyristor is shoed at a figure 1.3.4,a.

At last years GTO-thyristors were modified and there was created a new type of a device – GCT – Gate Commutated Thyristor or IGCT – Integrated Gate Commutated Thyristor. A times of commutation are lowing on already an order at this thyristors, because the all current turn on / turn off is commutated through the control electrode; hence, a commutating losses are lowing, too. It allowed to create IGCT on a 3 kA, 3,5 kV already today. Attached to this for this thyristor it is not required the application of snubbers – special outside networks, forming a trajectory of a working point attached to a turn off a thyristor. In a simple case it is a condenser, limiting the speed of increasing of a forward voltage at a thyristor attached to it`s turn off. A not big active resistance is turned on consecutively with a condenser for a limiting a condenser current. A conventional designation of a IGCT-thyristor is showed at a figure





1.3.4,b.

figure 1.3.4,a

figure 1.3.4,b

There are continued an investigations of the gate-off thyristors with a field-effect control (without a consuming of a current) – MCT (MOS Controlled Thyristor), which will press GTO-thyristors because of a simplification of a control attached to a condition of a comparability of their limit electric parameters.

1.3.1.2.2. The transistors

A principal difference of the transistors from the gate-off and usual thyristors, turned on and turned off by a short control impulses, is that it is necessary a presence of a control signal at it for an all time of conducting a forward current through a transistor. A limit electric parameters of a transistor, determine a possibilities of it application at a power electronic devices, are depending from a type of a transistor.

Bipolar transistors (BPT). This transistors are three-layer semiconductor structures of p-n-p and n-p-n types, in which there two p-n junctions: a base-emitter and a base-collector.

A bipolar transistor allows to control by a current more then before in tens of times, flowing through outside junction base-collector, displaced in the reverse direction, because of the changing of the base current of a p-n junction a base-emitter, displaced in a forward direction. There is getting a big amplification at a transistor on a voltage, hence, a very big (in a hundreds and thousands of times) power amplification because a reverse voltage at a collector (outside) junction may be more then a forward voltage at an inside junction base-emitter in a tens of times .

A conventional designation and outside voltage-ampere characteristics of a bipolar transistor are considered at a line 1 of a table 1.3.1.

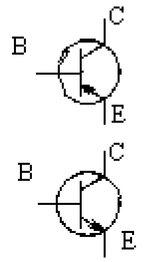
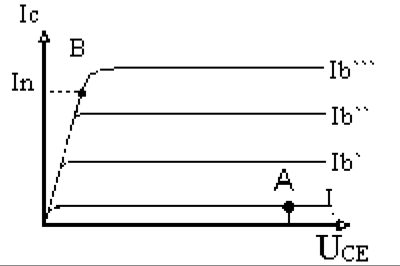
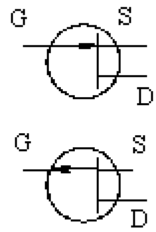
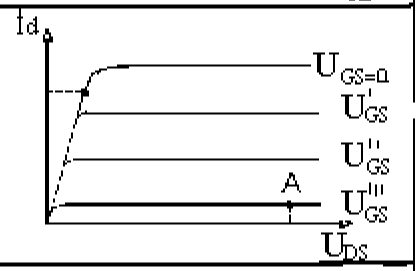
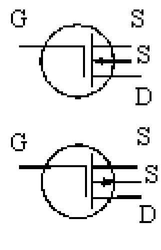
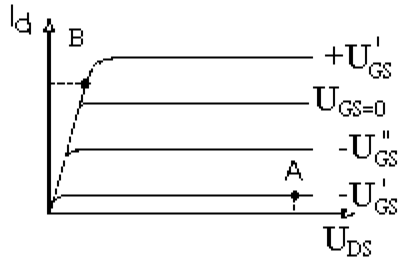
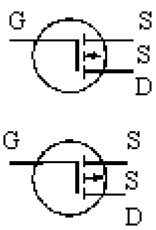
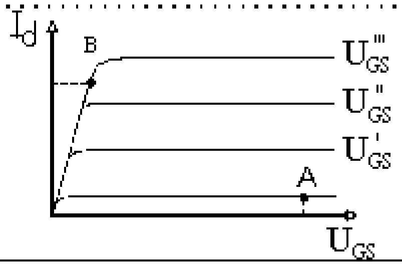
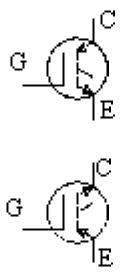
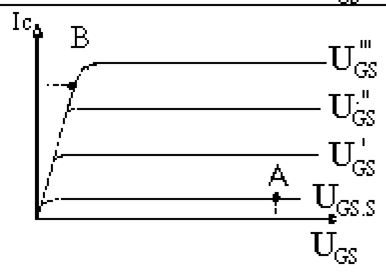
This possibility of a transistor attached to a work in a key (as a thyristor) conditions allows to use it at a power electronic devices for a control by flows of an energy with an aim of it converting. A key condition of the transistor work is provided by a corresponding control. A base current is equal to zero at a state «turn off» of a transistor (a point A at a outside characteristics), that is a key is turned off; attached to this we neglect by a little not control current of a collector at a down voltage-ampere characteristic. At a state «turn on» of a transistor a base current is established not lower such level $i_\delta{}^m$, to a work point of a transistor with a given outside network of an amount of the load current i_L was at the state B, corresponding to a minimal possible voltage at a transistor attached to this current, for a decreasing the power losses at a transistor.

An industry produces a power bipolar transistors calculated on the currents till hundreds of amperes with a voltage in hundreds of volts and with a maximal frequencies of a switching attached to this till units of kilohertz. The general shortcomings of a bipolar transistors are connected together with a noticed power expenditures for a control (a current control on a base) and with not enough fast-action , determining a speed of a junction of the work point of a transistor from the state A to the state B and reverse.

Field-effect transistors. A field-effect transistors use a one (unipolar) type of a current carrier in the difference from a bipolar transistors, working with two types of a current carriers- electrons and a holes. A conductivity of a canal between a source and a drain (a determined analogies with an emitter and a collector of a bipolar transistor) is modulated by an electric field, supplied to a canal in a diametrical direction by a third electrode – a gate (a control electrode). A canal may be of the two types: n-type or p-type. A conventional designations of a field-effect transistors with a gate in a kind of a reverse displaced p-n junction and their outside voltage-ampere characteristics (for a n-type canal) are considered at a line 2 of a table 1.3.1. Now a control parameter for an output characteristics is a gate voltage (at a input of a transistor), and not an input current, as at a bipolar transistors. An input network of a field-effect transistor is a very high ohm and it almost doesn't consume a current, that is a field-effect transistor control happens without a power expenditure. A field-effect transistor with a p-type canal has analogical properties and characteristics, just it is necessary to change a polarities of a voltage at a drain and a gate (relatively of a source) at a reverse at a last transistors.

The second kind of field-effect transistors – *transistors with an insulated gate.* A gate is divided by a dielectric pellicle from a canal at this transistors and because there is no current even theoretically at a input network of a transistor. Besides, such dividing of a gate from a canal allows making a canal by two variants: as a fitted (constructive) or as a induced (brought attached to a current conducting) canal of a p-type or a n-type. Conventional designations of such transistors and the output characteristics for a n-type canal are considered at a line 3 of a table 1.3.1. This transistors are called MOSFET or FET-transistors (Metal-Oxide-Semiconductor-Field-Effect Transistor) abroad, what is corresponding to our designation MOS (MDS) transistor (metal-oxide-semiconductor), where a metal means a gate electrode, an oxide means a dielectric, dividing the gate from a semiconductor canal between the source and the drain.

The general merits of a field-effect transistors – an absence of the power expenditures for a control and a high fast-action in a result of a current carrying in it by a carriers of a one sign (by a general carriers), in a difference from a bipolar transistors, where the current at a middle part of a device (at a base) is carrying in general by a slow not general carriers.

NN	Attributes / Type of transistor	Designations	Output VAC
1	**Bipolar** **p-n-p** **n-p-n**		
2	**Field-effect(FET)** **p-n junction** **n-type channel** **p-type channel**		
3	**Field-effect of the (MOS)-type (with isolated gate)** **With a fitted channel of n-type** **p-type**		
	With inductanced channel of **n-type** **p-type**		
4	**A combined (IGBT-transistor)** **n-type channel** **p-type channel**		

But the field-effect transistors are noticeably inferior to a bipolar on limit values of an output voltage and a current, what determines the niche of their using at power electronic devices with a low voltage and with high frequencies of the processes of a converting an electric energy.

A combined transistor. At the last years there appeared a combined device, constructively uniting a field-effect transistor with an insulated gate (at input) and a bipolar transistor (at output) and it is called a transistor IGBT (Isolated Gate Bipolar Transistor). It has a high input resistance and doesn't need the power for a control, as a field-effect transistor. Parameters of the output voltage and current are the same , as a bipolar transistor has, that is meaningfully higher, then a field-effect transistor has. In according to a four types of a field-effect transistors with an isolated gate there are possible four types of IGBT transistors, a conventional designations of which and the output voltage-ampere characteristics for a transistor with an induced channel of n-type are showed at a line 4 of a table 1.3.1.

There are produced IGBT-transistors of the fourth generation with the output currents till 1200 A and the voltage till 3500 V at a real time abroad.

The feature of the all types of the transistors in comparison with an another their «competitor» between a valves with a full control – GTO thyristor is that it is necessary the control signal at input for an all time of a current conducting at a output network of a device for the transistors. Besides some types of the transistors, as it is considered from the output voltage-ampere characteristics at a table 1.3.1, demand the presence of a direct voltage source at a control network for a providing the turn off of a transistor at a points A of the corresponding (down) voltage-ampere characteristic.

There are necessary the control impulses of a reverse polarity at the moments of the turning on and the turning off of a device for GTO-thyristors.

The oblasts of the preferable using of the different types of the semiconductor valves in 1980-1990 years and in 2000 y. are represented at the diagram's of a figure 1.3.5 (from given facts of Chip News 1999, №1).

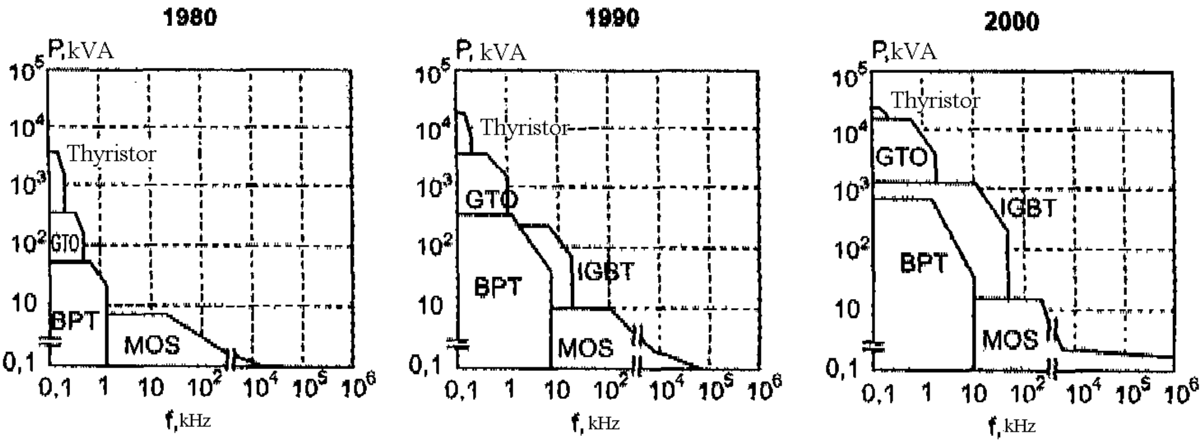


Fig. 1.3.5

The next development of the semiconductor element base of power electronic devices is a creating of whole fragments of power electronic devices in one semiconductor crystal or in one hybrid construction, that is in a module. It is a totality of some power semiconductor devices, united to a scheme of a typical device (a power integral scheme – PIS), or a power element with a device of a control and defense (Smart, Intelligent – *an intellectual scheme*). The examples of such modules will be considered at corresponding parts.

Power semiconductor devices, which gave a life to a new electric-technical branch – to a semiconductor power electronics, are the main elements in a basic cells of the converters of an electric energy. A basic cells, if it is necessary, are supplied by the additional elements-by transformers, reactors, condensers for an increasing a possibilities of the cells of a converting, for a better quality of a converting of an electric energy and providing electromagnetic compatibility of the converters with a maim.

1.3.2. The transformers and reactors

A nomenclature of the transformers of a n industrial manufacture for a semiconductor converters is a meaningfully molester in a difference from a wide nomenclature of a power semiconductor devices, accounting a thousands of varieties, which differ by a type and parameters. It is connected with a limit line of the values of the industrial voltages both for networks of an electric supply, and for typical consumers of an electric energy. It determines a necessity of the issue of a power transformers with a fixed transformation coefficients, which in a composition of a converter of a determined type with the converting coefficient on a voltage provide a transmission of a converting energy from a level of a maim voltage to a level of a consumer voltage. It brought to that tactically were unified converting transformers only for a one class of the converters – rectifiers (and naturally commutating inverter) in a former SSSR. Transformers for another types of converters, working at a not standard frequency 50 Hz, at a not standard form of a voltage (no sinusoidal), with a not standard transformation coefficient, were projected and made usually there, where the converter in a whole. Because it is unified at a level of a factory, and not a branch.

The inquiry facts on unified converting transformers (for a voltage rectifiers with a frequency 50 Hz) are represented in the works [25,27,36,37]. This indexes are changing depending on a transformer power, are decreasing with an increasing of a power, and for transformers of the type TSP with a power 10…200 kVA often there are

M_s=10-5 kG/kVA, M_v=4-2 dm^3/kVA.

A coefficient of a useful action (CUA) or efficiency of a converting transformers of this power diapason is equal to 0,96…0,98.

Reactors at valve converters are used for a current limitation, filtration in networks of an alternating current and for a smoothing a current in networks of a direct current. In the case of making a valve converter without an input transformer there are established the current limitation reactors at an input of a converter, which need to limit currents attached to a short circuit in a load of a converter or inside of it. A full power of this reactor S_{or} determines the equal voltage of a short circuit of a hypothetical input transformer with a typical power St :

$$U_k\% = \frac{S_{op}}{S_T} 100\,\% \qquad (1.3.3)$$

Here the specific indexes because of a little power of a current limiting reactors of a type RTST (1…7 kVA) are meaningfully worse, then transformers have:

M_S = 25-20 kG/kVA, M_V = 50-35 dm^3/ kVA.

For a smoothing reactors of a type SROS, working in the networks of a direct current with a little relative value of a current pulsation (of a magnet flow at a magneto conductor), the specific indexes are meaningfully better and are equal for the reactors powers 10…200 kVA

M_S = 3,5-1,5 kG/kVA, M_V =1,7-0,7 dm^3/ kVA.

1.3.3. The capacitors

In according to two kinds of electric energy (alternating current and direct current) the capacitors are differ on a purpose. For the alternating current networks there are intended the «cosinusoidal» (compensation) capacitors, producing a reactive current at a source of a reactive power (SRP), outstripping a sinusoidal voltage at a quarter of a period, and a filter capacitors, intended for a filtration (decreasing) of a higher harmonics, presented in a networks with the converters. For a direct current networks are intended polar capacitors (usually electrolytic), intended to smooth a pulsation of a direct voltage.

The real mass-sized indexes of capacitors are essentially depending, besides constructive-technological features, from parameters of a regime of an electric network else, in which it will be used. A regime determines a level of an active power losses at it, hence, and a degree of it permissible electric loading depending on a frequency and a voltage (current) form of a network.

As it is known, the power losses in a capacitor are proportional to a tangent of an angle of a dielectric losses tgδ. So in the case of a no sinusoidal voltage at a capacitor the resulting losses on a base of the superposition method of the regimes on the separate harmonics will be equal to

$$\Delta P_c = C\omega_1 \sum_{n=1}^{\infty} nU_{(n)}^2 \operatorname{tg} \delta_{(n)} \qquad (1.3.4)$$

Usually there brings a dependence of tgδ from a frequency at a reference books on the condensers, what allows to calculate the losses attached to a work of a condenser at a network of a sinusoidal current of a known frequency and at a network of a no sinusoidal current. Calculated losses at a both cases increase with an increasing a frequency, what will demand a decreasing a voltage at a condenser for a limitation an increasing of the losses. This will bring to a decreasing of real values of a specific mass-sized index of a capacitor. Here there are represented the values of the indexes for a line of kinds of not polar and polar capacitors. For the home capacitors of a type MBGT at f =50 Hz an index of a specific mass 0,15 dm^3/

Dg, for the capacitors of a type K72-11 at 500 Hz – 0,6 dm^3/kVAR, for electrolytic capacitors of a type K50-27 this index is already equal 0,002…0,005 dm^3/Dg. It is possible to determine the indexes of a specific mass of the capacitors through an index of a specific weight of the capacitors, which is typically equal about 2 kG/dm^3. The foreign capacitors have the indexes better at 1-2 orders.

1.4. THE KINDS OF VALVE CONVERTERS OF AN ELECTRIC ENERGY

The aim of the given part is a consideration with an existing admission of a basic converting cells and a power interfacing cells, providing an electromagnetic compatibility of a basic converting cells with a main and a load.

All possible kinds of the converters of an electric energy from a one kind (determined by a generating side) to another (determined by a consuming side) are showed schematically at a figure 1.4.1. This final great number of the converter kinds consists from the next basic cells.

- A converters of an alternating (two-direction) current to a direct (a one-direction) current, called a *rectifiers*, which it is comfortably to design AC-DC on a known shortening at the English literature AC-DC (Alternating Current – Direct Current).

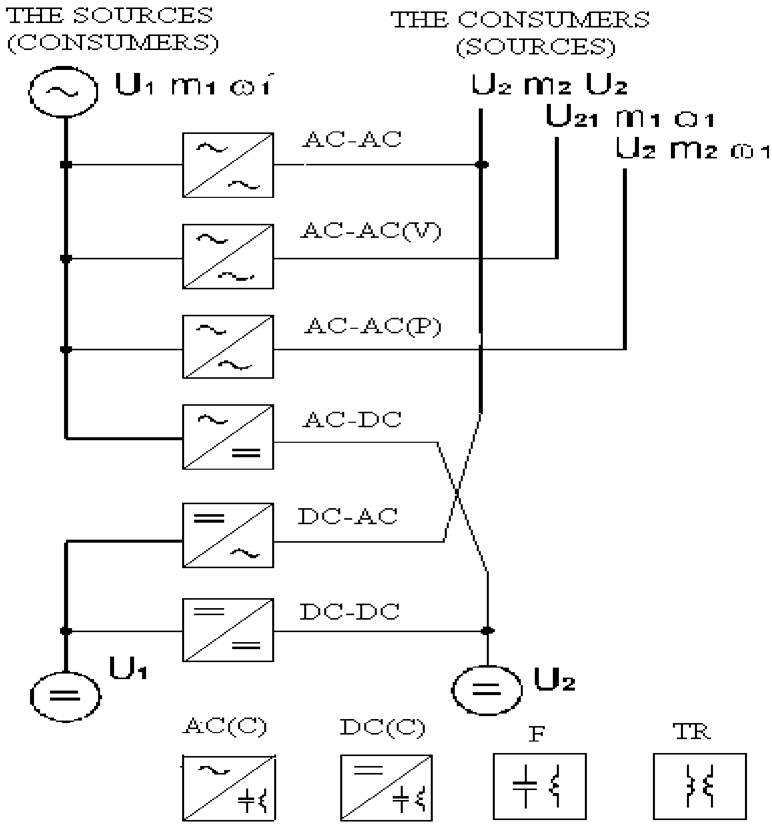


figure 1.4.1

- * A converters of an alternating current of a one frequency to an alternating current of the another frequency, it is possible that with an another number of phases, called a *frequency converters*, which we will design, following to the logic, as AC-AC on a foreign technical designation.
- Converters of an alternating current with a one number of phases to an alternating current of the same frequency with the another number of phases, called *the converters of a number of phases* and presenting, on an essence, a private case of the last type of the converters and because designated at the next AC-AC (P).
- Converters of an alternating current of a one frequency to an alternating current of another frequency, differing in a fixed number of phases *M* from an issuing frequency, called *the multipliers of a frequency,* presented the another private case of a frequency converter and because designated at the next AC-AC (F).
- Converters of an alternating voltage to a variable alternating voltage of the same frequency, called *the regulators of an alternating voltage* and designated AC-AC (V).
- Converters of a direct current into an alternating current, called *the invertors,* which we will design DC-AC on a foreign designation.
- Converters of a direct current to a direct, called *the regulators of a direct current* (electric «transformers»), which we will design DC-DC on a corresponding foreign shortening.
- A regulated *sources of a reactive (not active) power,* designated SRP, allowing to introduce the additional (to a reactive powers of the sources) reactive powers of a shear SRP (S) to a system of an electric supply, of the distortion SRP (D), of the nonsymmetrical SRP (N) with an aim of a compensation of a corresponding powers of the not quality consumers and so of an improvement of an electric energy quality at a system of the electric supply. There are possible two variants of the

connection of a SRP to a network: to a node (a diametrical compensation because of the giving the additional current at a node of a network (SRPC)) and between the nodes (a longitudinal compensation because of the giving the additional voltage between the noodles of a network (SRPV)). Depending on the kind, the way of a turn on and the algorithm of a control the SRP can make the functions of the reactive power compensator of a shear, of the regulator of a voltage at a node, of *the active filter* (by introducing to a network an influence with a spectrum, which is a reverse to a spectrum of a indignation of a normal network regime).

A full converting device consists, besides the basic cell, from the input and output transformers else (cell T) attached to a presence of the networks of an alternating current, and usually from the input and output filters (cells F, figure 1.4.1).

A transformer is intended, at the first, for the concordance of the demanded output voltage level of a basic cell with a given level of a maim voltage, at the second, for a possibility of the increasing the number of **phases of an** alternating voltage at a secondary side of a transformer, at the third, for the creating a galvanic (conductive) insulation of the networks of the input and output of a converter. The last fact , providing a wireless connection (only through the electromagnetic field of a transformer) of the input and output networks of the converter, excludes a possibility of a dangerous hit of a voltage from the side, having the more high potential, to a side, having the more low potential attached to the turn off a transformer at a one of the sides.

The converting of an electric energy at a basic cells happens with a help of the sharply not linear elements – a valves, which may be found only at the one from the two states- turned on (conducting) or turned off. As a result both the energy consuming by the cell from the main and the transmission of it at the cell output to a consumer happens decremently, what brings to the decreasing a quality of a converting and converted electric energy. The filters at the input and output of a valve cell are intended for a decreasing and a smoothing the consequences of a decrementing of a process of the energy converting. That is, this filters provide an electromagnetic compatibility of a converting cell with a main and a load. Under *the electromagnetic compatibility* at the electric technique understands an ability of the different electro technical devices, connected by the networks of an electric supply and an electric distribution, to synchronously function at a real conditions of the using attached to a presence of the not premeditated signals-noise at this networks and not create the not permissible electromagnetic signals-noise at a network to another devices, connected to this network. Figuratively saying, the situation is alike with a human compatibility of the many lodgers of the communal flat at the common kitchen, forced to use by the common communal services, not creating the not permissible noises to each other.

All considered basic cells are characterized by the singling of the converting of an electric energy and have a determined admission of the properties, which will be considered later at the corresponding parts. For the expansion or modification of the properties of the electric energy converters it is possible to make it from the basic cells as from the admission of the constructor, creating already the basic structures, characterizing by a repeated (usually two-, three-multiple) converting of the electric energy kind at the way from the input to output of the converter. For example, it is possible to convert an alternating current to a variable direct current not only with help of a basic cell AC-DC, but in the next compound structures (the cells of a transformer and filters here are not considered):

- AC-AC(V) - AC-DC (at first the regulation of the alternating voltage amount, then the rectifying without a regulation);
- AC-DC – DC-DC (at first the rectifying without a regulation, then the regulation of the direct voltage);
- AC-DC – DC-AC – AC-DC (at first the rectifying without a regulation, then a converting to an alternating voltage of the high frequency with a regulation of a voltage, then a rectifying without a regulation again) and so on.

The properties of such compound converting structures you can get from the totality of the properties of basic cells, as it will be showed at the third part of the book.

The proper process of a converting of the electric current kind happens at a valve cell, which is a determined structure from the valves. Attached to a permissible idealization of the valve by a key at the first stage of the analysis its function, which is to commutate a voltage and a current, may be described by a periodical explosive single function Ψ_v, called *the commuting function* of the valve and showed at a fig.1.4.2.To a state «turn on» (conducting) of

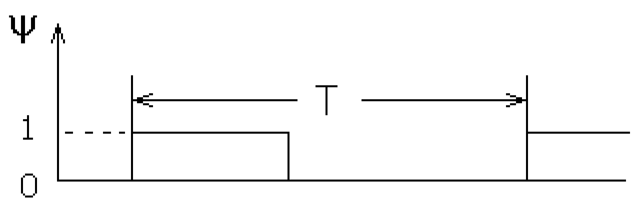


Figure 1.4.2

the valve is corresponding a level of the single value of the function, to a state «turn off» - the level of a zero Attached to this a corresponding alternating value y (a voltage, a current, a momentary power) after the key it is expressed by an evident figuration through the same alternating value u till the key, that is:

$$y=u\Psi_v=u(\Psi_0+\Psi_{(1)}+\Psi_{hh})=u\Psi_0+u\Psi_{(1)}+u\Psi_{hh} \qquad (1.4.1)$$

Here Ψ_0- a direct component of the commuting function;

Ψ_{hh}- a higher harmonics of the commuting function attached to the expansion it into the row of a Fourier.

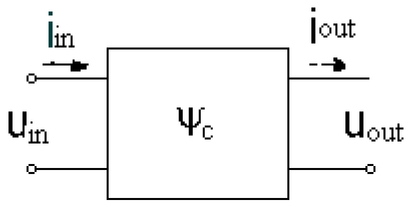


Figure 1.4.3

If the input alternating value u is a harmonic function (a voltage of a maim), there will be maintenance already a direct component at the component $u\Psi_{(1)}$, indicative of happened converting of a current kind from the alternating to the direct output current with a pulsation (a components $u\Psi_0$, $u\Psi_{hh}$ and an alternating component $u\Psi_{(1)}$). It is possible to change a direct or an useful alternating component of the converted current kind by the regulation of a phase, of a relative duration and in a common case of a impulses frequency of the commuting function. At least this converting and regulation of the parameters of the converted energy comes to a modulation of the turn on and turn off moments of valves.

From the commuting functions attached to the known connection to a structure it is may to determine *the commuting function of the valve group* of the converter Ψ_C, connecting the input and output voltages and currents of the valve group, considered as a four-pole in according to a figure 1.4.3.

Attached to a connection of such four-pole to a source of emf the equations of the bond have a such kind:

$$U_{out} = \Psi_c u_{in}, \qquad i_{in} = \Psi_c i_{out}, \qquad (1.4.2)$$

and attached to a connection to a source of the current:

$$i_{out} = \Psi_c i_{in}, \qquad u_{in} = \Psi_c u_{out}. \qquad (1.4.3)$$

The one-linear mathematical model (1.4.2) or (1.4.3) is increasing by the corresponding figure attached to a many-phased input or output of the valve group of the converter.

So, all the basic structures of the electric energy converting have a one-type ideal models, having a difference only on a kind of the commuting function of a converter, which determines a kind and a quality of the electric energy converting.

1.5. THE METHODS OF A CALCULATION OF THE ENERGETIC INDEXES

The aim of the given part is a comparative studying of the three existing approaches to a calculation of the energetic parameters of the valve converters.

1.5.1. The mathematical models of the power converters

A kind of the mathematical model of the power converter essentially determines the choosing of a calculation way of the electromagnetic processes at it. The way of a calculation determines a labor-intensiveness of a calculation, a size and a kind of a got result. Because a choosing of the mathematical models of the valves and the converter and the calculation way of the processes at a converter it is necessary to make concordancely.

A periodic valves commutation at a converter attached a valve model in a kind of a key brings to two kinds of the mathematical models of a converter. A valve converter together with an input source is replaced by the source of a voltage or a current of the explosive form in according to the first equations of the systems (1.4.2) and (1.4.3), if at an input of the valve converter are used models of the ideal sources of emf and current, and inside of the valve group there is no passive elements of an electric network (resistances, condensers, reactors). So the processes in a load are described by the differential equations with the constant coefficients and the explosive right part. The processes in a load and in the input networks of the converter are described by the differential equations with a variable periodic (explosive) coefficients, if at the input or inside of the valve converter there are passive elements (for example, the elements of the filters). The analysis of the processes at a converter becomes more difficult in a case of such kind of the models [20,38].

The next three calculation methods of the energetic indexes of the converters are applied for the both forms of the given mathematical models of the valve converters:

1) Integral;
2) Spectral;
3) Forward.

1.5.2. The calculation methods of the energetic indexes of the converters

1.5.2.1. The integral method

All the absolute amounts, which come in a determination of the indexes, are expressed in a form of the determined integrals from the corresponding currents, voltages and their combinations in the integral calculation method of the relative energetic indexes. It is the effective values of the currents and voltages

$$I = \sqrt{\frac{1}{T}\int_0^T i^2 dt}\,, \qquad U = \sqrt{\frac{1}{T}\int_0^T u^2 dt}\,. \tag{1.5.1}$$

This is *an active* power

$$P = \frac{1}{T}\int_0^T uidt \tag{1.5.2}$$

a reactive power of the displacement (attached to a sinusoidal form of a voltage or a current)

$$Q = \frac{1}{T}\int_0^T u\frac{di}{dt}dt = -\frac{1}{T}\int_0^T i\frac{du}{dt}dt \tag{1.5.3}$$

a full power

$$S = U \cdot I = \sqrt{\frac{1}{T}\int_0^T u^2 dt}\ \sqrt{\frac{1}{T}\int_0^T i^2 dt} \tag{1.5.4}$$

For a profound characteristic of the nonsinusoidal energetic processes it is may to use else a lot of another *partial components of the full power*, the common expression for which in integral form have a look [20,21,39].

$$M_j = \frac{C_j}{T}\int_0^T N_j\{u\} \cdot L_j\{i\}dt\,. \tag{1.5.5}$$

Here a look of the converting operators of a voltage $N_j\{u\}$ and a current $L_j\{i\}$ determines a kind of that or another partial component M_j of the full power S.

It is necessary to know the changing lows of the momentary values of the corresponding alternating amounts for a calculation of the all given integrals. You can find it only from the decision of the differential equations, compound for that electric network, at which you determine the energetic indexes. This fact determines the consumering *properties of the integral method*.

1. The method is universal, because it is possible to solve the differential equations always, if not analytically, so numerically.
2. The calculation method of the energetic indexes becomes a numerical attached to an absence of the analytical decision of the differential equation. It does not allow making a common investigation in the analytical form of a dependence of the energetic indexes from the parameters of the electric network.
3. The method becomes a very labor-intensive and is accessible only for IBM attached to a high order of the differential equations (higher 2-3) and attached to a presence at a period of a many points of the violation of a not interruption of the functions, called by a spasmodic switching of the valves.

1.5.2.2. The spectral method

At *a spectral method of a calculation* of relative energetic indexes all the absolute amounts, which come in determination of the indexes, are expressed in a form of the infinite rows, which are getting from the rows of Fourier (spectrums) of the corresponding currents and voltages. So, the effective values of the voltages and currents in according to a formula of Parseval from the theory of the rows of Fourier

$$I = \sqrt{\sum_{k=1}^{\infty} I_{(k)}^2}\ ,\ U = \sqrt{\sum_{k=1}^{\infty} U_{(k)}^2}\ , \qquad (1.5.6)$$

where *I (k), U (k)* – the harmonics of a k-order of the current and voltage correspondingly.

The active power

$$P = \sum_{k=1}^{\infty} U_{(k)} I_{(k)} \cos \varphi_{(k)} \qquad (1.5.7)$$

The reactive power of Budeany (of a displacement) attached to a nonsinusoidal voltages and currents

$$Q_B = \sum_{k=1}^{\infty} U_{(k)} I_{(k)} \sin \varphi_{(k)} \qquad (1.5.8)$$

The full power

$$S = UI = \sqrt{\sum_{k=1}^{\infty} U_{(k)}^2}\ \sqrt{\sum_{k=1}^{\infty} I_{(k)}^2} \qquad (1.5.9)$$

It is necessary to know the spectrums of the voltage and current at an electric network for a calculation of the all mentioned amounts. A spectrum of a voltage you can find on a known form of a voltage curve by a decomposition of it to a row of Fourier. A spectrum of a network current is calculated through a spectrum of a voltage and founded on a scheme of a network full resistance of a network on each harmonic of a spectrum. This procedure determines a consumering properties of the spectral method.

1.The method does not need a composition and a decision of the differential equations, what allows to use less time.

2. The energetic indexes are represented by expressions, maintained the infinite rows. The practical truncation of the row always brings a error into a calculation, which it is difficult to estimate.

3. The network parameters come in each member of the row, what makes the investigation more difficult at the analytical form of a degree of the influence of the separate network parameters on each energetic index, making a calculation procedure a numerical.

1.5.2.3. The forward method

The forward method – a method of an algebraization of the differential equations. Under the forward calculation methods of an energetic indexes at the networks with a nonsinusoidal voltages and currents understands the methods, not demanding the foundation of the momentary current values (as in integral method), the foundation a spectrum of a current (as in spectral method). Calculating formulas for an energetic indexes are brought through the coefficients of the differential equation and the parameters of an applied voltage of a current (as in spectral method). Calculating formulas for an energetic indexes are brought through the coefficients of the differential equation and the parameters of an applied voltage of a current (as in spectral method). Calculating formulas for an energetic indexes are brought through the coefficients of the differential equation and the parameters of an applied voltage of a current (as in spectral method). Calculating formulas for an energetic indexes are brought through the coefficients of the differential equation and the parameters of an applied voltage of a current (as in spectral method). A calculating formulas for an energetic indexes are brought through the coefficients of the differential equation and the parameters of an applied voltage at a version of a forward method, considered here, called a method of an algebraization of the differential equations (ADE) [21]. As a such parameters it is used an admission *of the integral coefficients of the voltage harmonics*, which are the broadening of a determination of a traditional coefficient of the voltage harmonics, as it is showed lower.

We will consider the calculation procedure by a method ADE at the examples of a calculation of the networks of the first and second order, to which it is possible to bring mathematical models of a majority, studied at a course basic cells of the converters. Attached to it it is possible to apply the method ADE for a calculation:

- Of an effective value of a nonsinusoidal current (a method ADE1);
- Of an effective value of the higher current harmonics (a method ADE2);
- Of the first harmonic of a current (a method ADE (1));
- Of the powers, created by the current curve (a method ADEP1), the high frequency component (a method ADEP2), the first harmonic (a method ADEP(1)).

1.5.2.3.1. The method ADE1

The considered lower procedure of a conclusion by the forward method of a final formulas for a calculation of the effective current value at a network with a converter is important for a potential elaborators of the new calculation methods, but it does not demanded for the users of this formulas. It is enough to have ready calculating correlations, which got by an applying the method, for the users.

A procedure of the algebraization of a differential equation consists from the next steps.

1.A composition of the equal electric scheme of a converter replacing, in which the converter is represented by a source of a voltage (current) of the given nonsinusoidal form. This scheme is represented at a figure 1.5.1 for the converter with a LC-filter and an active load R .

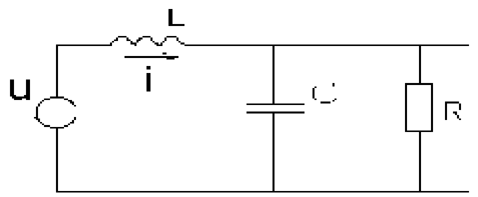


Figure 1.5.1

2.The getting of a differential equation for the interesting alternating value, here – a load current, with a help, for example, of the symbolic method.

$$I(p) = \frac{U(p)}{pL + \dfrac{R\dfrac{1}{pC}}{R + \dfrac{1}{pC}}} \cdot \frac{R\dfrac{1}{pC}}{R + \dfrac{1}{pC}} \cdot \frac{1}{R} = \frac{U(p)}{p^2 LCR + pL + R}$$

or attached to $p = \dfrac{d}{dt}$, $LCR\dfrac{d^2 i}{dt^2} + L\dfrac{di}{dt} + Ri = u$ (1.5.10)

Generalizing the differential equation for a network of the second order of the any configuration, we will have

$$a_2 \frac{d^2 i}{dt^2} + a_1 \frac{di}{dt} + a_0 i = b_2 \frac{d^2 u}{dt^2} + b_1 \frac{du}{dt} + b_0 u \qquad (1.5.11)$$

For the considered network

$a_2 = LCR$, $a_1 = L$, $a_0 = R$, $b_2 = b_1 = 0$, $b_0 = 1$.

3. The converting of a differential equation DE into an integral equation IE by an integrating of it so times, what an order of the differential equation (here two times). Writing the integral equation, we use the symbols

$$\bar{i}^{(q)} = \underbrace{\int \dots dt \int i dt}_{q\text{-times}}, \qquad \bar{u}^{(q)} = \underbrace{\int \dots dt \int u dt}_{q\text{-times}}, \qquad (1.5.12)$$

$$a_2 i + a_1 \bar{i} + a_0 \bar{i}^{(2)} = b_2 u + b_1 \bar{u} + b_0 \bar{u}^{(2)}. \qquad (1.5.13)$$

The got integral equation is just attached to a realization of the two conditions. At the first, there is no a constant component of a voltage at the output voltage of a source *u*. The calculation is made separately for the alternating component on the described here method and for the constant component (we get an elementary scheme of the replacing for it from the issue, closing the branches with the inductance and disconnecting the branches with the capacities). At the second, the constants of an integration, giving an account to a damping transitional components of a process, are lowing, and the aim of a converting is a getting of an integral equation for an established regime, for which the effective values of the alternating amounts are determined.

4. A converting of the integral equation IE relatively a momentary values of the alternating amounts into the algebraic equation AE relatively an effective values of the alternating amounts by the next operator: IE brings to square and averages at a period of a voltage, that is

$$\frac{1}{T}\int_0^T (\mathrm{IE})^2 dt \Rightarrow \mathrm{AE}. \qquad (1.5.14)$$

It is enough to consider in detail a procedure of an algebraization conformably only to the left part of the equation, because left and right parts of the differential equation (1.5.13) are analogical exactly to the designations.

In according to (1.5.14) we get

$$\frac{1}{T}\int_0^T \left(a_2 i + a_1 \bar{i} + a_0 \bar{i}^{(2)}\right)^2 dt = \frac{a_2^2}{T}\int_0^T i^2 dt + \frac{a_1^2}{T}\int_0^T \left(\bar{i}\right)^2 dt + \frac{a_0^2}{T}\int_0^T \left(\bar{i}^{(2)}\right)^2 dt +$$
$$+ \frac{2a_2 a_1}{T}\int_0^T i\bar{i} dt + \frac{2a_1 a_0}{T}\int_0^T \bar{i}\bar{i}^{(2)} dt + \frac{2a_2 a_0}{T}\int_0^T i\bar{i}^{(2)} dt \qquad (1.5.15)$$

It is showed at the common theory of the method ADE [21], that the average values from the product of the integrated a different number of times alternating periodic functions (here - currents) submit to the next producing correlation:

Attached to (q_1+q_2) – not even

$$\frac{1}{T}\int_0^T \bar{i}^{(q_1)}\bar{i}^{(q_2)}dt = \begin{cases} 0 \\ \pm\left|\bar{I}^{\left(\frac{q_1+q_2}{2}\right)}\right|^2 & \text{attached to } (q_1+q_2) - \text{even} \end{cases} \qquad (1.5.16)$$

The sign plus is taken in cases, when a difference of the indexes q_1 and q_2 is multiple to four, and the sign minus – when a difference of the indexes q_1 and q_2 is multiple to two.

Taking into account this fact, the fourth and fifth integrals at (1.5.15) will be equal to zero, and the integral equation, converted into the algebraic equation, will take a look:

$$a_2^2 I^2 + \left(a_1^2 - 2a_2a_0\right)\left(\bar{I}\right)^2 + a_0^2\left(\bar{I}^{(2)}\right)^2 = b_2^2U^2 + \left(b_1^2 - 2b_2b_0\right)\left(\bar{U}\right)^2 + b_0^2\left(\bar{U}^{(2)}\right)^2, \qquad (1.5.17)$$

where at the right part of the equation there are used a designations for the effective values of the integrals of a voltage, which are analogical to designations at the left part:

$$\left(\bar{U}\right)^2 = \frac{1}{T}\int_0^T \bar{u}^2 dt, \qquad (1.5.18)$$

$$\left(\bar{U}^{(2)}\right)^2 = \frac{1}{T}\int_0^T \left(\bar{u}^{(2)}\right)^2 dt \qquad (1.5.19)$$

and in the common case

$$\left(\bar{U}^{(q)}\right)^2 = \frac{1}{T}\int_0^T \left(\bar{u}^{(q)}\right)^2 dt. \qquad (1.5.20)$$

5. The additional equation of an algebraic equation (1.5.17), maintaining three unknown quantities (I, $\bar{I}$, $\bar{I}^{(2)}$), by two additional equations for getting a solvable system of the algebraic equations relatively an effective value of a current I. The missing algebraic equations got within the bounds of a first level of an approximation N = 1 attached to an assumption, that

$$\bar{i} \approx \bar{i}_{(1)}, \qquad \bar{I} \approx \bar{I}_{(1)} = \frac{I_{(1)}}{\omega} \qquad (1.5.21)$$

$$\bar{i}^{(2)} \approx \bar{i}^{(2)}_{(1)}, \qquad \bar{I}^{(2)} \approx \bar{I}^{(2)}_{(1)} = \frac{I_{(1)}}{\omega^2} \qquad (1.5.22)$$

This assumption is physically substantiated by that the integration of an alternating nonsinusoidal function brings to an improvement of the sinusoidal degree of the integrated signal, because attached to an integration all the k-harmonics are decreasing in k times relatively the first harmonic. Another saying, a mathematical operation of an integration of the distorted sinusoid means technically the filtration, bringing to an improvement a signal/noise (first harmonic/higher harmonics). In a case of the distorted currents the first level of an approximation attached to a calculation may be not enough and so the decision is building for the higher levels of an approximation [21].

6. A decision of the got system of an algebraic equations (1.5.17), (1.5.21), (1.5.22), taking into account that a foundation of the first harmonic of a current I(1) do not represents a problem and may be executed by the method ADE(1),too. It is necessary to convert before a writing of a decision the amounts $\bar{U}$, $\bar{U}^{(2)}$ to the more obvious form, evident from the next expressions.

$$U = \sqrt{\sum_{k=1}^{\infty} U_{(k)}^2} = U_{(1)}\sqrt{1 + \sum_{k=2}^{\infty}\left(\frac{U_{(k)}}{U_{(1)}}\right)^2} = U_{(1)}\sqrt{1 + K_h^2} \qquad (1.5.23)$$

or in the common case

$$\overline{U}=\sqrt{\sum_{k=1}^{\infty}\left(\overline{U}_{(k)}\right)^2}=\frac{U_{(1)}}{\omega}\sqrt{1+\sum_{k=2}^{\infty}\left(\frac{U_{(k)}}{kU_{(1)}}\right)^2}=\frac{U_{(1)}}{\omega}\sqrt{1+\left(\overline{K}_{\text{h}}\right)^2},$$

$$\overline{U}^{(2)}=\sqrt{\sum_{k=1}^{\infty}\left(\overline{U}^{(2)}_{(k)}\right)^2}=\frac{U_{(1)}}{\omega^2}\sqrt{1+\sum_{k=2}^{\infty}\left(\frac{U_{(k)}}{k^2U_{(1)}}\right)^2}=\frac{U_{(1)}}{\omega^2}\sqrt{1+\left(\overline{K}^{(2)}_{\text{h}}\right)^2},$$

$$\overline{U}^{(q)}=\sqrt{\sum_{k=1}^{\infty}\left(\overline{U}^{(q)}_{(k)}\right)^2}=\frac{U_{(1)}}{\omega^q}\sqrt{1+\sum_{k=2}^{\infty}\left(\frac{U_{(k)}}{k^qU_{(1)}}\right)^2}=\frac{U_{(1)}}{\omega^q}\sqrt{1+\left(\overline{K}^{(q)}_{\text{h}}\right)^2}\qquad(1.5.24)$$

where

$\overline{K}^{(q)}_{\text{h}}=\sqrt{\sum_{k=2}^{\infty}\left(\frac{U_{(k)}}{U_{(1)}k^q}\right)^2}$ - *the integral coefficient of the harmonics* of a q-order.

It differs from an usual coefficient of harmonics K_h , considered at electrical engeneering, by that it product a suspended (on the number of a harmonic) summation of the harmonics, what allows to modulate an action of the amplitude-frequency characteristic of the ideal electric network of the corresponding order and to forecast at this base a quality of a current at the network without a calculation of the current, as it will be seen later.

A *differential coefficient of the harmonics of a voltage* of the q-order is introduced analogically, if to give formally to a negative values q the meaning of a not operation of an integration of a signal already, but an operation of the differentiation. We will use for this case the second designation (with a tick), more narrow (only for operation of the differentiation), but and more comfortable (in this cases).

$$\overline{u}^{(-q)}=\breve{u}^{(q)}=\frac{d^qu}{dt^q}$$

$$\breve{U}^{(q)}=\overline{U}^{(-q)}=\omega^qU_{(1)}\sqrt{1+\sum_{k=2}^{\infty}\left(\frac{U_{(k)}}{U_{(1)}}k^q\right)^2}=\omega^qU_{(1)}\sqrt{1+\left(\breve{K}^{(q)}_{\text{h}}\right)^2}\qquad(1.5.25)$$

where $\breve{K}^{(q)}_{\text{h}}=\sqrt{\sum_{k=2}^{\infty}\left(\frac{U_{(k)}}{U_{(1)}}k^q\right)^2}$ - *differential coefficient of the harmonics* of a q-order.

Here a summation of the corresponding harmonics, intensified in k^q times, characterizes their underlining at the ideal differential network of a q-order.

A formula for a calculation of the effective value of a current at the considered network of the second order attached to an action of a nonsinusoidal voltage of the arbitrary, but a known form within the bounds of the first level of an approximation have a look (issue from (1.5.17), taking into account (1.5.24) and that the coefficients $b_2=b_1=0$ for the concrete considered scheme at a figure 1.5.1).

$$I^2=\frac{U^2_{(1)}}{a_2^2}\left\{\frac{b_0^2\left[1+\left(\overline{K}^{(2)}_{\text{h}}\right)^2\right]^2}{\omega^4}-\left(a_1^2-2a_2a_0\right)\frac{I^2_{(1)}}{U^2_{(1)}\omega^2}-a_0^2\frac{I^2_{(1)}}{U^2_{(1)}\omega^4}\right\}.\qquad(1.5.26)$$

Attached to it a parameters of the electric network (C,L,R) «are hidden» at the coefficients of the differential equation (a_2 , a_1 , a_0 , b_2 , b_1 , b_0),and parameters of the nonsinusoidal voltage are represented by an admission $U_{(1)}$,K_h , $\bar{K}_h$, $\bar{K}_h^2$.

There is founded a dependence of an effective current value only from two parameters of the voltage – $U_{(1)}$ and $\bar{K}_h^2$ for the considered scheme at a figure 1.5.1. So, really, the integral coefficient of the harmonics of a voltage of the second order determines here the quality of a current, without a foundation of the current or the spectrum.

A conclusion of the common correlation for a calculation of the electric network of the arbitrary order attached to any level of an approximation is given in [21].

1.5.2.3.2.The method ADE2

There is demanded a calculation of an effective value of the first harmonic of a current $I_{(1)}$ at a process of a calculation of an effective value of a current I by the method ADE1. But attached to a founded value $I_{(1)}$ the effective value of a current it is possible to get from the correlation

$$I = \sqrt{I_{(1)}^2 + I_{\mathrm{hh}}^2} \tag{1.5.27}$$

where I_{hh} – an effective value of that part of the current curve, which is conditioned by the presence of the higher harmonics in it (distortions of a sinusoid). The method ADE2 gives a possibility to calculate values of a current I_{hh} , attached to it the formulas for this component of a current it is may to get meaningfully simpler then the formulas for a calculation of the whole current I, because the members, answering for the determination $I_{(1)}$, are excluded from this formulas.

A procedure of an algebraization of the differential equations for the high frequency component of the current i_{hh} is analogical to a procedure, considered higher for all current i. As there are possible a decomposition of a process and an applying of a method of the superposition at a linear scheme of a replacement, so

$$i = i_{(1)} + i_{hh} \tag{1.5.28}$$

1. A composition of the equal electric scheme of a replacement for the higher harmonics of a figure 1.5.1. In a case of the linear load with the constant parameters this scheme of a replacement has the same topology, as an issue scheme of a replacement for all process at a figure 1.5.1. In a case of the linear load with the alternating parameters, as, for example, the asynchronous engine has the equal scheme of a replacement, a scheme of a replacement for the higher harmonics will differ by the topology and parameters [21].
2. A getting the differential equation for the high frequency component of the interesting alternating amount, here i_{hh} . In the common case of the system of a second order, analogically to (1.5.11), we will have

$$a_2\frac{d^2 i_{\mathrm{hh}}}{dt^2} + a_1\frac{di_{\mathrm{hh}}}{dt} + a_0 i_{\mathrm{hh}} = b_2\frac{d^2 u_{\mathrm{hh}}}{dt^2} + b_1\frac{du_{\mathrm{hh}}}{dt} + b_0 u_{\mathrm{hh}}\,. \tag{1.5.29}$$

3.A converting of a differential equation into an integral equation by the double integration.

$$a_2 i_{\mathrm{hh}} + a_1\bar{i}_{\mathrm{hh}} + a_0\bar{i}_{\mathrm{hh}}^{(2)} = b_2 u_{\mathrm{hh}} + b_1\bar{u}_{\mathrm{hh}} + b_0\bar{u}_{\mathrm{hh}}^{(2)} \tag{1.5.30}$$

4.A converting of an integral equation into an algebraic equation in according to
the operator (1.5.14). As the left and right parts of the equation (1.5.29) are identical again with an exactness to the designations, we will consider in detail a procedure of an algebraization only of the left part, and a result for the right part we will write on the analogy.

$$\begin{aligned}\frac{1}{T}\int_0^T \left(a_2 i_{\mathrm{hh}} + a_1\bar{i}_{\mathrm{hh}} + a_1\bar{i}_{\mathrm{hh}}^{(2)}\right)^2 dt = \frac{a_2}{T}\int_0^T i_{\mathrm{hh}}^2 dt + \frac{a_1}{T}\int_0^T \left(\bar{i}_{\mathrm{hh}}\right)^2 dt + \frac{a_0}{T}\int_0^T \left(\bar{i}_{\mathrm{hh}}^{(2)}\right)^2 dt + \\ + \frac{2a_2a_1}{T}\int_0^T i_{\mathrm{hh}}\bar{i}_{\mathrm{hh}}dt + \frac{2a_1a_0}{T}\int_0^T \bar{i}_{\mathrm{hh}}\bar{i}_{\mathrm{hh}}^{(2)}dt + \frac{2a_2a_0}{T}\int_0^T i_{\mathrm{hh}}\bar{i}_{\mathrm{hh}}^{(2)}dt\end{aligned} \tag{1.5.31}$$

Taking into account a producing correlation (1.5.16), applied to the left (1.5.31) and right parts of the algebraic equation, we will get

$$a_2 I_{\text{вг}}^2 + \left(a_1^2 - 2a_2 a_0\right)\left(\bar{I}_{\text{вг}}\right)^2 + a_0^2\left(\bar{I}_{\text{вг}}^{(2)}\right)^2 =$$

$$= b_2 U_{\text{вг}}^2 + \left(b_1^2 - 2b_2 b_0\right)\left(\bar{U}_{\text{вг}}\right)^2 + b_0^2\left(\bar{U}_{\text{вг}}^{(2)}\right)^2 \qquad (1.5.32)$$

5. The additional equation of an algebraic equation (1.5.32), maintained three unknown quantities I_{hh} , I_{hh} , $I_{hh}^{(2)}$, by two additional equations. Issue from the same physical facts, as attached to the additional equation at the method ADE1, on the base of (1.5.21) and (1.5.22) we get

$$i = i_{(1)} + \bar{i}_{\text{hh}} \approx \bar{i}_{(1)}, \quad \bar{i}_{\text{hh}} = 0, \quad \bar{I}_{\text{hh}} = 0, \qquad (1.5.33)$$

$$\bar{i}^{(2)} = \bar{i}_{(1)}^{(2)} + \bar{i}_{\text{hh}}^{(2)} \approx \bar{i}_{(1)}^{(2)}, \quad \bar{i}_{\text{hh}}^{(2)} = 0, \quad \bar{I}_{\text{hh}}^{(2)} = 0. \qquad (1.5.34)$$

6. The decision of the got system from three algebraic equations. As a result we have the next formula for a calculation of an effective value of the high frequency component of the current:

$$I_{\text{hh}}^2 = \frac{1}{a_2^2}\left[b_2 U_{\text{hh}}^2 + \left(b_1^2 - 2b_2 b_0\right)\left(\bar{U}_{\text{hh}}\right)^2 + b_0^2\left(\bar{U}_{\text{hh}}^{(2)}\right)^2\right] =$$

$$= \frac{U_{(1)}^2}{a_2^2}\left[b_2 K_{\text{h}}^2 + \left(b_1^2 - 2b_2 b_0\right)\left(\frac{\bar{K}_{\text{h}}}{\omega}\right)^2 + b_0^2\left(\frac{\bar{K}_{\text{h}}^{(2)}}{\omega^2}\right)^2\right] \qquad (1.5.35)$$

Comparing this formula with a formula (1.5.26) of the method ADE1, we saw the simplification, which will be more essential for the considered scheme, where $b_2 = b_1 = 0,\ b_0 = 1$, that gives

$$I_{\text{hh}} = \frac{U_{(1)}}{a_2 \omega}\left(\bar{K}_{\text{h}}^{(2)}\right)^2, \qquad (1.5.36)$$

that is a degree of the distortion of a current at the active load with a filter of the second order is a forward proportional to an integral coefficient of the harmonics of a voltage of the second order. A brought result one more time illustrates a possibility of the new quality indexes of the nonsinusoidal voltage (current) – of the integral coefficients of the harmonics of a voltage (current) of the high orders, which determine forwardly a quality of a current (voltage) at the networks of the corresponding orders without an analysis of the processes in the network.. So, a system of the classic indexes of the processes quality, represented at a part 1.2.1, must be added by this new.

The decisions, got at the methods ADE1 and ADE2, are the asymptotically approximated, the exactness of which depends from an exactness of the made assumptions at the stage of an additional equation of the algebraic equations and increases with an increasing of a level of the approximation. At the majority of the power electronic devices a character of the electromagnetic alternating amounts is such, that usually it is enough only the first level of an approximation for getting the engineer error of a calculation in 10…20 %

1.5.2.3.3.The method ADE(1)

A calculation of the processes at the networks with a valve converters on a first harmonic is not only necessary for the methods ADE1 and ADE2, but and has a self-dependent value for a calculation of the sinusoidal processes. It is possible to make the calculation within the bounds of the common methodology of an algebraization of the differential equations – at the method ADE(1). A procedure of the algebraization of the differential equation for the first harmonic

$$a_2 \frac{d^2 i_{(1)}}{dt^2} + a_1 \frac{di_{(1)}}{dt} + a_0 i_{(1)} = b_2 \frac{d^2 u_{(1)}}{dt^2} + b_1 \frac{du_{(1)}}{dt} + b_0 u_{(1)} \qquad (1.5.37)$$

is staying the same and brings to the next evident result:

$$a_2 I_{(1)}^2 + \left(a_1^2 - 2a_2 a_0\right)\left(\bar{I}_{(1)}\right)^2 + a_0^2\left(\bar{I}_{(1)}^{(2)}\right)^2 = b_2 U_{(1)}^2 + \left(b_1^2 - 2b_2 b_0\right)\left(\overline{U}_{(1)}\right)^2 + b_0^2\left(\overline{U}_{(1)}^{(2)}\right)^2 \qquad (1.5.38)$$

Taking into account (1.5.21) and (1.5.22) and analogically

$$\overline{U}_{(1)} = \frac{U_{(1)}}{\omega}, \qquad \overline{U}_{(1)}^{(2)} = \frac{U_{(1)}}{\omega^2}$$

from the equation (1.5.38) we get a formula for a calculation of an effective value of the first harmonic of a current through the first harmonic of a voltage and coefficients of a differential equation

$$I_{(1)}^2 = U_{(1)}^2 \frac{b_2^2 + \left(b_1^2 - 2b_2 b_0\right)\frac{1}{\omega^2} + \frac{b_0^2}{\omega^4}}{a_2^2 + \left(a_1^2 - 2a_2 a_0\right)\frac{1}{\omega^2} + \frac{a_0^2}{\omega^4}} \qquad (1.5.39)$$

Attached to it from the (1.5.39) it is evident the expression for a module (in a square) of a full resistance of a network on a first harmonic $Z_{(1)}$ or on any k-harmonic, if to change *w* to *kw*:

$$Z_{(1)}^2 = \frac{a_2^2 + \frac{a_1^2 - 2a_2 a_0}{\omega^2} + \frac{a_0^2}{\omega^4}}{b_2^2 + \frac{b_1^2 - 2b_2 b a_0}{\omega^2} + \frac{b_0^2}{\omega^4}}\,. \qquad (1.5.40)$$

If at the methods ADE1 and ADE2 the got decisions were approximated, so here a decision will be an exact, because there were no any assumptions at the stage of an additional equation of the algebraic equations.

In that case, if there is demanded to know a phase besides the module of the first harmonic of a current, so the fourth stage of the procedure of an algebraization will change, and now it is executed twice. At a process of the first converting the converting operator is aimed for getting the first algebraic equation with the effective value of a cosinusoidal component of a first harmonic $J_{(1)\,cos}$ of the row of Fourier, that is

$$\frac{2}{\sqrt{2}\,T}\int_0^T (\mathrm{IE})\cos\omega t\,dt = \mathrm{AE}_1\,. \qquad (1.5.41)$$

At a process of the second converting the converting operator is aimed already for getting the second algebraic equation with the effective value of a sinusoidal component of a first harmonic $J_{(1)\,sin}$ of the row of Fourier, that is

$$\frac{2}{\sqrt{2}\,T}\int_0^T (\mathrm{IE})\sin\omega t\,dt = \mathrm{AE}_2\,. \qquad (1.5.42)$$

There are determined, on a founded from a decision of two algebraic equations components $J_{(1)\,cos}$ and $J_{(1)\,sin}$ of the first harmonic of a current, the phase angle

$$\varphi = \operatorname{arctg}\frac{I_{(1)\cos}}{I_{(1)\sin}} \qquad (1.5.43)$$

and the resulting effective value of the first harmonic of the current

$$I_{(1)} = \sqrt{I_{(1)\sin}^2 + I_{(1)\cos}^2} \qquad (1.5.44)$$

1.5.2.3.4.The methods ADEP1,ADEP2,ADEP(1)

There was established at the part 1.2, that for a determination of an energetic indexes of the converting quality of an energy at the power electronic devices it is necessary a determination of the different powers else besides of a calculation of an effective and average values of the different alternating amounts. The last it is possible to do at versions of the methods of an algebraization of the differential equations for the powers – at the methods ADEP1, ADEP2, ADEP(1). The meaning of the figures at the designations of this methods is the same, as a meaning of the figures at the designations of the methods ADE1, ADE2, ADE(1).

A procedure of a junction from the algebraic equations to formulas for the unknown quantities powers is staying the same, as at the methods ADE, except for the stage number 4 of a converting of the integral equation into the algebraic. The look of the converting operator here is changing in according to a common integral determination of the partial component of the full power on a (1.5.5).

We will limit here by the short considering of the method ADEP2. It is necessary to have a differential equation for the higher harmonics of a current of this source i_{shh} in the case of the same scheme of a replacement at a figure 1.5.1 for a determination of the active power on the higher harmonics, taken away from a source of a voltage *u*. This equation is got with a help of the symbolic method, as an equation (1.5.10), and has a look

$$LC\frac{d^2 i_{u.\text{hh}}}{dt^2} + L\frac{di_{u.\text{hh}}}{dt} + Ri_{u.\text{hh}} = RC\frac{du_{\text{hh}}}{dt} + u_{\text{hh}}\,. \qquad (1.5.45)$$

A junction to a common form of the differential equation of a network of the second order (1.5.11) here is provided attached to

$a_2 = \text{LC}$, $a_1 = \text{L}$, $a_0 = \text{R}_1$ $\qquad$ $b_2 = 0$, $b_1 = \text{RC}$, $b_0 = 1$.

The integral equation, got from the common form of a differential equation

$$a_2 i_{u.\text{hh}} + a_1\bar{i}_{u.\text{hh}} + a_0\bar{i}_{u.\text{hh}}^{(2)} = b_1\bar{u}_{\text{hh}} + b_0\bar{u}_{\text{hh}}^{(2)}\,, \qquad (1.5.46)$$

.is converting into the algebraic equation relatively an active power of the source of a current. For it the equation (1.5.46) is multiplied from left and right at the high frequency component of a source voltage and the result is averaged (in according to an integral determination of the active power on a (1.5.2)), that is

$$\frac{1}{T}\int_0^T (\text{IE})\, u_{\text{hh}} dt \Rightarrow AE\,, \qquad (1.5.47)$$

what brings to the next correlation:

$$\frac{a_2}{T}\int_0^T i_{u.\mathrm{hh}} u_{\mathrm{hh}} dt + \frac{a_1}{T}\int_0^T \bar{i}_{u.\mathrm{hh}} u_{\mathrm{hh}} dt + \frac{a_0}{T}\int_0^T \bar{i}_{u.\mathrm{hh}}^{(2)} u_{\mathrm{hh}} dt = \\ = \frac{b_1}{T}\int_0^T \bar{u}_{\mathrm{hh}} u_{\mathrm{hh}} dt + \frac{b_0}{T}\int_0^T \bar{u}_{\mathrm{hh}}^{(2)} u_{\mathrm{hh}} dt \,. \tag{1.5.48}$$

It was substantially taken (1.5.33) and (1.5.34) attached to the additional equation within the bounds of a first level of an approximation.

So from the equation (1.5.48) we get

$$P_{u.\mathrm{hh}} = -\frac{b_0}{a_2}\left(\bar{U}_{\mathrm{hh}}\right)^2 = -\frac{b_0}{a_2} U_{(1)}^2 \left(\bar{K}_{\mathrm{h}}\right)^2 = -\frac{1}{LC} U_{(1)}^2 \left(\bar{K}_{\mathrm{h}}\right)^2 \tag{1.5.49}$$

The additional selection of a power from the source here is proportional to a square of the integral coefficient of the harmonics of a voltage of a first order of the input source.

1.5.2.3.5. The final observations

1.The methods of an algebraization of the differential equations, excluding a labour-intensive and not often possible at an analytic form procedure of the foundation of the momentary values of an alternating amounts, allow to establish a formula for a calculation an effective value of a current and a demanded power at a network with the nonsinusoidal voltage forwardly on the coefficients of the issue differential equation.

2.The methods are an asymptotically approximated; the exactness (and a difficulty) increases with an increasing of a level of the approximation. A calculation of an effective values of the currents and values of the powers of a nonsinusoidal processes it is preferably (on a labour-intensiveness and exactness) to do within the bounds of a second version of the method and a composition of a common decision in according to (1.5.27) for a current and .$M = M_{(1)} + M_{b2}$ for the powers. An increasing of the exactness happens here because of that the components of a decision on the first harmonic ($I_{(1)}$,, $M_{(1)}$) are determined by the forward methods exactly, and just they are usually dominating components at the common decision.

A division of a calculation procedure of the characteristics of the power electronic devices, working with nonsinusoidal currents, is justified in a methodical plan. Usually at first there is executed a draft calculation of an energetic processes on the smooth components (on the first harmonics – at the networks of an alternating current and on the average values – at the networks of a direct current), determining the elementary look of a device. Later there are calculated the processes on the higher harmonics, characterizing a quality of the converting and converted energy. Then the characteristics of the energetic processes are corrected on the smooth components, taking into account an additional influence of the distortions on the higher harmonics.

3. A considered form of the mathematical model of an electric network at a look of a differential equation of a n-order allows to determine an energetic characteristic of one alternating amount for a one circle of a calculation. A using of the mathematical model of an electric network as a system of the differential equations of the first order allows by the matrix converting for a one circle of a calculation to determine the energetic characteristics of all alternating amounts of the network state together, attached to it a mathematical model of a network may have an alternating coefficients [20].

QUESTIONS TO THE CHAPTER 1

I shall be intoxicated by a harmony at times again,
I shall be in a flood of tears over the imagination.

A. S. Pushkin

1. What is the factor, forming a system, for a majority of the elements?
2. What is a procedure of a system analysis of the technical systems?

3. What is the principal difference of the system properties from a totality of the properties of the elements, forming it?
4. By what admission of the majorities a system is determined formally?
5. What kinds of problems of a systems investigation are known?
6. Enumerate the energetic criterions of a quality of the electromagnetic processes.
7. Enumerate the energetic criterions of a quality of the converting device of an electric energy at the semiconductor valves.
8. Enumerate the quality criterions of a construction of the valve converter.
9. What is the difference of the completely control valves from the not completely control valves?
10. By what current parameters a valve with not full control is characterized?
11. By what voltage parameters a valve with not full control is characterized?
12. By what parameters the dynamic properties of the valves with not full control are characterized?
13.What parameter of a turned off thyristor characterizes the ability to turn on?
14. Enumerate types of the power transistors.
15.What are the advantages of IGBT- transistors before MOSFET- transistors?
16. Enumerate the basic types of valve converters.
17. What is the commuting function of a valve?
18. What is the commuting function of a valve group of a converter?
19. Enumerate the methods of an analysis of the energetic criterions.
20. What are merits and demerits of the integral method of an analysis?
21. What are merits and demerits of the spectral method of an analysis?
22. What are merits and demerits of a method of the forward algebraization of the differential equations?
23. What is a calculation procedure at the method ADE1?
24. What is a calculation procedure at the method ADE2?
25. What is a calculation procedure at the method ADE (1)?
26. What is a calculation procedure at the method ADEP1?
27. What is a calculation procedure at the method ADEP (1)?

PROBLEMS

And fingers are just asking to a pen,
A pen - to a paper…
A. S. Pushkin

1.* To calculate a coefficient of the harmonics and a coefficient of the distortion for a function at a look of a meander.
2.* To calculate a coefficient of the harmonics and a coefficient of the distortion for a function at a look of a module of a sinusoid.
3. A converting device has the specific constructive indexes M_s = 5 kg/kVA and V_s = 3 dm^3/kVA. To determine an index of the specific weight of the device M_v.
4. A converting device with the output power P = 400 kWt has a CUA 0,9, an input coefficient of a power 0,8. A weight of a converter 300 kg, a size 200 dm^3. To determine all the specific indexes of a converter.
5.* To deduce a formula for a coefficient of the harmonics of a current at a consecutive RL-network attached to an influence of the alternating nonsinusoidal voltage of any form.
6.* To deduce a formula for a coefficient of a current distortion at a consecutive RL-network attached to an influence of a function at a look of a module of a sinusoid.
7. To calculate a integral coefficient of the harmonics of voltage for a function at a look of a meander.
8. To calculate a differential coefficient of the harmonics of voltage for a function at a look of a trapeze.
9. To deduce a formula of connection for a coefficient of the harmonics of a current with coefficient of current distortion.
10. To deduce a formula of connection of a power loss with efficiency of rectifier.

2. THEORY OF CONVERTING ALTERNATING CURRENT INTO DIRECT ATTACHED TO IDEAL PARAMETERS OF CONVERTER

We all studied a little
To anything and somehow...
A.S. Pushkin

I am studying to take
pleasure in a truth.
A.S. Pushkin

2.1 RECTIFIER AS SYSTEM. GENERAL DETERMINATIONS AND DESIGNATIONS

An aim of the given part is the examination of a structure and variable quantities of a first type of a basic cell of an electric energy converting-a rectifier from the system positions and a determination what statements of the tasks of the research are possible.

The determination of any valve converter RS as a system it is may to do by a giving of the next admission of the multitudes of his describing:

$$\mathbf{PC = \{Z,S,P,V,X,Y\}} \qquad (2.1.1)$$

where **Z**- a set of the system purposes having a special aim;

S= {**PS;CS}** – a set of a system structure describing, consisting from a structure describing of a power scheme of a valve converter **PS** and a structure describing of a control system **CS**, given as a flour chart, a principal scheme, a graph, an incidence matrix and etc , and also a describing of a type of the structure elements ;

P = {PP;CP} – a set of the parameters of the elements of a power scheme **PP** and a control system **CP**;

V – a set of the input actions at a system (energetic inputs , inputs of the control task, actions of the environment);

X – a set of the interior variable quantities (voltages, currents, powers);

Y – a set of the output variable quantities (energetic outputs, signal outputs to connect with a subsystem, actions at an environment (electromagnetic, thermal and etc).

There are possible four statements **of the research task of a valve converter** depending on the aims of a research.

1. *A task of an analysis.* There are given a structure and parameters of the system and the output variable quantities, i.e. **S, V, P**. It is necessary to find the interior and output variable quantities of the system, what allows formulating properties of the given converter depending on the results. It is the first and a general task attached to a learning of the valve converters. This task is always decidable.
2. *A task of an optimization.* There are given a structure, input and output variable quantities, i.e. **S, V, Y**. It is necessary to determine a set of the parameters of the elements **P**, providing the extremism of any purposes **Z** attached to given limitations at a set of the interior variable quantities $\mathbf{X_{lim}}$. Also they call this task as a task of a parametrical optimization, and her decision is determined by a dimension of the task («a dimension damnation» at the complicated systems). A task is appearing attached to a projecting or a modernization of the known (given) converter schemes and usually it admits a precision or an approximated decision.
3. *A task of a synthesis.* There are given sets of the input and output variable quantities, i.e. **V,Y** or just a set of the output variable quantities **Y**. It is necessary to determine a structure, a set of element parameters and a set of the interior variable quantities, i.e. **S, P, X** (in the first case and also additionally **V** – in the second case). A procedure of a decision is formalizing not completely and consists from heuristic receptions besides mathematical operations, what makes a decision not one-valued. The most difficult is the determination of a system structure, demanding a presence of your experience, an intuition and expert advice. The problem is important for a working up the

systems of the automatized projecting of the valve converters, for a generation the new converter schemes.

4. *A task of identification.* There are given sets of the input and output variable quantities, i.e. **V, Y.** It is demanded to determine a structure and parameters of a system **S** and **P**, considered as a «black box». It is a task of a structural and parametrical identification. The task of a structural identification is not decided in general case. The task of a set determination of the parameters becomes a task of a parametrical identification if the system structure is given, i.e. a determination of interior system parameters, what allows to find his interior parameters depending on the results of a measuring of input and output variable quantities of a «black box», i.e. to make a «black box» limpid («white»).

In general case a rectifier structure is showed at a flour chart of a figure 2.2.1 at a level of the elementary cells.

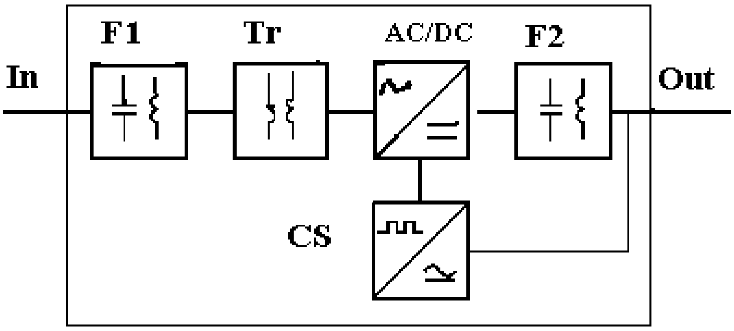


Fig. 2.2.1

Besides a basic cell AC/DC, where a converting of an alternating current into a direct (rectified) goes on, there are a cell T of the input transformer, a cells of the input F1 and output F2 filters and a cell CS of a control system of valves of a cell AC/DC, signals from voltage and current data units of rectifier output and input go to the inputs of a cell AC/DC, if it is necessary.

A purpose of rectifiers having a special aim in this part is a converting of an alternating voltage into a direct not controlled by the basic rectifying cells at ideal elements. There will be showed a common possibility of a direct voltage regulation at all basic rectifying cells at end of the part.

A common rectifier structure {S} is given in this part by a flour chart, in the next parts the concrete basic cells structures will be imagined by principal schemes of cells.

A set of parameters values of the power scheme elements {PP} within the bounds of this part consists only from two parameters values of elements: A zero and an infinity because all elements of a scheme are ideal at this stage of an analysis. A set of parameters of the control system elements {CP} will be given at part 2 too.

For tasks of sets of input, interior and output variable quantities we shall take a designation system in a rectifier.

There will be used the next designation system of variable quantities attached to an electromagnetic processes analysis in a rectifier. All the instantaneous values of emf, voltages and currents are designated by small letters *e, u, i,* and integral values of this variable quantities (effective, average, extreme) are designated by capital letters *E, U, I.* All variable quantities, having a relation to a maim, an input filter and primary winding of a transformer, are designated with an index 1 (u_1, i_1), variable quantities, having a relation to a secondary winding of a transformer, are designated with an index 2 (u_2 ,i_2), variable quantities, having a relation to rectified current network, are designated with an index *d* (u_d, i_d) (from an English word «direct»). A phase number of an alternating current is designated as *m*, a frequency of an alternating voltage *f* and $\omega=2\pi f$. Powers are designated as S - a full power, P – an active power, $s = ui$ – an instantaneous power, *Q* – a reactive power, calculated as a geometrical bound between a full and an active powers

$$Q = \sqrt{S^2 - P^2}.$$

A set of input actions of a not control rectifier is determined by a set of phase voltages of a main. Currents and voltages of transformer windings determine a set of the interior variable quantities of a rectifier, by currents and voltages of valves, by currents and voltages of elements of input and output rectifier filters.

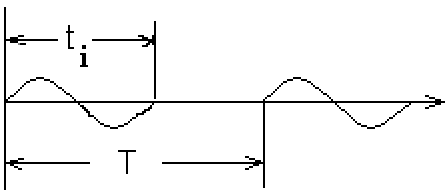


Fig. 2.1.2

They are differing two types of the basic rectifying cells: *half – wave* and *full – wave*. Half – wave schemes are used to rectify (to take out a power from an alternating current main) only a one half – wave of an alternating voltage from two in each period. Full – wave schemes are used to rectify both half - waves in each period of an input alternating voltage. The conditional designations for the schemes are correspondingly $q = 1$ and $q = 2$. For characterizing a number of used half – waves of an input multiphase alternating voltage for a period there is introduced a *pulseplicity* of a rectifier $p = qm_2$, determining also a number of pulsations of rectified voltage for a period of a maim voltage.

To exclude from the calculating correlation for a frequency rectifier of a maim voltage ω_1 and give to the correlations the universal character there is
introduced a *dimensionless time* $\theta = \omega_1 t$. So the formulas to calculate the effective values F_e and average F_{av} values of functions *f(t)* will take such a look.

$$F_e = \sqrt{\frac{1}{T}\int_0^T f^2(t)\,dt} = \sqrt{\frac{1}{\omega T}\int_0^{\omega T} f^2(\vartheta)\,d\vartheta} = \sqrt{\frac{1}{2\pi}\int_0^{2\pi} f^2(\vartheta)\,d\vartheta} \tag{2.1.2}$$

$$F_{av} = \frac{1}{T}\int_0^T f(t)\,dt = \frac{1}{\omega T}\int_0^{\omega T} f(\vartheta)\,d\vartheta = \frac{1}{2\pi}\int_0^{2\pi} f(\vartheta)\,d\vartheta \tag{2.1.3}$$

The feature of all electromagnetic variables at a valve converter is a piecewise smooth character, often with zero pauses, because of the discrete valve work. In this case, such function becomes a standard function with known for her effective $F_{e.e}$ and average $F_{ev.e}$ values if to exclude the zero pauses from it, the formulas (2.1.2) and (2.1.3), taking in account the designations of a figure 2.1.2, become the simple correlations by not difficult converting

$$F_E = \sqrt{\frac{1}{T}\int_0^T f^2(t)dt\frac{t_i}{t_i}} = \sqrt{\frac{t_i}{T}\frac{1}{t_i}\int_0^{t_i} f^2(t)dt} = \frac{F_{E.i}}{\sqrt{q_o}}, \tag{2.1.4}$$

$$F_{av} = \frac{1}{T}\int_0^T f(t)dt\frac{t_i}{t_i} = \frac{1}{t_i}\int_0^{t_i} f(t)dt = \frac{F_{av.i}}{q_o}, \tag{2.1.5}$$

where $q_0 = T/t_i$ *– a porosity of an impulse function.*

A rectifier, made from not control valves, is called *a not control rectifier* and is purposed to get a direct voltage of not alternating value. A rectifier, made from control valves, is called *a control rectifier* and is purposed to get a regulated and (or) stabilized direct voltage.

2.2. MECHANISM OF CONVERTING ALTERNATING CURRENT INTO RECTIFIED AT BASIC CELL AC/DC

The aim of this part is an acquaintance with a common converting mechanism of an alternating (bi-directional) current into an onedirectional pulsing (direct) by only valves without using another scheme elements, what allows «in a clear look» to show a specificity of converting an energy of this kind.

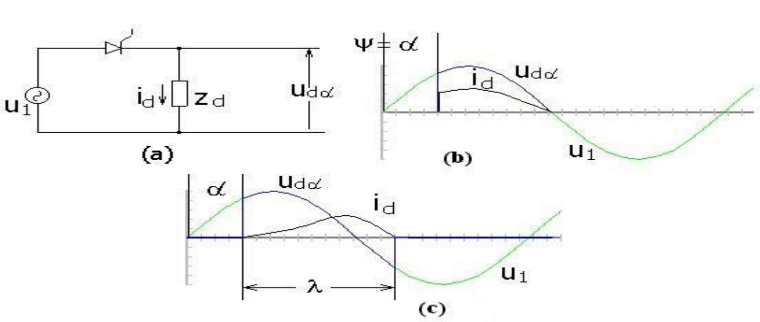


Fig.2.2.1

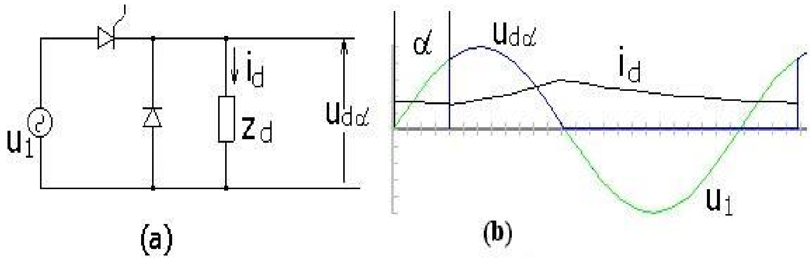


Fig.2.2.2

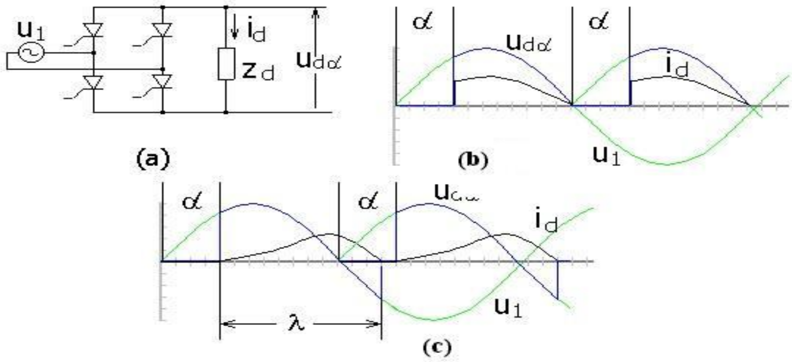


Fig.2.2.3

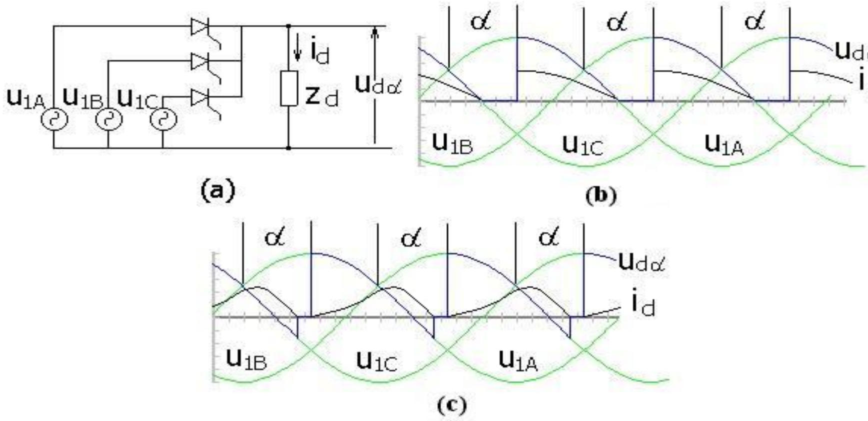


Fig.2.2.4

A scheme of the most simple (one-valve) basic cell of the half – wave control rectifying of an one-phase current is showed at a figure 2.2.1, a, temporary diagrams of a rectified voltage u_d and rectified current i_d are showed at a figure 2.2.1,b for a case of an active load, and at a figure 2.2.1,c – for an active-inductive load.

An inductance at a network of rectified current L_d may consist from its own load inductance (winding) and a filter inductance for a pulsation smoothing of rectified current and then not divide to the component parts. A current at a load continue to flow because of the inductance presence and after changing a sign of a maim voltage current flows reverse to it because of an energy, accumulated at a magnetic field of an inductance L_d as long as it does not expenditure at a load resistance R_d and does not come back to a maim.

It is characteristically that a rectified current has an interrupted character, i.e. current impulses are divided by the zero pauses. *An interrupted rectified current* of a rectifier, as it will be seen from the next analysis, lead to a distortion of all general rectifier characteristics and, as a rule, is not desirable. It is necessary to reduce the existence field or to eliminate it completely:

- a supplying of *a zero valve* V_o, as it is showed at a figure 2.2.2, a;
- an increasing of a rectifying semi period from $q = 1$, as it was considered some time ago, till $q = 2$ (a changing of so called zero rectifying schemes or schemes with an output of a zero power supply point, as they are called also, at a bridge), as it is showed at a figure 2.2.3, a;

- a phase number increasing of an alternating voltage of a rectifier, as it is showed at a figure 2.2.4, a
- an increasing of a time constant of a load because of an increasing of a filter inductance L_d.

At a scheme with a zero valve V_o it begins to work attached to a maim voltage polarity changing and lead conduct a load current during an interval T2because of an energy, accumulated at a magnetic field of a filter inductance L_d.

Valves 1,2 conduct a current at a scheme of a bridge rectifying at a positive half-wave of a maim voltage, and at a negative –valves 3,4, because an impulse frequency of a rectified current attached to a full-wave rectifying increases in two times in comparison with a half-wave, represented at a figure 2.2.1, a.

The next pulsation frequency increasing of a rectified current till $f_p = qm_2f_1$ is provided attached to a phase number increasing of a maim voltage, as you can see from a figure 2.2.4 for a three-phase maim. Pulsations of a rectified voltage, which are estimated by a pulsation coefficient of a voltage K_p, reduce attached to it.

So as a regime of an interrupted rectified current is not very qualitative for a consumer, it is necessary to determine his limits at a space of a rectifier parameters, i.e. at a function R_d, L_d, α, m.

Evidently, that valves work independently from each over at this regime, because a differential equation will have such a look for a rectified current attached to one conducting valve (attached to an ideal valve)

$$Xd\,\frac{\mathrm{did}}{d\vartheta} + id \cdot Rd = e_2 = \sqrt{2}E_2 \sin\vartheta \tag{2.2.1}$$

because

$$Ld\,\frac{\mathrm{did}}{dt}\frac{\omega}{\omega} = \omega Ld\,\frac{\mathrm{did}}{d(\omega t)} = Xd\,\frac{\mathrm{did}}{d\vartheta}$$

Its decision

$$id = id_{forc} + id_{free} = \frac{\sqrt{2}E_2}{Z_d}\sin(\vartheta - \varphi) + A_1 e^{-\frac{\theta-\psi}{\omega\tau}}, \tag{2.2.2}$$

$$where\_Zd = \sqrt{Xd^2 + Rd^2},$$

$$\varphi = \mathrm{arctg}\,\frac{Xd}{Rd}$$

$$\omega\tau = \frac{Xd}{Rd}$$

$$A_1 = -\frac{\sqrt{2}\,E_2}{Zd}\sin(\psi - \varphi)$$

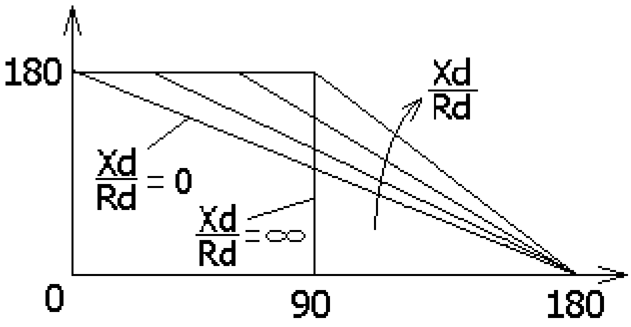
An integration constant A_1 is determined (on 2.2.2) from the first condition $i_d = 0$ attached to $\theta = \psi$. (2.2.3)

So the decision (2.2.2) will look so

$$id = \frac{\sqrt{2}\,E_2}{Zd}\left[\sin(\vartheta - \varphi) + \sin(\psi - \varphi)e^{-\frac{\vartheta-\psi}{\omega\tau}}\right] \tag{2.2.4}$$

From (2.2.4) we get an equation for a valve current flowing duration λ, if to take $i_d = 0$ attached to $\theta = \psi+\lambda$:

$$\sin(\psi + \lambda - \varphi) + e^{-\frac{\lambda}{\omega\tau}}\sin(\psi - \varphi) = 0 \tag{2.2.5}$$



$$\psi = \frac{\pi}{2} - \frac{\pi}{qm} + \alpha$$

Fig.2.2.5

This equation is transcendental relatively λ, because its decision by a numerical method gives dependence graphics $\lambda = f(\psi, Xd/Rd)$, represented at a figure 2.2.5.

A bond of a regulation degree α, counted down from
a point of a natural ignition (a

point of intersection of positive half-waves of a maim voltage), with a degree of a valve work begin ψ, counted down from a zero of a main voltage, is evident from a figure 2.2.4:

(2.2.6)

An integration constant A_1 at (2.2.1) is determined from a condition of an established regime for a regime of a continuous rectified current:

$$id\Big|_{\vartheta=\psi} = id\Big|_{\vartheta+\frac{2\pi}{qm}=\psi}$$

(2.2.7)

So a decision (2.6) will take a look:

$$id = \frac{\sqrt{2}\,E_2}{Zd}\left[\sin(\vartheta - \varphi) + \frac{\sin\left(\psi + \frac{2\pi}{qm} - \varphi\right) - \sin(\psi - \varphi)}{1 - e^{-\frac{2\pi}{qm\omega\tau}}} \cdot e^{-\frac{\vartheta-\psi}{\omega\tau}}\right]$$

(2.2.8)

Issue from a rectifying mechanism of an alternating voltage, proved by got analytical decisions for a rectified current at an interrupted (2.2.4) and continuous (2.2.8) regimes, there are the following **conclusions:**

1. A rectifier as a direct current supply, besides a regime of a continuous current, characteristic for traditional direct current supply (accumulators, direct current generators), has also a specific regime of an interrupted current even attached to a stationary load.
2. An inductance L_d at a regime of an interrupted rectified current at a load network influence on not only a pulsation value of a rectified current, but on the average value.
3. An inductance L_d at a regime of a continuous rectified current at a load network influence on only a pulsation value of a current, but does not influence on its constant component, i.e. its average value.
4. Pulsations of a rectified current at a continuous regime reduce when Rd reduces attached to the same value L_d, i.e. when an average current value increases, hence, and rectifier power. Because using of an inductive filter to smooth a rectified current at a power rectifiers (with a small value Rd) is practically the one acceptable way. Conductive smoothing filter properties (for a rectified voltage) will be considered at a part 3.3.5.
5. A valve cells addition by input transformers changes a character of electromagnetic processes at a rectifier input, because then there are considered basic cells of rectifiers in the conditions of the same assumptions to compare it between each over and determine rational fields of it applying. This **assumptions are the next:**

- A transformer is ideal, i.e. it is characterized by only the one parameter – a *transforming coefficient* $K_t = U_1/U_2$;
- Valves are ideal, i.e. they are changed by keys, having a zero resistance at a state «turn on» and an infinite resistance (disconnected network) at a state «turn off», as a result we take an influence of the concrete valve parameters on the rectified current parameters;

- An input filter is absence, an output filter is ideal, i.e. attached to an inductive filter $L_d\ (X_d) = \infty$, a rectified current does not consist from pulsations, as a result we take an influence of a concrete filter parameters on a rectified current parameters.

The using of considered assumptions at a mathematical model of a rectifier allows do not to use a complicated correlations between variables, got from using a differential equations method, as it was showed higher. Besides, if we do not take real parameters of rectifier elements we can find properties of converting process of an alternating current into a direct. Simple calculated correlations , got at this stage of an analysis, for rectifier elements in the next, at a second stage of a rectifier analysis taking into account real parameters of rectifier elements, will be just corrected , but not destroyed.

2.3. TWO-PHASE RECTIFIER OF ONE-PHASE CURRENT (m_1=1, m_2=2, q=1)

The same rectifying schemes not always are called the same (and correctly) at different books, because a formal scheme code is brought. There follows from it, that input transformer converts an one-phase maim voltage into two-phase, which is rectified at a half-wave rectifying scheme, showed at a figure 2.3.1.

An aim of a rectifier analysis is an establishment of its properties. An analysis of the scheme goes on into two stages. There is made a qualitative analysis of the electromagnetic processes at a scheme by the time diagrams of the instantaneous values for voltages and currents at the first stage. Then at the second stage on these diagrams there is made a quantitative analysis, which allows to get calculated correlations for all the scheme elements and to make conclusions about properties and acceptable field of the rectifier applying at its base.

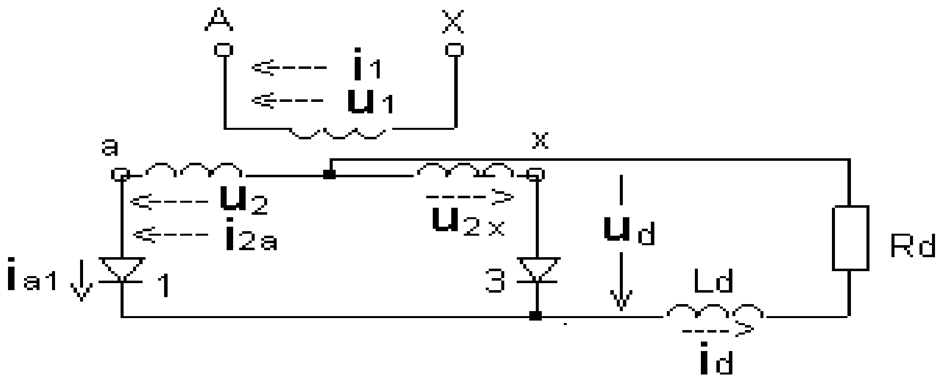


Fig. 2.3.1

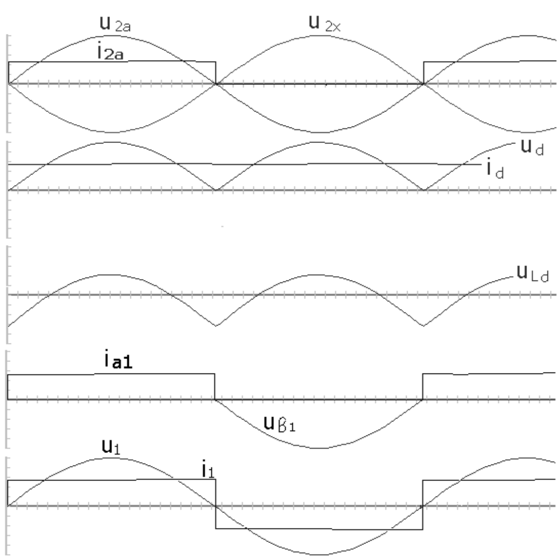


Fig. 2.3.2

Time diagrams of the rectifier voltages and currents attached to a regulating absence (α=0) are showed at a figure 2.3.2.

There are represented at a first diagram two-phase voltages of the secondary transformer windings u_{2a}, u_{2x} and a current at one secondary winding, a building methodic of which is described lower after a diagrams building of valves anode currents. **A rule of determination the conducting valve at a cathode valve group** (valves, connected by cathodes) is the next: that valve conducts a current, a potential of which anode is the most positive.

Valves of the cathode group are designated by not even numbers. There are represented at the second diagram the curves of rectified voltage u_d and rectified current i_d

attached to a made assumption of an ideal filter $X_d = \infty$. A curve of a rectified voltage repeats curves of the secondary voltages on intervals of conductivity of the corresponding valves. A rectified current does not consist from pulsations, and its instantaneous values are the same as its average value i_d. There is represented at a third diagram a voltage curve at a smoothing reactor, which takes all pulsation at an ideal case (alternating component) of the rectified voltage. There are showed at a fourth diagram an anode current of the first valve i_{a1} and a reverse voltage at it u_{b1}. When a valve 3 conducts a current, when u_{2x} is positive, through it a through-phase voltage u_{2a}-u_{2x} is supplied to a valve 1, i.e. a double amplitude value of a phase voltage u_2. A considered higher rule of conducting valve determination at a cathode group becomes an evident from this diagram: when an anode valve voltage is not the most positive, so a reverse voltage is supplied to it and it cannot conduct a current. From the other side, when a valve conducts a current, so a forward voltage drop is absence at it attached to a made assumption about an ideality of the valves.

If you know the valves anode currents, you can build currents at transformer secondary windings. Because there is connected one valve 1 at a secondary winding with a voltage u_{2a}, a form of winding current is the same as a form of a valve anode current, i.e. $i_{2a} = i_{a1}$, what is represented at the first time diagram. A form of the secondary current at a winding with a voltage u_{2x} is determined the same way, i.e. $i_{2x} = i_{a3}$, which, evidently, is an analog of the secondary current form i_{2a}, but it is biased at a time at a half of period of a maim voltage. And, finally, on the known currents forms at the secondary transformer windings at fifth time diagram there are build a current curve at a primary winding i_1 of the ideal transformer and a voltage curve of a primary winding u_1, a voltage u_{2a} of the transformer secondary winding is synphase with which an according with taken positive directions of the windings voltages, designed by arrows. A building methodic of a secondary current follows from equations for mdf of the transformer windings, bonded by a Kirchhoff law for magnetic networks:

$$i_1 w_1 = i_{2a} w_2 - i_{2x} w_2$$

$$i_1 = \left(i_{2a} - i_{2x}\right)\frac{w_2}{w_1} = \frac{i_{2a} - i_{2x}}{K_т},$$

(2.3.1)

where w_1, w_2 – a turns number of the primary and secondary windings correspondingly.

At a given scheme a current of a primary winding is equal to algebraic summa of the secondary windings currents, taken with a transforming coefficient K_t.

It is necessary to notice **a characteristic feature** of a half-wave rectifying scheme – **a one-direction of the currents** at the transformer secondary windings, what shows a presence of the constant components at it. But as a magnetic system (a rod from a transformer steal) of one-phase transformer is one-contour, so there will be no constant dc magnetization at a result magnetic flow at a rod, because the currents at two secondary windings have a different direction.

It is may notice also the next **features of the transformer using** on results of the qualitative analysis of electromagnetic processes at a researched rectifier. At first, a current forms difference at the secondary and primary transformer windings and, at second, its non-sinusoidal character. The first feature is bonded with a valves presence at the transformer secondary windings, while an alternating current flows at a primary winding, connected to an alternating voltage supply. The second feature is bonded with that a valve cell for a sinusoidal voltage network represents a not linear load, a current form at which depends on a kind of this non-linearity.

A second stage of rectifier analysis is a mathematical. At first, here, it is necessary to get the energetic indexes of the arrangement elements quality, i.e. calculated correlation to determine parameters of a transformer, valves, a filter through parameters of the direct current network, which are given attached to a projecting. At second, it is necessary to calculate the energetic indexes of a processes quality at a rectifier input and output. **A methodic of an analysis** attached to a first level of assumptions – assumptions about an ideality of the scheme elements – consists from the next **fifteen steps.**

1. A bond is established between an average value of a not control valve rectified voltage U_{do} with an effective voltage value of the transformer secondary winding from the corresponding time diagram at a figure 2.3.2.

$$U_{d0} = \frac{1}{\pi}\int_0^{\pi} u_d \, d\vartheta = \frac{1}{\pi}\int_0^{\pi}\sqrt{2}\, u_2 \sin\vartheta \, d\vartheta = \frac{2\sqrt{2}}{\pi} U_2 = 0.9U_2$$

(2.3.2)

from where

$$U_2 = \frac{U_{d0}}{K_{s}} = \frac{\pi}{2\sqrt{2}} U_{d0} = 1.11 U_{d0}$$

2. An average anode current value of a valve I_a is calculated

$$I_a = \frac{1}{2\pi} \int_0^{2\pi} i_a \, d\vartheta = \frac{1}{2\pi} \int_0^{\pi} I_d \, d\vartheta = \frac{I_d}{2}$$

(2.3.3)

3. An effective anode current value of a valve $I_{a.e}$ is calculated

$$I_{a.e} = \sqrt{\frac{1}{2\pi} \int_0^{2\pi} i_a^2 \, d\vartheta} = \sqrt{\frac{1}{2\pi} \int_0^{\pi} I_d^2 \, d\vartheta} = \frac{I_d}{\sqrt{2}}$$

(2.3.4)

A form coefficient of a valve anode current

$$K_F = \frac{I_{a.e}}{I_a} = \sqrt{2}.$$

(2.3.5)

4. A maximal value of a valve anode current is calculated

$$I_{a.max} = I_d$$

(2.3.6)

An amplitude coefficient of an anode current

$$K_a = I_{a.max}/I_a = 2 \qquad (2.3.7)$$

5. A maximal value of a reverse voltage at a valve relatively U_{do} is calculate (2.3.8)

$$U^*_{b.\max} = \frac{U_{b.\max}}{U_{d0}} = \frac{2U_{2.\max}}{U_{d0}} = \frac{2\sqrt{2}\,U_2}{U_{d0}}$$

6. An established valves power with not full control (thyristors) is calculated

$$S^*_{b1} = \frac{S_{b1}}{P_{d0}} = \frac{nU_{b\max}I_a}{P_{d0}} = \pi$$

(2.3.9)

with a full control (transistors, turn off thyristors)

$$S^*_{b2} = \frac{S_{b2}}{P_{d0}} = \frac{nU_{b\max}I_{a\max}}{P_{d0}} = 2\pi$$

(2.3.10)

7. An effective current value at a transformer secondary winding

$$I_2 = I_{a.e} = \frac{I_d}{\sqrt{2}}$$

(2.3.11)

8. An effective current value at a transformer primary winding

$$I_1 = \frac{I_d}{K_T}$$

(2.3.12)

a converting coefficient of a rectifier on a current is determined (2.3.13)

$$K^1_{CC} = \frac{I_d}{I_1} = K_T$$

9. A full power of the transformer secondary windings is calculated

$$S_2 = 2U_2 I_2 = 2\frac{\pi}{2\sqrt{2}}U_{d0}\frac{I_d}{\sqrt{2}} = \frac{\pi}{2}P_{d0} \qquad S_2^* = \frac{S_2}{P_{d0}} = \frac{\pi}{2}$$

(2.3.14)

where P_{do} – an active power at a not control rectifier output.

10. A full power of the transformer primary windings is calculated

$$S_1 = U_1 I_1 = K_{\text{т}}\frac{\pi}{2\sqrt{2}}U_{d0}\frac{I_d}{K_{\text{т}}} = 1.11P_{d0} \qquad S_1^* = \frac{S_1}{P_{d0}} = 1.11$$

(2.3.15)

11. There is calculated a typical established power of a transformer (having different windings full powers), determined in this case as

$$S_{\text{т}} = \frac{S_1 + S_2}{2} = \frac{1.11 + 1.57}{2}P_{d0} = 1.34P_{d0} \qquad S_{\text{т}}^* = \frac{S_{\text{т}}}{P_{d0}} = 1.34$$

(2.3.16)

12. The demanded value of a smoothing reactor L_d at a direct current network and its conventional established power are estimated.

Here we shall not take into account at this level of an analysis the made assumptions about an ideality of a rectified current smoothing, ($L_d = \infty$) to estimate the reactor expenditures. It is may to take a rectified current as an almost direct attached to a harmonic presence at a current (current pulsation) at a stage of some percents from an average current value with an engineer exactness.

We use a method ADE2 attached to a task of a harmonic coefficient of a rectified current K_{hc} for a calculation of a necessary reactor inductance. We suppose that all pulsation of the rectified voltage is applied to a filter (to a reactor), so a differential equation for a high-frequency current component will take a look

$$L_d\frac{di_{d.hh}}{dt} = u_{d.hh}.$$

After its algebraization

(2.3.17)

$$I_{d.hh} = \frac{1}{L_d}\overline{U}_{d.hh} = \frac{U_{do}\overline{K}_h}{\omega L_d},$$

where *an integral harmonic coefficient of a voltage at a direct current network*

$$\overline{K}_{\text{г}} = \sqrt{\sum_{k=1}^{\infty}\left(\frac{U_{d(k)}}{kU_{d0}}\right)^2} \qquad (2.3.18)$$

A harmonic coefficient of a rectified current taking into account (2.3.17) will look so

(2.3.19)

$$K_{hc} = \frac{I_{d.hh}}{I_d} = \frac{U_{do}\overline{K}_h}{\omega L_d I_d}$$

Reverse, a necessary reactor inductance is

$$L_d = \frac{U_{do}\overline{K}_h}{I_d K_{hc}\omega}$$

(2.3.20)

So a maximal energy value of a smoothing reactor is

$$W_L = L_d I_d^2 = \frac{P_{do}}{\omega}\frac{\overline{K}_h}{K_{hc}}.$$

(2.3.21)

To provide a possibility of an expenditures comparison to a smoothing reactor, working at a direct current network, with expenditures to a filter reactor, working at an alternating current network (as a transformer), we shall introduce a conventional established power of a reactor. It is a reactive power of this reactor, equal to a full power (there is no an active power at an ideal reactor), which it could have with a given current and inductance at an alternating current network. It is known from an electrical engineering that it is may to express a reactive power of a reactor as a product of a degree frequency ω and a maximal value of a reactor energy, what, taking into account (3.2.1), leads to a such result:

(2.3.22)

For a rectifier with $qm_2 = 2$, $\bar{K}_h = 0,24$.

Of course, conditions of a magneto-conductor work of a smoothing reactor are easier than of a filter

$$S^*_{T.L} = \frac{S_{T.L}}{P_{do}} = Q^*_L = \frac{\omega W_L}{P_{do}} = \frac{\bar{K}_h}{K_{hc}}.$$

reactor magneto-conductor, because its variable component of a magnetic flow, conditioned by only pulsations of a rectified current, consists from some percents from a constant component of a flow. And because an established power of a smoothing reactor, determined by a brought higher way, is called as a conventional and is used just attached to a comparison of the different rectifying schemes on the conventional expenditures to the smoothing reactors.

Attached to a task of pulsation coefficient of a rectified current K_{pc} it is not difficult to show that a conventional established power of a reactor

$$S_{T.L} = \frac{\omega W_L}{P_{do}} = \frac{K_C}{K_{CC}},$$

(2.3.23)

i.e. is determined by a correlation of the pulsation coefficients of a rectified voltage K_p and a rectified current K_{pc}. Here $K_p = 0,67$.

13.An input coefficient of a rectifier power is calculated

$$\chi = \frac{P_1}{S_1} = \nu_I = \frac{1}{S_1^*} = 0.9 \qquad \cos\varphi_{1(1)} = 1$$

(2.3.24)

what gives for $K_{hc} = 0,48$.

14. A coefficient of a rectifier converting on a voltage (on the smooth components) is calculated

(2.3.25)

$$K_{CV} = \frac{U_{do}}{U_1} = \frac{2\sqrt{2}}{\pi K_T}.$$

$$K_{CC} = \frac{I_d}{I_{1(1)}} = \frac{I_d}{\nu_I I_1} = \frac{K_T}{\nu_I} = 1,11 K_T.$$

15. A coefficient of a rectifier converting on a current (on the smooth components) is calculated

(2.3.26)

Sometimes they determine a coefficient of a rectifier converting on a current as

$$K'_{CC} = \frac{I_d}{I_1} = K_T = \nu_I K_{CC}.$$

(2.3.27)

A valve type is chosen on a reference book on the calculated values I_a $(I_{a.max})$, $U_{b.max}$. A ready transformer is chosen on a reference book on the calculated values U_2, I_2, I_1, S_t, and attached to a transformer absence – a task is given to project a transformer on this given values. A ready reactor is chased or a new transformer is projected on a value of a smoothing reactor inductance and on a current at it.

It is may to do the next **conclusions** on the results of a second stage of a rectifier analysis.

- A rectifier is characterized by a bad transformer using, because S_t^* >1 on 34%. It is conditioned by bad current forms at transformer windings, especially at the secondary because of a rectifying half-wave.

- A rectifier is characterized by a bad valves using on a reverse voltage, which more than a demanded rectified voltage at π times.
- A rectifier is characterized by a bad quality of a rectified voltage (pulsations are compared with a constant component of a rectified voltage).
- A low input coefficient of a rectifier power.

Usually rectifiers of an one-phase current attached to U_1 = 220 V are applied till powers $P_{do} \approx 3\ldots5$ kWt and attached to a rectified voltage about till 300 V for a given scheme attached to a condition of a valves access with a working voltage not higher 15 class.

2.4. RECTIFIER OF ONE-PHASE CURRENT ON THE BRIDGE SCHEME ($m_1 = m_2 = 1, q = 2$)

A rectifier scheme is showed at a figure 2.4.1.

A valve bridge consists from two groups of valves – a cathode (not even valves) and anode (even valves).Two valves conduct a current at the same time at a bridge scheme – one from a cathode group and one from an anode group. A rule of a determination of a conducting valve at a cathode group is formulated at last part. **A rule of determination of a conducting valve at an anode group** – that valve conducts a current, a potential of which cathode is the most negative.

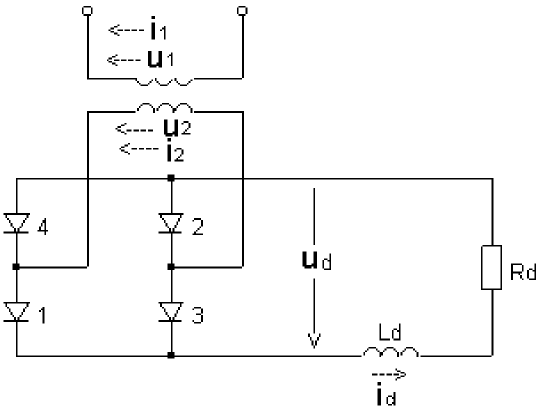


Fig.2.4.1

A task of an analysis is the same that for the last basic scheme, i.e. a determination of the scheme properties and creating the advises on the field of a rectifier appliance. An analysis methodic is the same, i.e. at first a qualitative analysis of electromagnetic processes by the time diagrams, and at the second stage – a quantitative analysis with an aim of getting the calculated correlation.

Time diagrams of scheme voltages and currents are showed at a figure 2.4.2 at the same order as for the last scheme.

A methodic of their building is the same. Diagrams are different because of a reverse voltage value at a valve and a current form at a transformer secondary winding. A voltage of a transformer secondary winding u_2 is applied to a valve 1 at a reverse direction attached to conducting valves 3,4 of the bridge. A current form at a transformer secondary winding is determined by a summa of valve currents, connected to this winding, for example, 1 from a cathode group and 4 from an anode group. A current presence at a winding at a positive and negative half-wave of a voltage shows that a rectifying process is full-wave and, hence, a constant component at a secondary current is absent.

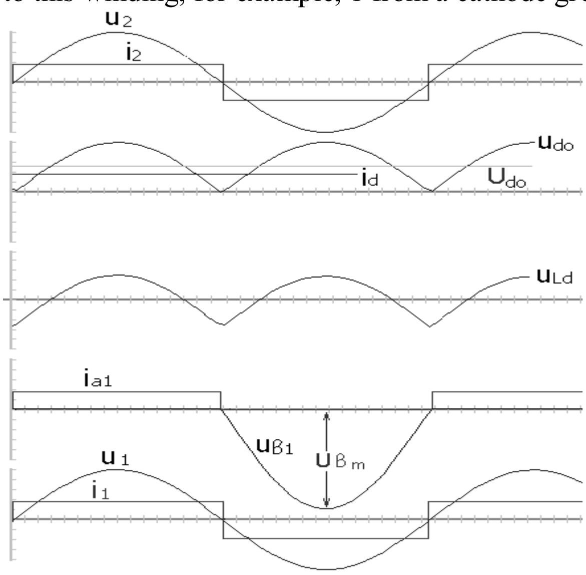


Fig. 2.4.2

The most diagram analogy at this and last schemes provides an analogy of the corresponding calculated correlations. The differences of calculated correlations, brought lower, are conditioned by a showed difference of two time diagrams – a reverse voltage curve and transformer secondary current. A maximal value of a reverse voltage at a valve here is

$$U^*_{B.T} = \frac{U_{B.T}}{U_{d_0}} = \frac{\sqrt{2}\,U_2}{E_{d_0}} = \frac{\sqrt{2}}{U_{d_0}} \cdot \frac{\pi}{2\sqrt{2}} U_{d_0} = \frac{\pi}{2}$$

An effective current value at a transformer secondary winding is determined so:

$$I_2 = \sqrt{\frac{1}{2\pi}\int_0^{2\pi} i_2^2\, d\vartheta} = \sqrt{\frac{1}{\pi}\int_0^{\pi} I_d^2\, d\vartheta} = I_d$$

A full power of the transformer secondary windings, taking into account this, is changing

$S^*_2 = 1{,}11.$

$$S_2 = U_2 I_2 = \frac{\pi}{2\sqrt{2}} U_{d0} I_d = 1{,}11\, P_{dn}$$

$$S_T = S_2 = S_1 = 1{,}11\, P_{d0} \qquad S_T^* = 1{,}11$$

As a result a typical transformer power is

All other energetic indexes are the same as at the last scheme.

So a using of the same processes at the given and last rectifying schemes allowed to economy not only a paper and a time, but and thinking.

It is may to do the next **conclusions** on the results of an analysis:

- A transformer using at a full-wave rectifying scheme is better than at a half-wave scheme because of better (more close to sinusoid) curve of a transformer secondary current;
- A valve using on a reverse voltage at a bridge scheme better in two times than at a zero rectifying scheme (a scheme with a transformer zero point output);
- A quality of a rectified voltage at a considered and last rectifying schemes is the same because they have the same pulseplicity $p=qm_2=2$;
- A demerit of the bridge scheme is a rectified current flowing through two valves, turned on consecutively, what leads to double loses of a voltage and power at valves with real parameters, reducing a CUA (a coefficient of useful action) of a rectifier attached to low rectified voltage values.

So, at a base of formulated bridge rectifying scheme properties there follows a conclusion that this scheme is better than a zero scheme attached to average values of a rectified voltage and, of course, is more rational attached to high values of a rectified voltage (without the bounds of recommendations on the zero rectifying scheme using).

2.5. RECTIFIER OF THREE-PHASE CURRENT WITH SCHEME OF TRANSFORMER WINDINGS CONNECTING « TRIANGLE-STAR» WITH ZERO OUTPUT *($m_1 = m_2 = 3$, $q = 1$)*

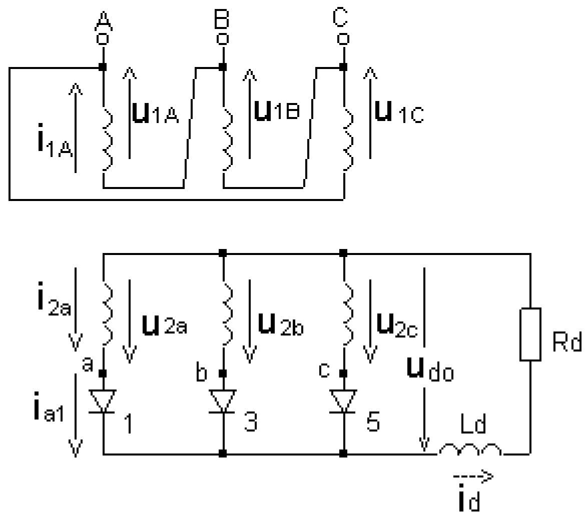


Fig.2.5.1

The common observations on a three-phase current rectifying. An input currents of an one-phase current rectifier are higher than strain limit permissible values 16…25 A for domestic consumers attached to active load powers P_{do} higher than 3…5 kWt (this limits may be some more for industrial one-phase consumers, depending on a maim). It is necessary a rectifier power from a three-phase maim in this cases. Attached to it there appear a lot of rectifying schemes depending on a connecting way of a rectifier input transformer primary and secondary windings (a triangle, star, zigzag, double zigzag), a learning of which two variants compounds **the aims at this and the next parts.**

A first variant of three-phase current half-wave rectifier, demanding to connect transformer secondary windings into a star to get a common zero wire and transformer primary windings – into a triangle, a necessity of what is showed lower, represented at a figure 2.5.1.

An aim of analysis of this basic rectifying scheme is the same – a determination of its properties. An analysis methodic is the same: at first a qualitative analysis of the electromagnetic processes by the time diagrams of voltages and currents, and at the second stage – a quantitative analysis to get the calculated correlation and determine the given rectifier properties on it. The time diagrams of characteristic voltages and currents of a rectifier are represented at a figure 2.5.2 at the same consecution as for the last schemes. At a first diagram there is showed a three-phase voltage system of a transformer secondary windings u_{2a}, u_{2b}, u_{2c} and intervals of a conducting state of a cathode group valves are showed, they are determined according to formulated higher conducting rule of a cathode group valves. The intersection points of a secondary voltage positive half-waves, beginning from which there appears a forward voltage at valves, are called *points of a natural (ignition) commutation* (a term is introduced at a pre-semiconductor age of the gas-discharged valves, when its entry to a work happened because of a discharge «ignition» at it). It is necessary to take into account that points of a natural ignition and passage points of the secondary voltages through zero values are not the same attached to a number of secondary voltage phases about three or more, because a countdown of a delay when valves begin to work relatively corresponding zeros of the secondary voltages was designed by a ψ degree at a figure 2.2.4, and a countdown of a delay when valves begin to work relatively the points of a natural ignition is designed by α degree at the control rectifiers.

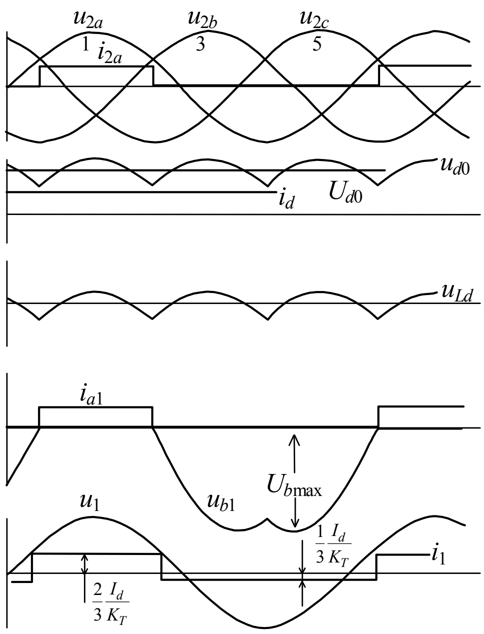


Fig.2.5.2

There is built a curve of a rectified voltage u_{do} as a totality of the secondary voltages areas on the conducting intervals of valves and a curve of a rectified current i_d for a case $X_d = \infty$ at the second diagram. At the third diagram there is represented a voltage form at a smoothing reactor, receiving an alternating component (pulsation) of a rectified voltage. There are showed an anode current diagram of the first valve i_{a1} and a curve of a reverse voltage at it u_{b1} at a fourth diagram. The last is determined as a difference between the instantaneous voltage values at a valve anode (u_{2a}) and a rectified voltage u_{do}, counted down relatively a common (zero) point of a transformer secondary windings. A valve anode current is equal to a rectified current at a conducting interval of one valve. Evidently, at this scheme a current at a transformer secondary winding i_{2a} repeats a form of a valve anode current i_{a1}, connected to a winding consecutively, what is showed at the first diagram. A one-directional character of a secondary winding current make to pay attention at it again, i.e. a presence of a constant component at it $I_{2(=)}$, numerically equal to an average value of this current, i.e. to an average value of an anode valve current I_a. Taking into account this fact, a secondary winding current of a transformer may be conventionally divided into a summa of a constant component $I_{2(=)}$ and an alternating (which stayed after a subtraction of $I_{2(=)}$) component $i_{2(\sim)}$:

$$i_2 = i_{2(\sim)} + I_{2(=)}. \qquad (2.5.1)$$

It is may to formulate here **an empiric build rule of a transformer primary current** at a basis of this dividing on the founded secondary current (not taking into account a current of a transformer magnetization). So as only an alternating component of a current may transform to a primary winding from the secondary, hence, subtracting a constant component from a secondary current curve and taking into account a transforming coefficient, we get:

$$i_1 = \frac{i_2 - I_{2(=)}}{K_T} \qquad (2.5.2)$$

It is necessary to notice that a strict mathematical correlation for a transformer primary current may be got from the equations for magnetizing powers, compounded on a second Kirchhoff law for the magnetic networks so as it was made at a part 2.3 and will be made at a common look at a part 3.5.

At the fifth time diagram there are built a voltage curve of a primary transformer winding and a primary current curve at this winding, placed at a rod of a phase A of a transformer magneto-conductor.

There appears a non-compensated *one-directional flow of the forced transformer dc magnetization* because of a presence of a constant component at transformer secondary windings current in each from three rods of three-phased transformer magneto-conductor.

This fact leads to a corresponding bias B_{magn} of the issue state of a working point at a magneto-conductor magnetization curve, limiting a permissible diapason of an inductance changing of a magneto-conductor till values $dB = B_{sat} – B_{magn}$, lower than an inductance value, corresponding to a saturation limit B_{sat}. As a result it is necessary to increase proportionally a magneto-conductor section to save an alternating component of a flow at the same level, demanded by the given voltage at the primary windings, i.e. its mass and size (here this increasing will be equal to 1/3 in according to that a constant component of a flow is equal to a third part from a result flow amplitude).

The summa expenditures to a transformer (here we understand a cost of a copper and a magneto-conductor or its mass, or its size at a construction), attached to a condition of an equality of expenditures to a cooper and magneto-conductor at summa expenditures, will increase at 1/6, i.e. at 16,5 % at this conditions. Taking into account equality of the considered components of the expenditures it is possible to speak about an increasing of a typical power of transformer at 16,5 % in this case, because used typical methodic of its calculation not taking into account a forced dc magnetization of a transformer by the one-directional flow.

Also it is may to consider qualitatively a question about a connecting scheme of the transformer primary windings. If to make the same analysis of the electromagnetic processes at a transformer attached to a connecting of its primary windings into a star, it is may to show that there appear alternating flows of a forced dc magnetization additionally from the harmonics at a transformer magneto-conductor attached to a presence of the harmonics at the secondary currents, multiple to three (regimes with $X_d \neq \infty$), because these harmonics will absence at the primary windings because of an absence of a flowing way for them. Because the primary windings are connected into a triangle, which makes a counter for a harmonic flowing, multiple to three, what compensate flows from these harmonics at the secondary currents, deleting by this way a forced dc magnetization of a magneto-conductor by these harmonics [8].

It is may to make a stage of a quantitative analysis of the processes at a rectifier now. A task, assumptions and an analysis methodic are the same as at a calculation of the one-phase current rectifiers, what allows comparing the results, got at the same conditions. The same **fifteen calculation points** have the next maintenance here.

1. An average value of a rectified voltage of a not control rectifier U_{do}

$$U_{d_0} = \frac{1}{\frac{2\pi}{3}} \int_{-\frac{\pi}{3}}^{\frac{\pi}{3}} \sqrt{2}\, U_2 \cos\vartheta \, d\vartheta = \frac{3\sqrt{6}}{2\pi} U_2 = 1{,}17\, U_2 \tag{2.5.3}$$

from where

$$U_2 = \frac{U_{d_0}}{1{,}17} = 0{,}84\, U_{d_0} \tag{2.5.4}$$

We shall notice that a beginning of a time countdown is chosen only from how to calculate more simply attached to a recording of an issue calculated integral and does not influence at a calculation result.

2. An average value of a valve anode current

$$I_a = \frac{1}{2\pi} \int_0^{2\pi} i_a d\vartheta = \frac{1}{2\pi} \int_0^{2\pi/3} I_d d\vartheta = \frac{I_d}{3} \tag{2.5.5}$$

3. An effective value of a valve anode current

$$I_{a.e}=\sqrt{\frac{1}{2\pi}\int_0^{2\pi} i_a^2 d\vartheta}=\sqrt{\frac{1}{2\pi}\int_0^{2\pi/3} I_d^2 d\vartheta}=\frac{I_d}{\sqrt{3}}$$

(2.5.6)

A coefficient of form

$$K_F=\frac{I_{a.e}}{I_a}=\sqrt{3}.$$

(2.5.7)

4. An amplitude value of an anode current

$$I_{a\max}=I_d$$

(2.5.8)

An amplitude coefficient

$$K_a=\frac{I_{a\max}}{I_a}=3$$

(2.5.9)

5. A maximal value of a reverse voltage at a valve

$$U^*_{b\max}=\frac{U_{b\max}}{U_{d_0}}=\frac{\sqrt{2}\,\sqrt{3}\,U_2}{U_{d_0}}=\frac{\sqrt{2}\,\sqrt{3}\,2\pi}{3\sqrt{6}}=\frac{2\pi}{3}=2{,}09$$

(2.5.10)

6. An established power of the valves:

- with not full control

$$S^*_{b1}=\frac{S_b}{P_{d_0}}=n\frac{U_{b\max}I_a}{P_{d_0}}=3\frac{2\pi\cdot U_{d_0}I_d}{3\cdot 3}=\frac{2\pi}{3}$$

(2.5.11)

- with a full control

$$S^*_{b_2}=S^*_{b1}K_a=3\frac{2\pi}{3}=2\pi$$

(2.5.12)

7.An effective value of a current at a transformer secondary winding

$$I_2=I_{a.e}=\frac{I_d}{\sqrt{3}}.$$

(2.5.13)

8. An effective value of a current at a transformer primary winding

$$I_1=\sqrt{\frac{1}{2\pi}\int_0^{2\pi} i_1^2 d\vartheta}=\sqrt{\frac{1}{2\pi}\left[\int_0^{2\pi/3}\left(\frac{2}{3}\frac{I_d}{K_T}\right)^2 d\vartheta+\int_{2\pi/3}^{2\pi}\left(-\frac{1}{3}\frac{I_d}{K_T}\right)^2 d\vartheta\right]}=\frac{\sqrt{2}}{3}\frac{I_d}{K_T}$$

(2.5.14)

9. A full power of the secondary windings

$$S_2=3U_2I_2=\frac{3\cdot 2\pi}{3\sqrt{6}}U_{d_0}\;\frac{I_d}{\sqrt{3}}=\frac{2\pi}{3\sqrt{2}}P_{d_0}=1{,}48\;P_{d_0}$$

(2.5.15)

10. A full power of the transformer primary windings

$$S_1=3U_1I_1=3K_TU_2\frac{\sqrt{2}\,I_d}{3K_T}=3\frac{2\pi}{3\sqrt{6}}U_{d_0}\cdot\frac{\sqrt{2}}{3}I_d=\frac{2\pi}{3\sqrt{3}}P_{d_0}=1{,}21\;P_{d_0}$$

(2.5.16)

11. A typical transformer power

(2.5.17)

$$S_T=\frac{S_1+S_2}{2}=\frac{1{,}21+1{,}48}{2}P_{d_0}=1{,}345\;P_d \qquad S^*_T=1{,}345$$

This calculated value does not take into account a forced dc magnetization of a transformer by a one-directional flow. Taking into account made higher qualitative influence estimations of a one-directional flow of a forced dc magnetization

$$S^*_{tf} = 1{,}165 S^*_t = 1.57 \tag{2.5.18}$$

12. An inductance of a smoothing reactor is estimated the same way on a formula (2.3.19), and its relative conventional established power – on a formula (2.3.21) or (2.3.22). Here

$$\overline{K}_h = 0{,}06, \qquad K_c = 0{,}25. \tag{2.5.19}$$

13. An input power coefficient is calculated the same as (2.3.23)

$$X = \nu_I = \frac{1}{S_1^*} = \frac{1}{1{,}21} = 0{,}79 \tag{2.5.20}$$

14. A coefficient of a rectifier converting on a voltage

$$K_{cv} = \frac{U_{do}}{U_1} = \frac{3\sqrt{6}}{2\pi} K_t = 1{,}17 K_T \tag{2.5.21}$$

15. Coefficients of a rectifier converting on a current

$$K_{CC} = \frac{K_T}{\nu_I} = 1{,}21 K_T \qquad K'_{CC} = \frac{K_T}{\sqrt{2}}. \tag{2.5.22}$$

We shall compare a considered scheme of a half-wave rectifying of a three-phase current with an analyzed higher scheme of a half-wave rectifier of a one-phase current (*m_1=1, m_2=2, q=1).*

1. There happened an additional worsening of a transformer using on a typical power S^*_{TR} at a considered scheme in comparison with the last scheme because of a presence of a transformer magneto-conductor dc magnetization by a direct current.
2. A valve using on a reverse voltage at a considered scheme is better in 1,5 times than at the last. Correspondingly, it reduced an established power of the not full control valves.
3. A quality of a rectified voltage at a considered scheme is higher in 4 times on a criterion $\overline{K}_h$ and in 2,5, times on a criterion K_c than at the last scheme. It bounded with a pulseplicity increasing of a rectifying in 1,5 times (from *qm_2=2 till qm_2=3)*, i.e. with a pulsation frequency increasing and with pulsation amplitude reducing almost in 2 times. It is seen that talks about a quality of a rectified voltage on a known criterion K_c are not enough, because it does not take into account a voltage pulsation frequency, also influencing at a quality of a rectified current. A criterion $\overline{K}_h$ takes into account a pulsation frequency, because it forwardly determines a quality of a rectified current and an established power of a smoothing reactor.
4. An input power coefficient here is much lower than at the last scheme, what, as it will be showed at a part 3.13, means a big reverse negative influence of a rectifier on a main.

So, taking into account these scheme properties, it has a limited own using (only attached to low rectified voltage values with a not high quality), but is a component of the more complicated and more qualitative rectifiers (look part 2.7.).

2.6. THREE-PHASE RECTIFIER WITH TRANSFORMER WINDINGS « STAR-ZIGZAG WITH ZERO» (*m_1=m_2=3, q=1)*

A transformer using worsening at the last scheme of a half-wave rectifying, bounded with a presence of a not compensated one-directional flows of a forced magnetization at each magneto-conductor rod, created by the constant current components of the transformer secondary windings, may be deleted. **A deleting mechanism of a forced one-directional dc magnetization** is evident – to place two secondary windings at each transformer rod, one-directional currents of which to direct each to other. It have got by a natural way at a two-phase rectifier of a one-phase current because of a converting of a one-phase voltage into two-phase at a transformer by two transformer secondary windings. It demands a presence of the second system of the transformer secondary windings at a half-wave rectifier of a three-phase current. Different bound variants of these winding systems between each other and with valves create different rectifier schemes of a three-phase current with compensated one-directional flows of the forced magnetization. A special connecting (by a zigzag) of these winding systems between each other gives a scheme, considered at this part. A connecting of the second winding system, turned on against-phase to a first system, with a second valve group, with the next parallel or consecutive connecting of these winding complexes and valve groups gives correspondingly a rectifying scheme with an equalizing reactor, considered at the next part, and a cascade rectifying scheme [8], not used already at new projects.

A half-wave rectifier scheme of a three-phase current with a connecting of two systems of the transformer secondary windings into a zigzag is showed at a figure 2.6.1. **An analysis aim** of a new scheme is the same – a determination of the scheme properties within the bounds the same assumptions. A determined intellectual voltage, bonded with an analysis beginning of each new rectifying scheme, it is may to make much less if to see a pro-image of the known already

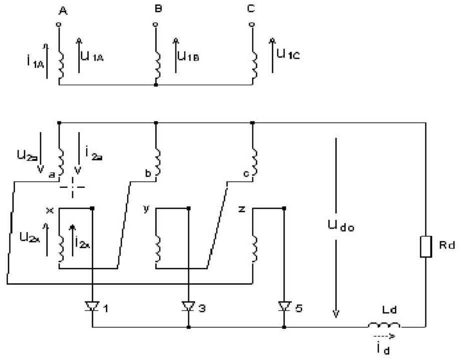


Fig. 2.6.1

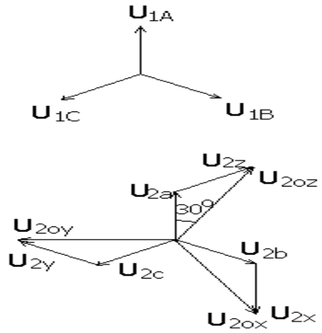


Fig.2.6.2

scheme at a new. A vector diagram for the result secondary voltages U_{20} of a transformer is showed at a figure 2.6.2. Let's compare alternating voltage systems at the secondary side of transformers, demanding a rectifying, from these positions at the last and at a considered rectifier schemes. There follows from a diagram that a star of three-phase voltages is rectified here, too, vectors of which are only longer than winding voltage vectors in 1,732 times and turned at 150° at the late side relatively anode voltages of the last scheme valves.

Hence, an analysis of the electromagnetic processes at this scheme it is may to begin from a time diagram building of these result three-phase secondary voltages u_{20a}, u_{20b}, u_{20c}, as it is showed at a figure 2.6.3 at a first diagram. A second diagram with a rectified voltage and current, a third diagram with a smoothing reactor voltage, a fourth diagram with a valve anode current and a reverse voltage at it are the same qualitatively as corresponding diagrams of the last rectifying scheme.

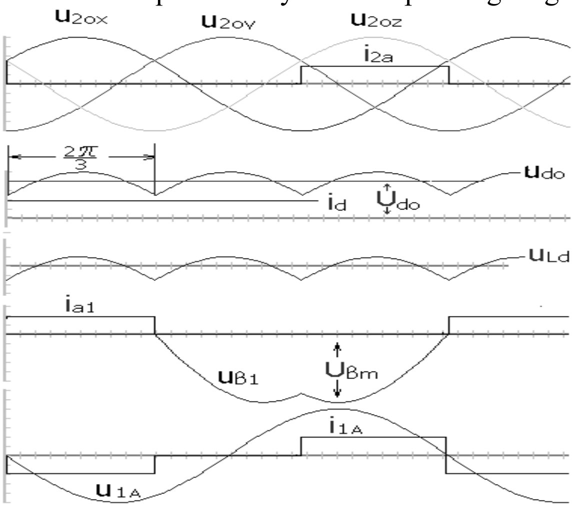


Fig. 2.6.3

At a fifth diagram there is showed a primary voltage of a phase A, leaving behind a result secondary voltage U_{20x} at 150°, as you can see it from a vector diagram at a figure 2.8.2.

A primary winding current of a phase A i_{1A} it is may to build by that empiric rule (2.5.2), which was used attached to a building of a primary current at the last scheme, supplying it to two secondary windings *a* and *x*, placed at the same magneto-conductor rod of a transformer, what gives here:

(2.6.1)

$$i_{1A} = \frac{i_{2a} - i_{2x}}{K_T} = \frac{i_{a5} - i_{a1}}{K_T}$$

A strict explanation of a primary current curve it is may to get from decision of the equations, compounded on the second Kirchhoff low for closed magnetic networks. We shall get two such equations for a contour from three rods A-B and A-C, getting round them against an hour arrow, a third

$$\begin{vmatrix} -1 & -1 & 0 \\ -1 & 0 & 1 \\ 1 & 1 & 1 \end{vmatrix} \begin{vmatrix} i_{1A} \\ i_{1B} \\ i_{1C} \end{vmatrix} = \frac{W_2}{W_1} \begin{vmatrix} -2 & 1 & 1 \\ -1 & -1 & 2 \end{vmatrix} \begin{vmatrix} i_{a1} \\ i_{a3} \\ i_{a5} \end{vmatrix}$$

equation – for the primary currents

(2.6.2)

A decision of this equation system (look a part 3.5) will give the same result for a primary current here as an empiric rule.

A quantitative analysis stage of the processes at the considered scheme gives the same results for that elements, which have the same time diagrams as the last scheme (a rectified voltage network, a smoothing throttle, valves), having differences only for a transformer.

An average value of a rectified voltage is expressed through U_2 :

$$U_{d_0} = \frac{3\sqrt{6}}{2\pi} U_{20} = 1{,}17\, U_{20} \qquad U_{20} = \sqrt{3}\, U_2$$

An effective value of a primary current, calculated on (1.1.3), will look so

$$I_1 = \frac{I_d}{K_T} \sqrt{\frac{2}{3}}$$

A full power of six transformer secondary windings

$$S_2 = 6 U_2 I_2 = 6 \frac{2\pi U_{d_0}}{3\sqrt{6}\sqrt{3}} \frac{I_d}{\sqrt{3}} = \frac{4\pi}{3\sqrt{6}} P_{d_0} = 1{,}71\, P_{d_0}, \quad S_2^* = 1.71\,.$$

A full power of the transformer primary windings

$$S_1^* = \frac{3 U_1 I_1}{P_{d_0}} = \frac{3 K_T U_2 I_d}{P_{d_0}} \sqrt{\frac{2}{3}} = \frac{2\pi}{3\sqrt{3}} = 1{,}21$$

As a result a typical transformer power is

$$S_T = \frac{S_1 + S_2}{2} = \frac{1{,}21 + 1{,}71}{2} P_{do} = 1{,}48 P_{do}$$

A distortion coefficient of a rectifier input current

$$\nu_J = \frac{1}{S_1^*} = 0{,}827$$

A converting coefficient of a rectifier on a voltage

$$K_{CV} = \frac{E_{do}}{E_1} = \frac{3\sqrt{6}}{2\pi}\sqrt{3} K_T = 1{,}17\sqrt{3} K_T$$

A converting coefficient of a rectifier on a current

$$K_{CC} = \frac{I_d}{I_{1(1)}} = \frac{I_d}{I_1 \nu j} = \frac{2\pi}{3\sqrt{2}} K_T = 1{,}48 K_T, \qquad K'_{CC} = \frac{I_d}{I_1} = \sqrt{\frac{3}{2}} K_T = 1{,}225 K_T\,.$$

As a result a similarity of the electromagnetic processes and calculated correlation at both half-wave rectifying schemes of a three-phase current makes fields of their using more closer. A connecting of the secondary windings into a zigzag provides the best transformer using on a magneto-conductor because of an absence of its one-directional dc magnetization. But a geometrical summa of the secondary winding voltages at a result voltage worse a transformer using on winding copper. A practice showed that a transformer becomes smaller at a considered rectifying scheme attached to I_d >85…120 A, and attached to currents, lower than pointed, a transformer smaller at the last rectifying scheme.

2.7. SIX-PHASE RECTIFIER OF THREE-PHASE CURRENT WITH CONNECTING OF TRANSFORMER SECONDARY WINDINGS « STAR - REVERSE STAR WITH EQUALIZING REACTOR» (m_1=3, m_2=2×3, q=1)

A considered rectifier (a figure 2.7.1) made from two three-phase half-wave rectifiers, turned on at a parallel work on an output through an equalizing reactor.

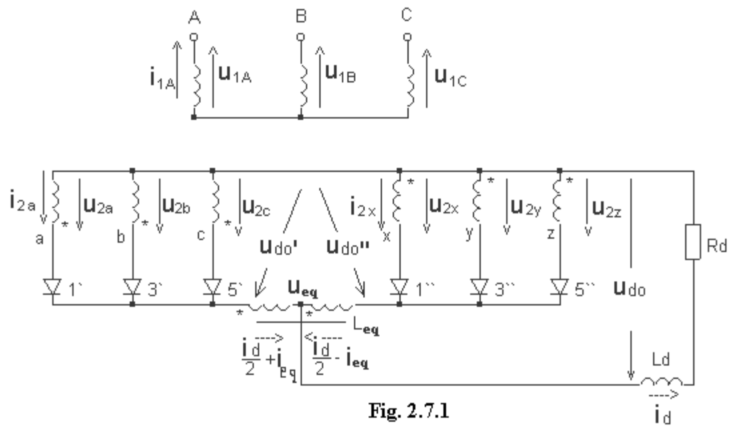


Fig. 2.7.1

To provide a one-directional flows compensation of a forced magnetization two stars of the transformer secondary voltages are made by against-phase voltages of the windings u_{2a} and u_{2x}, u_{2b} and u_{2y}, u_{2c} and u_{2z}, placed by pairs at the corresponding three rods of a magneto-conductor. It is reached by an unification of one star of the windings beginnings at the zero point, and at the zero point of a second star – windings endings. Attached to it, in spite of currents one-direction at each winding pair, placed at the corresponding magneto-conductor rods, a result magnetic flow of each rod does not contain a constant component, i.e. a forced dc magnetization by a one-directional flow is absence. Given scheme also is called *a double three-phase* because of a symbiosis of two three-phase groups of a rectifying.

An analysis task of a rectifier is the same: to get a calculated correlation for the scheme elements and to notice possible using fields of a scheme at a base of a comparison of it with the analogy correlation for earlier analyzed rectifying schemes.

An analysis feature of this scheme is a presence of two work regimes:

- of a double three-phase rectifying, which is a general;
- of a six-phase half-wave rectifying, appearing attached to small values of load,
- close to an idling.

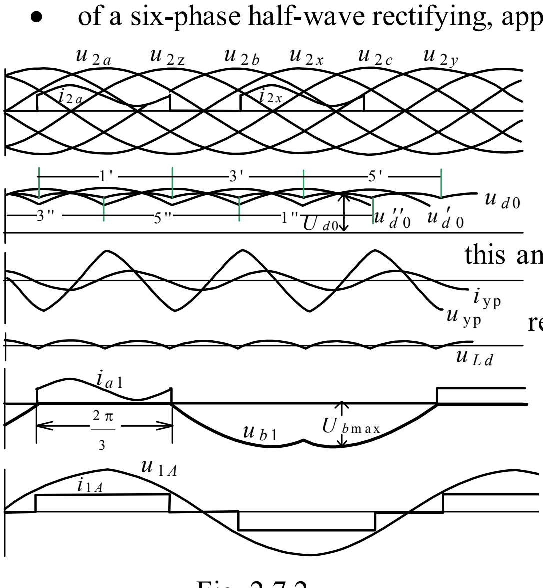


Fig. 2.7.2

A special attention is paid, of course, to a general work regime at a made lower analysis – to a double three-phase regime, when two half of a scheme work independently each to other. There is paid attention to the characteristic practical regime specific of a six-phase rectifying at end of this analysis – a voltage increasing at a rectifier output at 15% in comparison with a regime of a double three-phase rectifying.

A qualitative analysis of the electromagnetic processes anticipates a quantitative analysis, as usually, by the time diagrams, showed at a figure 2.7.2.

At the first diagram there are built two three-phase systems of the secondary voltages for two secondary winding systems, which are against-phase, concretely, the
systems from u_{2a}, u_{2b}, u_{2c} and u_{2x}, u_{2y}, u_{2z}. There are brought here the current diagrams i_{2a} and i_{2x} at the secondary windings *a* and *x* of a transformer, placed at one magneto-conductor rod. They, as it will be seen from the next, repeat anode currents of valves 1′ and 1″.

At the second diagram with the same two systems of the secondary voltages are designated intervals of a valve conducting state at two cathode groups on a known rule for a cathode group. A round of the positive half-waves of the first three-phase system voltages gives a curve of a rectified voltage u'_{do} of a left scheme side, and, on an analogy, a round of a second three-phase system – a curve of a rectified voltage u''_{do} of a right scheme side. Although average values of the rectified voltages of both scheme sides are the same, the instantaneous values of rectified voltages are different because of their pulsation displacement at a half of pulsation period, as it is seen from a diagram. A pulsation difference at two three-phase rectifiers demand to turn on them at a parallel work through a reactor, which is called an equalizing. This reactor, at first, receives a pulsation difference at the rectified voltages and limits an equalizing current between three-phase rectifiers and, at second, allows to get at a load, connected to an average point of an equalizing reactor, a voltage u_{do}, equal (on a method of superposition) to a half-summa of rectified voltages of each scheme sides. A voltage at a load has a six-multiple, i.e. a double pulsation frequency, because of a pointed displacement of their pulsation at a half of own period and $p=qm_2=6$. The rectified current curves i_d do not contain pulsation attached to an assumption about a filter is ideal ($X_d=\infty$).

A curve of an equalizing voltage u_{eq}, equal to a difference of the rectified voltages of a left and right rectifier sides, is brought at the third diagram. An equalizing current form, flowing at a contour, made by rectified voltages of both scheme sides, passing a load, is determined by an integral from an equalizing voltage. Because an integration of a non sinusoidal curve, as it was showed at a part 1.3.3, means a reducing at a result curve of the highest harmonics, so it is taken attached to an equalizing current building that it has a sinusoidal form and displaced at a quarter of a period at a late side from a rectified voltage. Usually an equalizing reactor inductance is chosen to limit (it is not useful for a load and parasite for a transformer) an equalizing current at the level 1-2% from a nominal value of a rectified current. An equalizing current is showed as some bigger because it is not possible to notice it at a rectified current level at the diagrams of an anode and secondary currents at the third time diagram.

At the fourth time diagram there is a curve of an anode current of a valve 1, not taking into account pulsation from an equalizing current (because it is very little), put on a half from a rectified current, divided at two half at two networks of a equalizing reactor. Attached to it a dividing of a rectified current goes on because of an inter-induction voltage of an equalizing reactor at a dynamics. There is represented a curve of a reverse voltage at a valve of the same form as the last three-phase current rectifiers here.

Currents of the transformer secondary windings, which are the same as the corresponding anode currents at half-wave rectifying schemes, are built at the first diagram after a form of valve anode currents is determined.

There is showed a voltage form at a primary winding of a transformer phase A and a current curve at this winding i_{1A} at the fifth diagram. You can build it on an empiric rule of a formula (2.5.2), used to two secondary currents i_{2a} and i_{2x} of one phase. Attached to it secondary current pulsations from an equalizing reactor are absence at a primary current, because these pulsations are against-phase at two pointed secondary currents and absence at a result magnetic flow of a magneto-conductor rod.

Calculated correlations for a general rectifier work regime – a regime of a double three-phase rectifying are got by built time diagrams of currents and voltages at the same fifteen-step analysis procedure.

1. An average value of a rectified voltage at this scheme is the same as the scheme sides (halfs) have, because an average voltage value at an equalizing reactor is equal to zero, i.e.

$$U_{d0} = \frac{3\sqrt{6}}{2\pi} U_2$$

2. An average value of a valve anode current

$$I_a = \frac{I_d}{2 \cdot 3} = \frac{I_d}{6}$$

3. An effective value of a valve anode current, calculated also through a porosity on (2.1.5):

$$I_{a.e} = \frac{I_d}{2\sqrt{3}} \qquad K_F = \frac{I_{a.e}}{I_a} = \sqrt{3}.$$

4. A maximal value of an anode current

$$I_{a\max} = \frac{I_d}{2} \qquad K_a = \frac{I_{a\max}}{I_a} = 3$$

5. A maximal value of a valve reverse voltage

$$U_{B\max} = \sqrt{2}\,\sqrt{3}\,E_2 = \frac{2\pi}{3} U_{d0} \qquad U^*_{B\max} = \frac{2\pi}{3}$$

6. An established power of valves with not full control

$$S^*_{V1} = \frac{S_b}{P_{do}} = n\frac{U_{B.\max} I_a}{P_{do}} = 6\frac{2\pi}{3}\frac{1}{6} = \frac{2\pi}{3},$$

with a full control

$$S^*_{V2} = n\frac{U_{B.\max} I_{a.\max}}{P_{do}} = 6\frac{2\pi}{3}\frac{1}{2} = 2\pi$$

7. An effective current value at a transformer secondary winding

$$I_2 = I_{a.e} = \frac{I_d}{2\sqrt{3}}.$$

8. An effective current value at a transformer primary winding in according to (2.1.3) will look so

$$I_1 = \frac{I_d}{} \quad \sqrt{2}$$

$$S_2 = 6\,U_2 I_2 = 6\frac{2\pi}{3\sqrt{6}} U_{d0} \frac{I_d}{2\sqrt{3}} = \frac{2\pi}{3\sqrt{2}} P_{d0}, S^*_2 = 1{,}48$$

9. A full power of six secondary windings of a transformer

10. A full power of three primary transformer windings

$$S_1 = 3\,U_1 I_1 = 3\,K_T \frac{2\pi}{3\sqrt{6}} U_{d0} \; \frac{I_d}{2\,K_T}\sqrt{\frac{2}{3}} = \frac{\pi}{3} P_{d0}, \qquad S^*_1 = 1{,}045$$

11. A typical or an established transformer power

$$S_T = \frac{S_1 + S_2}{2} = 1{,}26\ P_{d0} \qquad S^*_T = 1{,}26$$

In comparison with the last schemes of three-phase rectifiers here there was appeared an additional element – an equalizing reactor, working at a threefold frequency (150 Hz) of a voltage. Expenditures at a reactor are determined by an established power value, which it is may to add to a transformer established power attached to a comparison of the different rectifier schemes, because a reactor, as a transformer, - an electromagnetic device, but just with a one winding. It is showed [8] that an established power of a reactor, working at a frequency 150 Hz and leaded to a frequency of a transformer work, i.e. 50 Hz, will look so

$$S_{eq} = 0{,}07 P_{do}, \qquad S^*_{eq} = 0{,}07.$$

12. An inductance of a smoothing reactor is determined on a correlation (2.3.19) depending on a demanding to a rectified current quality. A conventional established power of a smoothing reactor is calculated on (2.3.21) or (2.3.22), attached to it taking into account a six-multiple of a pulsation frequency of a rectified voltage

$$\overline{K}_h = 0{,}0067, \qquad K_C = 5{,}7\% = 0{,}057.$$

13. An input power factor

$$\chi = \frac{1}{S_1^*} 0{,}955$$

14. A converting coefficient of a rectifier on a voltage, evidently, is the same as for a three-phase half-wave rectifier value with a connecting of the secondary windings into a star

$$K_{CV} = \frac{U_{do}}{U_1} = \frac{3\sqrt{6}}{2\pi} K_T = 1{,}17 K_T .$$

$$K_{CC} = \frac{I_d}{I_{1(1)}} = 2\sqrt{\frac{3}{2}} k_T \cdot \frac{\pi}{3} = 2{,}56 K_T , \qquad K'_{CC} = \frac{I_d}{I_{1(1)}} = 2{,}45 K_T$$

15. A converting coefficient of a rectifier on a current is higher in two times because of a parallel connecting of two scheme sides

It is may do the next **conclusions** at a base of the got calculation results and a comparison of it with calculation results of two last schemes of three-phase rectifiers:

1. A considered scheme has a better transformer using on a typical power, than the last schemes.
2. A valve using on a reverse voltage and on an established power is the same at all three schemes of a half-wave rectifying. A feature of the given scheme is that a value of a scheme converting coefficient on a current is higher in two times and a relation of an average rectified current value to an average value of a valve anode current is higher in two times.
3. A rectified voltage quality is much higher here than at the last schemes because of a reducing of its pulsation amplitude (is characterized by an index K_p) and an increasing of the pulsation frequency in two times from three-multiple till six-multiple. Both this facts are characterized by an index $\overline{K}_h$, which is lower than at three-pulse rectifiers in 9 times. It means that an inductance of a smoothing throttle and its established power will be lower in the same number of times, too.
   An input coefficient of a rectifier power is the highest between all considered rectifiers:

$$\chi = \nu_I = \frac{1}{S_1^*} = 0{,}955$$

i.e. a quality of an input current is close to a sinusoidal, which has $\nu_I = 1$, from an energetic (and not

$$K_{hc} = \frac{I_{hh}}{I^1_{(1)}} = \sqrt{\frac{1}{\nu_J^2} - 1} = 0{,}3.$$

geometrical) position. Really, a harmonic coefficient of an input current

i.e. a part of an effective value of the current higher harmonics is equal to 30% from the first harmonic.

Now we shall consider **a specific character of the second regime** – *a regime of a six-phase half-wave rectifying.* It is impossible to create a demanded dc magnetization current of an equalizing reactor to provide its work as an equalizing attached to small values of a rectified current. Attached to it valves begin to work at the natural ignition points of a six-phase star of the secondary voltages, connected from left and right systems of three-phase stars. A round of the positive half-waves of voltage six-phase system becomes a curve of a rectified voltage, an average value of which

$$U_{do} = \sqrt{2}U_2 \frac{m_2}{\pi}\sin\frac{\pi}{m_2} = U_2 \frac{6}{\pi}\sin\frac{\pi}{6}U_2 = \frac{3\sqrt{2}}{\pi}U_2 = 1{,}35U_2.$$

In comparison with a regime of a double three-phase rectifying a voltage at a rectifier output increases in 15%.

Then just a changed value of a maximal reverse voltage at a valve is important, which now is equal to a double value of a secondary voltage amplitude, from another calculated correlation because a rectifier load by a current is very small

$$U_{Bmax} = 2\sqrt{2}\,U_2 = 2\sqrt{2}\cdot\frac{2\pi}{3\sqrt{6}}U_{d0} = \frac{4\pi}{3\sqrt{3}}U_{d0} = 2{,}42\,U_{d0}$$

i.e. is higher in 15%, too, than at the general regime. It leads to the same increasing of valve-established powers.

As a result, as all half-wave rectifying schemes, this scheme is rational also attached to low values of a rectified voltage, but high values of a rectified current, because a rectified current consists from the anode currents of six valves (but not three, as at all considered basic rectifiers of a three-phase voltage). Attached to it you must take into account a possibility of a voltage increasing at a rectifier output in 15% at regimes, close to an idling.

2.8. THREE – PHASE RECTIFIER ON BRIDGE SCHEME ($m_1=m_2=3$, $q=2$)

Full-wave rectifying schemes, characterized by an alternating current at the transformer secondary windings (on a determination), are less critical to a connecting scheme of the transformer primary and secondary windings. A connecting of the primary and secondary windings of a transformer into a star is the most spread, a such rectifier scheme is showed at a figure 2.8.1,a.

To make an analysis of a new scheme of a full-wave rectifying more simply **we shall use a way of a new scheme brining** to something known already again. It is may to image a bridge rectifier conventionally as a consecutive connection of two zero rectifying schemes, disconnecting the transformer secondary windings, as it is showed at a figure 2.8.1,b. First zero scheme is made by a cathode valve group (plus) and by a zero point of the transformer secondary windings, connected into a star (minus). A second zero scheme is made by an anode valve group (minus) and by a zero point of the same transformer secondary windings again (plus).

An analysis aim is the same as at all basic rectifying schemes: to learn scheme properties and to determine the rational using fields on it. A methodic of two-stage analysis is the same, too.

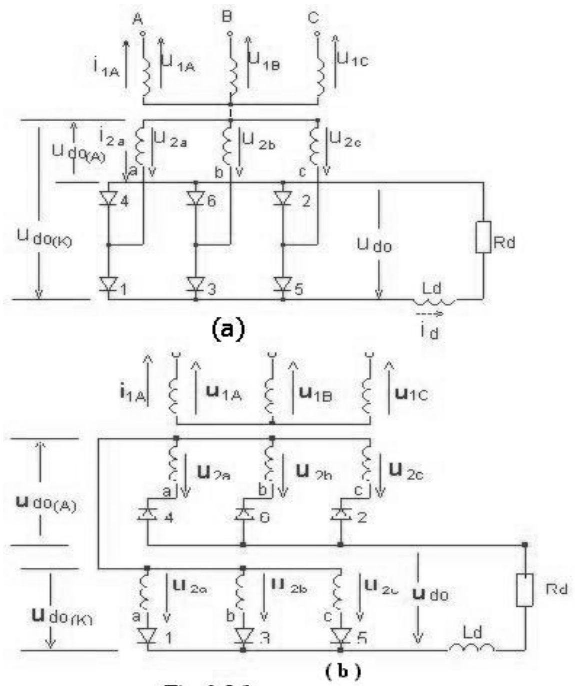


Fig. 2.8.1

There are showed time diagrams of the scheme voltages and currents at a figure 2.8.2. There is represented a three-phase system of transformer secondary winding voltages at the first diagram. Conducting intervals of valves of the cathode and anode valve groups are showed here, and also there are brought curves of rectified voltages of these groups $u_{d(k)}$, $u_{d(a)}$ relatively a zero point of the secondary windings.

It is seen that there is working one valve from a cathode group and one from an anode group at any time.

There are represented curves of a rectified voltage u_d and a rectified current i_d at the second diagram. A summa of two three-pulse rectified voltages $u_{d(k)}$ and $u_{d(a)}$, which have

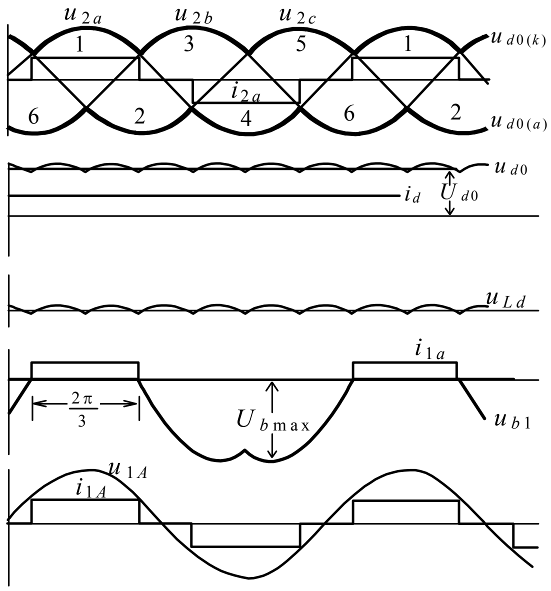


Fig. 2.8.2

displaced pulsation at a half of period, gives a six-pulse curve of a rectified voltage $u_{do.}$ In contrast to zero rectifier schemes, where phase voltages are rectified, between-phase voltages, i.e. linear, are rectified at a bridge scheme, as it is seen from a diagram.

There is represented a voltage curve at a smoothing reactor u_{Ld} at the third diagram.

At the fourth diagram there are represented a curve of a valve anode current and a curve of a reverse voltage at it, built on the same methodic as at the zero schemes. Knowing a form of valve anode currents, it is may to build currents at all transformer secondary windings now. So, a current at a secondary winding of a transformer phase *a* i_{2a} is equal to an algebraic summa (taking into account their direction) of the anode currents i_{a1} and i_{a4}, flowing on a winding correspondingly at a positive and negative half-waves of a secondary voltage on a determination of a full-wave rectifying, as it is showed at the first diagram. At the fifth diagram there are represented voltage curves of a primary wing of a phase A u_{1A}, given by a maim, and a current of the same winding i_{1A}. A current at the transformer secondary winding is an alternating (without a constant component), it is transformed to a primary winding with the same form. A strict explanation of this result again it is may to make by an equation creating for transformer magnetizing forces on a second Kirchhoff law for magnetic networks.

Using the corresponding process analogies at the given and earlier considered rectifiers of a three-phase current, it is not difficult to get necessary calculated correlation at the same fifth-step calculation procedure.

1. An average value of a rectified voltage at a bridge scheme is higher in two times, than at a zero scheme, i.e.

$$U_{d0} = 2\frac{3\sqrt{6}}{2\pi}U_2 = \frac{3\sqrt{6}}{\pi}U_2 = 2{,}34\,U_2$$

2,3,4. An average, effective and maximal values of valve currents is the same as at a zero scheme, i.e.

5. But a relative value of a maximal reverse voltage at a valve is lower in two times here (because of an increasing of a rectified voltage in two times)

$$I_a = \frac{I_d}{3}, \quad I_{a.e} = \frac{I_d}{\sqrt{3}}, \quad I_{a.\max} = I_d.$$

$$U^*_{B.\max} = \frac{U_{d.\max}}{U_{do}} = \frac{\sqrt{2}\sqrt{3}U_2\pi}{3\sqrt{6}U_2} = \frac{\pi}{3} = 1{,}045.$$

6. A valve established power

- with not full control

$$S^*_{V1} = n\frac{I_a U_{B.\max}}{P_{do}} = 6\frac{I_d \pi U_{do}}{3\cdot 3P_{do}} = \frac{2\pi}{3}$$

- with a full control

$$S^*_{V2} = n\frac{I_{a.\max}U_{B.\max}}{P_{do}} = 6\frac{I_d \pi U_{do}}{3P_{do}} = 2\pi .$$

7,8. The same current form at the transformer primary and secondary windings (with an exactness to K_t) means a similarity of the calculated correlation for effective values of these currents, taking into account (1.1.3):

$$I_1 = \frac{I_2}{K_T} = \frac{I_d}{K_T}\sqrt{\frac{2}{3}}$$

9,10,11. An equality of voltage forms at primary and secondary transformer sides and current forms at it means an equality of full winding powers of primary and secondary sides, i.e.

$$S^*_2 = S^*_1 = S^*_T = \frac{3U_2I_2}{P_{d0}} = 3\frac{\pi\sqrt{2}}{3\sqrt{6}\sqrt{3}} = \frac{\pi}{3} = 1{,}045$$

12. A quality of a rectified voltage is the same as at the last six-pulse rectifying scheme with an equalizing reactor, i.e.

$$\overline{K}_h = 0.0067, \qquad K_C = 0.057.$$

13. An input coefficient of a rectifier power is high, too

$$\chi = \frac{1}{S^*_1} = 0{,}955$$

14. A converting coefficient of a rectifier on a voltage is higher in two times

$$K_{C.V.} = \frac{U_{do}}{U_1} = 2{,}34K_T$$

15. A converting coefficient of a rectifier on a current

It is may to do the next **conclusions** at a base of the made formal analysis, comparing the got results

$$K_{CC} = \frac{I_d}{I_{1(1)}} = \sqrt{\frac{3}{2}}K_T\frac{\pi}{3} = 1{,}282K_T, \quad K'_{CC} = \frac{I_d}{I_1} = 1.225K_T$$

with the analysis results of considered earlier rectifiers of a three-phase current.

- A bridge rectifying scheme of a three-phase current has the best using of a transformer established power between all schemes.
- A quality of an output voltage and an input current of a rectifier is the same here as at a six-pulse scheme with an equalizing reactor.
- A valve using on a reverse voltage at a full-wave (bridge) rectifying scheme is better in two times than at all half-wave (zero) rectifying schemes of a three-phase current, what has an analogy with a situation of half-wave and full-wave rectifying schemes of a one-phase current.
- A specific character of a bridge scheme is a rectified current flowing through two connected consecutively valves and because there are double losses of a voltage and a power in comparison with half-wave rectifying schemes.

So, a totality of merits of a three-phase bridge rectifying scheme makes it a **prima-scheme** between all rectifying schemes and provides a wide spread using of it, besides cases with a small values of a rectified voltage and vary big values of a rectified current.

A resulting properties table of basic rectifiers schemes is represented at a part 4.1.

2.9. CONTROL RECTIFIERS.
REGULATION CHARACTERISTIC

A control rectifier – is a scheme of a basic rectifier, made at control valves. There are possible two ways of a regulation of a rectified voltage average value at the rectifiers with not completely control valves: a phase regulation, a relay regulation.

2.9.1. PHASE REGULATION

Attached to a phase regulation a changing *of a regulation degree* α of the control valves at the basic rectifying schemes gives a possibility of a regulation of a rectified voltage average value. **An aim of our analysis** is a dependence determination of a rectified voltage average value from the control parameters.

A rectified voltage curve at a common case of a m_2-phase rectifier is showed at the first diagram of a figure 2.9.1, a diagram of a valve anode current and a reverse voltage at it is represented at the second diagram. From these diagrams analysis at a qualitative stage follows that a merit of a control rectifier, bonded with a possibility of an output voltage regulation, is accompanied by **three facts.**

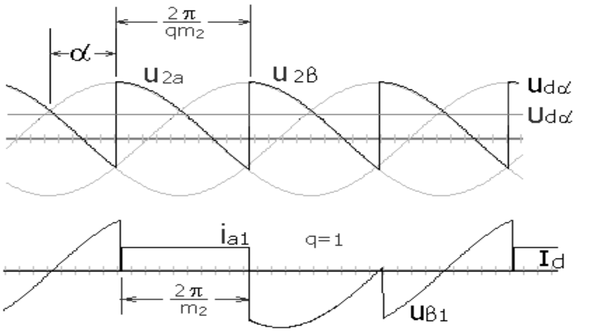


Fig.2.9.1

1.An average value regulation of a rectified voltage is reached because of a deformation (distortion) of an instantaneous curve form of a rectified voltage, i.e. is bonded with a worsening of a rectified voltage quality (a quantitative increasing K_p and $\bar{K}_h$ will be showed at a part 3.7) and, hence, leads to an increasing of a smoothing reactor inductance.

2. An increasing of a regulation degree α attached to a rectified voltage lowing is accompanied by the same increasing of an anode current displacement relatively an alternating voltage at a rectifier input. Currents of transformer windings, which were determined through the anode valve currents attached to a building, are displaced on a phase on an analogy. And currents, late on a phase relatively a voltage (as people, late from leaders at a society) reduce their full «efficiency», because a given away active power in according with (1.3.7) is reducing attached to it at a cosinus function of a displacement degree. An estimate of a regulation influence at an input power coefficient will be given at a part 3.10.

3. After a reverse voltage is applied to a valve, during of which it must restore the control properties, a forward voltage is applied to a valve. A valve must be turned off attached to this voltage till when a signal to turn on a valve at its control electrode is applied.

Now we shall determine a quantitative dependence of an average value of a rectified voltage of an ideal rectifier $U_{d\alpha o}$ from a regulation degree α, which is called a *regulation characteristic* of a control rectifier. In accordance with a diagram $u_{d\alpha}$ we have at a figure 2.9.1

where (2.9.1)

$$U_{d_{\alpha 0}} = \frac{1}{\frac{2\pi}{qm_2}} \int_{-\frac{\pi}{qm_2}}^{\pi/qm_2} \sqrt{2}\, U_2' \cos\vartheta \, d\vartheta = U_2' \frac{\sin\frac{\pi}{qm_2}}{\frac{\pi}{qm_2}} \cos\alpha = U_{d_0} \cos\alpha$$

$$U_2' = \begin{cases} U_2 & attached.to.q = 1 \\ \sqrt{3}U_2 & attached.to.q = 2 \end{cases}$$

A relation of a rectified voltage average value of a control rectifier to a rectified voltage average value of a not control rectifier is called *a regulation degree of a rectified voltage* and designated as C_p. So an equation of a regulation characteristic at relative units has such a look

$$C_P = \frac{U_{d\alpha o}}{U_{do}} = \cos\alpha..$$

(2.9.2)

A graphic of this dependence is showed at a figure 2.9.2.

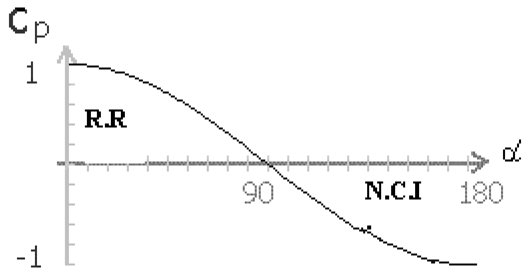


Fig. 2.9.2

A rectifying regime of a scheme work at the ideal elements appears attached to 0 < α < 90°, attached to 90° < α < 180° - a regime of a naturally commuting inverting, which will be considered at a part 3.3.4.

2.9.2. RELAY REGULATION

A second way of a regulation of a voltage average value at a rectifier output is provided by *a relay (cyclical) control algorithm.* Attached to it a voltage at a rectifier output at a control period (cycle period) takes two values: a maximal rectified (attached to α = 0) or a zero value, as it is showed at the time diagram of a figure 2.9.3 for a two-pulse rectifier. If to make a rectifier on a half-wave (zero) rectifying

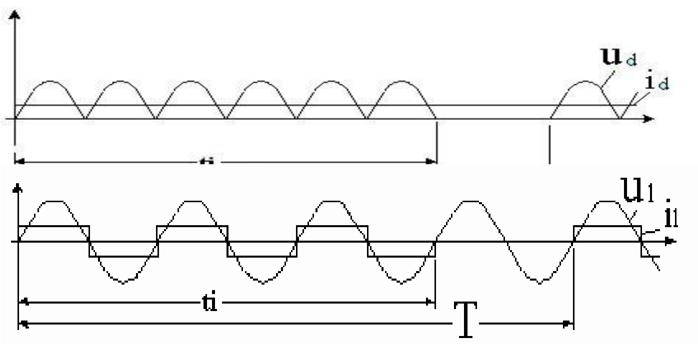


Fig. 2.9.3

scheme a zero value of a rectified voltage, if to save a flowing possibility a constant rectified current ($X_d = \infty$) at a load, is provided by an addition of a zero valve at a rectifier output, on an analogy with a showed at a figure 1.2.1, a. If to make a rectifier on a full-wave (bridge) scheme, hence, a zero valve function is made by two consecutive valves of a bridge scheme one arm.

An average value of a rectified voltage is regulated attached to this correlation changing of voltage presence duration at a load with a period duration T (cycle). A regulation characteristic equation is expressed here through porosity by an evident way [look (1.1.4)]

$$C_P = \frac{t_i}{T},$$

(2.9.3)

and a regulation degree of a rectified voltage is a linear control function.

In comparison with a considered higher phase regulation way of a rectified voltage a relay way has the merit that a rectifier input current is always in phase with a maim voltage and reactive power of a rectifier displacement is equal to zero. Hence, there increases an input power coefficient, which now will equal to, taking into account (2.9.3)

$$\chi = \frac{P_1}{S_1} = \frac{E_{d\alpha} I_d}{m_1 U_1 I_1} = \frac{C_P E_{do} I_d}{m_1 U_1 \dfrac{I_d}{K_T}\sqrt{\dfrac{t_i}{T}}} = \frac{K_T K_{C.V.}}{m_1}\sqrt{C_P}. \tag{2.9.4}$$

A demerit of this control way is an appearing of *sub-harmonics* (harmonics of a more low frequency, than usual) at a rectified voltage and a primary current, what is bonded with a period increasing of all electromagnetic processes at a scheme with $T_{1/p}$ for a rectified voltage and T_1 – for an input current till a cycle period T, which is more higher usually than a maim voltage period T_1. To save small pulsation at a rectified current an electromagnetic constant of a load network (with a filter) must be much more than a cycle period T.

So, a relay control, as more simple, is applied when a rectifier load is windings of electromagnets, electric machines, having a corresponding electromagnetic inertia.

GUESTIONS

I shall be intoxicated by a harmony at times again,
I shall be in a flood of tears over the imagination.

A.S.Pushkin

1. What elements does a control rectifier flour chart consist from?
2. What rectifier work regime is called a regime of a load discontinuous current?
3. How it is may to decrease a discontinuous current zone at a rectifier?
4. What are the differences between the full-wave rectifying schemes and half-wave rectifying schemes?
5. In what rectifying scheme of a one-phase voltage valves are used better:
   on a reverse voltage,
   on an anode current?
6. What can you say about a presence of a transformer rod forced dc magnetization by a one-directional flow at one-phase voltage rectifiers?
7. By what criterion the using zones of a three-phase current rectifiers with the schemes of winding connections ∇/λ_0 and λ/Z_0 are limited?
8. What can you say about a presence of a transformer rod dc magnetization by a one-directional flow at six-pulse rectifiers of a three-phase current?
9. What can you say about a presence of a transformer rod dc magnetization by a one-directional flow at three-pulse rectifiers of a three-phase current?
10. When is it rational to apply a three-phase current rectifier with an equalizing reactor?
11. When is it rational to apply a three-phase bridge rectifying scheme?
12. From what time moment a regulation degree α is counted down and why?
13. What new qualities has a control rectifier in comparison with a not control rectifier?
14. What does a control rectifier regulation characteristic determine? In what diapason it is need to change a regulation degree α to change a rectified voltage from a maximal value till zero?
15. What is the difference between a relay regulation way of a rectified voltage and a phase regulation way?

PROBLEMS

And fingers are just asking to a pen,
A pen - to a paper…
A.S.Pushkin

1. Calculate diode parameters at a one-phase bridge without-transformer rectifying scheme attached to $X_d = \infty$ and $R_d = 20$ Ohm.
2. Calculate diode parameters at a one-phase bridge without-transformer rectifying scheme attached to $X_d = 0$ and $R_d = 20$ Ohm.
3. Build a curve of a rectifier input current on a p.2 and calculate its power input coefficient.
4. On what scheme is it necessary to make a rectifier with U_{do} = 500 V and R_d =100 Ohm? What will be a transformer typical power equal to attached to $X_{d==\infty}$ and $X_d = 0$?

5*. Build a rectified voltage curve of a three-pulse rectifier with $X_d = \infty$ attached to a disconnecting of one diode and determine a voltage average value.

6. Build a curve of a current, consumed by a three-pulse rectifier from a maim attached to a connecting of transformer windings ∇/λ_0.

7*. Calculate a harmonic coefficient of a rectified voltage of a six-pulse rectifier.

8*. Calculate a first harmonic value of an equalizing voltage relatively to an average value of a rectified voltage at a scheme with an equalizing reactor.

9. Calculate a distortion coefficient of a rectified voltage of a control two-pulse rectifier.

10. Calculate a regulation degree value of a control rectifier to decrease a voltage in 10 times.

11*. Calculate a current distortion coefficient of a one-phase main attached to a relay regulation of a rectified voltage.

Chapter 3

A THEORY OF AC – DC CURRENT CONVERSION TAKING INTO ACCOUNT REAL PARAMETERS OF CONVERTER ELEMENTS

Прошла любовь, явилась муза,
И прояснился темный ум.

Alexander S. Pushkin

The chapter is devoted to a theory of an electric energy conversion at modes of a controlled rectification and dependent inversion at basic schemes considered at the previous chapter taking into account real parameters of converter elements.

Transformer real parameters determine a real external (load) curve of a rectifier (section 3.1). Real filter parameters at a converter dc-current link taking into account a presence of a load counter-emf cause possible real modes at a dc-current link – an interrupted or continuous pulsing current (section 3.2). A rectifier operation to a capacitor filter in a limit brings to an operation at a counter-emf (section 3.3). In the section 3.4 a specific of a converter transfer from a controlled rectification mode to a dependent inversion mode is considered. In the next six sections common patterns of investigated converters are considered: for instantaneous primary currents, their spectrums, spectrums of inverted and rectified voltages, optimal number of transformer phases, for a commutation influence at transformer processes, a performance index and converter power factor. Sections 3.11 and 3.12 are devoted to rectifier schemes at fully controlled valves and reversible valve converters, correspondingly. In the section 3.13 a problem being actual in the last time of reverse influence of valve converters at a mains is considered.

3.1. A COMMUTATION PROCESS AT A CONTROLLED RECTIFIER HAVING A REAL TRANSFORMER. AN EXTERNAL CURVE

At rectifiers at ideal elements considered at the previous chapter a current commutation process, i.e. a current transfer from a transformer phase with a valve finishing to operate to a transformer phase with a valve starting to operate, was instantaneous that was accompanied with a current step. In real circuits there is always an inductance (introduced or self-inductance) in which current steps are impossible, hence, an instantaneous commutation is impossible. Obviously, a real transformer will significantly influence to a commutation by it's reactive resistances that is task of our analysis.

A known T-type *transformer equivalent circuit* is represented at the fig. 3.1.1.

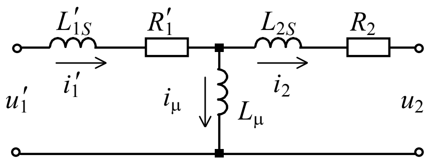


Fig. 3.1.1

In contrast to power engineering specialists which reduce transformer parameters to a primary, i.e. mains, winding, it's necessary in conversion technique to reduce equivalent circuit parameters to a secondary winding. It's connected with a reverse direction constructing electromagnetic processes at a rectifier transformer: firstly, as it was shown currents at the secondary transformer windings are formed up, then – at the primary windings.

A direct using of a full T-type transformer equivalent circuit with a three inductances at a calculated model of a rectifier will complicate an analysis so much (at the section 1.2.1 we took into account only one inductance) that calculation will be almost impossible in the analytical form. That's why there is necessary a rational according to the aims of the analysis an equivalent circuit simplification. A professional skill of a specialist is in his capability to find permissible simplification of a task mathematical model. The calculation itself after this simplification has in common a technical character and is permissible for all persons by applying computer aids. A simplification is based on a clear understanding of processes physics, dividing them on significant and not significant for analysis aims.

A current commutation process at a rectifier transformer is accompanied with turning-off and connecting secondary transformer windings to a load, primary transformer windings are all the time connected to mains. Hence, there is almost no basic magnet flux changing at a transformer while a commutation. A value of a magnetizing component of a power transformer i_μ is equal to some percents from the current conditioned by a load. That's why a magnetizing inductance L_μ can be excluded from an equivalent circuit at this stage. As a result of the first simplification step we obtain an equivalent circuit having a one inductance L_a that is equal to a sum of leakage inductances of reduced primary and secondary transformer windings and one active resistance R that is also equal to a sum of active resistances of reduced primary and secondary windings. A leakage inductance of a transformer reduced to the secondary (anode) transformer winding of a rectifier is called *an anode inductance.*

A necessity of the second step for a transformer calculated equivalent circuit simplification is connected with the circumstance that calculated correlations at *LR*-circuit with a valve as it's seen from the section 1.2.1 have a transcendent character that doesn't allow to obtain final analytical correlations. That's why how it's known from an experience that a reactive leakage resistance of transformers of an average and great powers is greater than an active resistance of windings in 3…5 times. We can neglect by the last of them. Since an influence of L_a at a rectified voltage through a commutation is equal to some percents then an influence of R_a will be equal to about one percent and second made simplification is also seriously.

Thus, we will make an influence estimation of a real transformer RT to a current commutation process at a rectifier by changing a real transformer to a set of an ideal transformer IT (as at the previous model) and a summary leakage inductance of windings reduced to the secondary side (L_a) as it's shown at the fig. 3.1.2.

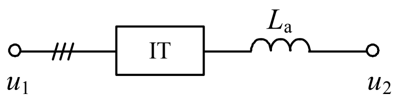


Fig. 3.1.2

At the fig. 3.1.3 there is represented a controlled rectifier scheme that real transformer parameters are considered by connection of reactive resistances $X_a = \omega L_a$ to the secondary windings. Valves and smoothing reactor as before are ideal to find out an influence of a one novel element – X_a. Obviously, that at connecting the next valve, for example VD_3, it's current will increase with a finite velocity, at a

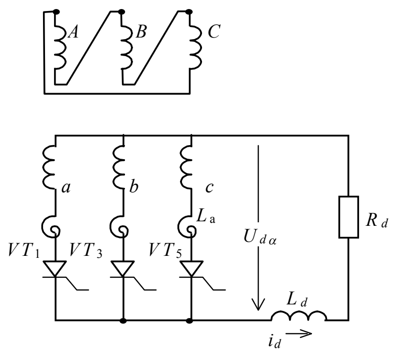


Fig. 3.1.3

valve VD_1 that finishes to operate it's current will decrease also with a finite velocity. Hence while a current commutation time a *commutation contour* is formed up from a phase voltage U_{2b}, it's X_a, valves VD_3 and VD_1, phase voltage U_{2a} and it's X_a.

In the common case a commutation contour is a closed contour formed by a branch with a valve that starts to operate and a branch with a valve that finishes to operate. At a commutation contour a presence of a (commutating) voltage supply is necessary. If it's a mains voltage (or a receiving mains in the case of naturally commutating inverter, section 3.3.4) then a such commutation is called *a natural commutation.*

A differential equation for a current at a commutation contour $i_к$ is the next

$$2X_a \frac{di_к}{d\vartheta} = u_{2a} - u_{2b}\,. \tag{3.1.1}$$

A phase-to-phase voltage vector $\mathbf{U}_{2к}$ at m_2-phase system under the influence of which is a commutation process is determined by a vector diagram represented at the fig. 3.1.4.

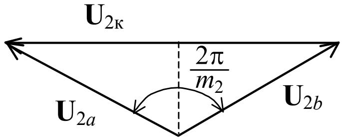


Fig. 3.1.4

Then a solution of the equation (3.1.1) for a commutation current that coincides with an anode current of the valve that starts to operate in condition that a time reference is placed into a point of a natural ignition is the next

$$i_к = i_{ai} = \frac{1}{X_a}\int \sqrt{2}U_2 \sin\frac{\pi}{m_2}\sin\vartheta\, d\vartheta = = -\frac{\sqrt{2}U_2}{X_a}\sin\frac{\pi}{m_2}\cos\vartheta + C_1\,.$$

(3.1.2)

An integration constant C_1 is determined from an initial condition $i_{ai} = 0$ at $\vartheta = \alpha$:

$$C_1 = \frac{\sqrt{2}\,U_2}{X_a}\sin\frac{\pi}{m_2}\cos\alpha\,. \tag{3.1.3}$$

Taking into account this constant a solution of the equation (3.1.2) is the next

$$i_{\mathrm{ai}} = \frac{\sqrt{2}U_2 \sin(\pi/m_2)}{X_{\mathrm{a}}}(\cos\alpha - \cos\vartheta). \tag{3.1.4}$$

A commutation process duration is determined by a condition of achieving by a current of VD_3 valve starting to operate a value equal to I_d at this a current of the valve finishing to operate decreases to a zero since at an ideal filter a commutation contour is broken.

$$i_{\mathrm{a1}} + i_i = i_d = I_d \tag{3.1.5}$$

An interval of a (relative) time during which at a commutation contour both valves taking part in a commutation process conduct a current is called *a commutation angle* and is denoted as γ. A condition $i_{\mathrm{ai}} = I_d$ at $\vartheta = \alpha + \gamma$ is substituted into the equation (3.1.4):

$$I_d = \frac{\sqrt{2}U_2 \sin\pi/m_2}{X_{\mathrm{a}}}\left[\cos\alpha - \cos\left(\alpha+\gamma\right)\right], \tag{3.1.6}$$

from here we obtain a formula to calculate a commutation angle

$$\gamma = \arccos\left[\cos\alpha - \frac{I_d X_{\mathrm{a}}}{\sqrt{2}U_2 \sin(\pi/m_2)}\right] - \alpha. \tag{3.1.7}$$

Thus, there are determined rules of current changing of a valve starting to operate (3.1.4) and a valve finishing to operate (3.1.5) at a commutation interval and the interval duration. A Character of changing of a rectified voltage instantaneous value at a commutation interval $u_{d\gamma}$ when two transformer phases conduct a current. Values u_{2a} and u_{2b} are determined by a syncopation supposing that a load network is a current supply I_d:

$$u_{d\gamma} = \frac{u_{2a} + u_{2b}}{2}. \tag{3.1.8}$$

A rectified voltage at an interval γ is formed as a half-sum of transformer phase voltages taking part in a commutation process.

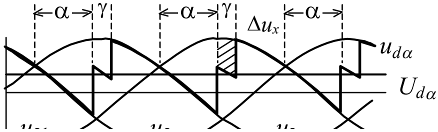


Fig. 3.1.5

At the fig. 3.1.5 time diagrams of a rectified voltage $u_{d\alpha}$ and anode currents taking into account a commutation are represented.

It's characteristic that at out-of-commutation interval, i.e. at an interval with one conducting valve, an instantaneous curve of a rectified voltage coincides with a transformer secondary voltage curve though a presence of an inductance L_a at a valve anode network. A self-induction voltage at a network from a current i_d flowing is equal to zero since a derivative of an ideally smoothing rectified current is also equal to a zero.

From the diagram it's seen that a real commutation of currents led to losing of the area under a rectified voltage curve at a value of commutating voltage drop Δu_x (it's shaded at the diagram). It means a decreasing of an average value of a rectified voltage that depends on a commutation angle γ, hence, according to (3.1.7) on an average value of a rectified current at a constant regulation angle α. This response $U_{d\alpha} = f(I_d)\alpha = \text{const}$ is called *a rectifier external curve*. (A word "external" – mnemonic prompting of a load network location that is connected from the outside to a rectifier). An equation of a rectifier external curve is wrote down from an evident consideration: a voltage at a rectifier output at a presence of a load is equal to a difference of an idling voltage $U_{d\alpha 0}$ and a rectifier voltage drop from a commutation ΔU_x at appearing a load current:

$$U_{d\alpha} = U_{d\alpha 0} - \Delta U_x = U_{d0}\cos\alpha - \Delta U_x, \qquad (3.1.9)$$

where ΔU_x – an average value of a commutation voltage drop that in the common case is equal to

$$\Delta U_x = \frac{1}{2\pi/qm_2}\int_{\alpha}^{\alpha+\gamma}\Delta u_x d\vartheta = \frac{1}{2\pi/qm_2}\int_{\alpha}^{\alpha+\gamma} X_{\text{a}}\frac{di_{a3}}{d\vartheta}d\vartheta =$$

$$= \frac{1}{2\pi/qm_2}\int_{0}^{I_d} x_{\text{a}} di_{\text{a}3} = \frac{X_{\text{a}} I_d}{2\pi/qm_2}. \qquad (3.1.10)$$

After substituting an equation (3.1.10) into (3.1.9) we obtain in an explicit form a rectifier external curve equation taking into account a commutation

$$U_{d\alpha} = U_{d0}\cos\alpha - I_d\frac{X_{\text{a}}}{2\pi/qm_2}. \qquad (3.1.11)$$

Remarkable that external curves that are straight lines for the different values of a regulation angle α are parallel since a commutation voltage drop ΔU_x doesn't depend on it. Diagrams of external curves are shown at the fig. 3.1.6.

Another characteristic result is that an influence of an inductance L_a through a commutation process to an average value of a rectified voltage formally is analogical to an influence of an equivalent inherent quasi-active resistance $R_{inh.eq}$:

$$R_{inh.eq} = \frac{X_a}{2\pi / qm_2}. \tag{3.1.12}$$

It allows to represent a rectifier on an output by an equivalent circuit containing a dc voltage generator of a value equal to $U_{d0} \cos\alpha$ and a quasi-active resistance $R_{inh.eq}$ connected with it in series (fig. 3.1.7).

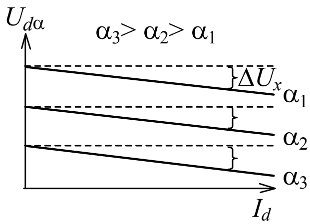


Fig. 3.1.6

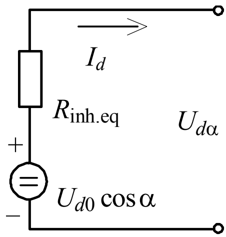


Fig. 3.1.7

The resistance is called quasi-active because though a voltage drop on it equal to ΔU_x there are no active power loss in it and there is no active power loss in commutation process itself conditioned by a reactivity L_a.

In the same time an obtained result (3.1.12) prompt us that the active resistances at current flowing contour inside a rectifier to external clamps will exert to a rectified voltage the same influence as a quasi-active inherent equivalent resistance $R_{inh.eq}$. It allows to combine into a summary inherent resistance of a rectifier R_r active resistances of transformer windings reduced to the secondary side $(R_1' + R_2)$, a dynamic valve resistance at a direct direction R_{dyn}, an active resistance of a smoothing reactor winding of an output filter R_f, and also $R_{inh.eq}$:

$$R_{\mathrm{r}} = R_{\mathrm{inh.eq}} + R_1' + R_2 + R_{\mathrm{dyn}} + R_{\mathrm{f}}. \qquad (3.1.13)$$

Taking into account an equation (3.1.13) a generalized equation of an external curve has the next view

$$U_{d\alpha} = U_{d0}\cos\alpha - I_d\left(\frac{X_{\mathrm{a}}}{2\pi/qm_2} + R_1' + R_2 + R_{\mathrm{dyn}} + R_{\mathrm{f}}\right) - q\Delta U_0 .$$

(3.1.14)

The last item considers the second parameter of a real valve at a conducting state – a cutoff voltage of a voltage-ampere curve line ΔU_0.

For electric power engineering specialists the more convenient term is a such parameter of a transformer as *it's short circuit voltage* $U_{1\text{к}}\%$, but not an inductive leakage resistance of a transformer reduced to the secondary side X_{a}. Two these parameters are connected by an evident formula:

$$X_{\mathrm{a}} = \frac{X_{1\text{к}}}{K_{\text{т}}^2} = \frac{U_{1rat}}{I_{1\mathrm{rat}}K_{\text{т}}^2}\frac{U_{1\text{к}}\%}{100} = \omega L_{\mathrm{a}} . \qquad (3.1.15)$$

Thus, at the second stage of an analysis – investigation of a rectifier having real elements – there are taken into account real parameters of a transformer, valves, output filter (besides an assumption $L_d = \infty$) that extending results of a first analysis stage with ideal elements widen application boundaries of a theory to practical tasks. Finally, we have to escape the last assumption and to find out an influence of an inductance finite value at a load network L_d first of all at two rectifier curves basic for user – an external and regulating curves. We will do it in the next section.

3.2. A RECTIFIER OPERATION AT A COUNTER-EMF AT A FINITE VALUE OF A SMOOTHING INDUCTANCE

A rectifier operation at a counter-emf at a load network is the most common case of real loads. Such loads containing counter-emf are:

1) an anchor network of a dc current motor containing else a motor rotation emf at an equivalent circuit besides *RL*-parameters of an anchor winding;

2) accumulators substituted for a emf supply having a not significant active inherent resistance;

3) galvanic tubs at a chemical and metallurgical productions having an oncoming emf of a solution or flux;

4) electric arcs of a weld, gas-discharge lighting devices, plasma plants and so on.

Conditionally a rectifier operation to an active-inductive load (field windings of electric motors, relay windings and so on) can be relative to the such mode at a stage of a current decay at a winding (an extinction of a winding field). At this a magnetic field electric energy of a winding returns (recuperates) to a mains. A rectifier has also a proper inherent counter-emf according to an equation (3.1.14) equal to $q\Delta U_0$ that significantly influences at a rectified current at small value of a rectified voltage (low-voltage rectifiers or rectifiers regulated to small values of voltages having a great rated rectified voltage).

A model scheme of a controlled rectifier comprising all the pointed cases is represented at the fig. 3.2.1.

The task of this section is consideration of an influence of an inductance finite value at a load network L_d at two basic curves of a rectifier: an external and regulating curves. If it's necessary a result of electromagnetic processes analysis at this model can be spread to analysis of an influence of a finite value L_d at processes at another circuits besides an output circuit on that approach which was used above while analyzing basic rectification cells.

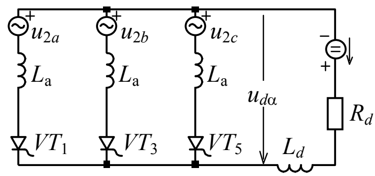


Fig. 3.2.1

At a finite value of an inductance at a load network there may appear a qualitatively novel mode of a rectifier operation – a mode of *an interrupted rectified current* as it was noticed at the section 2.2. That's why we will make an analysis of modes of an interrupted current, boundary limit-continuous current, continuous current in series.

3.2.1. A MODE OF AN INTERRUPTER CURRENT ($\lambda < 2\pi/qm_2$)

We consider time diagrams of a rectified voltage and current represented at the fig. 3.2.2 for a qualitative analysis of electromagnetic processes.

In this mode a current of a conducting valve decreases to a zero earlier than a control impulse achieves the next valve and a null pause is formed at a rectified current. A separate operation of all the valves means a dependency of their modes from each other that's why an equivalent circuit of any rectifier at an interval of a valve conducting state has a view shown at the fig. 3.2.3.

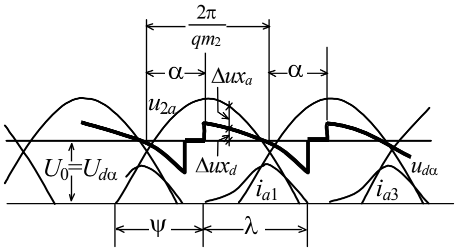


Fig. 3.2.2

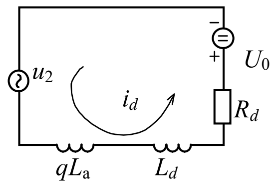


Fig. 3.2.3

If it's necessary we can combine into a resistance R_d all inherent active rectifier resistances that are in (3.1.13). Identically we can combine cutoff voltages of a voltage-ampere curve line of conducting valves $q\Delta U_0$ into a load counter-emf. A transformer leakage inductance L_a is with a multiplier q since at $m_2 = 3$, $q = 2$ a current flows at two transformer phases.

As it was shown at the section 2.2.2 at an interrupted rectified current mode at RL-load a current flowing duration λ is determined from a solution of a transcendental equation that means a impossibility to obtain an analytical expression for an average value of a rectified current. It, in their turns, means an absence of a closed analytical expression for an external rectifier curve. A rectifier operation analysis for a such common case is made at works of A. A. Bulgakov [40] and C. V. Zaharevich [41], it's a complicated enough, that's why we consider only a case without an active resistance at a rectified current network ($R_d = 0$). At this an external curve can be obtained as equations at a parametric form:

$$U_{d\alpha} = f_1(\alpha, \lambda), \qquad I_d = f_2(\alpha, \lambda). \qquad (3.2.1)$$

Then taking values equal to $\lambda_j < 2\pi/qm_2$ we can calculate on an equation (3.2.1) their corresponding values $U_{d\alpha_j}$, I_{d_j} at $\alpha = \text{const}$ and, thus, to obtain external curve points.

Firstly we determine an average value of a rectified voltage integrating a curve of it's instantaneous value while a current flowing duration λ (while a current is equal to a zero instantaneous values of a rectified voltage and counter-emf coincide, hence, their average values at a pause interval coincide, too) according to the diagram represented at the fig. 3.2.2:

$$U_{d\alpha} = \frac{1}{\lambda}\int_0^{\lambda} u_{d\alpha}\, d\vartheta = \frac{1}{\lambda}\int_0^{\lambda}\left(u_2 - qX_\text{a}\frac{dia}{d\vartheta}\right)d\vartheta = \frac{1}{\lambda}\int_{\psi}^{\psi+\lambda}\sqrt{2}U_2 \sin\vartheta\, d\vartheta =$$

$$= \frac{\sqrt{2}U_2}{\lambda}[\cos\psi - \cos(\psi+\lambda)] = f_1(\alpha,\lambda), \qquad (3.2.2)$$

where $\psi = \pi/2 - \pi/qm_2 + \alpha$ – an angle when a valve starts to operate. The angle is counted relatively a secondary voltage null where a time reference is placed.

Since an average value of a rectified voltage is balanced by an average value of a counter-emf, the next follows from an equation (3.2.2):

$$U_0 = U_{d\alpha} = \frac{\sqrt{2}\,U_2}{\lambda}\left[\cos\psi - \cos\ (\psi+\lambda)\right] = f_1(\alpha,\lambda)$$

(3.2.3)

To calculate an average value of a rectified current it's necessary to find out an it's instantaneous value expression that is foun out as a differential equation solution for a current that according to a calculated equivalent circuit (fig. 3.2.3) has the next view

$$(qX_\text{a} + X_d)\frac{di_d}{d\vartheta} + i_d R_d = u_2 - U_0\,. \qquad (3.2.4)$$

A direct integration of this equation at R_d = 0 gives the next equation

$$i_d = \frac{\sqrt{2}\ U_2'}{qX_\text{a} + X_d}[-\cos\vartheta - \tau\vartheta] + C_1\,, \qquad (3.2.5)$$

where $\tau = \dfrac{U_0}{\sqrt{2}U_2'}$ – a relative value of a counter-emf,

$$U_2' = \begin{cases} U_2 & at \quad m_2 = 3, \quad q = 1, \\ \sqrt{3}\,U_2 & at \quad m_2 = 3, \quad q = 2. \end{cases}$$

An integration constant is determined from the condition $i_d = 0$ at $\vartheta = \psi$

$$C_1 = \frac{\sqrt{2}U_2^3}{qX_a + X_d}[\tau\psi + \cos\psi]\,.$$

Then a solution (3.2.5) will have the next view

$$i_d = \frac{\sqrt{2}U_2'}{qX_a + X_d}\left[\cos\psi - \cos\vartheta + \tau\left(\psi + \vartheta\right)\right]. \qquad (3.2.6)$$

From the equation (3.2.6) we find out an average value of a rectified current

$$I_d = \left(\frac{2\pi}{qm_2}\right)^{-1}\int_0^{\lambda} i_d\, d\vartheta =$$

$$= \frac{\sqrt{2}U_2' q m_2}{(qX_a + X_d)2\pi}\int_{\psi}^{\psi+\lambda}[\cos\vartheta - \cos(-\psi_0 + \alpha) + \tau(\psi - \vartheta)]\,d\vartheta =$$

$$= \frac{\sqrt{2}U_2' q m_2}{(qX_a + X_d)2\pi}\left[\sin(\psi + \lambda) - \sin\psi + \cos\psi - \tau\,\frac{\lambda^2}{2}\right] = f_2'\ (\alpha,\ \lambda,\ \tau).$$

(3.2.7)

If to substitute into (3.2.7) a value of τ determined from the equation (3.2.3) then we obtain a dependence in a clear form

$$I_d = f_2'(\alpha,\lambda,\tau) = f_2'(\alpha,\lambda,f_1(\alpha,\lambda)) = f_2(\alpha,\lambda)\,. \qquad (3.2.8)$$

Usually they are limited by the dependence (3.2.7) for a rectified current.

3.2.2. A LIMIT-CONTINUOUS CURRENT MODE ($\lambda = 2\pi/qm_2$)

Correlations for average values of a rectified voltage and current at the pointed boundary condition are obtained from the equations (3.2.2) and (3.2.7), correspondingly taking into account (3.2.3) at substituting into them a value equal to $\lambda = 2\pi/qm_2$:

$$U_{d\alpha \text{г}} = \sqrt{2}\ U_2' \frac{qm_2}{\pi} \sin\frac{\pi}{qm_2} \cos\alpha = B\cos\alpha\,, \qquad (3.2.9)$$

$$I_{d\alpha \text{г}} = \frac{\sqrt{2}\,U_2'}{qX_\text{a} + X_d} \frac{1}{\pi} \sin\frac{\pi}{qm_2} \left(1 - \frac{\pi}{qm_2} \operatorname{ctg}\frac{\pi}{qm_2}\right) \sin\alpha = A\sin\alpha\,. \qquad (3.2.10)$$

Correlations (3.2.9) and (3.2.10) determine an ellipse arc equation at a parametric form at external curves graphic shown at the fig. 3.2.4 and separate *areas of continuous and interrupted* rectified *currents*.

3.2.3. A CONTINUOUS CURRENT MODE ($\lambda > 2\pi/qm_2$)

In the mode of a continuous rectified current it's period consists of two subintervals: out-of-commutating and commutating. At an out-of-commutating interval one valve conducts a current (at half-wave rectification schemes), at a commutating interval two valves conduct a current. A number of differential equations for a current becomes greater than at an interrupted current mode in three times (three equations) that leads to formulas complication, they become complicated for engineering calculations. That's why it can be used an approximate methodic of an external curve forming if values of X_d are significantly greater than values of X_a. The more precise but more complicated methodic is represented in [8]. This correlation between X_d and X_a values is executed since a value of a smoothing reactor reactance is chosen from a condition of rectified current pulsations obtaining at a level of some percents from it's average value. At a calculation with an engineering precision in this case we can neglect by rectified current pulsations, i.e. we can suppose that the current is ideally smoothing as at $X_d = \infty$. Then a counter-emf at a load network can be changed by an equivalent active resistance of a load $R_{d\,\text{eq}}$:

$$R_{deq} = \frac{U_0}{I_d}\,, \qquad (3.2.11)$$

at which a rectified current makes the same constant voltage drop as U_0. A rectifier operation mode at this at a static mode isn't changed, i.e. a rectifier "doesn't feel" such load changing.

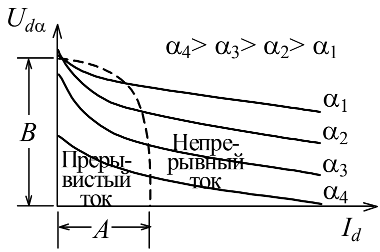


Рис. 3.2.4

A rectifier external curve equation with an active-inductive load at $X_d = \infty$ was obtained in the form (3.1.11) and in the more common rectifier model – in the form (3.1.14). Diagrams of resulting external curves of a rectifier loaded at a counter-emf are represented at the fig. 3.2.4.

An abrupt fall down of curves is significant in the range of an interrupted rectified current. It's conditioned by an abrupt dependence of a current flowing duration λ from a contour-emf changing and a limitation of a current impulse value by reactance's X_a and X_d. At a continuous current mode this limitation is determined by a commutation process in which only a reactance X_a takes part.

Another f e a t u r e of a rectifier operation at a counter-emf in the case of a finite value of X_d is a possibility of an operation mode appearing with *a forced regulation angle* α_{forc}. In this case a valve starts to operate not with a regulation angle α given on a control channel but with a forced regulation angle α_{forc} determined by a moment of a direct voltage appearing at a valve as it's shown at the time diagram at the fig. 3.2.5.

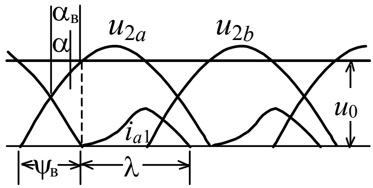


Fig. 3.2.5

A value of an angle ψ_{forc} (α_{forc}) is determined by a correlation between a secondary voltage of a transformer and a load counter-emf:

$$\psi_{\mathrm{forc}} = \frac{\pi}{2} - \frac{\pi}{qm_2} + \alpha_{\mathrm{forc}} = \arcsin\frac{U_0}{\sqrt{2}\,U_2'} . \qquad (3.2.12)$$

In these cases if $\alpha_{\mathrm{forc}} > \alpha$ then at formulas for an external curve calculation it's necessary to change α to α_{forc}.

Thus, a final value of a smoothing reactor inductance at a rectified current network leads to an appearing of an interrupted load current mode causing:

– a significant nonlinear distortion of a rectifier external curve that worsens rectifier properties as an element of an automatic control system;

– an abrupt quality worsening of a rectified current having in this mode a great exceeding of current impulse amplitudes over it's average and rms values (an increasing of an amplitude factor and crest-factor);

– a significant worsening of a rectifier input power factor at a range of λ that are small relatively $\lambda = 2\pi/qm_2$ (section 3.10);

– a fast-action increasing of a rectified current regulation to one period of it's pulsations while at a continuous current mode it's changing velocity is determined by an electromagnetic time constant of a load network (a smoothing reactor).

The second task of an investigation being formulated at this section is analysis of influence of a finite X_d value to rectifier regulation curves. Formally a dependence of a rectified voltage average value from a regulation angle α can be considered for an interrupted current mode on the equation (3.2.3). A dependence of a rectified voltage not only from α but and a duration λ of rectified current impulses flowing is evident. Thus, a regulation curve statement becomes dependent from a mode at a rectified current network, i.e. regulation curves are also distorted because of a not single-value of their statement from a regulation angle α.

In general an interrupted current mode is unfavourable for an operation since it worsens all it's basic rectifier curves (besides a fast-action that vice versa is improved). We can decrease an area of interrupted currents by increasing a smoothing reactor inductance, an equivalent number of phases of a rectified voltage qm_2 and a maximal value limitation of a regulation angle α.

3.3. A RECTIFIER OPERATION WITH A CAPACITOR SMOOTHING FILTER

A smoothing filter of an inductance L_d connected in series to a load at a rectified current network has a smoothing influence to a pulsing. In the cases when a load represented at a calculated model as rectified current in general while a rectified voltage $U_{d\alpha}$ as before is a constant or alternating active resistance R_d requires a dc voltage, it's necessary to use a smoothing capacitor C_d connected in parallel at a load network as it's shown at the fig. 3.3.1 at the example of a one-phase current rectifier.

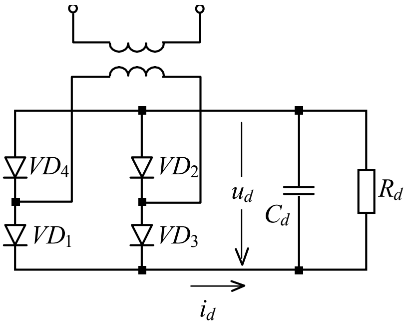


Fig. 3.3.1

A simplified analysis of a capacitor capacitance influence to a rectified voltage is an aim of this section.

A reality of transformer parameters according to results obtained at the section 3.1 is expressed by adding to an ideal transformer secondary winding a leakage inductance L_a reduced to the secondary side, i.e. L_a isn't external scheme element, it's a parameter of a transformer equivalent scheme. Only at an input transformer absence there is required a connection of a corresponding input reactor.

If a time constant of a load network $\tau_d = C_d R_d$ is high enough and we can neglect by voltage pulsations at a capacitance C_d then this scheme operation mode becomes like a rectifier operation mode at a counter-emf considered at the previous section. Let's estimate a required value of a smoothing capacitor capacitance depending on a load P_d:

$$P_d = \frac{U_d^2}{R_d}. \tag{3.3.1}$$

Let's require that a time constant of a load network to be a significantly greater than pulsations period of a rectified voltage

$$\tau_d = C_d R_d = C_d \frac{U_d^2}{P_d} >> \frac{T_1}{qm_2} = \frac{0,02}{qm_2}, \tag{3.3.2}$$

then

$$C_d >> \frac{0,02}{qm_2} \frac{P_d}{U_d^2} = \frac{0,02}{qm_2 R_d}. \tag{3.3.3}$$

For example, at an uncontrolled rectifier of a mains voltage equal to 220 V with a transformer less input an average value of a rectified voltage will be close to a mains voltage amplitude (at low loads), i.e. it's approximately equal to 300 V. Then we obtain the next expression from the equation (3.3.3) taken a tenfold exceeding of a time constant over pulsations period:

$$C_d = 0,11 \cdot 10^{-6} P_d \; [\mathrm{F}]. \tag{3.3.4}$$

Thus, required great values of a smoothing capacitor capacitance usually limit a power of one-phase rectifiers with a such filter that value is equal to about units of kilowatts. Application of a smoothing filter having an inductance L_d is rational vice versa at low values of R_d having their place at power rectifiers that are fed from a three-phase mains (at $P_d >> 3 \ldots 5$ kW). At an intermediate diapason of rectifier powers (from hundreds of watts to 3…5 kW) combined types of $L_d C_d$-filters are used at a rectifier output that a re carried out on Г-, П- and T-type schemes [10, 11].

On the other hand, at low power (tens of watts) rectifiers at a transformer equivalent scheme active winding resistances are prevail over reactive resistances of winding leakage inductances. For this case a rectifier operation analysis at an active load having a capacitor filter is made at the paper [11].

3.4. AN ACTIVE POWER FLOW TRANSFORMATION AT A CONVERTER. A NATURALLY COMMUTATING INVERSION MODE

Considered devices of ac-dc current conversion are characterized by an active power transfer from an ac mains into dc current network – a load network. In such devices an energy recuperation from dc current network into an ac current network is often required. At an electrical engineering it's actual at electric energy conversions by a dc current. A similar situation appears in those cases when a rectifying device feed an anchor network of dc current motor at an electric drive system of a some carrier or a load-lifting. At a transport motion at incline or a load-lifting downward (with a cargo) a dc current motor transfers from a motion operation mode into a generator mode because of a mechanical energy led to it from an executing mechanism. This energy can be used usefully transforming it into an electric energy and returning it through a (reversible) valve converter into an ac current mains (section 3.12). At this at converter there is a changing of an active power flow direction to a reverse direction that is called *an inversion.* A converting process of dc-ac current energy at a presence of an ac mains made by another energy supply of ac current is called *a naturally commutating inversion.* This process investigation at naturally commutating one-phase and three-phase inverters is an aim of the present section.

It's evident that changing of an active power flow direction at a dc current link at storing it's direction unchanged because of valves presence is possible only by a voltage polarity changing. This changing is provided according to a regulation curve equation of a controlled valve converter (2.9.2) at regulation angles equal to $\alpha > 90^{o}$. At this a current curve is displaced at a transformer primary winding, hence, it's fundamental is displaced, too, at an angle equal to $\varphi_{1(1)} = \alpha$. Then according to the equation (1.2.8) at $\varphi_{1(1)} > 90^{o}$ an active power sign is changed at ac current network of a valve converter, i.e. there will actually occur a power transfer to an ac mains but not it's consumption from a mains as in the case of a controlled rectification.

Let's draw attention to that now a term "a valve converter" is used instead of a term "a controlled rectifier" since we consider two possible operation modes of the same device – a mode of a controlled rectification and a mode of a naturally commutating inversion. In those cases when a mode of a naturally commutating inversion is an only (prolonged) a such device of dc-ac voltage conversion, a frequency, a form and a value of which a re determined by another existing mains is called *a naturally commutating inverter.*

A destination of a naturally commutating inverter in this case is led to a supplementary active power supply into an existing system of an ac voltage.

Also an operation of the considered valve converter to some windings of magnetic systems (a field coil of electric motors, electric magnets, superconducting storages) leads to a momentary appearing of a naturally commutating inversion mode. In the cases when there is required to lead out a stored energy fast and effectively from windings by a current chop into it, a winding voltage polarity is to be changed to a reverse that is also provided at a valve converter by a regulation angle α increasing greater than 90º. At a current slop moment to a zero a naturally commutating inversion mode, of course, will be finished since a temporary energy source will disappear at a dc current link.

Thus, rectifiers and naturally commutating inverters have the same conversion schematic circuits but naturally commutating inverters can't be made at uncontrolled valves. To make sure in it's sufficient to consider a naturally commutating inversion mode for a one scheme of a one-phase valve converter and a one scheme of a three-phase valve converter.

3.4.1. A ONE-PHASE NATURALLY COMMUTATING INVERTER (m_1=1, m_2=2, q=1)

It's necessary to notice that naming of primary and secondary transformer windings at a valve converter are stored independently from it's operation mode to eliminate a confusion with their numeration at changing of an active power flow direction through a transformer.

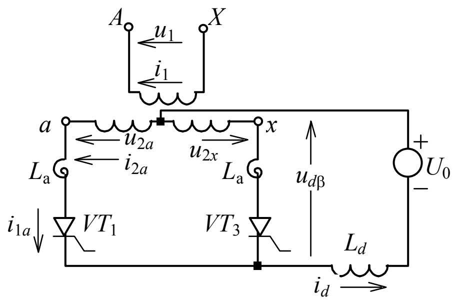


Fig. 3.4.1

A naturally commutating inverter scheme is shown at the fig. 3.4.1. A real transformer (section 2.9) is represented as a totality of an ideal transformer and leakage inductances of L_a windings reduced to the secondary side.

In contrast to a rectifier operation at an external counter-emf at a dc voltage link of an inverter an external emf polarity is changed to reverse. It's one of the conditions as it was shown above at a qualitative level to transfer a controlled rectifier operating at a counter-emf to a naturally commutating inverter mode. The second condition is that a regulation angle α has to be greater than 90º.

A task of the analysis is obtaining of a naturally commutating inverter basic curves. To calculate it them it's more appropriate to use instead of a regulation angle α a regulation angle β that supplements with an angle α to 180º:

$$\alpha + \beta = 180^{\circ}. \qquad (3.4.1)$$

It makes all the correlations of curves from an angle β at an inverter similar to correlations of the corresponding curves from an angle α at a rectifier.

To more visually represent features of electromagnetic processes at a naturally commutating inverter over a controlled rectifier loaded at a counter-emf time diagrams are represented at the fig. 3.4.2: for a rectification mode (*a*) and a naturally commutating inversion mode (*b*).

A methodic of time diagrams forming is the same as at a rectifying operation mode within the bounds of an assumption of $X_d = \infty$. For an inverted mode two features of time diagrams are typical.

At first, a significantly less interval duration of a reverse voltage applying to a valve:

$$\delta = \beta - \gamma \geq \delta_{r}, \qquad (3.4.2)$$

that has to be greater than a registration certified *recovery time of control properties of not fully controlled valves* (thyristors) δ_r.

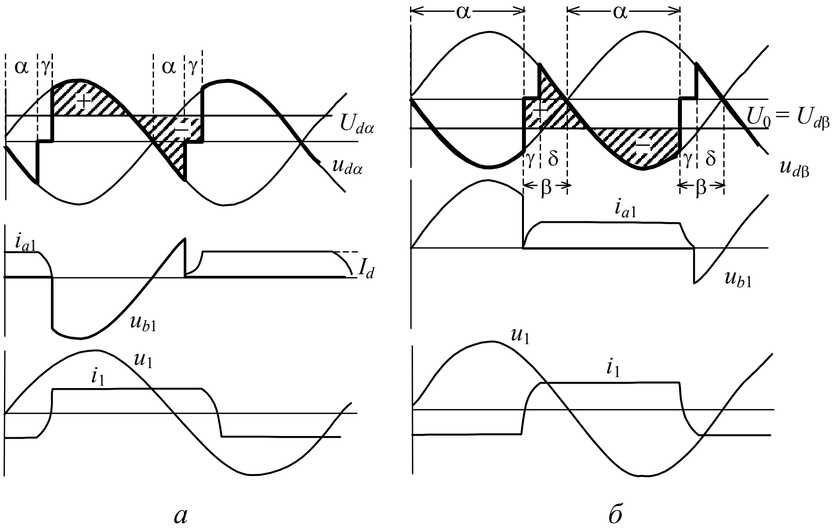


Fig. 3.4.2

This circumstance limits a minimum possible value of a regulation angle β at an inverted mode by a value equal to

$$\beta_{min} = \gamma_{\,max} + \delta_r. \tag{3.4.3}$$

At a rectifier a minimum value of a regulation angle α can be equal to a zero. Hence, a maximum possible active power of a valve converter at a rectified mode always would be greater than a maximum possible active power at an inverted mode.

Secondly, a changing rule of valve anode currents at commutation intervals at an inverted mode according to the equation (3.1.4) at $\alpha > \pi/2$ and $\vartheta > \pi/2$ is the next: at an increase interval a current curve has a prominent nature, at a slope interval it has a concave nature, i.e. a reverse nature relatively a changing at a rectified mode.

As it would be seen from the further inquiry a formal analysis of an inverted mode of a valve converter operation is more appropriate to do in the reverse consequence relatively a rectified mode. At first, we obtain equations of basic inverter curves – an input, regulating and limiting curves then using them we find out features of calculated correlations for valve converter scheme elements.

An input curve. A converter input at a naturally commutating inversion mode is a dc current network that's why there is a significant correlation of an average value of an inverted voltage U_{dx} from an average value of an inverted current I_d at a constant regulating angle β that is called *a naturally commutating inverter input curve*. Formally it's equation is obtained from the external curve equation of a controlled rectifier loaded at a counter-emf (3.1.9) at substituting at it α to β on the equation (3.4.2):

$$U_{d\beta} = U_{d0}\cos\alpha - \Delta U_x = -(U_{d0}\cos\beta + \Delta U_x)\,. \qquad (3.4.4)$$

A minus sign at a voltage $U_{d\beta}$ verifies a voltage polarity changing at a dc current link of an inverter over a rectifier. A sign changing at an average value of a commutating voltage drop ΔU_x is evidence of that input inverter curves rise while a current increases with the same tilt with that rectifier external curves comprise. Diagrams of input curves are represented at the fig. 3.4.3.

In the case if a valve converter operates in turns at a rectified and inverted modes their external and input curves are shown at a joint graphic at the first and fourth quadrants, correspondingly (fig. 3.4.4).

Regulating curve. Regulating curve of the dependent inverter turns out from the regulating curve of the controlled rectifier (2.9.1) replacement α on β on (3.4.1) for mode Id = 0:

$$U_{d\beta} = U_{d0}\cos\alpha = -U_{d0}\cos\beta\,. \qquad \mathbf{(3.4.5)}$$

A minus sign is an evidence of a reverse voltage polarity at a dc current link of a naturally commutating inverter over a controlled rectifier. A diagram of a joint regulating curve for both modes is shown a the fig. 2.9.2.

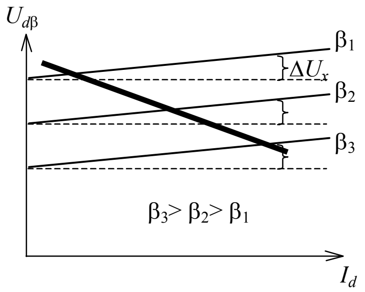


Fig. 3.4.3

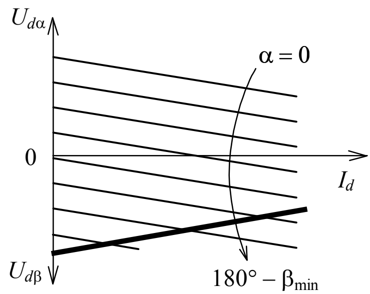


Fig. 3.4.4

A limiting curve. A curve peculiar to only a naturally commutating inverter is called *a limiting curve,* it determines a correlation of a maximum permissible average value of an inverted voltage from a maximum permissible average value of an inverted current. These limitations are conditioned by a presence of a limit to an inverter permissible current $I_{d\,\max}$ determining a maximum permissible commutating angle $\gamma_{\max}$ at a given regulating angle β according to the equation (3.4.2):

$$\gamma_{\max} = \beta - \delta_{\mathrm{r}}\,. \tag{3.4.6}$$

The greater an angle β is the greater a permissible commutating angle is, hence, an inverted current is greater. Drawing values of an inverted current at input curves corresponding to this angle β (i.e. at this determining $U_{d\beta\ \max}$), we can build through obtained points a limiting curve.

A formal equation of a limiting curve becomes evident if to look at the time diagram of an inverted voltage $u_{d\beta}$ at the fig. 3.4.2, *б* from under a time axis in a reverse direction. From positions of a such consideration a curve of $u_{d\beta}$ is similar to a curve of $u_{d\alpha}$ represented at the fig. 3.4.2, *a* if to suppose that a regulating angle α is an angle δ_{r}, a sum of angles $\alpha + \gamma$ is an angle β. An equation of a limiting curve is obtained from an external curve equation of a controlled rectifier at a control angle equal to δ_{r} a which diagram is set to a family of inverter input curves at the fig. 3.4.3 (bold line) and corresponds to the equation

$$U_{d\beta\max} = U_{d0}\cos\delta - \Delta U_x = U_{d0}\cos\delta - \frac{X_{\mathrm{a}}}{\pi} I_{d\max}\,. \tag{3.4.7}$$

An effective range of a naturally commutating inverter is an area under a limiting curve. An area of *an inverter "upset"* is disposed higher than this curve. At a current overload a time led to a recovery of valve control properties is less than a required value and a valve again starts to conduct a current from a time instant of a direct voltage appearing at it, i.e. at a regulating angle equal to α = 0 as it's seen from a time diagram for a reverse voltage (fig. 3.4.2, *b*). A voltage curve polarity of a valve converter $U_{d\beta}$ is "turned over" to a reverse polarity as at uncontrolled rectifier and it's voltage is turned-on not secondly but in accord with an external source voltage. It's shown at equivalent circuits on average values of variables at a dc current link for a normal operation mode of a naturally commutating inverter (fig. 3.4.5, *a*) and it's turn over mode (fig. 3.4.5, *b*).

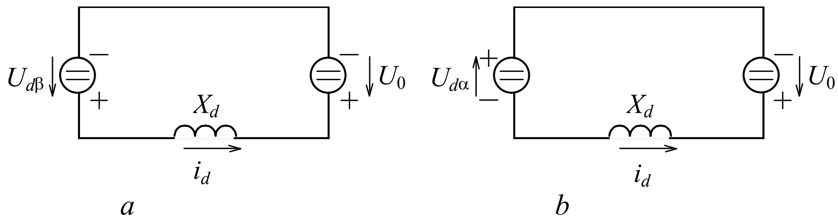


Fig. 3.4.5

There appears a great emergency current limited by only small active loss resistances of scheme elements.

At a finite value of a smoothing reactor inductance L_d at a naturally commutating inverter all it's curves can be obtained from corresponding curves of a controlled rectifier loaded at a counter-emf (section 3.2) at substituting 180 ° – β instead of α and a sign at a counter-emf at a dc voltage link U_0 «-» instead of «+».

Features of calculated correlations for inverter elements. A methodic of a naturally commutating inverter calculation is identical to a methodic of a rectifier calculation with only the feature that a minimum regulating angle β_{min} at an inverted mode can't be equal to a zero while a calculated rectifier mode was made at α = 0.

β_{min} value and a rms value of a transformer secondary voltage U_2 are mutually dependent that's why they are determined by a joint solving of two equations having these variables – the equations (3.4.7) and (3.1.6). From the equation (3.4.7) a value of U_{d0} is determined, hence, a value of U_2 is determined in the next way:

$$U_{d0} = \frac{U_{d\beta_{\min}} + \Delta U_x}{\cos\ \delta_r}, \tag{3.4.8}$$

$$U_2 = \frac{U_{d0}}{K'_U} = \frac{U_{d\beta_{\min}} + \Delta U_x}{K'_U \cos \delta_r}, \tag{3.4.9}$$

where $K'_U = 2\sqrt{2} / \pi$.

From the equation (3.1.6) at substituting 180 ° – β instead of α

$$I_d = \frac{\sqrt{2}\, U_2}{X_a} [\cos \delta_r - \cos \beta] \tag{3.4.10}$$

a value of cos β_{min} is calculated for $I_{d\ max}$ taking into account the equation (3.4.9):

$$\cos\beta_{\min} = \cos\delta_{\text{r}} - \frac{X_{\text{a}} I_{d\max}}{\sqrt{2}\,U_2}. \qquad (3.4.11)$$

As a result of the first made step at a fifteen-step procedure (that isn't repeated here) of a valve converter calculation at a rectified (section 2.3) and at an inverted mode a value of U_2 is determined, hence, a value of $K_{\text{т}}$ is determined that are necessary for the further steps of the calculation. At this obtained correlations for variables $U^*_{b\max}$, K_{I}, S^*_2, S^*_1, $S^*_{\text{т}}$, S^*_b, $S^*_{\text{т}.L}$, χ, $K_{\text{c.c.}}$ will depend on a recovery time of a valve control properties δ_{r} and a reduced transformer leakage inductance L_{a} or a value of a transformer short circuit voltage $U_{\text{к}}$ % that is determined by the inductance and is more habitual parameter for power engineering specialists (3.1.15). At small values of δ_{r} and $U_{\text{к}}$ % guaranteeing a smallness of $\beta_{\min}$ all the pointed calculated indexes for a naturally commutating inverter will be close to their corresponding values for a rectifier. That's why recommendations on application range of valve converter basic cells at a rectified mode operation will be correct for an inverted mode.

3.4.2. A THREE-PHASE NATURALLY COMMUTATING INVERTER ($m_1 = 3$, $m_2 = 3$, $q = 2$)

A scheme of a three-phase bridge naturally commutating inverter taking into account L_{a} parameter of a real transformer is shown at the fig. 3.4.6, the time diagrams at an assumption of $X_d = \infty$ are represented at the fig. 3.4.7.

At a small values of regulating angles β at an inverted mode angles of valve operating starting α are close to 180º. Taking into account this fact all the time diagrams are built on the same methodic as for a controlled rectifier (section 3.2).

To obtain calculated correlations for the given naturally commutating inverter on a standard fifteen-step procedure at the first step of calculation U_2 is determined again on the equation (3.4.9) at $K'_v = 3\sqrt{6}/\pi$ and $\beta_{\min}$ from the equation for an inverted current I_d.

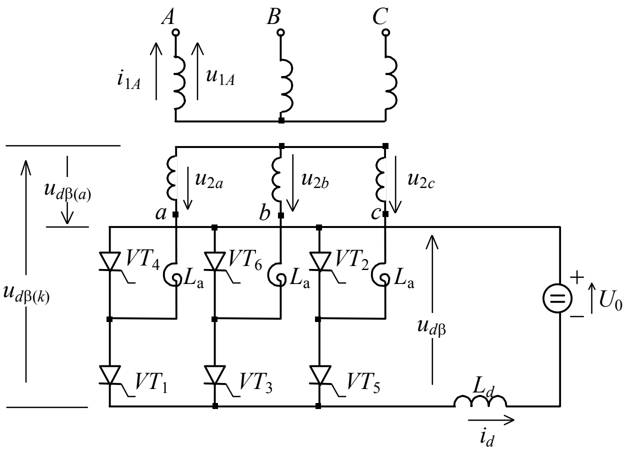


Fig. 3.4.6

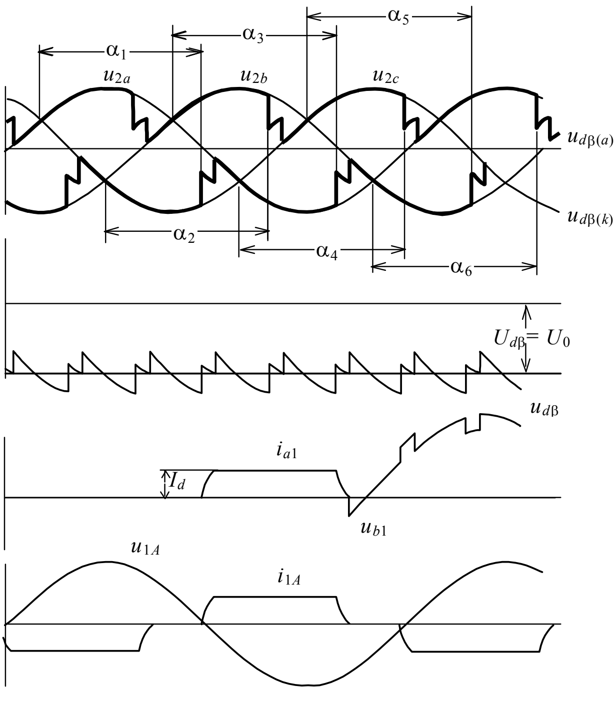


Fig.3.4.7

For a three-phase bridge scheme an equation for $I_{d\ max}$ current has the next view:

$$I_{d\max} = \frac{\sqrt{2}\sqrt{3}U_2}{2X_{\text{a}}}\sin\frac{\pi}{3}\left[\cos\delta_{\text{r}} - \cos\beta_{\min}\right], \tag{3.4.12}$$

then

$$\cos\beta_{\min} = \cos\delta_{\text{r}} - \frac{2X_{\text{a}}I_{d\max}}{\sqrt{2}\sqrt{3}\ U_2}. \tag{3.4.13}$$

Taking into account observations made at the end of the section 3.4.1 and properties of a three-phase bridge rectification scheme marked at the section 2.7 we can make a conclusion that the given scheme is the best from all the basic schemes of valve converters for a naturally commutating inverter mode.

3.5.* A COMMON CORRELATION OF A RECTIFIER PRIMARY CURRENT FROM AN ANODE AND RECTIFIED CURRENT (CHERNYSHEV LAW)

At electromagnetic processes analysis at rectifier basic cells time diagrams of cell primary currents according to the considered methodic were made in the last turn after making all the diagrams. As a rectification scheme analysis is time-consuming enough it's necessary to find a common methodic of a primary current forming at any rectification scheme without it's detail analysis procedure. Such methodic can be formulated based on *Chernyshev law* for rectifier primary currents obtaining which equation is an aim of the analysis.

A fragment of a three-phase rectifier to obtain a Chernyshev law is shown at the fig. 3.5.1. A valve anode voltage that has to be rectified in the common case is summed up geometrically from secondary winding voltages placed at all transformer rods. A vector diagram of a resulting voltage at anodes is represented at the fig. 3.5.2.

A rotation angle of a resulting vector relatively a primary winding voltage of a transformer A phase is denoted as

$$\sigma = \sigma_0 + (n-1)\frac{2\pi}{m_2}, \tag{3.5.1}$$

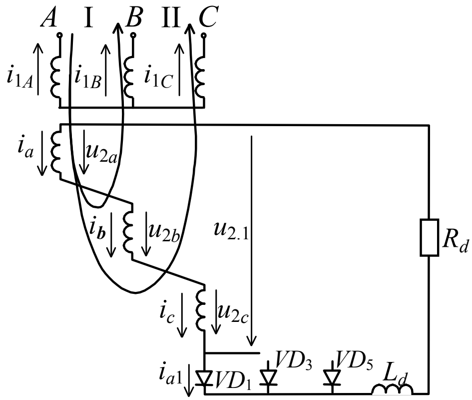


Fig. 3.5.1

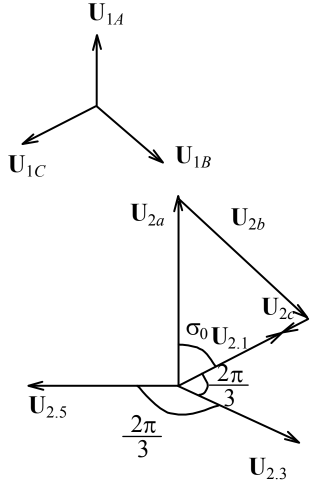


Fig. 3.5.2

where σ_0 – a vector rotation angle of the first valve; n – numbers of conducting valves ($n = 1, 2, ..., m_2$).

To find out three variable currents at transformer primary windings i_{1A}, i_{1B}, i_{1C} three equations are necessary. Two equations are obtained according to the second Kirchgoff rule for magnetic networks at detour of contours formed by transformer rods A and B, A and C at the direction pointed at the scheme at the fig. 3.5.1. The third equation is obtained on the first Kirchgoff rule, as a result we have the next equations

$$i_{1A} + i_{1B} = i_{\text{a}1}\frac{w_{\text{a}}}{w_1} - i_{\text{a}1}\frac{w_b}{w_1},$$

$$i_{1A} + i_{1C} = i_{\text{a}1}\frac{w_{\text{a}}}{w_1} - i_{\text{a}1}\frac{w_c}{w_1}, \qquad (3.5.2)$$

$$i_{1A} + i_{1B} + i_{1C} = 0.$$

A solution for a primary current on a Kramer rule has the next view

$$i_{1A} = \frac{2}{3}i_{\text{a}1}\left(\frac{1}{K_{\text{т}a}} - 0,5\frac{1}{K_{\text{т}b}} - 0,5\frac{1}{K_{\text{т}c}}\right) =$$

$$= \frac{2}{3}\,\frac{i_{\text{a}1}}{U_1}\left(U_{2a} + U_{2b}\cos 120 + U_{2c}\cos 240\right).$$

The equation in brackets at the right part of the equation is a sum of projections of secondary voltage vectors at a phase *A* vector direction and can be substituted by a projection of a resulting voltage vector $\mathbf{U}_{2.1}$ at the same direction:

$$i_{1A} = \frac{2}{3} i_{a1} \frac{\mathbf{U}_{2.1}}{U_1} \cos\sigma_0 = \frac{2}{3} \frac{i_{a1}}{K_{\text{т.рез}}} \cos\sigma_0 . \qquad (3.5.3)$$

An obtained correlation characterizes a contribution to a primary current of *A* phase from an anode current of the first conducting valve. Let's summarize the equation (3.5.3) for conducting intervals of remaining valves. A common correlation of a primary current from valve anode currents at a discrete and continuous form obtains the next view:

$$i_1 = \frac{2}{3K_{\text{т}}} \sum_{h=1}^{m_2} i_{ah} \cos\left[\sigma_0 + (h-1)\frac{2\pi}{m_2}\right] = \frac{2}{3K_{\text{т}}} i_a \cos\sigma , \qquad (3.5.4)$$

where σ – a distortion angle between a voltage vector of a primary winding at which the current is and an anode voltage resulting vector of the conducting valve having a current i_a.

The equation (3.5.4) is *a Chernyshev law* for rectifier primary currents that is valid at the next conditions:

1. Transformer primary windings are connected into a star. If they a re connected into a triangle then the rule allows to find out a linear current, a transformation factor $K_{\text{т}}$ and σ_0 angle are determined relatively voltages of primary voltages equivalent star into which primary voltages triangle is transferred.
2. Transformer secondary windings can be connected on any scheme.
3. A rectification scheme is half-wave. For full-wave rectification schemes there are necessary a preliminary reduction of them to an equivalent totality of half-wave rectification schemes as it was shown at a three-phase bridge rectification scheme example at the section 2.8 and the next application of Chernyshev rule and syncopation to them.

Then a methodic of a diagram construction of a rectifier primary current can be formulated straight away on the scheme omitting a construction of all remaining time diagrams and to illustrate it at a three-phase rectifier example at that transformer windings are connected on a triangle-star with a null outlet scheme at the assumption that $X_d = \infty$, $X_a = 0$ (section 2.5).

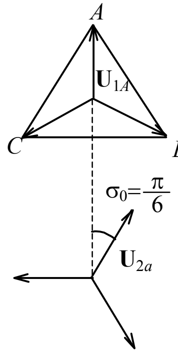


Fig. 3.5.3

1. A triangle of primary voltages is transformed into an equivalent star and a transformation factor $K_т$ and angle σ_0 equal to $\pi/6$ (fig. 3.5.3) are determined for it.

2. An auxiliary *construction cosinusoid* having an amplitude equal to $(2/3)(I_d/K_т)$ is drawn on the diagram since $i_a = I_d$ at an interval of any valve conducting.

3. At a construction cosinusoid points corresponding to rotation angles of valve anode voltage vectors equal to $\pi/6$, $(\pi/6) + 2(\pi/3)$, $(\pi/6) + 2(2\pi/3)$ and so on are noted. These points corresponding to middles of valve conducting sections determine heights of steps that duration is equal to $2\pi/m_2$ (in our case – $2\pi/3$) by a primary current curve that is obtained at connecting step ends by vertical lines as it's shown a the fig. 3.5.4. A linear current (mains current) waveform is such for the given scheme that can be made at time diagrams (fig. 2.5.2) as a geometrical sum of primary currents of two windings.

In the case of a controlled rectifier at $X_d = \infty$ a primary current waveform doesn't depend on an angle α but is displaced only on a phase at an angle α. Commutating angles at curves of valve anode currents become apparent also at a primary current curve. According to a construction cosinusoid pulsations of a rectified (anode) current in the case of a finite value of a smoothing reactor inductance ($X_d \neq \infty$) become apparent at a primary current curve.

A Chernyshev rule representing a common analytical expression for a primary current curve of a rectifier allows to determine a rms value of a primary current in the common case of half-wave rectification schemes at $X_d = \infty$.

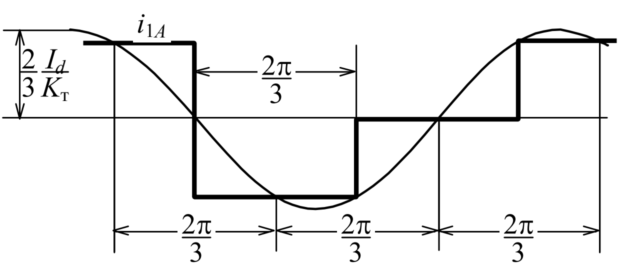


Fig. 3.5.4

Based on a rectangular-step nature of a primary current curve we can write the next expression

$$I_1 = \sqrt{\frac{1}{2\pi} \sum_{h=1}^{m_2} \int_0^{2\pi/m_2} A_m^2 \cos^2\left[\theta_0 + (h-1)\frac{2\pi}{m_2}\right] d\sigma} = \frac{A_m}{\sqrt{2}}, \qquad (3.5.5)$$

where $A_m = (2/3)\,(I_d / K_т)$ is an amplitude of a construction cosinusoid, i.e. a rms value of a construction cosinusoid determines a rms value of a rectifier transformer primary current.

Thus, Chernyshev rule allows to find out primary currents curves of known and novel rectification schemes and a rms value of this current without a procedure of electromagnetic processes detail analysis at schemes. Besides, this rule can be effectively applied to find out rectifier input current spectrums in the common view as it's shown below.

3.6. PRIMARY CURRENT SPECTRUMS OF RECTIFIER TRANBSFORMERS AND NATURALLY COMMUTATING INVERTERS

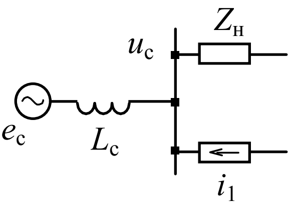


Fig. 3.6.1

Valve converters drawing a nonsinusoidal current a from mains have a significant reverse negative influence on a mains. A degree of an influence depends on *a primary current spectrum* of a rectifier a determining of which is an aim of this section. At the fig. 3.6.1 there is shown a one-linear equivalent scheme of a mains containing a emf e_m supply (a synchronous generator) having an inherent inductance L_m, a mains load as a linear complex resistance Z_{load} and a nonsinusoidal current I_1 supply that is an equivalent of a valve converter on it's input. A nonlinear load presence at a mains as a valve converter makes the next problems:

1. A mains voltage u_m waveform is distorted since

$$u_m = e_m - X_m \frac{d}{d\vartheta}\left(i_{load} + i_1\right), \qquad (3.6.1)$$

that at a current i_1 nonsinusoidality will lead to a voltage u_m nonsinusoidality. A nonsinusoidal mains voltage will influence negatively on “good” (linear) electric energy consumers that equivalents is resistance Z_{load}.

2. Additional loss of an active power appear at mains elements from current higher harmonics that can cause these elements (transformers, cosinus capacitors, electric motors) overheat.

3. Mains over voltages appear because of resonant events at a coincidence of primary current harmonic frequencies of a valve converter having proper resonant frequencies of a mains that is really a system having distributed *LC*-parameters. These over voltages can cause false protective disconnection or a death of mains elements.

In the common pointed problems (and many others) are problems of an electromagnetic compatibility at electric mains. *An electromagnetic compatibility of electric-technical devices* that are connected by a common mains is called as their capability to normally functioning at real exploitation conditions at a presence of unpremeditated disturbances at a mains and at this not to make impossible electromagnetic disturbances at a mains for other devices [20] (in detail look a chapter 6).

To calculate indexes characterizing a degree of a reverse influence of a valve converter to a mains by a spectral method it’s necessary to know a spectral composition of valve converter input currents. In order not to do a calculation of input current spectrums of all the basic cells many times it’s necessary to make a spectral analysis in the common view. It can be made by using a connection rule of a primary current and anode currents (Chernyshev rule).

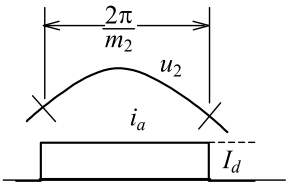


Fig. 3.6.2

At the fig. 3.6.2 an anode current waveform of a valve is shown at an assumption that $X_d = \infty$, $X_a = 0$.

An amplitude of n-order harmonic of a Fourier row of this curve is equal to

$$I_{\text{a.max}(n)} = \frac{2 \cdot 2}{2\pi} \int_0^{\pi/m_2} I_d \cos n\vartheta \, d\vartheta = \frac{2}{\pi}\frac{1}{n}\sin\frac{\pi}{m_2} n I_d\,, \qquad (3.6.2)$$

an expression for a harmonic instantaneous value

$$i_{\text{a}(n)} = \frac{2}{\pi}\,\frac{\sin\dfrac{\pi}{m_2}}{n} I_d \cos n\vartheta = I_{\text{a}\max(n)} \cos n\vartheta\,, \qquad (3.6.3)$$

where n is a number of a harmonic relatively a mains frequency.

Then an instantaneous value of n-order harmonic of a valve converter primary current according to a discrete formula of the equation (3.5.4) of Chernyshev rule valid for single harmonics (because of a connection linearity of i_1 and $i_{1(n)}$) for h-order input phase is equal to the next

$$i_{1(n)} = \frac{2}{3K_{\text{т}}} \sum_{h=1}^{m_2} I_{\text{a}\max(n)} \cos n\left[\vartheta - (h-1)\frac{2\pi}{m_2}\right] \cos\left[\sigma_0 + (h-1)\frac{2\pi}{m_2}\right] =$$

$$= \frac{1}{3K_{\text{т}}} \sum_{h=1}^{m_2} I_{\text{a}\max(n)} \left\{ \cos\left[\sigma_0 + n\vartheta - (n-1)\frac{2\pi}{m_2}h\right] + \right.$$

$$\left. + \cos\left[n\vartheta - \sigma_0 + \frac{2\pi}{m_2} - h\frac{2\pi}{m_2}(n+1)\right]\right\}. \qquad (3.6.4)$$

Sums of cosinusoids of the same frequency but of the different phase shifts of m_2-phase vectors star represented at the fig. 3.6.3, *a*, are different from a zero only in that case if all the vectors of m_2-phase system are in phase as it's shown at the fig. 3.6.3, *b*.

It's possible at the next conditions (for a distinctness $h = 1$)

$(n - 1)(2\pi/m_2) = 2\pi k,$

$(n + 1)(2\pi/m_2) = 2\pi k.$

As a result for numbers of n harmonics at the primary current have the next expression

$$n = km_2 \pm 1, \qquad (3.6.5)$$

where k are integer positive numbers ($k = 2, 3, 4$ ...).

A junction from half-wave rectification schemes ($q = 1$) to full-wave rectification schemes ($q = 2$) as it's shown above leads to a commutation frequency doubling at a converter, hence, to decreasing of time intervals in two times between commutation moments. An increasing of a number of transformer secondary phases has the same effect.

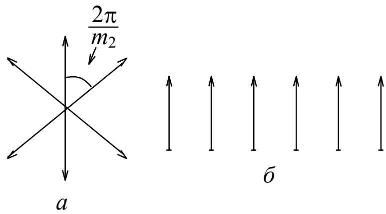


Fig. 3.6.3

Hence, the correlation (3.6.5) can be also generalized taking into account a scheme half-wave then it will have the next view ($p = qm_2$ is a pulseplicity)

$$n = kqm_2 \pm 1 = kp \pm 1. \tag{3.6.6}$$

At the primary current of any rectifier only harmonics which order is determined on (3.6.6) present. For example, for a three-phase bridge scheme and a scheme having an equalizing reactor at which $p = 6$ at the primary current besides the fundamental harmonics of 5, 7, 11, 13, 17, 19 ... order will present.

From the correlations (3.6.2) and (3.6.4) follows that a relative value of higher harmonics at the fundamental parts will be equal to

$$\frac{I_{1(n)}}{I_{1(1)}} = \frac{\sin n\dfrac{\pi}{m_2}}{n\sin\dfrac{\pi}{m_2}} = \frac{\sin\left(kqm_2 \pm 1\right)\dfrac{\pi}{m_2}}{n\sin\dfrac{\pi}{m_2}} = \frac{\cos kq\pi}{n} = \pm\frac{1}{n}. \tag{3.6.7}$$

The sign is an evidence of a synchronism or antiphase of the corresponding harmonic relatively a fundamental. With a number of harmonics increasing their relative value decreases droningly.

Based on a spectral analysis of transformer primary currents of rectifiers and naturally commutating inverters the next **conclusions** can be made:

1. With an increasing of electric energy conversion schemes pulseplicity frequencies of primary current higher harmonics increase and their relative value decreases, i.e. a current waveform is close to a sinusoidal, at a limit at an infinitely large pulseplicity a primary current becomes sinusoidal;

2. A regulation angle $\alpha(\beta)$ leading at the assumption that $X_d = \infty$ doesn't change harmonic frequencies, their relative values leading only to their phase shifting for n-order harmonics, correspondingly, at an angle $\varphi_{(n)}=n\alpha$ on a deviation to a mains voltage curve. It causes a consuming increasing of a reactive power proportional to sin $\varphi_{(1)}$ from a mains, hence, a decreasing (worsening) of a power factor as it will be shown a t the section 3.10;

3. A commutation angle γ taking into account eliminating primary current curve steps doesn't change a number of harmonic in it but decreases relative values of higher harmonics because a primary current waveform improvement. For an accurate taking into account of a commutation influence to a current because of a complicated

(nonlinear) current changing nature at commutation intervals (section 3.1) bulky formulas for current harmonics are necessary [8]. A current approximation at commutation intervals by a linear dependence (line) allows to obtain more simple formulas for it's harmonics having an error equal to some percents [42, 43].

3.7. SPECTRUMS OF A RECTIFIED AND INVERTED VOLTAGE OF A VALVE CONVERTER

Determining of *a rectified voltage spectrum* that is an aim of this section is necessary to calculate a rectifier output filter and estimate quality indexes of a rectified voltage through which a damage from a distorted quality of a converted energy is determined at a load.

At a rectified voltage curve shown at the fig. 3.7.1 sinus and cosinus components of a Fourier row present.

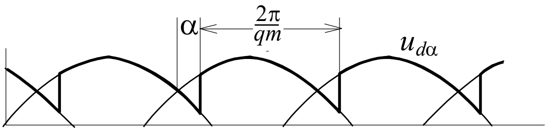


Fig. 3.7.1

Then placing a reference frame at a rectified voltage maximum u_2' determined as

$$u_2' = \begin{cases} \sqrt{2}U_2 \cos\vartheta & \text{при} \quad q=1, \\ \sqrt{2}\sqrt{3}\,U_2 \cos\vartheta & \text{при} \quad q=2, \end{cases}$$

for a rms value of n-order harmonic of a Fourier row sinus part of a rectified voltage we obtain the next expression

$$U^S_{d\alpha(n)} = \frac{2}{\frac{2\pi}{p}\sqrt{2}} \int\limits_{-\frac{\pi}{p}+\alpha}^{\frac{\pi}{p}+\alpha} \sqrt{2}U_2' \cos\vartheta \sin n\vartheta\, d\vartheta =$$

$$= U_2' \frac{p}{\pi} \frac{2kp}{p^2-1} \sin\frac{\pi}{p} \sin\alpha = U_{d0} \frac{2kp}{(kp)^2-1} \sin\alpha. \qquad (3.7.1)$$

At a rectified voltage curve there present harmonics of an order

$$n = kqm_2 = kp\,, \qquad (3.7.2)$$

where n – a number of harmonic at a rectified voltage relatively a mains voltage period; k – a number of harmonic at a rectified voltage relatively it's period.

Analogically a rms value of n-order harmonic of Fourier row cosinus part is equal to

$$U^{C}_{d\alpha(n)} = \frac{2}{\frac{2\pi}{p}\sqrt{2}} \int\limits_{-\frac{\pi}{p}+\alpha}^{\frac{\pi}{p}+\alpha} \sqrt{2}\; U_2' \cos\vartheta \cos n\vartheta \; d\vartheta = -U_{d0}\frac{2}{(kp)^2-1}\cos\alpha\,. \qquad (3.7.3)$$

A rms value of n-order resulting harmonic at a rectified voltage curve is equal to ($n = kp$)

$$U_{d\alpha(n)} = \sqrt{(U^{S}_{dn})^2 + (U^{C}_{dn})^2} = \frac{2}{(kp)^2-1}\sqrt{1+(kp\,\mathrm{tg}\,\alpha)^2}\,U_{d0}\cos\alpha\,. \qquad (3.7.4)$$

Graphics of quantitative dependencies of relative values of rectified voltage harmonics $U_{d\alpha(n)}/U_{d0}$ from a regulation angle α are represented at the fig. 3.7.2.

A value of qm_2 order harmonics at the corresponding rectification scheme determines such dc voltage quality index as a pulsation's factor Kp by (1.2.12) since at $n = p = qm_2$

$$U^{*}_{d\alpha(p)} = \frac{U_{d\alpha(p)}}{U_{d0}} = K_p C_r\,. \qquad (3.7.5)$$

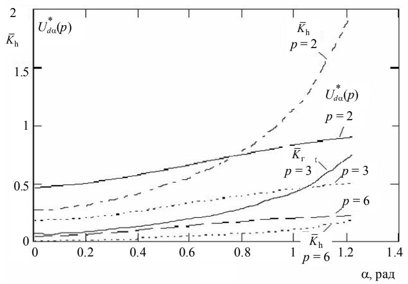


Fig. 3.7.2

At the fig. 3.7.2 there are also shown dependencies of another quality index of a rectified voltage which is called an integral harmonic factor $\bar{K}_h$ as it was shown at the chapter 2 appropriate to calculate a smoothing reactor inductance. It can be said that in both cases to obtain a rectified current of the same quality (on a harmonic factor value) a correlation of smoothing reactor inductance's at rectifiers having $qm_2 = 2, 3, 6$ has to be approximately equal to the correlation 36 : 9 : 1 at $\alpha = 0$.

A voltage spectrum at a valve converter dc-current link operating at a naturally commutating inverter mode is formally obtained on the same expressions (3.7.1) – (3.7.4) at substituting $180\,° - \beta$ instead of α at it.

A physical identity of a voltage spectral distribution of a rectifier and naturally commutating inverter at $\alpha = \beta$ is conditioned by an identity of these voltages time diagrams from observers positions considering them over a time axes at a time axes direction and under it at a direction opposite to a time axes direction, correspondingly.

A commutation effect at a valve converter taking into account real transformer parameters leads to an additional waveform distortion of a rectified and inverted voltage taking into account because of a commutation angle γ as it was shown at the sections 3.1, 3.4, 3.5. But since at this pulsation's period of a rectified voltage doesn't change numbers of harmonics at a rectified voltage don't change. The numbers of harmonics as earlier are determined on (3.7.2) only their relative components are changed. Formulas to calculate a voltage harmonic value at this mode that are more complicated than formulas (3.7.4) are represented at [8].

Thus, a rectified voltage quality increases while it's pulseplicity increases approaching to dc voltage without pulsation's at a tendency to an infinity of an equivalent number of transformer secondary phases. At his an ideal sinusoid is a transformer primary current.

3.8. AN OPTIMIZATION OF A NUMBER OF RECTIFIER TRANSFORMER SECONDARY PHASES. EQUIVALENT MULTI-PHASE RECTIFICATION SCHEMES

At rectifiers feeding from an industrial mains of a general use a number of transformer primary phases is given and equal to one or three. At the same time a number of transformer secondary phases as it was seen at considering rectifier basic cells can be greater than a number of primary phases. If it's necessary it can be any including a value not multiple to three that is achieved by a combination of transformer secondary winding voltages similar to a connection to a zigzag. That's why a question about an optimal number of transformer secondary voltage phases from positions of other criteria than at two previous sections namely, first of all, from a position of a full power relative value of transformer secondary windings S_2^*. As an analysis aim we determine an establishment of a common correlation of S_2^* from a number of transformer secondary phases, firstly, for half-wave rectification schemes having q = 1 then for full-wave schemes having q = 2.

At half-wave rectification schemes currents at the transformer secondary windings repeat valve anode currents connected with these windings. At a rectifier model with the assumptions that $X_d = \infty$, $X_a = 0$ (we will take into account a commutation in the next section) a rms value of the secondary current that duration is equal to $\lambda = 2\pi/m_2$ will be the next

$$I_2 = \frac{I_d}{\sqrt{m_2}} . \tag{3.8.1}$$

Then a full power of the transformer secondary windings at relative units taking into account (2.9.1) is equal to

$$S_2^* = \frac{S_2}{P_{d0}} = \frac{m_2 U_2 I_2}{P_{d0}} = \frac{m_2}{P_{d0}} \frac{U_{d0}}{(m_2/\pi)\sin(\pi/m_2)\sqrt{2}} \frac{I_d}{\sqrt{m_2}} =$$

$$= \frac{\pi}{\sqrt{2}\sin\pi/m_2\sqrt{m_2}} . \tag{3.8.2}$$

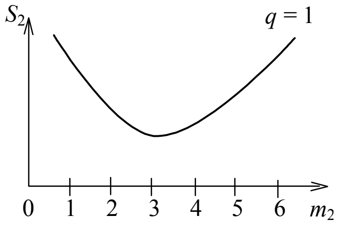


Fig. 3.8.1

This correlation graphic is shown as the fig. 3.8.1. Conditionally we suppose that m_2 is a continuous variable and mark out the points corresponding to real values of m_2 = 2, 3, 6, 12 at an obtained correlation. It's seen that a continuous curve extreme is close to m_2 = 3, this value is optimal value of a number of secondary phases for half-wave . rectifiers.

In the case of full-wave rectifiers two valves, one of them is from a cathode group and another is from an anode group, are connected to each transformer secondary winding. Then a rms value of a transformer secondary current is determined taking into account the fact that it is formed by two impulses of anode currents of a duration equal to $2\pi/m_2$ each of them (fig. 3.8.2):

$$I_2 = \frac{I_d\sqrt{2}}{\sqrt{m_2}} . \tag{3.8.3}$$

A full power of a one secondary winding

$$S_2' = U_2 I_2 = U_2 I_d \sqrt{\frac{2}{m_2}} . \tag{3.8.4}$$

An active power transferred by a secondary winding to a rectified current network is determined by an interaction of a secondary fundamental current with a secondary voltage sinusoid. A rms value of a fundamental Fourier row for a current curve shown at the fig. 3.8.2,

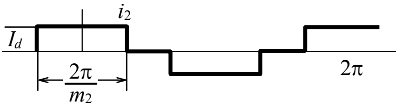


Fig. 3.8.2

is determined the next way at a time reference choosing at a current impulse middle:

$$I_{2(1)} = \frac{4 \cdot 2}{2\pi \sqrt{2}} \int_0^{\pi/m_2} I_d \cos\vartheta \; d\vartheta = \frac{2\sqrt{2}\; I_d}{\pi} \sin\frac{\pi}{m_2} . \tag{3.8.5}$$

An equation for an active power of a secondary winding will have the next view

$$P_2' = U_2 I_{2(1)} = U_2 I_d \frac{2\sqrt{2}}{\pi} \sin\frac{\pi}{m_2} . \tag{3.8.6}$$

Taking into account the expressions (3.8.4) and (3.8.6) we obtain a full power of transformer secondary windings at relative units as the follow

$$S_2^* = \frac{S_2'}{P_2'} = \frac{\sqrt{2}\,\pi U_2 I_d}{2 U_2 I_d \sqrt{m_2}\;\sqrt{2} \sin\frac{\pi}{m_2}} = \frac{\pi}{2\sqrt{m_2}\sin\frac{\pi}{m_2}} . \tag{3.8.7}$$

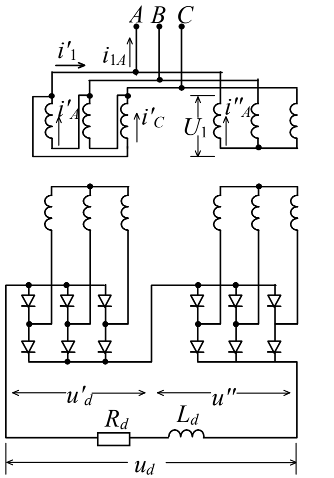
This correlation of a quality index from m_2 value to a multiplier $\sqrt{2}$ repeats the correlation (3.8.2).

Thus, optimal numbers of transformer secondary phases on a criteria S_2^* are the same for a full-wave and half-wave rectification scheme. At the same time an optimal number of transformer secondary phases tends to an infinity on quality criteria of a rectified voltage and rectifier primary current. Then it's evident that it's necessary to construct power rectifiers on such schemes at that at transformer secondary windings currents with it's optimal duration equal to 2π/3 flow as at three-phase schemes. These schemes would be analogical to multiphase rectification schemes on a number of a rectification pulseplicity. Such schemes are called as *equivalent multiphase rectification schemes*.

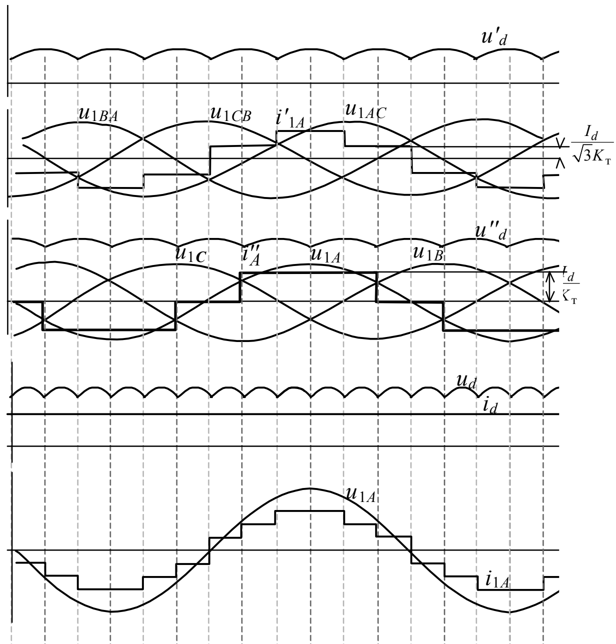
A scheme of an equivalent twelve-phase rectification based on two three-phase bridge schemes shown at the fig. 3.8.3, a became widespread for power high-voltage rectification schemes (first of all, at systems of an energy transfer by a dc current).

The two three-phase bridge rectification schemes are connected in parallel on an input, in series – on an output. To obtain 30° shift between six fold pulsation of rectified voltages of an each bridge primary (or secondary) transformer windings of a one bridge are connected into a triangle and feed from nonlinear mains voltages shifted at a required angle relatively phase voltages. Time diagrams of scheme voltages and currents shown at the fig. 3.8.3, *b*, are made at the assumptions that have rectifier basic cells considered at the chapter 2 ($X_d = \infty$, $X_a = 0$).

At the first diagram there are presented a three-phase system of a transformer primary voltages of the left bridge, this bridge rectified voltage curve u'_d and a transformer phase linear current curve i'_{1A}. At the second diagram the same is made for the right bridge. At the third diagram a resulting rectified voltage curve as a sum of rectified voltages u'_d and u''_d of separate bridges is formed. It's seen that a rectified voltage pulsations period u_d is equal to 30°, i.e. pulsations are twelve-fold relatively a mains voltage frequency. At the fourth diagram a resulting mains current curve is formed as an algebraic sum of primary currents i'_{1A}, i'_{1C} that form a linear current of the left transformer i'_{1A} and a current i''_{1A} of the right transformer


A B C
i′1
i1A
i′A
i′C
U1
i″A
u′d
u″
Rd
Ld
ud

a


u′d
u1BA
u1CB
i′1A
u1AC
Id/√3Kт
u″d
u1C
i″A
u1A
u1B
ud
id
u1A
i1A

b

Fig. 3.8.3

At this we took into account that a transformation factor of the left transformer $K'_\text{т}$ is greater than a transformation factor of the right transformer $K_\text{т}$ in $\sqrt{3}$ times, since

$$K'_\text{т} = \frac{\sqrt{3}\, U_1}{U_2} = \sqrt{3} K''_\text{т} = \sqrt{3} K_\text{т}\,.$$

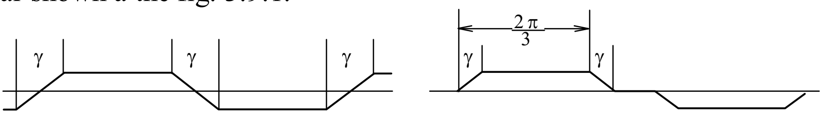
A resulting current curve at an input of an equivalent twelve-phase converter contains twelve steps per a period that according to Chernyshev rule (3.5.3) confirms that a rectification is twelve–phase.

Thus, jointing some three-phase rectification schemes having an optimal current duration equal to λ = 2π/3 at transformer secondary windings and combining schemes of primary and secondary windings connection to obtain an equivalent multiphase system of rectified voltages we can obtain equivalent 24-, 48- and even 96-phase rectifiers.

3.9.* A COMMUTATION INFLUENCE AT RMS VALUES OF TRANSFORMER CURRENTS AND IT'S TYPICAL POWER

A presence of transformer windings leakage inductance leads to an elimination of current steps at it and commutation intervals appearance with a fluent current changing at it as it was shown at the section 3.1. A current curve is filtered on higher harmonics that leads to it's harmonic content improvement (section 3.6). It causes a changing of rms values of currents, hence, and winding full powers which determining is an aim of this section.

To obtain at a simple form corrections from a commutation at formulas to calculate transformer currents rms values we substitute a linear nature of a current changing at a commutation interval γ on (3.1.4) instead of a nonlinear nature of changing. Curves of primary (and secondary at $K_\text{т} = 1$) currents of the considered basic schemes of a full-wave rectification of a one-phase (*a*) and a three-phase (*b*) current ar shown a the fig. 3.9.1.



a *b*

Fig. 3.9.1

For these current waveforms it's not difficult to show that their rms values taking into account a commutation $I_{1\gamma}$ are equal to:

for two-pulse schemes

$$I_{1\gamma}=\frac{I_d}{K_{\text{т}}}\sqrt{1-\frac{2}{3}\frac{\gamma}{\pi}}=I_1\sqrt{1-\frac{2}{3}\frac{\gamma}{\pi}}, \qquad (3.9.1)$$

for six-pulse rectification schemes

$$I_{1\gamma}=\frac{I_d}{K_{\text{т}}}\sqrt{1-\frac{\gamma}{2\pi}}\sqrt{\frac{2}{3}}=I_1\sqrt{1-\frac{\gamma}{2\pi}}. \qquad (3.9.2)$$

then a typical transformer power taking into account a commutation is the next for a full-wave one-phase rectifier

$$S_{\text{т}\gamma}=U_1 I_{1\gamma}=U_1 I_1\sqrt{1-\frac{2}{3}\frac{\gamma}{\pi}}=S_2\sqrt{1-\frac{2}{3}\frac{\gamma}{\pi}} \qquad (3.9.3)$$

and for a full-wave three-phase rectifier is the next

$$S_{\text{т}\gamma}=3U_1 I_{1\gamma}=3U_1 I_{1\gamma}\sqrt{1-\frac{\gamma}{2\pi}}=S_1\sqrt{1-\frac{\gamma}{2\pi}}=S_1\sqrt{1-\Psi(\gamma,m_2)}. \qquad (3.9.4)$$

The last equation is a summarizing for full-wave rectification schemes where a view of a function $\psi(\gamma)$ determining a value of a correction multiplier at S_1 depends on a number of rectifier phases.

Thus, a commutation improving a current waveform at transformer windings decreases it's typical power. This conclusion is fair for half-wave rectification schemes. Obviously, a commutation will change an input power factor of a rectifier (the next section).

A linear approximation of an input rectifier current at commutation intervals allows to simplify a calculation of a differential harmonic factor of this current that as it will be shown at the section 3.13 determines a degree of a reverse influence of a rectifier to a mains voltage distortion. According to the equation (1.5.25) a current differential harmonic factor can be calculated through a rms value of the current I' first derivative :

$$K'_{\text{h}}=\sqrt{\left(\frac{I'}{\omega I_{(1)}}\right)^2-1}. \qquad (3.9.5)$$

A value of I' is determined without any difficulties for current curves shown at the fig. 3.9.1. So for the current shown as a curve at the fig. 3.9.1, *b*, we have the next expression

$$\tilde{I} = \frac{I_d}{K_{\text{т}}\gamma}\,\frac{1}{\sqrt{\pi/\gamma}} = \frac{I_d}{K_{\text{т}}\,\sqrt{\gamma\pi}}\,. \tag{3.9.6}$$

A rms value of a fundamental current shown at the fig. 3.9.1, *b*, is determined exactly on [42], approximately it's determined on the next formula neglecting by γ influence

$$I_{(1)} = \frac{\sqrt{6}}{\pi}\frac{I_d}{K_{\text{т}}}\,. \tag{3.9.7}$$

Then

$$\tilde{K}_{\mathrm{h}} \cong \sqrt{\frac{\pi}{\gamma} - 1}\,. \tag{3.9.8}$$

This index as it will be shown at the section 3.13 integrally determines a degree of a converter reverse influence at a mains.

3.10. A PERFORMANCE INDEX AND A POWER FACTOR OF A POWER CONVERTER AT RECTIFICATION AND NATURALLY COMMUTATING INVERSION MODES

An aim of this section is studying of a correlation between two basic energetic factors of a valve converter, it's scheme parameters and a mode.

3.10.1. A PERFORMANCE INDEX

A performance index (PI) is determined by a correlation of an active power at a converter output to an a active power at it's input. Conformably to a rectification operation mode of a valve converter it means the next

$$\eta = \frac{P_d}{P_1} = \frac{P_d}{P_d + \Delta P}, \tag{3.10.1}$$

for a naturally commutating invertor mode –

$$\eta = \frac{P_1}{P_d} = \frac{P_d - \Delta P}{P_d}\,. \tag{3.10.2}$$

ΔP – active power loss inside a valve converter that are summed from loss at a transformer $\Delta P_{\text{т}}$, valves ΔP_{v}, filter ΔP_{f}, control system ΔP_{c}:

$$\Delta P = \Delta P_{\text{т}} + \Delta P_{\mathrm{v}} + \Delta P_{\mathrm{f}} + \Delta P_{\mathrm{c}}\,. \tag{3.10.3}$$

Transformer loss consist of a transformer steel loss and loss at winding cooper. The first can be compared to loss determined from an idling experimentation ΔP_{idl} when a magnetic flow is rated and there is no current in windings (neglecting by a magnetization current). The second at a rated load can be compared to loss determined from a short circuit experimentation ΔP_{sc} when at transformer windings rated currents flow and there is almost no magnetic flow at low values of transformer short circuit voltage applied at this experimentation to transformer primary windings. Then

$$\Delta P_{\text{т}} = \Delta P_{\text{idl}} + \Delta P_{\text{sc}} \, (I_d / I_{d\text{rat}})^2. \qquad (3.10.4)$$

Active power loss at valves are summed from loss formed by a direct anode current flowing through an opened valve ΔP_{dir}, a reverse current flowing through a closed valve ΔP_{rev}, switching loss connected with finite turn-on and turn-off times of a valve ΔP_{switch}:

$$\Delta P = \Delta P_{\text{dir}} + \Delta P_{\text{rev}} + \Delta P_{\text{switch}} \, . \qquad (3.10.5)$$

To simplify a calculation of ΔP_{dir} a nonlinear voltage-ampere curve of a valve at a direct direction is approximated by piecewise-linear curves as it's shown at the fig. 1.3.2. It leads to an equivalent scheme of a valve at a direct direction consisting of a dc voltage source ΔU_0 (a cutoff voltage) and an active dynamic resistance R_{dyn}. Then an active power obtained at a such network is equal to

$$\Delta P_{\text{dir}} = I_{\text{a}} \Delta U_0 + I_{\text{a.rms}}^2 R_{\text{dyn}} = I_{\text{a}} \Delta U_0 + I_{\text{a}}^2 K_{\text{f}}^2 R_{\text{dyn}} \, . \qquad (3.10.6)$$

Active power loss at a reverse voltage ΔP_{rev} at a valve, as a rule, are not significant because of a small reverse current of a valve.

Active power loss at a valve switching are also relatively small in comparison with ΔP_{dir} at switching frequencies (a supply voltage frequency) equal to not greater than 400 Гц. At valves operation at high frequencies these loss are significant or basic in common loss. In these cases a calculation of switching loss is significantly determined by waveforms of valve currents and voltages at the next chapters devoted to converting devices operation at high commutation frequencies. These calculation features will be considered.

An active power at dc current link P_d in the common case at a finite value of a smoothing reactor X_d is equal to a sum of active powers from an interaction of like voltage and current harmonics:

$$P_d = \sum_{k=0}^{\infty} P_{d(k)} = \sum_{k=0}^{\infty} U_{d(k)} I_{d(k)} \cos\varphi_{(k)} \, . \qquad (3.10.7)$$

At an ideally smoothed current ($X_d = \infty$) we obtain the next expression

$$P_d = U_d I_d \,. \qquad (3.10.8)$$

Knowing P_d and ΔP we can calculate a converter performance index depending on a load changing or at U_d regulating.

3.10.2. A POWER FACTOR

A power factor at ac current link of a valve converter (at a rectifier input and an inverter output) is determined by a correlation of an active power to a full power. For a rectifier it gives

$$\chi = \frac{P_1}{S_1} = \frac{m_1 U_1 I_{1(1)} \cos\varphi_{1(1)}}{m_1 U_1 I_1} = \nu_I \cos\varphi_{1(1)}\,, \qquad (3.10.9)$$

where ν_I – a correlation of a fundamental current rms value of a transformer primary winding to a primary current rms value that is called *a current distortion factor*.

A shift of a fundamental primary current relatively a curve of a primary voltage that has a sinusoidal waveform is conditioned by two reasons at a valve converter. At first, it's conditioned by a presence of a commutation angle γ, secondly, by a presence of a regulation angle α that allows to write the next expression approximately

$$\varphi_{1(1)} \cong \left(\frac{1}{2} \cdots \frac{2}{3}\right)\gamma + \alpha \,. \qquad (3.10.10)$$

1/2 factor appears at values of α close to 90°, 2/3 factor – at α close to small values of angle. At a linear approximation of a current commutation sector (the previous section) it's always necessary to use a factor equal to 0,5.

For a naturally commutating inverter mode we obtain the next expression analogically to (3.10.10)

$$\varphi_{1(1)} = \beta - \left(\frac{1}{2} \cdots \frac{2}{3}\right)\gamma \,. \qquad (3.10.11)$$

So, according to an equation (3.10.9) a power factor can be interpreted as a degree of a useful using of an electrical-technical equipment capacity that is chosen которое for a full power and an active power equal to $P_1 = \chi S_1$ will be passed through it to convert to other types of an energy. Besides, an input current quality determines a degree of a negative reverse influence of a valve converter at ac mains (section 3.13).

Especially the expression for a valve converter power factor is significant at the assumption that $X_a = 0$, $X_d = \infty$, when $\gamma = 0$, $\varphi_{1(1)} = \alpha$. Then (3.10.9) is converted taking into account (2.9.3) to the next equation:

$$\chi = \nu_I \cos\varphi_{1(1)} = \nu_I \cos\alpha = \nu_I C_p . \qquad (3.10.12)$$

This significant energetic curve of a converter shows how much is a price at an input of an output voltage regulation.

Thus, a valve converter power factor is linearly depends on a voltage regulation degree at a dc current link. It is "an Achilles heel" of all (considered) valve converters at not fully controlled valves (thyristors). A presence of a great part of a valve load at a mains intensifies for power engineering specialists a problem of a power factor conservation at a mains at a normative or an optimal levels (usually about 0,9). It makes a construction task of valve converters with improved energetic indexes (a power factor and a performance index) ways of solving which are considered at the next section and chapter 13 actual.

3.11. RECTIFIERS BASED ON FULL_CONTROLLED VALVES

The aim of this section is study of rectifiers carried out at fully controlled valves (gate-controlled thyristors, gate-turn-off thyristors, transistors).

The considered controlled rectifiers at not fully controlled valves are characterized by that at the next valve turn-on a reverse voltage is applied to a valve conducting a load current and it's turned off (blocked) by a natural way. That's why such current commutation from a valve to a valve is called *a natural commutation.* But a valve turn-on delay relatively naturally ignition points at α leads to a reactive power drawing by a rectifier from a mains and decreasing of it's input power factor while α increases.

A current commutation at rectification schemes at fully controlled valves capable to turn-on and turn-off by an influence over a control network at a direct voltage presence at a valve is called *a forced commutation.* (Earlier not fully controlled valves synthetically had properties of fully controlled valves because of a special schematic solution – a synthetic commutation node. Such commutation is called *synthetic* [11, 12, 16].) A forced commutation allows to regulate a rectified voltage by other ways that don't have a pointed feature of a lagging phase regulation. We will consider three such ways:

– a lagging phase regulation;
– a pulse-width regulation of a rectified voltage;
– a forced forming of a rectifier primary current curve.

3.11.1.* A RECTIFIER WITH A LAGGING PHASE REGULATION

A three-phase bridge rectifier scheme at gate turn-off thyristors is shown at the fig. 3.11.1, time diagrams are represented at the fig. 3.11.2.

A basic scheme of a three-phase bridge rectifier at gate turn-off thyristors is supplemented with a device for a storage energy dropping (SED)

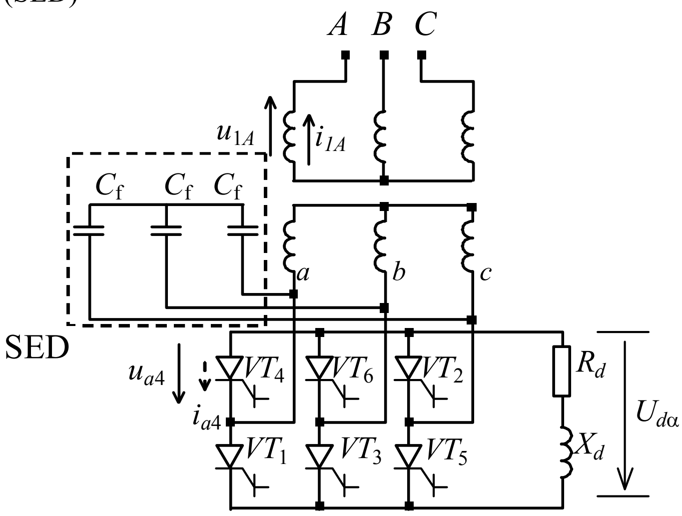


Fig. 3.11.1

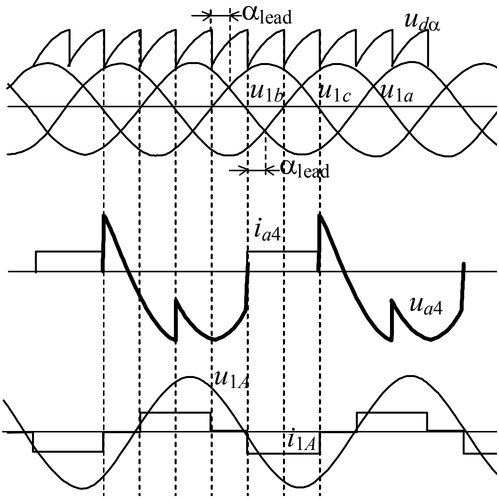


Fig. 3.11.2

from leakage inductances of a real transformer. This device consists of a three-phase block of capacitors C_f connected into a star or a triangle and connected at a valve block input. The next valve is turned-on at a moment of a control impulse supply with *a leading regulation angle* α_{lead} at it relatively a corresponding point of a natural commutation with a synchronous control impulse supply to a conducting valve turn-off. As a result a current at a turned-off valve will decrease to a zero stepwise (we neglect by processes of storage carriers resolution), a current at a turned-on valve will increase to a load current stepwise. A current commutation at transformer windings connected with these valve will last during a finite commutation time γ because of a presence of winding leakage inductances.

It's not difficult to make sure that an external, regulating and energetic rectifier curves with a leading phase regulation at the assumption that $X_d = \infty$ are obtained from the corresponding rectifier curves with a lagging phase regulation at changing α angle to α_{lead} angle and have an analogical view expect for curves for which a reactive power sign is important at a rectifier input. A rectifier input current leads a mains voltage at α_{lead} angle (we neglect by γ), i.e. a rectifier instead of a reactive power consumer became a reactive power generator. It allows to make a compound rectifier from two rectifier cells of the same type connected in parallel on inputs and in series (or in parallel) on an output. If to control by one cell with α angles, by another

– with $|\alpha_{lead}| = \alpha$ angles then, evidently, such compound rectifier will not consume a shifting reactive power on an input since a resulting input current will be in the phase relatively a mains voltage [16]. At this a rectified voltage waveform will be the same as at a single pulse-width regulation of a rectified voltage that is considered at the next section.

3.11.2.* A RECTIFIER WITH A PULSE-WIDTH REGULATION OF A RECTIFIED VOLTAGE

A considered rectifier consists of a basic cell of a three-phase voltage rectification on a bridge scheme and a SED as the previous scheme. Outwardly these two types of rectifiers are the same. The difference of their electromagnetic processes is conditioned by only a difference of valves control algorithms.

Time diagrams of voltages and currents of a rectifier with a pulse-width regulation (PWR) of a rectified voltage are shown at the fig. 3.11.3.

A voltage impulse forming at a rectifier output is provided by one valve turn-on at a cathode group and one valve – at an anode group as at a usual rectifier at not fully controlled valves and a phase regulation of a

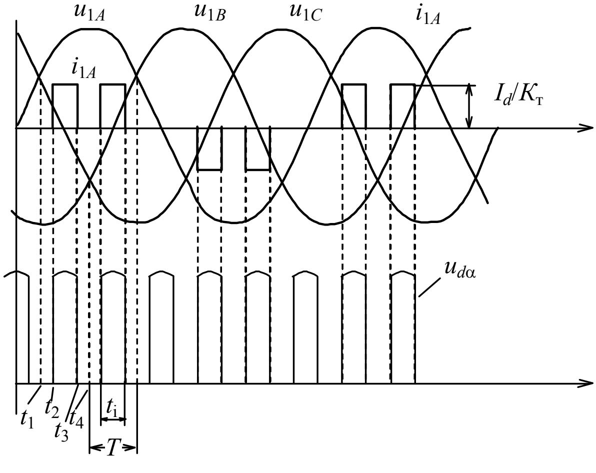


Fig. 3.11.3

rectified voltage, for example, by VT_1 and VT_6 valves turn-on at an interval t_2t_3. A voltage null pause forming at a rectifier output at an interval t_3t_5 is provided by turn-off of a gate-off thyristor VT_6 (or VT_1) on a control network with a simultaneous turn-on of an another thyristor of an operating scheme arm, i.e. VT_4 (or VT_2). At this a load current supported by a storage energy at a smoothing reactor having L_d inductance will flow through two conducting valves of one scheme arm. At a considered time interval they are VT_1 and VT_4 valves (or VT_2 and VT_6).

Energy storage at leakage inductances of transformer windings taking a part at a commutation (*a* and *в* phases) at first is dropped into capacitors C_f charging them and is taken away partly from them back to a mains and partly to a rectifier load.

It's evident that at load current closing intervals through valves of one bridge arm a rectifier is turned-off from a transformer and there will be no current at it's windings if to neglect by transformer magnetizing currents and filter capacitor currents C_f. Thus, transformer currents are also exposed to a pulse-width regulation as a rectified voltage as it's seen on i_{1A} curve .

Considered time diagrams of rectifier currents and voltages are related to the case when a frequency of rectifier output voltage impulses is greater than a mains voltage frequency in six times. To increase a fast-action of a rectified voltage and current regulation this frequency can be increased in 2, 3, 4 ... times then at an interval t_1t_4 there will be correspondingly 2, 3, 4 ... impulses at a primary current curve (i.e. there will be twelve-, eighteen-, …*n*-fold PWR instead of a six-fold PWR).

A PWR rectifier input power factor will be the next

$$\chi = \frac{P_1}{S_1} = \frac{P_d}{S_1} = \frac{U_{d\alpha} I_d}{3U_1 I_1} = \frac{C_p U_{d0} I_d}{3K_т E_2 \, \dfrac{I_d}{K_т} \sqrt{\dfrac{6T_т}{4t_i}}} = \frac{3}{\pi} \sqrt{C_r} \qquad (3.11.1)$$

taking into account a voltage conversion factor $K_{v.c.}$ and equations of a regulating curve at PWR (neglecting by a pulsation of impulse amplitudes)

$$C_r = \frac{U_{d\alpha}}{U_{d0}} = \frac{t_n}{T_т} . \qquad (3.11.2)$$

Thus, comparing a formula (3.11.1) with (3.10.12) it's seen that there is an input power factor improving at PWR. Besides a quantitative difference of an input power factor at a phase regulation and PWR there is a qualitative difference. At a lagging phase regulation a power factor worsening is conditioned by increasing of a primary current lagging relatively a mains voltage, i.e. by increasing a shifting reactive power consuming from a mains. At PWR a primary current is always in phase relatively a voltage, it's harmonic content worse at current impulses duration decreasing, i.e. a distortion power consuming increases from a mains.

3.11.3. A RECTIFIER WITH A FORCED CURVE FORMING OF A CURRENT DRAWN FROM A MAINS

At all rectification schemes considered above a current commutation at valves was accompanied with a current commutation at mains phases. At rectifiers at not fully controlled valves both commutations were realized in parallel, at rectifiers at fully controlled valves considered at this section, at first, a current commutation at valves was realized and then – a current commutation at phases. In both cases it led to an impulse nature of currents at input transformer phases and mains, i.e. to a reduced current quality in comparison with currents of electric energy linear consumers.

It's possible to significantly "correct" a valve converter nonlinearity on an input if to give to a valve converter a possibility to form it's input current curve. For this, obviously, firstly, it's necessary to a converter was realized at fully controlled valves and, secondly, after valves turn-off there have to be a path for continuing of a phase current flowing through an additional valve. A one-phase half-bridge scheme of a such converter at gate-off thyristors is shown at the fig. 3.11.4.

The additional valves are VD_1, VD_2 diodes. The second bridge arm is formed by C_1, C_2 capacitors from which there is obtained a dc voltage U_d simultaneously as from a rectifier output capacitance filter. An input reactor having an inductance L_f which is realized as a leakage inductance of an input transformer if it presents is meant for smoothing of pulsations conditioned by valve commutations at a continuous (without current pauses) of an input current curve.

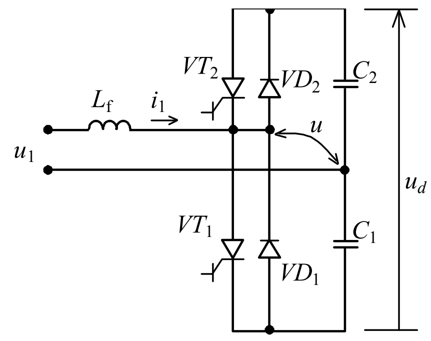


Fig. 3.11.4

We can modulate a conducting state duration of gate-off thyristors commutated at a high frequency at a sinusoidal law with a frequency equal to a mains voltage frequency by a pulse-width modulation (PWM). Then at the condition that U_d voltage is formed at a bridge output a pulse-width consequence of two-pole impulses u is formed at bridge input.

A positive voltage impulse at a bridge input u is formed at a turned-on state of a gate-off thyristor VT_2 or a diode VD_2 (depending on a current direction through this bridge arm), a negative voltage impulse u is formed at a turned-on state of a gate-off thyristor VT_1 or a diode VD_1. Under the influence of a difference of a mains voltage u_1 and a voltage u formed by a corresponding control a current limited by an inductance L_f i_1 will flow continuously with pulsations. At certain correlations between these voltages a fundamental current phase as it's seen from a vector diagram at the fig. 3.11.5 can be equal to a zero.

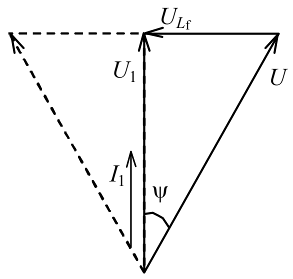


Fig. 3.11.5

At a sufficient increasing (in ten times or greater) of a thyristor commutation frequency over a mains voltage frequency current pulsations can be small, i.e. a rectifier input current will be an almost sinusoidal.

A scheme of a such rectifier fed from a three-phase mains is formed from three analogical valve arms as it's shown at the fig. 3.11.6. At this a necessity in a capacitive voltage divider that is at a one-phase scheme is no longer relevant.

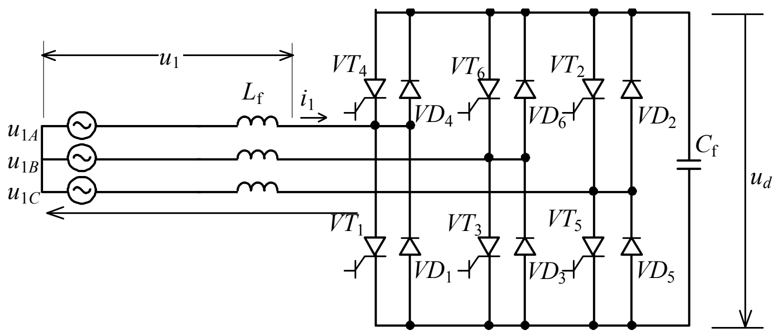
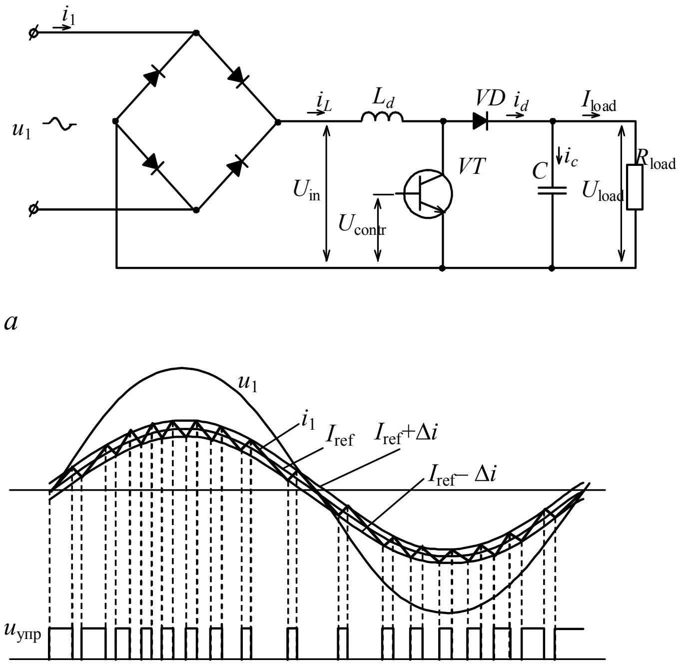


Fig. 3.11.6

As it's seen from a vector diagram part (fig. 3.11.5) shown by a hatch, at a negative sign of ψ angle and the same value of u voltage at a valve group input at a converter dc current link is in anti-phase relatively a voltage. It means that a PWM valve converter transfers to an inverting mode since an active power is transferred to an ac voltage link from a dc current link. By reducing a control angle ψ to a zero we can reduce an active power to a zero at a rectification and inverting modes. At this a dc current link voltage saves a sign and changes at finite limits that differs rectification-inverting modes at such PWM converter from a rectification-inverting modes at a converter at not fully controlled valves and a phase way of regulation (section 3.4).

The main characteristics of a such rectifier will be represented at the chapter 8 devoted to self-commutated inverters since this rectifier can be considered as a *reciprocal voltage inverter* (section 13.4).

If there is no requirement to a necessity of an energy recuperation from a rectifier dc current link, i.e. to provide a possibility of an inverting mode then a rectifier scheme with a forced input current forming is simplified and has the next view shown at the fig. 3.11.7, *a* for a one-phase mains, it's time diagrams are represented at the fig. 3.11.7, *b*.



a

b

Fig. 3.11.7

The scheme contains a one-phase bridge scheme of an uncontrolled rectifier, a storage reactor having an inductance L_d, a transistor (a fully controlled valve), a storage capacitor C with a separating diode *VD*. This part of the scheme after a diode rectifier as it will be shown a the chapter 7 is a sort of a boost dc-dc voltage converter. At a qualitative level it's operating mode is the next. At a conducting state of a transistor the whole rectified voltage of a diode bridge is applied to a storage reactor, it's current increases (интервал импульса управления $U_{упр}$ на рис. 3.11.7, *б*). At a transistor turn-off a leakage reactor current charges a storage capacitor C through an isolation diode *VD* and feeds a load network. Modulating a duration of a transistor conducting state with a frequency greater than a mains voltage frequency in many times, it's possible to form almost sinusoidal current half-waves at a storage reactor L_d synphase relatively a rectified voltage. A rectified current at a such one-phase scheme (at rectifier diodes conducting on a mains half-period a bridge commutating function Ψ_r is a rectangular oscillation) is an input current module on (1.4.2). Then an almost sinusoidal current that is in phase relatively a mains voltage is obtained at a rectifier input. At this an output voltage U_d has to be greater than a rectified voltage amplitude at a diode bridge output. It's necessary to provide a control of a current slump of a storage reactor having an inductance L_d at a transistor turn-off interval when a difference between pointed voltage in the direction reverse to that was at a current rising interval is applied to a reactor.

Formally this compound converter is formed by a cascade connection of two simple pointed valve converters and has to be considered by means of the methodic considered at the chapter 13 devoted to compound converters. But a widespread occurrence of this rectification scheme, first of all, to feed low-power loads by a stabilized voltage (control devices, TV-, radio- and a home equipment) warranties it's qualitative consideration. In the west this scheme is called *a power factor corrector* because of it's property to provide an input power factor almost equal to a unity. It appeared as a result of simplification of schemes of one-cascade rectifiers with a forced input current forming considered above. They have a capability to recuperate an energy from a load [19]. Taking down of this requirement allowed to transfer a function of a current curve forced forming from an a c current network (fig. 3.11.4 and 3.11.6) to dc current network as at the scheme represented at the fig. 3.11.7. A two-stage conversion scheme with a one controlled valve is cheaper than a one-stage conversion scheme with two controlled valves.

Thus, rectifiers at fully controlled valves (gate-off thyristors, power transistors) have the better input energetic curves in comparison with the case of their realization at fully controlled valves (thyristors). A novel circuit design takes a significant step on the way of an ideal rectifier construction that is characterized by the whole electromagnetic compatibility with a mains, i.e. regulated at the whole diapason by a dc voltage at an output and a sinusoidal current at an input synphase with a mains voltage (section 13.4).

3.12. A REVERSIBLE (DUAL) POWER CONVERTER (REVERSIBLE RECTIFIER)

The aim of this section is a consideration of valve converters that have a possibility to set any polarity combinations of dc voltage and current at an output.

A single power converter provides a possibility of a load voltage polarity reverse at storing a current direction at it (fig. 3.4.4). At the same time many areas of techniques and, first of all, an electric drive require a presence of a supply that can reverse not only a voltage but a load current that also requires a presence of four-quadrant external curves. For this a valve converter has to be capable to pass a dc current of any direction through itself analogically to other dc voltage supplies conventional for a power engineering such as an dc current electric drive generator or an accumulator. A similar regulated reversible supply can be obtained based on tow basic valve converters connected in a such way to provide a load current flowing in both directions. This system is called *a reversible valve converter (RVC)*. If each nonreversible valve converter that is inside a reversible converter is fed from a separate system of power transformer secondary windings a such scheme is called *a cross scheme,* at feeding both valve groups from one system of transformer secondary windings a scheme is called *an anti-parallel scheme.* The schemes allow to use integral modules of power valves of the same type, i.e. valve groups connected by cathodes (anodes) and made in one case. At realizing valve converters at thyristors most often an anti-parallel scheme is used. A scheme of a reversible valve converter at a three-phase half-wave rectification and an anti-parallel connection of valve groups is shown a the fig. 3.12.1.

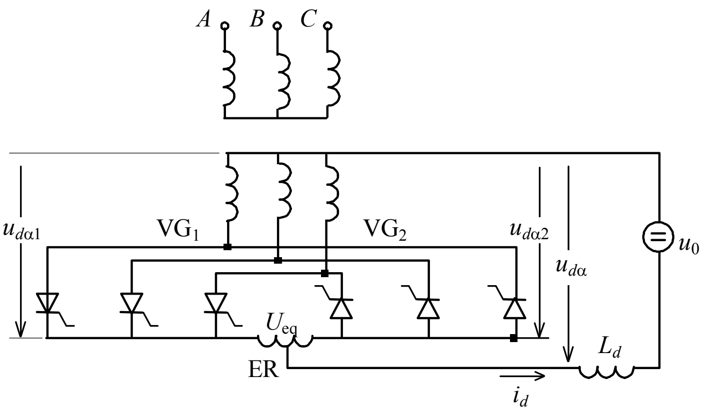


Fig. 3.12.1

A construction of a KMC by an anti-parallel connection of two valve groups VG_1 and VG_2 leads to appearing an additional contour for a current not including a load contour. This contour is formed by transformer windings and valves of VG_1 and VG_2 groups and is called an equalizing contour, a current flowing in it is called *an equalizing current*. A value of an equalizing current is determined by a difference of instantaneous values of voltages formed by valve VG_1 and VG_2 groups and a value of an equalizing contour resistance.

A practical absence of an active resistance at an equalizing contour requires to concord average values of valve group voltages to exclude a possibility of a continuous equalizing current appearing. For this average values of voltages of VG_1 and VG_2 groups have to meet the equation (at first, neglecting by ΔU_0 and a load current)

$$U_{d\alpha 1} = -U_{d\alpha 2}\,. \tag{3.12.1}$$

In other words, since valve groups are connected in parallel an equality of average values of their voltages is required. Taking into account the fact that they are connected in a counter way an opposition of signs of their own voltages is necessary, for this it's required that $\alpha_1(\alpha_2) < 90°$, $\alpha_2(\alpha_1) > 90°$ conditions were realized. Then taking into account the regulation curve equations (2.9.2), we will write the next expression

$$U_{d0}\cos\alpha_1 + U_{d0}\cos\alpha_2 = 0$$

or

$$2\cos\frac{\alpha_1+\alpha_2}{2}\cos\frac{\alpha_1-\alpha_2}{2} = 0\,,$$

where α_1 and α_2 are regulating angles of VG_1 and VG_2 groups, correspondingly. The equality can be realized at the next two conditions realizing:

$$\alpha_1 + \alpha_2 = 180°, \qquad (3.12.2)$$

$$\alpha_1 - \alpha_2 = 180°. \qquad (3.12.3)$$

The equation (3.12.2) is *a condition of a control concordance of RVC two valve groups*. At realizing it an equalizing current is limit continuous since a difference between instant voltage values of valve groups VG_1 and VG_2 in this case is a clearly alternating function.

The expression (3.12.3) in the case of realizing of RVC at not fully controlled valves is unrealized physically since there is required to $\alpha_1 > 180°$ or $\alpha_2 < 0$ that is impossible at a natural commutation. Only by using fully controlled valves it's possible to concord a control of two groups according to the condition (3.12.3). At this in the case of an even p a difference between instantaneous values of valve groups voltages that eliminates a reason of an equalizing current appearing at RVC as it's shown at [16].

The equation (3.12.2) is a condition of a precise equality of average values of valve groups voltages at an idling (without a load) that's why is can be called as a condition of a concordance at a joint (simultaneous) control of valve groups.

The concordance condition (3.2.12) of control angles α_1 and α_2 of valve groups means that at $\alpha_1 < 90°$, $\alpha_2 > 90°$ and, vice versa, when one valve group operates at a rectification mode then the second valve operates at naturally commutating inverter mode. It's control angle at an inversion mode taking into account (3.4.1) is equal to

$$\alpha_2 = 180° - \alpha_1 = \beta_1, \qquad (3.12.4)$$

i.e. factually when one valve group is controlled at a rectification mode with an angle $\alpha_1(\alpha_2)$ then the second valve group is controlled with an angle $\beta_1(\beta_2)$ equal to it at an inversion mode.

Average values valve group voltages $U_{d\alpha 1}$ and $U_{d\beta 2}$ are equal, instantaneous values are differed that's why an equalizing reactor (ER) perceiving this difference of voltages u_{eq} is connected between the groups. An equalizing current is parasitic since it additionally loads valves and transformer at a joint control.

Resulting external curves of a reversible rectifier are shown at the fig. 3.12.2. They are formed by two families

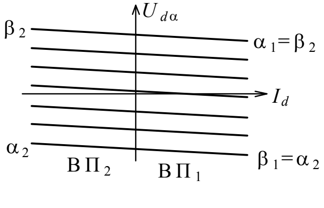


Fig. 3.12.2

of external curves such as are represented at the fig. 3.4.4 taking into account that the second valve group is inverse-parallel connected relatively the first valve group and provides another direction of dc load current.

Thus, a reversible valve converter is a universal dc current and voltage supply providing any combination of their polarities according to four quadrants of external curves, hence, a possibility of *an energy (returning) recuperation* from a load.

3.13. A REVERSE INFLUENCE OF A POWER CONVERTER AT A MAINS

An aim of this section is a consideration of a reverse influence of a power converter at a mains.

A specific of power electronics conversion devices realized at semiconductor controlled valves is bounded with a switch (discrete) nature of valves operation that determines a discretion as a process of an energy drawing from it's primary source by a converter since and a process of it's transfer to a consumer (load). A discrete energy consuming from an electric energy source by a converter leads to a significant reverse influence of a valve converter at a generated electric energy quality that consequences a converter itself and other consumers feeding from the same source feel.

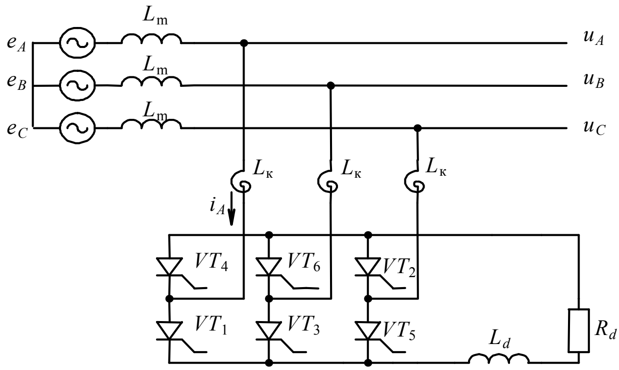
We will consider only a one side of a reverse negative influence of a valve converter at a mains that is a distortion of a mains voltage waveform from a nonsinusoidal nature of a valve converter input current.

A knowledge of a waveform and a spectrum content of input currents of typical power converters allows to calculate and to make a forecast a degree of a reverse influence of a valve converter at a mains of an autonomous system. For a such calculation it's necessary to have mathematical models of a mains and a valve converter on an input. A mathematical model of a mains can be obtained on the given topology of a mains and known parameters of it's elements. In the case of complicated structures of a mains a mathematical model is a frequency curve of a mains at a connection node. In the first approximation an equivalent of a mains is a emf source with an inductive mains reactance X_m, active mains resistances usually aren't taken into account. At the fig. 3.13.1 there are represented systems having a consumer as a three-phase bridge rectification scheme (*a*) and time diagrams of an input current i of a rectifier (at $X_d = \infty$), a mains emf e and a voltage at mains clamps u accessible to consumers (b).

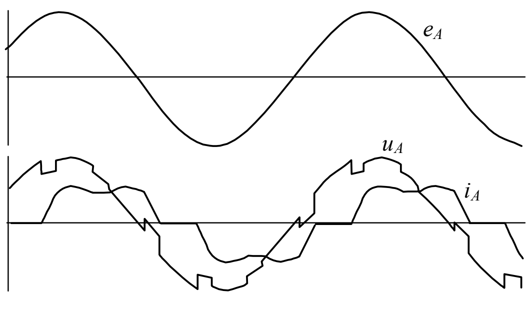
A presence a a transformer at a rectifier input is modulated by connecting a reduced leakage inductance of a transformer $L_к$. A mains voltage in this case is distorted at commutation intervals γ at a rectifier.

A mains is represented as a sinusoidal emf source e and a serial inductance L_m jointing all serial inductances of a network from a point of an electric energy obtaining to a point of it's consumption. Drops at a mains voltage curve are conditioned by the fact that the whole commutation voltage drop Δu_x (section 3.1) is divided between inductances L_m and $L_к$ and a voltage at a converter input is equal to

$$U = e - \Delta U_x \frac{L_m}{L_m + L_к} = e - L_m \frac{di}{dt}. \qquad (3.13.1)$$



a



b

Fig. 3.13.1

At this level of an assumption a valve converter on an input is changed substituted by a current supply of a known waveform. A calculated scheme of a source-converter system will look as it's shown at the fig. 3.13.2.

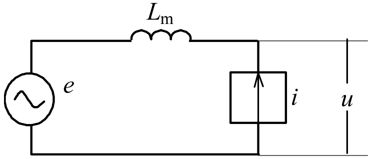


Fig. 3.13.2

From a differential equation for a mains voltage that is the next

$$u = e - L_{\mathrm{m}} \frac{di}{dt}, \tag{3.13.2}$$

we obtain a rms value of higher harmonics mains voltage by ADE2 method:

$$U_{\mathrm{h.h}} = L_{\mathrm{m}} \overline{I}^{(-1)}_{\mathrm{h.h}} = \omega L_{\mathrm{m}} I_{(1)} \overline{K}_{\mathrm{c.h}}, \tag{3.13.3}$$

and a rms value of a fundamental mains voltage by ADE method:

$$U_1 = \sqrt{E_1^2 - 2\omega L_{\mathrm{m}} E_1 I_1 \sin\varphi_1 + \omega^2 L_{\mathrm{m}}^2 I_1^2} \; .$$

As a result a mains voltage harmonic factor that doesn't have to overcome a value equal to $K_{\mathrm{mc.h}} = 0{,}08$ required by an All-Union State Standard 13109-97 will be equal to

$$K_{\mathrm{h}} = \frac{U_{\mathrm{h.h}}}{U_1} = \frac{\omega L_{\mathrm{m}} I_1 \overline{K}_{\mathrm{c.h}}}{\sqrt{E_1^2 - 2\omega L_{\mathrm{m}} E_1 L_1 \sin\varphi_1 + \omega^2 L_{\mathrm{m}}^2 I_1^2}} \le K_{\mathrm{mc.h}} \, . \tag{3.13.4}$$

Using a definition of a *short circuit current multiple factor* of a mains relatively a rated converter current K_{sc} equal to a correlation of a full power of a mains short circuit relatively a rated full power at a converter input

$$K_{\mathrm{sc}} = \frac{E_1}{\omega L_{\mathrm{m}}} \frac{1}{I_{1.rat}} \frac{E_1}{E_1} = \frac{E_1^2}{\omega L_{\mathrm{m}}} \frac{1}{E_1 I_{1.rat}} = \frac{S_{\mathrm{s.c}}}{S_{1.rat}}, \tag{3.13.5}$$

the expression (3.13.4) will be rewrote in the next form:

$$K_{\mathrm{h}} = \frac{\overline{K}^{(-1)}_{\mathrm{c.h}}}{\sqrt{K_{\mathrm{sc}}^2 - 2K_{\mathrm{sc}} \sin\varphi_1 + 1}} \le K_{\mathrm{mc.h}} \, . \tag{3.13.6}$$

A simple expression for K_{h} is obtained at neglecting by a difference between U_1 and E_1 at determining K_{h} on (3.13.4):

$$K_{\mathrm{h}} = \frac{U_{\mathrm{h.h}}}{E_1} = \frac{\omega L_{\mathrm{m}} I_1 \tilde{K}_{\mathrm{c.h}}}{E_1} = \frac{\tilde{K}_{\mathrm{c.h}}}{K_{\mathrm{sc}}} \le K_{\mathrm{mc.h}}\,. \qquad (3.13.7)$$

From (3.13.7) we find a limit power of a valve converter connected to a mains with a known short circuit power

$$K_{\mathrm{sc}} = \frac{S_{\mathrm{sc}}}{S_{\mathrm{rat}}} = \frac{\tilde{K}_{\mathrm{c.h}}}{K_{\mathrm{mc.h}}}, \qquad S_{\mathrm{rat}} = S_{\mathrm{sc}} \frac{K_{\mathrm{mc.h}}}{\tilde{K}_{\mathrm{c.h}}}\,. \qquad (3.13.8)$$

At other equal conditions this converter power is in inverse proportion to a differential harmonic factor of it’s input current. It allows for an each type of converter to determine it’s limit power without any difficulties at supplying from a mains with a given short circuit power. At electric energy quality European normative there is usually pointed (from an experience) a limit power of a connected valve converter (p = 6, 12) at parts of mains short circuit power.

In contrast with a considered case with one nonlinear consumer at ac current mains there is a lot of nonlinear consumers that resulting reverse influence at a mains can be summed and weaken at electric mains of common use. A reverse influence calculation for this case is given at the section 6.2.2.

Thus, a valve converter despite of a rule that is called “don’t bite a hand that feeds you” drawing an active power from a mains “pours out” at it higher harmonics power that spoils a mains voltage waveform and so complicates an operation of other electric energy consumers at a mains. To weaken this negative influence of valve converters at a mains besides a correlation limitation of converter and mains powers the next measures are taken:

1) increasing of a number of converter equivalent phases (sections 3.6 and 3.8);

2) applying of converter schemes with an improved input current waveform (sections 3.11.3);

3) a filtering of converter input currents, as a rule, by a parallel connecting serial LC-filters set at dominating input current harmonics at a mains (5, 7, 11, 13), (section 11.1);

4) using schemes of an input current active filtering compensating a deviation of a converter input current from a sinusoidal waveform (section 11.2).

CONTROL QUESTIONS

1. What is an equivalent circuit of a real transformer that is at a mathematical model of a rectifier?
2. Give a definition for a commutation angle of a rectifier.
3. What does an external curve of a controlled rectifier determine?
4. Why external curves of a controlled rectifier with an ideal filter are parallel at different values of α?
5. Write a generalized equation of an external curve of a rectifier.
6. Give a definition for an interrupted current mode of a rectifier.
7. What is an influence of an interrupted rectified current mode to external and regulating curves of a rectifier?
8. Which equivalent mode an operation mode of a rectifier at a load with a capacitive filter can be led to?
9. Which converter is called as a naturally commutating inverter?
10. How to transfer a rectifier loaded at a counter-emf to a naturally commutating inverter mode?
11. Which are the main curves of a naturally commutating inverter?
12. Why is an operation of a naturally commutating inverter with an angle equal to $\beta=0$ impossible?
13*. On what reasons a transfer of a naturally commutating inverter to an “upset” mode is possible?
14*. What does a Chernyshev rule at a rectifier set?
15. Which numbers of harmonics are there at rectifier primary currents?
16. Which numbers of harmonics are there at a rectifier rectified voltage?
17. At which number of rectifier transformer secondary phases is a transformer used optimally?
18*. What is an influence of a commutation angle to using of a rectifier transformer?
19. Give a definition for a quality factor of a rectifier and a naturally commutating inverter.
20. Give a definition for a power factor of a rectifier and a naturally commutating inverter.
21*. Why are fully controlled valves required at regulating of a rectified voltage by a leading angle α_{lead}?

22*. What is the difference between a rectifier input current with a pulse-width regulating and a rectifier input current with a phase regulating (lagging or leading)?

23. To which value an input power factor of a rectifier with a forced forming of a primary current curve is equal?

24. How a reverse influence of a valve converter at a mains is shown?

25. From which parameters of rectifier input current does a degree of it's reverse influence at a mains depend on?

EXERCISES

1.Lead out a correlation between a short circuit voltage of a transformer and it's leakage inductance (an anode inductance L_a) neglecting by active resistances of windings and taking into account them.

2. A one-phase bridge rectifier having I_d = 10 A is fed from a transformer having U_1 = 220 V, $K_т$ = 2, L_a = 0,01 H. Find an average value of a voltage ata load ($X_d = \infty$).

3.* From what condition can be obtained a correlation of an average value of a rectified current at a limit continuous mode from a angle λ at $m = 2$?

4.* Calculate a value of a forced regulation angle α_{forc} at a transformerless three-pulse rectification scheme at a value of a counter-emf equal to 200 V.

5.* A rectifier on a three-phase bridge scheme with a transformer having a factor $K_т$ = 2 and L_a = 0,005 H operates with a control angle α = 80° at a counter-emf. From which value of a rectified current a mode of a naturally commutating inversion will begin?

6. A three-phase rectifier having a scheme of transformer windings of Δ/λ_0 type is loaded ata counter-emf. A regulating angle α=30°,$K_т$ = 1, L_a =0,01 H, U_0=200 V. How much is an average value of a rectified current at it?

7. Which minimal value of a regulating angle β_{min} is it required to set at a transformerless one-phase bridge naturally commutating inverter if a recovery time of control properties of thyristors is equal to 200 μs ($X_d = \infty$)?

8.* Which maximal value of an inverted current is permissible ata three-phase bridge naturally commutating inverter at β = 30°, L_a

$= 0{,}005$ H,
$K_т = 2$ at thyristors having a null controllability recovery time?

9.* Calculate a harmonic factor of a mains current of an equivalent twelve-phase rectifier based on two three-phase bridges.

10. Calculate an input power factor of a six-pulse controlled rectifier operating at $\alpha = 60°$ and having $\gamma = 20°$.

11. At a six-pulse rectifier active power loss at idling equal to $\Delta P_{idl} = 200$ W are determined at a rated load current: at a transformer – 600 W, at a smoothing filter reactor – 200 W, at valves – 400 W at this a correlation between components of power loss at a valve dynamic resistance and a source modulating a cut-off voltage of a valve voltage – ampere curve line is equal to 1:1. A load power is equal to 10 кW. Determine a quality factor at a given and a half of a full loads of a rectifier taking a value equal to $U_{d\alpha} = \text{const}$.

12. Calculate an input power factor of a rectifier with a forced forming of a primary current curve if a fundamental current amplitude coinciding with a voltage on a phase is equal to 10 A, an amplitude of a high-frequency pulsation of this current is equal to 2 A.

13. Estimate a maximal power of a six-pulse rectifier connected to a mains and operating with a commutating angle equal to $\gamma = 20°$ if a short circuit power of a mains node is equal to 1000 кVA.

Chapter 4*

A MODEL EXAMPLE OF THE RECTIFIER DESIGNING

К беде неопытность ведет ...

И опыт, сын ошибок трудных,
И гений, парадоксов друг,
И случай, бог изобретатель.

Alexander S. Pushkin

The chapter is devoted to describing a designing of the controlled rectifier feeding a dc current motor that can be considered as an example for making a course paper and fragmentary as a set of topics of practical courses on the course.

As a matter of fact a heuristic stage of a rectifier scheme choice is based on the knowledge of properties of the basic rectification schemes. A rated stage on determining of scheme elements parameters is based on theoretical correlations obtained at the previous chapters. Calculation results are checked up by a mathematical modeling of the three-phase bridge scheme of the rectifier designed by means of Parus-ParGraph software implemented within the limits of the laboratory work (here – № 4) [22].

A task. It's necessary to design the rectifier to provide a run up of a dc current motor such as П2 where the current isn't greater than a rated armature current and to provide a sustained operation at a rated torque (current) at a rated number of revolutions at a constant excitation flux. Motor parameters are the next: P_{load} = 100 kW, $U_{\text{an.rat}}$ = 440 V, n_{load} = 1000 min^{-1}. Permissible pulsations of an anchor current aren't greater than 7 % from $I_{d\text{load}}$. A field coil voltage is equal to U_{fc} = 220 V.

It's necessary to determine parameters of a mains transformer, valves of anchor cable rectifiers and a field coil, smoothing reactors of rectifiers. A limitation requirement: a rectifier input power factor have to be not less than 0,8 at a rated mode. A three-phase mains is 220/380 V with an available neutral. A mains short circuit power is equal to S_{sc} = 5 000 кVAr, i.e. K_{sc} = 50 at a node of the converter connecting.

A novel rectifier designing includes two stages that are qualitatively different:

1) a structure synthesis stage at that a rectifier structure (schematic circuit) is determined;

2) a parametrical synthesis stage at that parameters of the selected rectifier structure (schematic circuit) elements are calculated.

4.1. A RECTIFIER SCHEME SELECTION (STRUCTURE SYNTHESIS STAGE)

For the time present there is no formal (mathematical) methods of synthesis of valve converter structures on a task requirement at the power electronics, although effort are carried out in this direction [19, 49] (section 13.3). That's why a synthesis procedure of a rectifier scheme is led to selecting from a set of known schemes basing on knowledge about their properties. Thus, a database on rectifier schemes is necessary. In those cases when it's impossible to select an appropriate rectifier scheme from a set of known schemes it's required a novel scheme invention or a correction of a rectifier designing task.

Basing on the analysis results of rectifiers basic schemes of one-phase and three-phase voltages a summary table is composed. Because of a many-dimensionality of an each scheme property vector based by parameters of table columns. A scheme selection at designing a novel rectifier with a required output parameters is potentially ambiguous and is difficult for a young specialist. That's why there is given an example of a rectifier scheme selection algorithm below based on three given parameters of the rectifier output (P_{d0}, U_{d0}, I_d) taking into account at a vector of scheme properties only two of them: using a standard transformer power and using valves on a reverse voltage. At this it's supposed that a designer has at his disposal valves with a maximum reverse voltage value (to 1000…1500 V), a load factor on valve voltages is equal to 1,5…2 at designing. In spite of a conditional character of this algorithm

Parameters of rectifier basic schemes

Scheme	VC in common						Valves					Transformer		
	qm_2	U_{d0} $K_{\text{т}}/U_1$	I_d $K_{\text{т}}/I_1$	χ	$K_{\text{c}(1)}$	$\bar{K}_{\text{h}}$	I_{a}^*	K_{f}	K_a	$U_{b\max}^*$	S_b/S_b'	S_2^*	S_1^*	$S_{\text{т}}^*$
$m_1 = 1, m_2 = 2, q = 1$	2	0,9	1,11	0,9	0,667	0,24	0,5	$\sqrt{2}$	2	3,14	3,14/6,28	1,57	1,11	1,34
$m_1 = m_2 = 1, q = 2$	2	0,9	1,11	0,9	0,667	0,24	0,5	$\sqrt{2}$	2	1,57	3,14/6,28	1,11	1,11	1,11
$m_1 = m_2 = 3, q = 1$ a triangle-star	3	1,17	1,21	0,79	0,25	0,06	0,33	$\sqrt{3}$	3	2,09	2,09/6,28	1,48	1,21	1,345
$m_1 = m_2 = 3, q = 1$ a star-zigzag	3	1,17	1,48	0,83	0,25	0,06	0,33	$\sqrt{3}$	3	2,09	2,09/6,28	1,71	1,21	1,46
$m_1 = 3, m_2 = 6, q = 1$ with a paralleling reactor	6	1,17	2,56	0,955	0,057	0,0067	0,166	$\sqrt{3}$	6	2,09	2,09/6,28	1,48	1,045	1,26+0,07 (УР)
$m_1 = m_2 = 3, q = 2$	6	2,34	1,28	0,955	0,057	0,0067	0,33	$\sqrt{3}$	3	1,045	2,09/6,28	1,045	1,045	1,045

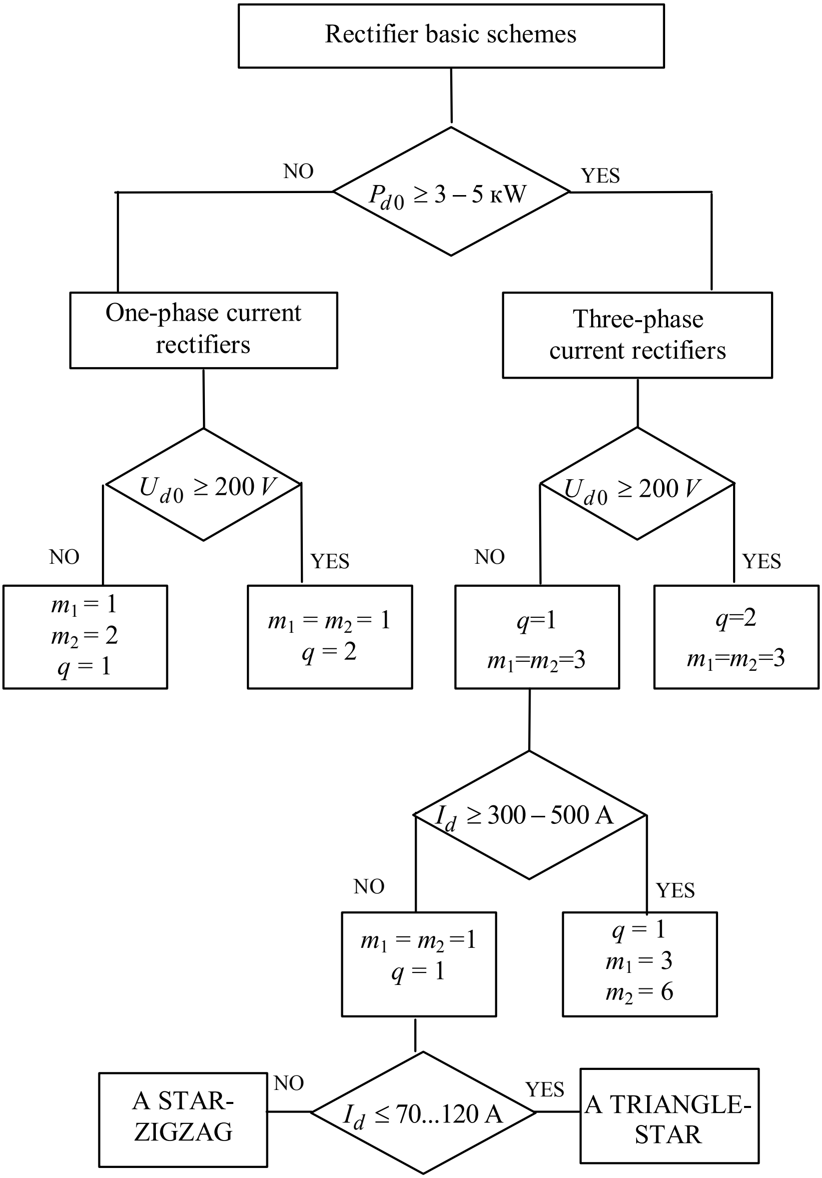

Rectifier basic schemes
$P_{d0} \geq 3-5$ кW
NO
YES
One-phase current rectifiers
Three-phase current rectifiers
$U_{d0} \geq 200\ V$
NO
YES
$m_1 = 1$ $m_2 = 2$ $q = 1$
$m_1 = m_2 = 1$ $q = 2$
$U_{d0} \geq 200\ V$
NO
YES
q=1 m_1=m_2=3
q=2 m_1=m_2=3
$I_d \geq 300-500$ A
NO
YES
$m_1 = m_2 = 1$ $q = 1$
$q = 1$ $m_1 = 3$ $m_2 = 6$
$I_d \leq 70...120$ A
NO
YES
A STAR-ZIGZAG
A TRIANGLE-STAR

Fig. 4.1.1

It will be useful as a possible example of an approach until a designer will have his own experience.

According to a designing task and a rectifier scheme selection algorithm on the fig. 4.1.1 the rectifier has to be three-phase (P_{d0} = 100 кW) and full-wave (a bridge scheme) since a high rectified voltage is required.

In the common case solutions of the similar tasks of a decision making can be formalized making a corresponding *expert system* based on power electronics knowledge as a program for a computer.

A field coil rectifier is also three-phase but because of a low value of a rectified voltage it can be made at a half-wave scheme. Since conversion factors on a voltage of selected rectifier schemes are differed in two times and their required rectified voltages are also differed in two times. A variant of feeding both the schemes from one system of transformer secondary windings is possible. Taking into account the fact that a transformation factor $K_т$ is greater that a unity but is close greater than a unity (buck transformer), a variant of rectifier feeding immediately from a mains is possible (without a rectifier transformer).

Thus, there are three alternative solutions for a designer and basing on calculation results we have to select one of them that requires to attract additional favour by a men which decides if a designer and he are different people.

4.2. A CALCULATION OF ELEMENT PARAMETERS OF A CONTROLLED RECTIFIER SCHEME (A PARAMETRIC SYNTHESIS STAGE)

A rectifier calculation for an anchor circuit taking into account real parameters of scheme elements basing on results represented at the chapter 3 requires knowledge of elements parameters. A rectifier calculation on ideal elements basing on results obtained at the chapter 2 doesn't require knowledge real elements parameters. That's why we are obliged to design a rectifier into two stages. At the first stage basing on the results represented at the chapter 2 types of elements are estimated for an ideal rectifier and we find their real parameters for these elements by means of reference books. At the second stage a correcting calculation is made for a rectifier taking into account real parameters of elements basing on the results obtained at the chapter 3.

4.2.1. AN ESTIMATION OF IDEAL RECTIFIER ELEMENTS

A mains voltage is determined by the electric energy quality standard [44]. It's maximal deviation from a rated value can achieve ± 10 %. That's why it's necessary to provide a rated rectified voltage at a minimum possible mains voltage. At this it's rational to have the regulation angle α equal to a zero at a rectifier. Then according to the section 2.8 taking into account the fact that $U_{\text{an.rat}} = U_{d0}$, we have the next expression

$$U_{2\min} = \frac{U_{d0}}{2,34} = \frac{440}{2,34} = 188 \text{ B},$$

Supposing that transformer windings are connected on a star-star scheme and a transformation factor of an input transformer is equal to

$$K_{\text{т}} = \frac{U_{1\min}}{U_{2\min}} = \frac{0,9 \cdot 220}{188} = 1,05.$$

Leaning on the correlations from the section 2.2.8 we can determine the others calculate values.

An average value of a rectified current is equal to

$$I_{drat} = \frac{P_{\text{an.rat}}}{U_{\text{an.rat}}} = \frac{100 \cdot 10^3}{440} = 227 \text{ A}.$$

An average value of a valve anode current is the next

$$I_{\text{a}} = \frac{I_{drat}}{3} = \frac{227}{3} = 76 \text{ A}.$$

A rms value of a valve anode current is equal to

$$I_{\text{a.rms}} = \frac{I_{drat}}{\sqrt{3}} = \frac{227}{\sqrt{3}} = 131 \text{ A}.$$

We choose a thyristor on an average value of an anode current taking into account that an amplitude factor is equal to $K_{\text{a}} = 3$. It's the thyristor T9-100 having the next parameters: $R_{\text{dyn}} = 0,002\ \Omega$, $\Delta U_0 = 1,3$ V [25, 26]. We determine a valve class on a voltage after more precise a maximal reverse voltage at a valve.

A rms value of a secondary transformer current

$$I_{2\text{rat}} = I_{drat}\sqrt{\frac{2}{3}} = 227\sqrt{\frac{2}{3}} = 185 \ \text{A}.$$

A rms value of a primary transformer current

$$I_{1rat} = \frac{I_2}{K_\text{т}} = \frac{187}{1{,}05} = 177 \ \text{A}.$$

A typical transformer power is determined taking into account that a mains voltage can be greater than a rated value:

$$S_\text{т} = S_2 = S_1 = 3U_{1\max} I_1 = 3 \cdot 1{,}1 \cdot 220 \cdot 177 = 143 \text{ кВА}.$$

According to a reference data contained in [25, 26] we have the next parameters for a transformer of the nearest great power of ТСП-160 type:

$$\Delta P_\text{sc} = 2{,}3 \text{ кW}, \qquad \Delta P_\text{idl} = 0{,}7 \text{ кW}, \qquad U_\text{к} = 6{,}2 \ \%.$$

If a standard industrial transformer isn't appropriate on a transformation factor $K_\text{т}$ then we have to design and produce a transformer that will have approximately the same parameters of the parameters which we are interested in. That's why we determine necessary elements parameters of T-type transformer equivalent circuit through these transformer parameters.

A short circuit impedance module of a transformer

$$Z_{1\text{k}} = \frac{U_\text{sc}}{I_{1\text{load}}} = \frac{0{,}062 \cdot 220}{177} = 0{,}077 \ \Omega.$$

An active transformer winding resistance reduced to a primary side is the next:

$$R_\text{tr} = R_1' + R_2' = \frac{\Delta P_\text{sc}}{3I_{1\text{load}}^2} = \frac{2300}{3 \cdot 177^2} = 0{,}022 \ \Omega.$$

A reactive leakage resistance of transformer windings that is reduced to a primary side:

$$X_{1k} = \sqrt{Z_{1k}^2 - R_\text{тr}^2} = \sqrt{(0{,}077)^2 - (0{,}022)^2} = 0{,}073 \ \Omega.$$

Then the same resistance reduced to the secondary transformer windings and called as an anode resistance X_a is the next

$$X_{\text{a}} = \frac{X_{1\text{к}}}{K_{\text{т}}^2} = \frac{0,073}{1,05^2} = 0,067\ \Omega, \qquad L_{\text{a}} = \frac{X_{\text{a}}}{\omega} = 0,2 \cdot 10^{-3}\ \ H\ .$$

Then we have to estimate parameters of a real smoothing reactor with an inductance L_d. The reactor calculation is made for the mode with a maximal possible regulation angle $\alpha_{\max}$ for the worst on a rectified current quality. This angle appears at a rectifier operation at a maximal mains voltage and is determined by a rectifier regulation curve

$$U_{d\text{rat}} = U_{d0\max} \cos \alpha_{\max} = 2,34\ U_{2\max}\ \cos \alpha_{\max}.$$

Then

$$\cos \alpha_{\max} = \frac{U_{drat}}{2,34 U_{2\min} \cdot 1,2} = \frac{440}{2,34 \cdot 188 \cdot 1,2} = 0,87, \quad \alpha_{\max} = 30^\circ\ .$$

A pulsation factor of a rectified current isn't worse than 0,07:

$$K_{\text{c.c}} = \frac{I_{d(6)}}{I_{drat}} \le 0,07\ , \quad I_{d(6)} = 0,07 \cdot 227 = 16\ \text{A},$$

where $I_{d(6)}$ – a fundamental amplitude of rectified current pulsations that is sixth harmonic relatively a mains voltage frequency at the six-pulse rectifier. This harmonic at a current is determined through a sixth harmonic at a rectified voltage that according to (3.7.4) is equal to $U_{d\alpha(6)} = 0,18 \cdot 556 = 102$ V at a maximal mains voltage equal to 242 V.

A required summary inductance of a rectified current contour is the next

$$L_{d\Sigma} = \frac{U_{d\alpha(6)}}{6\omega_1 I_{d(6)}} = \frac{102}{6 \cdot 314 \cdot 16} = 3,4 \cdot 10^{-3}\ \ H\ ,$$

then a smoothing reactor inductance is the next

$$L_d = L_{d\Sigma} - 2L_{\text{a}} = 3,4 \cdot 10^{-3} - 2 \cdot 0,2 \cdot 10^{-3} = 3 \cdot 10^{-3}\ \ H\ .$$

Then we choose an appropriate smoothing reactor for a current equal not less than 225 A by a reference book [26, 37]. It's a reactor of ФРОС-250 type. It has an active winding resistance equal to $R_{\text{F}} = 0,012$ Ω at an inductance equal to $3,2 \cdot 10^{-3}$ H.

Then we can correct a rectifier calculation taking into account real parameters of elements.

4.2.2. A RECTIFIER CALCULATION TAKING INTO ACCOUNT REAL PARAMETERS OF SCHEME ELEMENTS

A presence of real elements leads to an appearing of voltage loss ΔU inside a rectifier at a rectifier load that requires an overstating of a rectifier idling voltage that according to a generalized equation of an external curve (3.1.14) is the next (at a minimum mains voltage)

$$U_{d0\min} = \frac{U_{drat} + I_d\left[\frac{3X_{\mathrm{a}}}{\pi} + 2\left(R_1' + R_2\right) + qR_{\mathrm{dyn}} + R_{\mathrm{f}}\right] + q\Delta U_0}{\cos\alpha_{\min}} =$$

$$= 440 + 227\left[\frac{3}{\pi}\ 0{,}067 + 0{,}022\cdot 2 + 2\cdot 2\cdot 10^{-3} + 0{,}012\right] + 2\cdot 1{,}3 = 479\,V.$$

A voltage ΔU is dropped inside a rectifier:

$$\Delta U = U_{d0\min} - U_{drat} = 479 - 440 = 39\ \ V\ .$$

Then a rms value of a secondary transformer voltage corresponding to the voltage ΔU at a minimum mains voltage is equal to

$$U_{2\min} = \frac{U_{d0\min}}{2{,}34} = \frac{479}{2{,}34} = 204{,}7\ \ V$$

a transformation factor is equal to

$$K_{\text{т}} = \frac{U_{1\min}}{U_{2\min}} = \frac{0{,}9\cdot 220}{204{,}7} = 0{,}97\ .$$

Then it can be seen that a transformerless variant of a rectifier provides a possibility to store a load voltage at decreasing a mains voltage only at 7 % that corresponds to a voltage drop within a rate (±5 %) according to the All-Union State Standard 13109-97 [44]. At a maximum permissible mains voltage drop at 10 % a load voltage is decreased from a rated value approximately at 3 % in this case. It's "a payment" for an economy at an input transformer if not to use it.

A typical transformer power is the same if not to take into account a commutation influence on it. To estimate this influence according to (3.1.7) at first we determine a commutation angle γ for the case when a mains voltage is maximal:

$$\gamma = \arccos\left[\cos\alpha - \frac{I_d X_a}{\sqrt{2} U_2 \sin\dfrac{\pi}{m_2}}\right] - \alpha =$$

$$= \arccos\left\{\cos 30 - \frac{227 \cdot 0,067}{\sqrt{2} \cdot 230,5 \cdot 0,87}\right\} - 30 = 6°.$$

Taking into account corrections on a commutation according to (3.9.4) a transformer typical power changes not significantly and we can not to take it into account.

Then we can determine valve parameters on a reverse voltage that can achieve the next value at a maximal mains voltage:

$$U_{b\max} = 1,2 U_{d0\min} \cdot 1,045 = 1,2 \cdot 479 \cdot 1,05 = 670 \ V.$$

Taking into account possible pulse overvoltages inside a rectifier and in a mains we choose a valve taking a coverage voltage factor equal to 1,5…2. Finally we choose T9-100 valve of not lower than 10-th class. A valve class multiplied to 100 determines maximum permissible direct and reverse voltages at it.

A decreasing of a factor $K_т$ leads to a maximal value correction of a regulation angle α_{max}:

$$\cos\alpha'_{\max} = \frac{U_{drat}}{2,34 U_{2\min} \cdot 1,2} = \frac{440}{2,34 \cdot 208,4 \cdot 1,2} = 0,75, \qquad \alpha'_{\max} = 41°.$$

The sixth harmonic of a rectified voltage have to be determined taking into account an appeared commutation angle γ [8] and instead of 102 V it is equal to

$$U_{d\alpha(6)} = 0,24 \cdot 556 = 133,4 \text{ V}.$$

An inductance of a smoothing reactor L_d is increased proportionally at 30 %.

Finally we have to check a limitation of an input power factor task. To do it it’s necessary to know an active power at a rectifier input taking into account it’s loss inside a rectifier. Transformer power loss are equal to

$$\Delta P_{\text{тr}} = \Delta P_{\text{idl}} + \Delta P_{\text{sc}}\left(\frac{I_d}{I_{drat}}\right)^2 = 0,7 + 2,3 = 3 \text{ кW}.$$

Active power loss at valves are the next

$$\Delta P_{\text{v}} = 6\left(I_{\text{a}}\Delta U_0 + I^2_{\text{a.rms}}R_{\text{d.rat}}\right) = 6(76\cdot 1,3 + 131^2\cdot 2\cdot 10^{-3}) = 0,8 \text{ кW}.$$

Active power loss at a smoothing reactor are equal to

$$\Delta P_{\text{f}} = I_d^2 R_{\text{f}} = 252^2\cdot 12\cdot 10^{-3} = 0,7 \text{ кW}.$$

Common power loss inside a rectifier are equal to

$$\Delta P = \Delta P_{\text{тr}} + \Delta P_{\text{v}} + \Delta P_{\text{f}} = 3 + 0,8 + 0,76 = 4,56 \text{ кW}.$$

Then an input power factor of a rectifier is the next at a rated value of a mains voltage

$$\chi_{\text{rat}} = \frac{P_1}{S_1} = \frac{P_d + \Delta P}{3U_{1rat}I_{1rat}} = \frac{(100 + 4,56)\ 10^3}{3\cdot 220\cdot 177} = 0,9\,,$$

at a maximal mains voltage we have the next expression

$$\chi_{\text{min}} = \frac{P_1}{S_{1\max}} = \frac{P_d + \Delta P}{3U_{1\max}I_{1rat}} = \frac{104,56\cdot 10^3}{3\cdot 242\cdot 177} = 0,81,$$

i.e. it’s greater than a preset limitation.

At a maximal mains voltage a rectifier performance index is determined on the next formula

$$\eta = \frac{P_d}{P_1} = \frac{P_d}{P_d + \Delta P} = \frac{100}{104,56}\cdot 100\,\% = 95,6\,\%.$$

Thus, a designed rectifier meets all the task requirements.

Finally, we check whether a rectifier meets requirements of All-Union State Standard 13109-97 on an insertion distortion of a mains voltage at a connection node. A harmonic factor of a mains node voltage is determined on a correlation (3.13.7). The factor is conditioned by a nonsinusoidality of a rectifier input current. A differential harmonic factor of the first order of a rectifier input current at γ_{rat} = 10° (at $U_{1\text{rat}}$ = 220 V, α_{rat} = 15°) on a correlation (3.9.8) is equal to

$$\tilde{K}_{\text{c.h}} = \sqrt{\frac{\pi}{\gamma} - 1} = \sqrt{\frac{18\pi}{\pi} - 1} = 4,12,$$

since

$$\gamma_{\text{rat}} = \arccos\left\{\cos\alpha_{\text{rat}} - \frac{I_{drat} X_{\text{a}}}{\sqrt{2}\ U_{2rat} \sin\frac{\pi}{3}}\right\} - \alpha_{\text{rat}} =$$

$$= \arccos\left\{0,966 - \frac{227 \cdot 0,067}{\sqrt{2} \cdot 209 \cdot 0,866}\right\} - 15° = 10°.$$

Then a harmonic factor of a mains voltage that's *called as a nonsinusoidality factor* at an All-Union State Standard 13109-97 is equal to

$$K_{\text{h}} = \frac{\tilde{K}_{\text{c.h}}}{\text{K}_{\text{sc}}} = \frac{4,12}{50} = 0,08,$$

that is permissible according to this standard.

We can estimate mass-size factors of a designed rectifier on determined parameters of scheme elements on specific factors of element mass and size (section 1.2.2).

A rectifier of electric motor field coil is calculated analogically.

4.3. CALCULATION RESULTS CHECKING BY A MATHEMATICAL MODELING BY PARUS-PARGRAPH SOFTWARE

Calculation results accordance to a task requirement is checked by by a mathematical modeling by Parus-Pargraph software of a three-phase bridge rectification scheme. At the fig. 4.3.1 a model of an investigated rectifier is represented . The model is supplemented with a control system of thyristors on the method of multi-channel vertical control (chapter 12). Parameters of model elements are the next:

a mains $U_1 = 220$ V, $L_{\text{c}} = 4 \cdot 10^{-6}$ H,

a transformer $L_{\mu} = 0,1$ H, $L'_{\text{к}} = 0,4 \cdot 10^{-3}$ H, $R'_{\text{к}} = 0,022\ \Omega$,

valves $R_{\text{dyn}} = 0,002\ \Omega$, $\Delta U_0 = 1,3$ V, $\alpha_{\min} = \arccos 0,9 = 26°$,

a filter reactor $L_d = 3 \cdot 10^{-3}$ H, $R_{\text{f}} = 12 \cdot 10^{-3}\ \Omega$,

a counter emf $U_0 = 440$ V.

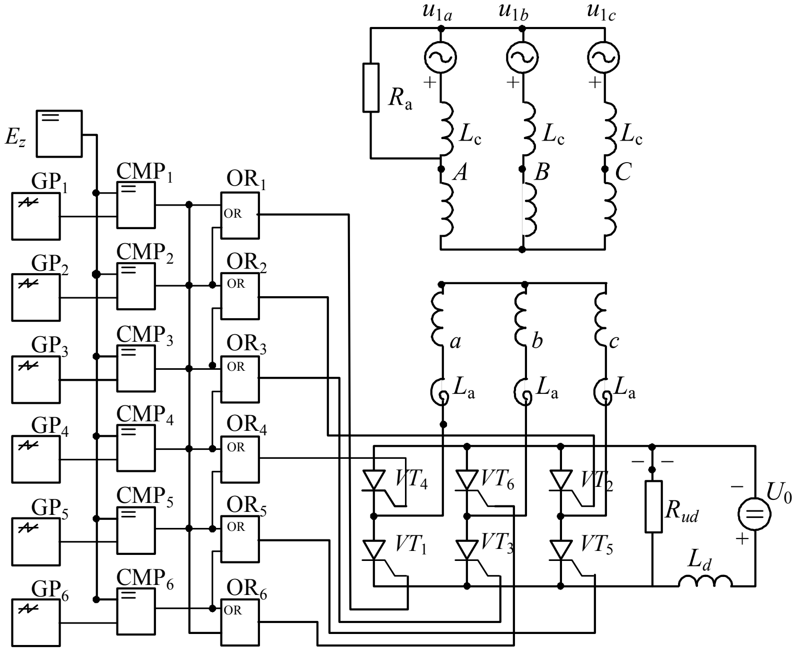


Fig. 4.3.1

Diagrams of scheme elements voltages and currents for a rated mode are represented at the fig. 4.3.2. An average value of a rectified voltage is equal to 425 V at the current equal to I_d = 100 A that differs from it's given value equal to 440 V to 3,4 %. It's an evidence of a sufficient

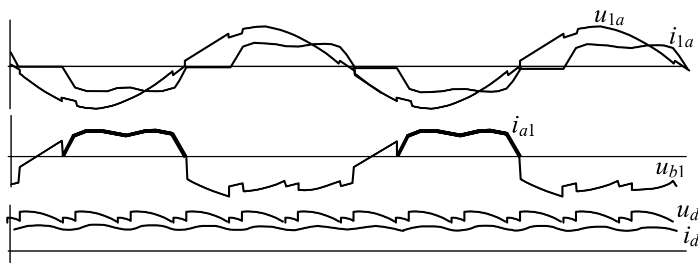


Fig. 4.3.2

accuracy of a rectifier calculation. A comparison is made on the condition of obtaining (stabilizing) of the given average value of a rectified current that is provided by changing of a rectified voltage.

Thus, an analytical designing of a valve converter checked by a mathematical modeling provides a required accuracy of a calculation.

CONCLUDING REMARKS

A theory and circuit design of rectifiers and naturally commutating inverters with a phase regulation that are the oldest types of electric energy valve converters are fully exist to the present time. The most in-depth statement of conventional rectification schemes theory simultaneously taking into account all the parameters of a transformer equivalent circuit is represented at [40, 41], taking into account parameters of an input synchronous generator – at [42]. A theory of power rectifiers having networks of longitudinal and cross capacitance compensation is considered at the book [43], having networks of an inherent compensation – at [45]. Many problems of a rectification appliance (a protection, cooling, diagnostics) are given at [46]. Questions of a classification, synthesis of novel rectifier schemes and theory of their external curves at the all diapason of a load changing to a mode of short circuit are considered in the paper [48] maintenance capacious where basic analysis results of 130 rectification schemes are represented. Traditional and novel methods of analysis and synthesis of valve converter schemes of all the classes and novel approaches to energy processes theory at valve systems are considered at monographs [47, 49] and [39], correspondingly and also in the book [50] from which at USSR searches of novel approaches to energy processes analysis methods at valve converters were begun.

A great amount of a rectifier load especially at common electric mains and, hence, it's significant reverse influence to an electric energy quality at a mains led to that, first of all, at countries of Western Europe there were accepted strict standards that strictly regulate a permissible degree of a nonlinear consumer current distortion. In the case of using conventional rectifier schemes with a phase regulation it requires to apply input *LC*-filters that weaken higher harmonics of input rectifier currents to permissible values. In the main cases active filters compensating distortions of an input rectifier current are applied (section 11.2).

Another way of an electromagnetic compatibility improvement of rectifiers and a mains was pointed by a power electronics itself. An appearance of a wide range of fully controlled valves permissible on a cost (transistors, gate turn-off thyristors (GTO)) close to ideal switches on properties allowed to make a novel technology for rectifiers with a leading phase regulation and pulse-width regulation of converted energy parameters that are especially constructive. Ideas of these novel approaches are considered at the section 3.11 and then will be evolved at the chapter 13. Nowadays these innovations are intensively evolved in the direction of a circuit engineering and in the direction of control algorithms perfection. Their final aim is an energy storing. Here there is a great range for a creative work of bachelors, masters of science, engineers, post-graduate students and persons working for doctor's degree.

Chapter 5*

DEVELOPMENT OF THEORETICAL METHODS FOR ANALYSIS OF POWER ELECTRONICS CONVERTERS

I verified the harmony with algebra

A.S. Pushkin

Generalization of direct analysis methods for mathematical models of power electronics converters as differential equation of n-th order is presented in section 5.1.Direct methods for mathematical models as a differential equations of 1-st order (in the state space) are developed in sections 5.2 and 5.3, including ones for 3-phase circuits (section 5.4). The possibilities to achieve precision solutions are shown in section 5.5 and also there are introduction to analysis of mathematical models of converters taking into consideration their discontinuity (differential equation methods and pulse systems theory) in section 5.6.

5.1.** GENERALIZATION OF DIRECT CALCULATION METHODS FOR POWER CONVERTERS

In section 1.5.2 new direct calculation methods for power factors of power electronics converters for 2-nd order circuits were discussed. Here these methods are generalized for electrical circuits of any order. The is also a derivation of productive correlation (5.1.5) and the on its basis a common solution for rms non-sinusoidal currents (voltages) in n-th order circuit is built using ADE1 method conceptions. General solution is derived using ADE2 conceptions and common formulae for circuit complex resistance modulus using differential equation coefficients. The methods of deriving of integral harmonics coefficients are also discussed in this section.

5.1.1. THE FUNDAMENTALS OF ALGEBRAIZATION OF DIFFERENTIAL EQUATION METHOD (ADE1)

THE IDEA OF THE METHOD.
DERIVATION OF PRODUCTIVE CORRELATION

Rms values (norms in $L_2(0, T)$ space, if functions are considered as vectors in infinite-dimensional space [51]) of reaction in linear system with periodical non-sinusoidal coercion without finding instant values could be obtained on the basis of the conception of transfer from differential equations for instant values of variables in steady state mode to linear algebraical equations for rms values of these variables. It is always possible to find a solution of such algebraical equation. As academician S.L. Sobolev defines, direct calculation methods in numerical integration of differential equation combine all the methods of their reduction to algebraical equations. Academician G.E. Puhov states that direct calculation methods in electrical engineering are ones that allow defining not all but some of the currents or voltages [21]. Method offered here uses algebraization of differential equations procedure that has another nature comparing to the discretization methods and only one current or voltage are defined directly. As a result the name of the offered method satisfies both definitions.

Obtained system of algebraic equations after algebraization procedure contains greater number of variables than number of equations. To achieve unambiguous solution we assume that we know the type (sine) of mostly integrated (smoothed) variables similarly to the harmonic balance method in automatic control theory, but here the number of these variables is $n + N - 1$, and their choice has degree of freedom (the assumption number N), but sought variable isn't in this number in contrast to harmonic balance method.

Initial differential equation of the system should be written in input-output shape (one input action and one output variable):

$$\sum_{l=0}^{n} a_l p^l x = \sum_{l=0}^{m} b_l p^l y . \tag{5.1.1}$$

During the algebraization process a calculation of scalar product is required

$$(\overline{x}^{(q_1)}, \overline{x}^{(q_2)}),\ (\overline{y}^{(q_1)}, \overline{y}^{(q_2)}) .$$

Let's determine these correlations that will be called *productive correlations.* Let's express the periodical solution of initial differential equation of any order as Fourier series:

$$x = \sum_{k} \sqrt{2} X_k \sin (k\omega t - \varphi_k). \tag{5.1.2}$$

Firstly let's consider the case where action and reaction are free of dc – components and then extend results for common case. So q-divisible integral of solution (5.1.2) could be written as follows:

$$\overline{x}^{(q)} = \int_0^t dt \underbrace{\int_0^t dt ... \int_0^t}_{q-times} x dt =$$

$$= \begin{cases} \pm \dfrac{1}{\omega^q} \sum_{k=1}^{\infty} \sqrt{2} \dfrac{X_k}{k^q} \cos(k\omega t - \varphi_k), \quad q - \text{odd}, \\ \pm \dfrac{1}{\omega^q} \sum_{k=1}^{\infty} \sqrt{2} \dfrac{X_k}{k^q} \sin(k\omega t - \varphi_k), \quad q - \text{even}. \end{cases} \qquad (5.1.3)$$

As a result constants of integration are not considered because the solution has to belong to the alternating periodic functions class (only steady states are considered).

Similarly the q – divisible derivative of solution (5.1.2) is (differentiation makes the sign of repetition factor index q negative):

$$\overline{x}^{|-q|} = \frac{d^{|q|} x}{dt^{|q|}} = \begin{cases} \pm \omega^{|q|} \sum_{k=1}^{\infty} \sqrt{2} X_k k \cos(k\omega t - \varphi_k), \quad q - \text{odd}, \\ \pm \omega^{|q|} \sum_{k=1}^{\infty} \sqrt{2} X_k k \sin(k\omega t - \varphi_k), \quad q - \text{even}, \end{cases} \qquad (5.1.4)$$

at that positive sign is used at $q = 4n + 1$ in first case and $q = 4n$ in second case, negative sign is used at $q = 4n + 3$ in first case and $q = 4n + 2$ in second case.

Considering the orthogonality of $\sin k\omega t$ and $\cos k\omega t$ for different values of k, it is easy to prove next ratio by direct calculation:

$$\frac{1}{T} \int_0^T \overline{x}^{(q_1)} \overline{x}^{(q_2)} dt = \begin{cases} 0 & \text{for} \quad (q_1 + q_2) \text{ odd}, \\ \pm \left| \overline{X}^{\left(\frac{q_1+q_2}{2}\right)} \right|^2 & \text{for} \quad (q_1 + q_2) \text{ even}, \end{cases} \qquad (5.1.5)$$

where $\left| \overline{X}^{\frac{q_1+q_2}{2}} \right|^2$ – squared rms value of $(q_1 + q_2)/2 > 0$ -times integrated or $(q_1 + q_2)/2 < 0$ -times differentiated reaction, the resulting sign is defined by the signs of efficients. The law of variation for efficients is shown above. In other words, if the difference ($q_2 - q_1$) is divisible by 2 we use negative sign; if the difference is divisible by 4 we use positive sign. This equation is similar to that for scalar product equation of integrals (derivatives) of harmonic function with corresponding geometrical interpretation of perpendicularity of vectors over a plane. If non-sinusoidal functions are represented as a vectors in L_2 - space [51] the zero value of producing equation stands for orthogonality of corresponding vectors of this space.

ALGEBRAIZATION PROCEDURE

The purpose of algebraization is to obtain the system of algebraic equations for rms values of required variable, usually current, and also rms values of current derivative and integral that define rms value if inductive element voltage and capacitor voltage correspondingly.

Firstly, we consider algebraization procedure to obtain rms value of a reaction. First step is integrating of initial differential equation (5.1.1) $n + N - 1$ – times and forming the system of integral equations from N last obtained equations.

$$\begin{array}{c} 1 \\ 2 \\ 3 \\ \cdots \\ N \end{array}\left|\begin{array}{cccccc} a_n & a_{n\text{-}1} & \cdots & a_0 & & \\ & a_n & a_{n\text{-}1} & \cdots & a_0 & \\ & & a_n & a_{n\text{-}1} & \cdots & a_0 \\ \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\ & & & & & \end{array}\right| \left|\begin{array}{c} x \\ \overline{x} \\ x^{(2)} \\ \cdots \\ \overline{x}^{(n+N\text{-}1)} \end{array}\right| =$$

$$= \left|\begin{array}{cccccc} b_m & b_{m\text{-}1} & \cdots & b_0 & & \\ & b_m & b_{m\text{-}1} & \cdots & b_0 & \\ & & b_m & b_{m\text{-}1} & \cdots & b_0 \\ \cdots & \cdots & \cdots & \cdots & \cdots & \cdots \\ & & & & & \end{array}\right| \left|\begin{array}{c} \overline{y}^{(n\text{-}m)} \\ \overline{y}^{(n\text{-}m+1)} \\ \overline{y}^{(n\text{-}m+2)} \\ \cdots \\ \overline{y}^{(n\text{-}m+N+1)} \end{array}\right|. \qquad (5.1.6)$$

The number N that is called *degree of assumption* in ADE method, is selected a priori to satisfy monoharmonic criteria (filter hypothesis) for variables $\overline{x}^{(N)}, \overline{x}^{(N+1)}, ..., \overline{x}^{(n+N-1)}$. This allows easy determination of values of "redundant" variables. The term "redundant" means that number of unknowns in (5.1.6) is $n + N$ when number of equations is N. The second step is squaring each equation of (5.1.6) and then averaging the result over a period considering productive equation (5.1.5); in such a way a system of algebraic equations are obtained:

$$\begin{array}{c} 1 \\ 2 \\ 3 \\ \dots \\ N \end{array}\left|\begin{array}{cccccc} a_n^2 & a_{n\text{-}1,k}^2 & \dots & a_0^2 & & \\ & a_n^2 & a_{n\text{-}1,k}^2 & \dots & a_0^2 & \\ & & a_n^2 & a_{n\text{-}1,k}^2 & \dots & a_0^2 \\ \dots & \dots & \dots & \dots & \dots & \dots \\ & & & & & \end{array}\right| \left|\begin{array}{c} X^2 \\ \overline{X}^2 \\ (X^{(2)})^2 \\ \dots \\ (\overline{X}^{(n+N\text{-}1)})^2 \end{array}\right| =$$

$$= \left|\begin{array}{ccccc} b_m^2 & \dots & b_0^2 & & \\ & b_m^2 & \dots & b_0^2 & \\ & & b_m^2 & \dots & b_0^2 \\ \dots & \dots & \dots & \dots & \dots \\ & & & & \end{array}\right| \left|\begin{array}{c} (\overline{Y}^{(n\text{-}m)})^2 \\ (\overline{Y}^{(n\text{-}m+1)})^2 \\ (\overline{Y}^{(n\text{-}m+2)})^2 \\ \dots \\ (\overline{Y}^{(n\text{-}m+N-1)})^2 \end{array}\right| \quad (5.1.7)$$

or

$$\mathbf{A_2 X_2} = \mathbf{B_2 Y_2}, \quad (5.1.8)$$

where

$$\begin{aligned} \mathbf{X} &= [X^2, (\overline{X})^2, (\overline{X}^{(2)})^2, \dots, (\overline{X}^{(n+N-1)})^2]^t, \\ \mathbf{Y} &= [(\overline{Y}^{(n-m)})^2,\ (\overline{Y}^{(n-m-1)})^2, \dots, (\overline{Y}^{(n-m+N-1)})^2]^t; \end{aligned} \quad (5.1.9)$$

$\mathbf{A}_2$, $\mathbf{B}_2$ – square banded matrixes of dimension N x $(n + N)$ и N x $(n + N)$.

Let's derive the rule of obtaining the coefficients of $\mathbf{A}_2$ and $\mathbf{B}_2$ matrixes. Accordingly to the producing equation at the integration of the squared equations of the system (5.1.6) considering equality

$$\left(\sum_{i=1}^{n} a_i \overline{x}^{(n-i)}\right)^2 = \sum_{i=1}^{n} a_i^2 (\overline{x}^{(n-i)})^2 + 2 \sum_{\substack{i,j=1 \\ i \neq j}}^{n} \overline{x}^{(n-i)} \overline{x}^{(n-j)} a_i a_j \quad (5.1.10)$$

we will have non zero value for only those cross products that are formed by efficients with indices with even number difference (2, 4, 6, …). Then

$$
\begin{aligned}
&a_{0k}^2 = a_0^2,\\
&a_{1k}^2 = a_1^2 - 2a_2 a_0,\\
&a_{2k}^2 = a_2 - 2a_3 a_1 + 2a_4 a_0,\\
&a_{3k}^2 = a_3^2 - 2a_4 a_2 + 2a_5 a_1 - 2a_6 a_0,\\
&a_{4k}^2 = a_4^2 - 2a_5 a_3 + 2a_6 a_2 - 2a_7 a_1 + 2a_8 a_0,\\
&\dots\dots\dots\dots\dots\dots\dots\dots\dots\dots\dots\dots\\
&a_{n-4,k}^2 = a_{n-4}^2 - 2a_{n-3}a_{n-5} + 2a_{n-2}a_{n-6} - 2a_{n-1}a_{n-7} + 2a_n a_{n-8},\\
&a_{n-3,k}^2 = a_{n-3}^2 - 2a_{n-2}a_{n-4} + 2a_{n-1}a_{n-5} - 2a_n a_{n-6},\\
&a_{n-2,k}^2 = a_{n-2}^2 - 2a_{n-1}a_{n-3} + 2a_n a_{n-4},\\
&a_{n,k}^2 = a_n^2
\end{aligned}
\tag{5.1.11}
$$

or in the general case

$$a_{n-l,k}^2 = a_{n-l}^2 + 2\sum_{r=1}^{n-l}(-1)^r a_{n-l+r}a_{n-l-r}, \tag{5.1.12}$$

at that $a_{n-1+r} = 0$ for $0 > \left(n-1 \pm r\right) > n$. The elements of $\mathbf{B}_2$ matrix are found in the same manner:

$$b_{m-1,k}^2 = b_{m-1}^2 - 2\sum_{r=1}^{m+1-l}(-1)^r b_{m-l+r}b_{m-l+r}. \tag{5.1.13}$$

To find unambiguous solution for system of equations (5.1.7) it is necessary to have square A2 matrix. To satisfy this condition as a third step we will transfer all "redundant" items that contain $\overline{\mathbf{X}}^{(q)}$ at $q \ge N$, to the right part of the equation, considering them known monoharmonic functions on condition that N-times integrating (smoothing) suppresses all higher harmonics in reaction

$$\overline{X}^{(q)} \approx \overline{X}_1^{(q)} = \frac{X_1}{\omega^q} = \frac{Y_1}{Z_1\omega^q}. \tag{5.1.14}$$

Here Z_1 – the modulus of 1-st harmonic transfer ratio (in case of channel action – reaction of the system), equal (in case of channel voltage - current) to the modulus of complex resistance for 1-st harmonic component.

Then system (5.1.7) will be written as

$$
\begin{array}{c}
N=1\\ N=2\\ N=3\\ \dots\\ N
\end{array}
\left|\begin{array}{ccccc}
a_n^2 & a_{n-1,k}^2 & a_{n-2,k}^2 & \dots & a_{n-N+1,k}^2\\
 & a_n^2 & a_{n-1,k}^2 & \dots & a_{n-N,k}^2\\
 & & a_n^2 & \dots & a_{n-N-1,k}^2\\
\dots & \dots & \dots & \dots & \dots\\
 & & & &
\end{array}\right|
\left|\begin{array}{c}
(X)^2\\ (\overline{X})^2\\ (\overline{X}^{(2)})^2\\ \dots\\ (\overline{X}^{(n+N-1)})^2
\end{array}\right| =
$$

$$
=\left|\begin{array}{c}
\sum\limits_{l=1}^{m+1} b_{l,k}^2(\overline{Y}^{(n-m+l-1)})^2-\sum\limits_{l=N}^{n} a_{n-l,k}^2(\overline{X}_1^{(l)})^2\\
\sum\limits_{l=1}^{m+1} b_{l,k}^2(\overline{Y}^{(n-m+l)})^2-\sum\limits_{l=N-1}^{n} a_{n-l,k}^2(\overline{X}_1^{(l+1)})^2\\
\sum\limits_{l=1}^{m+1} b_{l,k}^2(\overline{Y}^{(n-m+l+1)})^2-\sum\limits_{l=N-2}^{n} a_{n-l,k}^2(\overline{X}_1^{(l+2)})^2\\
\dots\\
\sum\limits_{l=1}^{m+1} b_{l,k}^2(\overline{Y}^{(n-m+N-2)})^2-\sum\limits_{l=1}^{n} a_{n-l,k}^2(\overline{X}_1^{(l+N-1)})^2
\end{array}\right|. \qquad (5.1.15)
$$

The solution for square rms value of reaction at N – th level of assumption is obtained using Kramer’s rule:

$$
X^2=\frac{1}{a_n^{2N}}\left(-1\right)^{i+1}\Delta_{i1}\left[\sum_{l=1}^{m+1} b_{l,k}^2(\overline{Y}^{(n-m+l+i-2)})^2-\right.
$$
$$
\left.-\sum_{l=N-i+1}^{n} a_{n-l,k}^2(\overline{X}^{(l+i-1)})^2\right]. \qquad (5.1.16)
$$

Here $(-1)^{i+1}\Delta_{i1}$ - adjuncts of matrix $\mathbf{A}_2$ (the matrix is edged for different N in general equation for this matrix in (5.1.15) for its expansion for the elements of the firs column): $a_n^{2N}=(a_n^2)^N$ - determinant of triangular matrix $\mathbf{A}$.

Considering (5.1.14) and the equation for rms values of action integrals using integral harmonic coefficients (1.5.24) the solution (5.1.16) transfers to:

$$\mathbf{X}_2 = \frac{1}{a_n^{2N}} \sum_{i=1}^{N} (-1)^{i+1} \Delta_{i1} \times$$

$$\times \left[\sum_{l=0}^{m} \frac{b_{l,k}^2 \left(Y_1\right)^2 [1 + (\overline{K}_{\Gamma}^{(q)})^2]}{\omega^{2q}} - \sum_{l=N-i+1}^{n} \frac{a_{n-l,k}^2}{\omega^{2(l+i-1)}} \frac{Y_1^2}{Z_1^2} \right] =$$

$$= \frac{1}{a_n^{2N}} \sum_{l=1}^{N} (-1)^{i+1} \Delta_{i1} \times \qquad (5.1.17)$$

$$\times \left[\sum_{l=1}^{m+1} \frac{b_{l,k}^2 Z_1^2 [1 + (\overline{K}_{\Gamma}^{(q)})^2]}{\omega^{2q}} - \sum_{l=N-i+1}^{n} \frac{a_{n-l,k}^2}{\omega^{2(l+i-1)}} \right] \frac{Y_1^2}{Z_1^2} = \frac{X_1^2}{\nu_x^2},$$

where $q = n - m + l + i$; ν_x – reaction distortion coefficient.

Thus, the equation for rms value of the reaction obtained that uses rms value of first harmonic component of the reaction and reaction (current) distortion coefficient that could be found using system parameters and the set of integral parameters of the action (integral harmonic coefficients) in close analytical form. Methods of how to calculate integral harmonic coefficients for non-sinusoidal action are considered in section 5.1.4.Methods of how to calculate the modulus of input-output (action-reaction) coefficient for 1-st (or any other) harmonic component are considered in section 5.1.3.

Derived algebraization procedure and the solution for rms value of the reaction shows that in order to find the solution for rms value of q – multiply integral or derivative of the reaction $\overline{X}^{(q)}$ ($q > 0$ for integral and $q < 0$ for derivative) system of the equations (5.1.15) is formed in such a way to have reaction component vector $\mathbf{X}_{2q} = [(\overline{X}_1^{(q)})^2, (\overline{X}^{(q+1)})^2, (\overline{X}^{(q+2)})^2, \ldots]^t$, at that action vector is $\mathbf{Y}_{2q} = [(\overline{Y}^{(n-m+q)})^2, (\overline{Y}^{(n-m+q+1)})^2, \ldots]^t$. Then the solution for $\overline{X}^{(q)}$ is obtained similarly to (5.1.16) with correspondent correction of integration indices in shown variables:

$$(\overline{X}^{(q)})^2 = \frac{1}{a_n^{2N}} \sum_{i=1}^{N} (-1)^{i+1} \Delta_{i1} \times$$

$$\times \left[\sum_{l=1}^{m+1} b_{l,k}^2 (\overline{Y}^{(n+q-1)})^2 - \sum_{l=N-i+1}^{n} a_{n-l,k}^2 (\overline{X}^{(l+i+q-1)})^2 \right]. \qquad (5.1.18)$$

The same correction has to be applied to (5.1.17) if named form of the solution is used.

Derived common equations are utilized in the special problem - oriented software [21] that allows computing systems which order limited only by CPU performance. At the same time most of engineering tasks have low dimension order. In these cases using concrete formulas one can use a simple microcalculator to perform the tasks.

For power conversion devices the parameters of continuous (filtering) part of the system are usually within the bounds that provide filtering purposes for output coordinate. In these cases the assumption of 1-st or 2-nd order (single or double stage mathematical integration of output coordinate) provides calculation error not more that several percents on the condition that there is no resonant phenomena occurs.

5.1.2. SECOND VERSION OF METHOD OD ALGEBRAIZATION OF DIFFERENTIAL EQUATION (ADE2)

Further development of direct calculation methods for rms values of reactions in systems with non-sinusoidal action by the algebraization of differential equations is based on the orthogonality of two instant components of the reaction. They are 1-st (fundamental) harmonic component and higher harmonic components of the reaction. Both these components (reactions) are produced by two analogous orthogonal components of the action. This property defines correspondent geometrical addition of components rms values to a total rms value of the reaction:

$$X = \sqrt{X_1^2 + X_b^2} \quad \text{at} \quad x = x_1 + x_b. \tag{5.1.19}$$

It is easy to find rms value of 1-st harmonic value of the reaction; the generalization of these procedure is shown in section 5.1.3.

It is easy to find the rms value of the first harmonic component of the reaction; general description of this procedure could be found in section 5.1.3. Parameters calculation relative to first harmonic component is usually done before the calculation relative to real (nonsinusiodal) shape of the system parameters. In this case the calculation of whole value of the reaction by ADE1 method is redundant (due to the presence of first harmonic component rms value in the overall result), and therefore is rather complicated. In ADE2 method the purpose is to find the rms value of the component that is consisted of higher harmonic components (reaction without first harmonic component); this makes the solution less complicated.

The procedure of calculation of X_b is identical to that described in ADE1 method, but the assumptions are:

1. One should write canonical differential equation of n-th order for higher harmonics of the reaction on the condition that the action y_b has only higher harmonic components (1-st harmonic components is absent):

$$\sum_{l=0}^{n} a_l p^l x_b = \sum_{l=0}^{m} b_1 p^l y_b. \tag{5.1.20}$$

2. TH equation of n-th order should be integrated $(n + N - 1)$ – times, where N – assumption number, and n – the order of the equation system composed of integral equations per se.

3. Obtained system is further algebraized by squaring each equation integrating in within the period. Producing equations (5.1.5) that are used to calculate the integrals of cross multiplied variables of integral equations remain valid for X_b.

4. The compatibility of the algebraic equation system is provided by equating all the elements of $N + 1$ – th column and all other following elements of left coefficient matrix with zero; this action makes the matrix square. Named zero placing is based on the following conception: integration decreases high frequency component of the reaction and rms value of N – times smoothed high frequency reaction tends to zero.

Considering above mentioned items the system of algebraic equations (5.1.7) transforms to:

$$\begin{array}{c} 1 \\ 2 \\ 3 \\ \dots \\ N \end{array}
\left|\begin{array}{cccccc}
a_n^2 & a_{n-1,k}^2 & \dots & a_0^2 & & \\
 & a_n^2 & a_{n-1,k}^2 & \dots & a_0^2 & \\
 & & a_n^2 & a_{n-1,k}^2 & \dots & a_0^2 \\
\dots & \dots & \dots & \dots & \dots & \dots \\
 & & & & & a_n^2
\end{array}\right|
\left|\begin{array}{c}
X_b^2 \\
(\bar{X}_b)^2 \\
(\bar{X}_b^{(2)})^2 \\
\dots \\
(\bar{X}_b^{(n+N-1)})^2
\end{array}\right| =$$

$$
= \begin{vmatrix} b_m^2 & \dots & b_0^2 & & \\ & b_m^2 & \dots & b_0^2 & \\ & & b_m^2 & \dots & b_0^2 \\ \dots & \dots & \dots & \dots & \dots \\ & & & & \end{vmatrix} \begin{vmatrix} (\overline{Y}_b^{(n-m)})^2 \\ (\overline{Y}^{(n-m+1)})^2 \\ (\overline{Y}^{(n-m+2)})^2 \\ \dots \\ (\overline{Y}^{(n-m+N-1)})^2 \end{vmatrix}. \tag{5.1.21}
$$

Solution for norm X_b is obtained for different N on the condition that all the elements of matrix $\mathbf{A}_2$ (the dimension of this matrix is defined by N) that do not belong to its bordered part by (5.1.8) are absent in the solution because their correspondent solutions $\overline{X}_b^{(N)}, \overline{X}_b^{(N+1)}, ..., X_b^{(n+N-1)}$ are equal to zero accordingly to the assumptions made above.

As for ADE1 method if the filter hypothesis is satisfied these assumptions are hold true and $\overline{X}_b^{(q)} = 0$ because of sinusoidal shape of $\overline{X}^{(q)}$.

Then rms value of energetic disturbance of the reaction at N - th level of assumption from (5.1.21) by Kramer rule is equal to

$$
\begin{aligned} X_b^2 &= \frac{1}{a_n^{2N}} \left\{ \sum_{i=1}^{N} (-1)^{i+1} \Delta_{i1} \sum_{l=0}^{m} b_{lk}^2 (\overline{Y}^{(n-l+i-1)})^2 \right\} = \\ &= \frac{Y_1^2}{a_n^{2N}} \left\{ \sum_{i=1}^{N} (-1)^{i+1} \Delta_{i1} \sum_{l=0}^{m} \frac{b_{lk}^2 (\overline{K}_\Gamma^{(n-l+i-1)})^2}{\omega^{2(n-l+i-1)}} \right\}. \end{aligned} \tag{5.1.22}
$$

Obtained equation is simpler than equation (5.1.17) for rms value of whole reaction.

Common equations for algebraic adjuncts Δ_{i1} expressed by coefficients of initial equation depending on the assumption number N and system order n are obtained in [21].

Thus, ADE2 method is characterized by commonality and compactness of the results that provides easy software realization of (5.1.22) formulae even with small microcontroller. Another advantage of ADE2 method is connected with the conception with underlies the basis of this method i.e. natural separation of the motion into “slow” (1- st harmonic component) and “fast” (higher) harmonics. This simplifies solution in those cases when differential equations of the object (its equivalent circuit) are different for these two components, i.e. in situation where it is impossible to use ADE1 method. Such situation is observed for model of asynchronous machine.

And, finally, there is another possibility to ease the calculation and simplify final equations for rms values of sought variables – it is their expression using open algebraic adjuncts Δ_{i1} that are the coefficients of differential equations. This possibility utilizes a new assumption connected with small values of circuit active resistance for equivalent circuit for higher harmonics comparing to reactive ones. For converters where “carrier” frequency (commutation frequency) is much greater than output voltage frequency (naturally of force commutated direct frequency converters, PWR and PWM inverters) such a phenomenon usually takes place.

5.1.3. DERIVATION OF COMMON EQUATIONS FOR COMPLEX RESISTANCE MODULUS

Equations for rms values of the currents in ADE1 method contain modulus of complex resistance of the circuit relative to 1-st harmonic components of applied voltage. Let’s formalize the its calculation principles using primary circuit parameters reasoning from the same mathematical model of the system viz from the differential equation written in “input-output” manner. Generally, the object doesn’t have to have electrical nature. In this case the modulus of complex resistance acts as a proportional coefficient between 1-st harmonic component of the reaction and 1-st harmonic component of the action, but here we will name this coefficient the modulus of complex resistance for convenience.

F i r s t m e t h o d. The equation of equivalent of complex resistance (for any harmonic component) could be obtained from differential equation (5.1.1) considering $p = j\omega$:

$$\underline{Z}(j\omega) = \frac{Y(j\omega)}{X(j\omega)} = \frac{\sum\limits_{l=0}^{m} b_l (j\omega)^l}{\sum\limits_{l=0}^{n} a_l (j\omega)^l} = \frac{A_n + jB_n}{C_n + jD_n}. \qquad (5.1.23)$$

Then the modulus of complex coefficient

$$\left|\underline{Z}_n\right|^2 = \frac{A_n^2 + B_n^2}{C_n^2 + D_n^2}, \qquad (5.1.24)$$

and its phase

$$\varphi_n = \operatorname{arctg}\frac{B_n C_n - A_n D_n}{A_n C_n + B_n D_n}. \qquad (5.1.25)$$

Equations for A_n, B_n, C_n, D_n are formed by sums of correspondent coefficients a_l, b_l set up by the induction method (for odd n on the left, for even n on the right):

$$A_n = \sum_{l=0}^{n} a_l (j\omega)^l, \qquad A_n = \sum_{l=0}^{n-1} a_l (j\omega)^l,$$

l – even $\qquad$ l – even

$$B_n = \sum_{l=1}^{n-1} a_l (j\omega)^l, \qquad B_n = \sum_{l=1}^{n} a_l (j\omega)^l,$$

l – odd $\qquad$ l – odd $\qquad$ (5.1.26)

$$C_n = \sum_{l=0}^{n} b_l (j\omega)^l, \qquad C_n = \sum_{l=0}^{n-1} b_l (j\omega)^l,$$

l – even $\qquad$ l – even

$$D_n = \sum_{l=1}^{n-1} b_l (j\omega)^l, \qquad D_n = \sum_{l=1}^{n} b_l (j\omega)^l,$$

l – odd $\qquad$ l – odd

Degrees $(j\omega)^l$ have repetition period for its sign, and this period is divisible by 4:

$$\begin{aligned} &(j\omega)^1 = j\omega, && (j\omega)^5 = j\omega^5, \\ &(j\omega)^2 = -\omega^2, && (j\omega)^6 = -\omega^6, \\ &(j\omega)^3 = -j\omega^3, && (j\omega)^7 = -j\omega^7, \\ &(j\omega)^4 = \omega^4, && (j\omega)^8 = \omega^8. \end{aligned} \qquad (5.1.27)$$

For n = 1...5 (that are mostly spoken about) the coefficients A_n, B_n, C_n, D_n have following values:

n = 1, $A_1 = a_0$, $\quad B_1 = \omega^2 a_1$, $\quad C_1 = b_0$, $\quad D_1 = \omega\, b_1$,

n = 2, $A_2 = A_1 = \omega^2$, $\quad B_2 = B_1, C_2 = C_1 - \omega^2 B_2$, $\quad D_2 = D_1$,

n = 3, $A_3 = A_2$, $\quad B_3 = B_2 - \omega^3 a_3$, $C_3 = C_2$, $\quad D_3 = D_2 - \omega^3 B_3$,
(5.1.28)

$n = 4,\ A_4 = A_2 + \omega^2 a_4,\ B_4 = B_3, C_4 = C_3 + \omega^4 B_4,\quad D_4 = D_3,$

$n = 5,\ A_5 = A_4,\qquad B_5 = B_4 - \omega^5 a_5, C_5 = C_4,\quad D_5 = D_4 - \omega^5 B_5.$

Equations (5.1.25) and (5.1.28) are the basis of applied software packet RV1 (Fourier series), which is use to verify accuracy of calculations that are performed with ADE method [21] .

S e c o n d m e t h o d. Considered method of complex resistance calculation for 1-st or any other (k-th) harmonic component (by substitution of $j\omega$ by $jk\omega$) provides an opportunity not only to define modulus of complex resistance using differential equation coefficients, but also phase pf the resistance and therefore its phase shift of correspondent current harmonic components relative to voltage one. In those cases when current phase value is unnecessary datum a much simpler method to define modulus of complex resistance could be proposed. This method doesn't utilize differential equation, but it works with algebraic equation obtained from differential one by ADE1 method. First equation of algebraic equations system (5.1.7) considering (5.1.14) and (5.1.24) for $\overline{K}_{\Gamma}^{(q)} = 0$ (q=1,2, …, q) could be written as ($n = m$):

$$\left(a_n^2 + \frac{a_{n-1,k}^2}{\omega^2} + \frac{a_{n-2,k}^2}{\omega^4} + \ldots + \frac{a_{1,k}^2}{\omega^{2(n-1)}} + \frac{a_0^2}{\omega^{2n}} \right) X_1^2 =$$

$$= \left(b_n^2 + \frac{b_{n-1,k}^2}{\omega^2} + \frac{b_{n-2,k}^2}{\omega^4} + \ldots + \frac{b_{1,k}^2}{\omega^{2(n-1)}} + \frac{b_0^2}{\omega^{2n}} \right) Y_1^2. \qquad (5.1.29)$$

From above equation we obtain the equation of (squared) complex resistance modulus for 1-st harmonic component:

$$Z_1^2 = \frac{Y_1^2}{X_1^2} = \sum_{l=0}^{n} \frac{a_{n-l,k}^2}{\omega^{2l}} \Big/ \sum_{l=0}^{n} \frac{b_{n-l,k}^2}{\omega^{2l}}. \qquad (5.1.30)$$

The comparison of this equation with formulae (5.1.24) of previous method shows significant decrease in number of multiplication and adding efforts, therefore this formulae is preferable for calculations where current phase value is unnecessary.

D e f i n i t i o n o f $\mathrm{Re}\,Z_k$. The ADE method by allowing to directly express rms current and voltage values using circuit and action parameters provider an opportunity to calculate apparent power directly. Active power at the branchy electrical circuit in this case coul be calculated from energy conversation law:

$$P_{\text{in}} = \sum_j I_j^2 R_j, \tag{5.1.31}$$

where I_j - rms current values that flow through the branches with ohmic resistance R_j.

To define the coefficient of active power output $\eta_в$ it is necessary to know active power for 1-st (useful) or correspondent k-th harmonic component. In this case the known equation from electrical circuitry theory is used:

$$P_k = \frac{1}{2} I_k^2 \operatorname{Re} Z_k, \tag{5.1.32}$$

where $\operatorname{Re} Z_k$ is calculated from (5.1.23):

$$\operatorname{Re} Z_k = \frac{A_n C_n + B_n D_n}{C_n^2 + D_n^2}. \tag{5.1.33}$$

5.1.4. METHODS TO CALCULATE INTEGRAL HARMONIC COEFFICIENTS

Integral harmonic coefficients accordingly to (1.5.24) directly connected with rms values of integrals of nonsinusoidal function (action); that is why our purpose is to calculate rms values of these integrals for given function. Direct calculation of rms value by integrating analyzed function is possible, but usually when action has simple shape in other cases it is necessary to develop special calculation methods.

Harmonic synthesis method. Calculation of rms values of integrals of action accordingly to (1.5.24) comes to the closed calculation of harmonic components sum

$$\sum_{k=1}^{\infty} \left(\frac{U_k}{U_1 k^q} \right)^2 = \left(\frac{\overline{U}^{(q)}}{U_1} \right)^2 \omega^{2q}, \tag{5.1.34}$$

this requires knowledge of relative harmonic composition of the action and defined by the possibility to synthesize obtained series in closed form [52]. The simplest case is observed when shape of instance is fixed and doesn't depend on the regulation; in the method considered below this corresponds to the case with voltage source inverter with amplitude regulation (refer to chapter 8).

Thus, for pulse-amplitude modulation relative harmonic component in the output voltage of single-phase inverter is defined by the equation $U_k / U_1 = 1/k$, where k – odd harmonics, and do not divisible by 3 in case of 3 – phase bridge inverter. Then for single phase inverter considering known equations we can write following equation

$$\sum_{k=1}^{\infty}\left(\frac{U_k}{U_1 k^q}\right)^2 = \sum_{k=1}^{\infty}\left(\frac{1}{k^{q+1}}\right)^2 = \sum_{n=1}^{\infty}\frac{1}{(2n-1)^{2(q+1)}} = \\ = \left(\frac{\pi^{2(q+1)}2^{2(q+1)}}{2[2(q+1)]}\mathrm{B}_{q+1}\right)^{\frac{1}{2}}, \tag{5.1.35}$$

where B_{q+1} – Bernoulli numbers, at that $\mathrm{B}_1 = 1/6$, $\mathrm{B}_2 = 1/30$, $\mathrm{B}_3 = 1/42$. Then

$$(\bar{U})^2 = \frac{\pi}{4\sqrt{6}}\left(\frac{U_1}{\omega}\right)^2 = 1{,}0073\left(\frac{U_1}{\omega}\right)^2, \quad K_{\text{г}} = 0{,}08544;$$
$$(\bar{\bar{U}})^2 = \frac{\pi^3}{24}\left(\frac{U_1}{\omega^2}\right)^2\sqrt{\frac{63}{105}} = 1{,}0072\left(\frac{U_1}{\omega^2}\right)^2, \quad \bar{\bar{K}}_{\text{г}} = 0{,}02683; \tag{5.1.36}$$
$$(\bar{U}^{(3)})^2 = \frac{\pi^4}{48}\sqrt{\frac{51}{105}}\left(\frac{U_1}{\omega^3}\right)^2 = 1{,}000077\left(\frac{U_1}{\omega^3}\right)^2, \quad \bar{K}_{\text{г}}^{(3)} = 0{,}0088.$$

In case of 3-phase inverter the harmonic components divisible by 3 are absent, then

$$\sum_{k=1}^{\infty}\left(\frac{U_k}{U_1 k^q}\right)^2=\sum_{k=1}^{\infty}\left(\frac{1}{k^{q+1}}\right)^2=\sum_{n=1}^{\infty}\frac{1}{(2n-1)^{2(q+1)}}-$$

$$-\sum_{n=1}^{\infty}\frac{1}{\left[3(2n-1)\right]^{2(q+1)}}=\left[1-\frac{1}{3^{2(q+1)}}\right]\sum_{n=1}^{\infty}\frac{1}{(2n-1)^{2(q+1)}}= \tag{5.1.37}$$

$$=C_q\sum_{n=1}^{\infty}\frac{1}{(2n-1)^{2(q+1)}},$$

i.e. this sum also comes to the previous one, at that $C_q = 80/81$, $C_q = 728/729$, $C_q = 6560/6561$ for $q = 1, 2, 3$ correspondingly, and $\overline{K}_\text{г} = 0{,}0849$, $\overline{\overline{K}}_\text{г} = 0{,}0268$, $\overline{K}_\text{г}^{(3)} = 0{,}0087$.

Values of integral coefficients of harmonics of the series of typical input (output) currents (voltages) of valve converters are placed in Table 5.1.1.

It is possible to show that for controlled rectifier integral harmonic coefficients of rectified voltage defined relative o dc component of rectified voltage are calculated using formulae (developed by M.V. Martinovich)

$$\overline{K}_\text{г}^{(q)} = \sqrt{A^2 + B^2 \operatorname{tg}^2 \alpha},$$

where coefficients A and B for 3-phase and 6-phase rectifiers are taken from the Table. 5.1.2.

In case of complex spectrums, as it takes place in PWM autonomous inverters (refer to chapter 8), the summarizing of harmonic components is performed using PC, at that harmonic components are defined with help of software based on the method of academician A.N. Krilov; in case of step function, as in inverter, harmonics are defined by the values of function steps at the points of discontinuity. This method is convenient to define $\overline{K}_\text{г}^{(q)}$ in follow-up control algorithms (see chapter 12).

Table 5.1.1

Integral harmonic coefficients

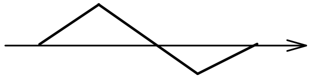
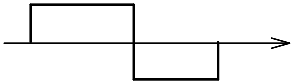
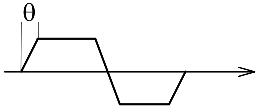
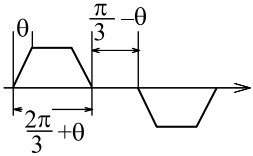
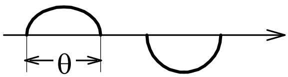
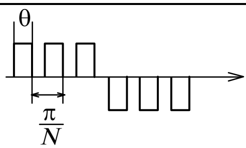
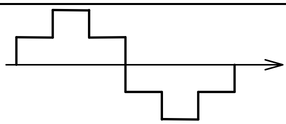
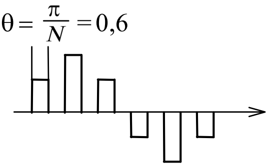
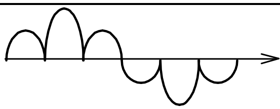
Variable parameter	Function	$\tilde{K}_{\text{г}}$	$K_{\text{г}}$	$\overline{K}_{\text{г}}$	$\overline{K}_{\text{г}}^{(2)}$
—		0,483	0,121	0,038	0,012
—		—	0,45	0,121	0,038
$\theta=\frac{\pi}{6}$ $\theta=\frac{\pi}{3}$		1,377 0,77	0,319 0,225	0,106 0,075	0,034 0,025
$\theta=\frac{\pi}{6}$ $\theta=\frac{\pi}{3}$		1,722 1,077	0,556 0,45	0,249 0,218	0,122 0,108
$\theta=\frac{\pi}{2}$ $\theta=\frac{\pi}{3}$		2,108 4,016	0,623 0,974	0,202 0,288	0,067 0,091
$N=3,\ \theta=\frac{\pi}{6}$ $N=6,\ \theta=\frac{\pi}{12}$		—	1,111 1,141	0,228 0,158	0,06 0,042
—		—	0,31	0,046	0,009
$N=3$		—	0,848	0,141	0,027
—		0,547	0,104	0,02	0,004

Table 5.1.2

Coefficients for 3-phase and 6-phase rectifiers

m	q	A	B
3	0	$6.76 \cdot 10^{-2}$	$8.57 \cdot 10^{-1}$
	1	$7.04 \cdot 10^{-3}$	$6.76 \cdot 10^{-2}$
	2	$7.73 \cdot 10^{-4}$	$7.04 \cdot 10^{-3}$
	3	$8.52 \cdot 10^{-5}$	$7.73 \cdot 10^{-4}$
6	0	$3.52 \cdot 10^{-3}$	$1.89 \cdot 10^{-1}$
	1	$9.22 \cdot 10^{-5}$	$3.52 \cdot 10^{-3}$
	2	$2.40 \cdot 10^{-6}$	$9.22 \cdot 10^{-5}$
	3	$2.52 \cdot 10^{-7}$	$2.40 \cdot 10^{-6}$

Fourier integral method. The procedure of summarizing of infinite series to calculate rms value of action integral, that is considered above, could be substituted by the procedure of calculation of definite integral in case if Fourier series is substituted by Fourier spectral density and Raleigh formulae:

$$\overline{U}^{(q)} = \frac{1}{T}\int_0^T (\overline{u}^{(q)})^2 dt = \frac{1}{T}\frac{1}{2\pi}\int_{-\infty}^{\infty} \overline{u}_T^{(q)}(j\omega)(\overline{u}_T^{(q)}(j\omega))^* d\omega. \qquad (5.1.38)$$

Spectral density of action integral could be found using spectral density of the action and theorem of integral Fourier transform about function integral representation via unction representation:

$$\overline{u}(j\omega) = \frac{u(j\omega) - u(0)}{j\omega},$$

$$\overline{\overline{u}}(j\omega) = \frac{\overline{u}(j\omega) - \overline{u}(0)}{j\omega} = \frac{u(j\omega)}{(j\omega)^2} - \frac{u(0)}{(j\omega)^2} - \frac{\overline{u}(0)}{j\omega}, \qquad (5.1.39)$$

$$\overline{u}^{(q)}(j\omega) = \frac{u(j\omega)}{(j\omega)^q} - \sum_{l=0}^{q-1} \frac{u^{-l}(0)}{(j\omega)^{(q-1)}}.$$

Application of this method is rational when different PWR algorithms realized in the inverter (see chapter 8); in this situation it is possible to obtain quite compact expressions for spectral density of the voltage [21], especially when common voltage vector is employed. As q - increases the difficulty of calculations increases too because accordingly to the (5.1.39) the number of integrals $(q+1)^2$ increases, i.e. the number of parts of (5.1.38) becomes larger. But (5.1.36) and (5.1.37) show that integral harmonics coefficients tend to zero as q increases.

The method of productive equation. Equation that used to define rms values of action integrals could be transformed with help of productive equation (5.1.5):

$$\overline{U}^{(q)} = \frac{1}{T}\int_0^T \overline{u}^{(q)}\overline{u}^{(q)}\, dt = \frac{1}{T}\int_0^T \overline{u}^{(2q)} u\, dt = \frac{1}{T}\int_0^T \overline{u}^{(2q+1)} u'\, dt. \qquad (5.1.40)$$

Considering specific character of inverter output voltage shape, i.e. its step character whose derivative is equal to the aggregate of delta-functions at points of discontinuity, the equation (5.1.40) on the basis of delta function filtering properties could be rewritten as:

$$\overline{U}^{(q)} = \frac{1}{T}\int_0^T \overline{u}^{(2q+1)} \sum_{t_r} \delta(t_r)\, dt = \frac{1}{T}\sum_{t_r} \overline{u}^{(2q+1)}(t_r)\, c_r. \qquad (5.1.41)$$

Thus, knowing the values of the function steps c_r at the points discontinuity (in the linear inverter mode the values of all steps are equal to the inverter output voltage)and the values of discrete (2q+1)-times integrated output voltage, the rms values of the function integral could be found. Discrete voltage values $\overline{u}^{(2q+1)}$ are precisely found using difference equations method (refer to section 5.6) for equal sections between adjacent t_r (PWR algorithms) [21] or in more difficult way when named intervals are modulated (PWM algorithms). If $\overline{u}^{(2q+1)}$ is sinusoidal then the calculations using (5.1.41) are elementary.

5.2. DEVELOPMENT OF DIRECT CALCULATION METHODS IN STATE SPACE RELATIVE TO FIRST HARMONIC COMPONENT

Direct calculation methods considered above were built for mathematical model of converter such as *single input – single output (SISU)*. Model of the converter for instant value of variable of interest is a differential equation of n-th order. After the algebraization of such differential equation an algebraic equation was obtained for correspondent integral value of the sought variable (rms value, average value, first harmonic component). Each new integral variable requires to repeat the algebraization procedure again.

Multiply algebraization procedure could be avoided in n-th order valve converter system if mathematical models of the converter such as *single input – multiply output (SIMU)* or *multiply input – multiply output (MIMU)* are employed.

This requires to write mathematical model in state space, i.e. as the system of the differential equations such as

$$\dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}\mathbf{u}, \qquad (5.2.1a)$$

$$\dot{\mathbf{y}} = \mathbf{C}\mathbf{x} + \mathbf{D}\mathbf{u}, \qquad (5.2.1б)$$

where $\mathbf{x} = (x_1, x_2, x_3, ...x_n)^t$ – variables state vector (currents of inductances and voltages of capacitors); $\mathbf{u} = (u_1, u_2, u_3, ...u_n)^t$ – vector of input actions (e.m.f. of power sources, currents values of current sources); **A** –matrix of the system that in general case looks as

$$\mathbf{A} = \begin{vmatrix} a_{11} & a_{12} & a_{13} & \cdots & a_{1n} \\ a_{21} & a_{22} & a_{23} & \cdots & a_{2n} \\ a_{31} & a_{32} & a_{33} & \cdots & a_{3n} \\ \vdots & & & & \\ a_{n1} & a_{n2} & a_{n3} & \cdots & a_{nn} \end{vmatrix}; \qquad (5.2.2)$$

y – vector of output variables; **C, D** –matrixes of correspondent dimensions.

Because rms value of first harmonic component is presented almost in all the important power factors definitions it is rational to begin studying algebraization procedure (5.2.1) from the calculation of the first harmonic components of the state variables. Because matrix equation (5.2.1b) is similar to matrix equation (5.2.1a), the algebraization procedure could only be done for (5.2.1a). Except that, first harmonic calculation is valuable as a calculation of a smooth component in quasiharmonical systems.

5.2.1. CONSTANT COEFFICIENTS OF SYSTEM MATIX A

To make algebraization method (5.2.1) obvious, first we apply this method to 2-nd order system and then generalize the result to n-th order.

In expanded form the differential equations for the 2-nd order system look as

$$\begin{aligned} \dot{x}_{1(1)} - a_{11}x_{1(1)} - a_{12}x_{2(1)} &= b_1 u_{1(1)}, \\ \dot{x}_{2(1)} - a_{21}x_{1(1)} - a_{22}x_{2(1)} &= b_2 u_{2(1)}. \end{aligned} \tag{5.2.3}$$

Here all the variables are considered as harmonic functions, because the calculation is performed relative to first harmonic component and all the variables have correspondent index:

$$\begin{aligned} &u_{1(1)} = \sqrt{2}U_{1(1)} \sin \omega t = U_{1(1)\mathrm{a}} \sin \omega t, \\ &u_{2(1)} = \sqrt{2}U_{2(1)} \sin(\omega t - \psi_2) = \sqrt{2}U_{2(1)} \left[\sin \omega t \cos \psi_2 - \sin \psi_2 \cos \omega t\right] = \\ &\qquad = U_{2(1)\mathrm{a}} \sin \omega t - U_{2(1)\mathrm{p}} \cos \omega t, \\ &x_{1(1)} = \sqrt{2}X_{1(1)} \sin(\omega t - \varphi_1) = X_{1(1)\mathrm{a}} \sin \omega t - X_{1(1)\mathrm{p}} \cos \omega t, \\ &x_{2(1)} = \sqrt{2}X_{2(1)} \sin(\omega t - \psi_2 - \varphi_2) = X_{2(1)\mathrm{a}} \sin \omega t - X_{2(1)\mathrm{p}} \cos \omega t, \\ &\frac{dx_{1(1)}}{dt} = \omega\sqrt{2}X_{1(1)} \cos(\omega t - \varphi_1) = \omega X_{1(1)\mathrm{a}} \cos \omega t + \omega X_{1(1)\mathrm{p}} \sin \omega t, \\ &\frac{dx_{2(1)}}{dt} = \omega\sqrt{2}X_{2(1)} \cos(\omega t - \psi_2 - \varphi_2) = \omega X_{2(1)\mathrm{a}} \cos \omega t + \omega X_{2(1)p} \sin \omega t. \end{aligned} \tag{5.2.4}$$

In order to avoid complex calculation of harmonic phases after algebraization procedure of (5.2.1) the first harmonic components in (5.2.4) were split into orthogonal active and reactive components. Then during algebraization procedure equations (5.2.3) are consequently. Now during the algebraization we should consequently multiply equations (5.2.3) and $\cos\omega t$ и $\sin\omega t$ and average them within first harmonic component period. The result is a system composed of 3 equations (items that contain define integrals of sine and cosine multiplications equal to zero are removed):

$$\frac{\omega}{T}\int_0^T X_{1(1)\text{a}}\cos^2\omega t dt+\frac{a_{11}}{T}\int_0^T X_{1(1)\text{p}}\ \cos^2\omega t dt+\frac{a_{12}}{T}\int_0^T X_{2(1)\text{p}}\ \cos^2\omega t=0,$$

$$\frac{\omega}{T}\int_0^T X_{2(1)\text{a}}\cos^2\omega t dt+\frac{a_{21}}{T}\int_0^T X_{1(1)\text{p}}\cos^2\omega t dt+\frac{a_{22}}{T}\int_0^T X_{2(1)\text{p}}\cos^2\omega t=$$

$$=-\frac{b_2}{T}\int_0^T U_{2(1)\text{p}}\cos^2\omega t dt,$$

$$\frac{\omega}{T}\int_0^T X_{1(1)\text{p}}\sin^2\omega t dt-\frac{a_{11}}{T}\int_0^T X_{1(1)\text{a}}\sin^2\omega t dt-\frac{a_{12}}{T}\int_0^T X_{2(1)\text{a}}\sin^2\omega t dt=$$

$$=\frac{b_1}{T}\int_0^T U_{1(1)\text{a}}\sin^2\omega t dt,$$

$$\frac{\omega}{T}\int_0^T X_{2(1)\text{p}}\sin^2\omega t dt-\frac{a_{21}}{T}\int_0^T X_{1(1)\text{a}}\sin^2\omega t dt-\frac{a_{22}}{T}\int_0^T X_{2(1)\text{p}}\sin^2\omega t dt=$$

$$=\frac{b_2}{T}\int_0^T U_{2(1)\text{a}}\sin^2\omega t dt. \tag{5.2.5}$$

After calculation of define integrals the system of algebraic equations relative to sought amplitudes of first reaction harmonic components $X_{1(1)\text{a}}$, $X_{1(1)\text{p}}$, $X_{2(1)\text{a}}$, $X_{2(1)\text{p}}$, that in matrix form looks like:

$$\begin{vmatrix} \omega & 0 & | & a_{11} & a_{12} \\ 0 & \omega & | & a_{21} & a_{22} \\ - & - & + & - & - \\ -a_{11} & -a_{12} & | & \omega & 0 \\ -a_{21} & -a_{22} & | & 0 & \omega \end{vmatrix} \begin{vmatrix} X_{1(1)\text{a}} \\ X_{2(1)\text{a}} \\ X_{1(1)\text{p}} \\ X_{2(1)\text{p}} \end{vmatrix} = \begin{vmatrix} 0\cdot U_{1(1)\text{p}} \\ -b_2 U_{2(1)\text{p}} \\ b_1 U_{1(1)\text{a}} \\ b_2 U_{2(1)\text{a}} \end{vmatrix}. \tag{5.2.6}$$

By dividing coefficient matrix at the sought vector of variables by dashed lines into four submatrixes and considering system matrix (5.2.2) type, the equation (5.2.6) could be written as:

$$\begin{vmatrix} \omega\mathbf{E} & | & \mathbf{A} \\ - \ - & + & - \ - \\ -\mathbf{A} & | & \omega\mathbf{E} \end{vmatrix} \begin{vmatrix} \mathbf{X}_{(1)\mathrm{a}} \\ --- \\ \mathbf{X}_{(1)\mathrm{p}} \end{vmatrix} = \begin{vmatrix} \mathbf{U}_{(1)\mathrm{p}} \\ --- \\ \mathbf{U}_{(1)\mathrm{a}} \end{vmatrix} \tag{5.2.7}$$

where **E** – unitary diagonal matrix.

It is evident that matrix equation (5.2.7) is correct for systems of higher order than 2-nd one if dimension of the matrix A is increased correctly accordingly to (5.2.2) as n x n matrix; similar situation is for all others submatrixes of the equation (5.2.7).

This fact requires to write the solution of (5.2.7) as a solution of a system of n-th order by multiplying left side of the equation by the matrix reverse to coefficient matrix at sought vector of coefficients, this results in

$$\begin{vmatrix} \mathbf{X}_{(1)\mathrm{a}} \\ --- \\ \mathbf{X}_{(1)\mathrm{p}} \end{vmatrix} = \begin{vmatrix} \omega\mathbf{E} & \mathbf{A} \\ -\mathbf{A} & \omega\mathbf{E} \end{vmatrix}^{-1} \cdot \begin{vmatrix} \mathbf{U}_{(1)\mathrm{p}} \\ --- \\ \mathbf{U}_{(1)\mathrm{a}} \end{vmatrix} . \tag{5.2.8}$$

Resulting amplitude of j-h variable harmonic component is defined by its modulus

$$X_{j(1)}^2 = \sqrt{X_{j(1)\mathrm{a}}^2 + X_{j(1)\mathrm{p}}^2} \tag{5.2.9}$$

and phase

$$\varphi_{j(1)} = \operatorname{arctg} \frac{X_{j(1)\mathrm{p}}}{X_{j(1)\mathrm{a}}} . \tag{5.2.10}$$

Thus, a formula (5.2.8) provides a solution for amplitude values of 1-st harmonic components for all state variables. Comparing to symbolic calculation method for linear electrical circuits with sine currents described method doesn't require difficult procedure of calculation of modulus and phases of partial complex resistances between circuit input and other partial element; this procedure should be done n-times in n-th order system. Except that the solution is presented as a function of differential equation coefficients; this fact makes it common equitable for systems of any order. In comparison with operational method of electrical circuits calculation using Laplace transformation here the knowledge of roots of characteristic equation is not required. It is widely known that calculation of these roots for equations of order higher than

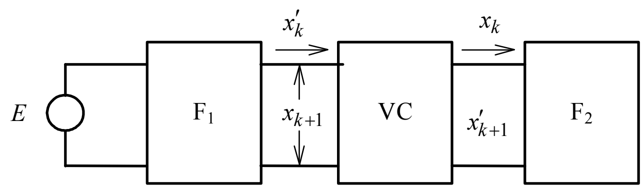
3 is impossible in general case, but necessary to perform reverse Laplace transform.

To distinguish between direct calculation method for first harmonic component described in section 1.5 suitably for mathematical model of SISO and direct calculation method for first harmonic for mathematical models SIMO, MIMO first method will be denoted as *ADESS(1)*; added letters *SS* mean *State Space*

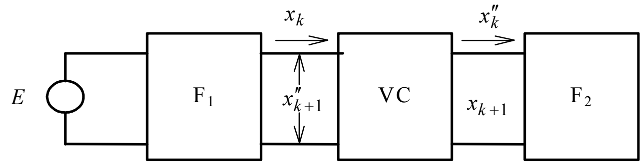
5.2.2*. VARIABLE COEFFICIENTS OF SYSTEM MATRIX A

Mathematical model of a converter in a differential equation form with constant coefficients of system matrixes A and C in (5.2.1) is suitable when converter is fed from ideal power sources (emf with zero internal impedance, current sources with infinite internal impedance). At that current source and valve converter without any R, L, C elements in mathematical model represented as discontinuous (nonsinusoidal) input actions named as **Bu** and **Du** due to presence of discontinuous commutation functions of valves in action coefficient matrixes **B** and **D**. Then the analysis of electromagnetic processes in system adds up to the analysis of discontinuous actions (right part of differential equations (5.2.3)) on the system with constant parameters (left part of differential equations (5.2.3)).

When valve converter is fed from real power source that has internal impedance and when the converter has input filter the key converter is connected between two continuous parts of the system as shown in Fig.5.2.1; case *a* for converters of voltage source type, case *b* for converters of current source type.



a



б

Fig. 5.2.1

Here the voltage source type of the converter means that converter forms its output voltage curve; current source type – its output current. Then for first type of the converter (Fig. 5.2.1,a) it is possible to write connection equation between its input and output variables considering (1.4.2) as

$$x'_k = \psi_n x_k, \qquad x'_{k+1} = \psi_n x_{k+1}. \tag{5.2.11a}$$

State variables are continuous variables x_{k+1} (voltage on the input filter capacitors for voltage converter) and x_k (load inductance current); auxiliary variables x'_{k+1} and x'_k describe output voltage of the converter and its input current correspondingly, they are discontinuous functions (refer to VSI in chapter 8). For the second type of the converter (Fig. 5.2.1,b) similar equations are correct but state variables characterize current of input inductance (x_k) and output capacitance voltage (x_{k+1}) (refer to CSI in chapter 8). Auxiliary variables x''_k and x''_{k+1} define discontinuous output current and discontinuous input voltage of the converter correspondingly

$$x''_k = \psi_n x_k, \qquad x''_{k+1} = \psi_n x_{k+1}. \tag{5.2.11b}$$

Auxiliary variables that present in state equations for each discontinuous part of the system will vanish in process of building of united space equation system using (5.2.11). Indeed, state equations for first continuous part of the system (F_1) of k-th order look like

$$\begin{aligned} \dot{x}_1 &= a'_{11}x_1 + a'_{12}x_2 + \ldots + a'_{1k}x_k + a'_1 x'_{k+1} + b'_1 e \\ &\vdots \\ \dot{x}_k &= a'_{k1}x_1 + a'_{k2}x_2 + \ldots + a'_{kk}x_k + a'_k x'_{k+1} + b'_k e. \end{aligned} \tag{5.2.12}$$

State variables for second continuous part of the system (F_2) of m-th order look like

$$\begin{aligned} \dot{x}_{k+1} &= a''_{k+1}x'_k + a''_{11}x_{k+1} + a''_{21}x_{k+2} + \ldots + a''_{k1}x_{k+m} + b'_{k+1}x'_k \\ &\vdots \\ \dot{x}_{k+m} &= a''_{k+m}x'_k + a''_{m1}x_{k+1} + a''_{m2}x_{k+2} + \ldots + a''_{mm}x_{k+m} + b'_{k+m}x'_k \text{''}. \end{aligned} \tag{5.2.13}$$

Then the united equation system of state variables for converter of $n = k + m$ order considering (5.2.11)-(5.2.13) could be written as:

$$\begin{vmatrix} \dot{x}_1 \\ \dot{x}_2 \\ \vdots \\ \dot{x}_k \\ \dot{x}_{k+1} \\ \vdots \\ \dot{x}_{k+m} \end{vmatrix} = \begin{vmatrix} a'_{11} & a'_{12} & \dots & a'_{1k} & | & a_1\Psi_n & & & \\ a'_{21} & a'_{22} & \dots & a'_{2k} & | & a_2\Psi_n & & & \\ \vdots & & & & | & & & & \\ a'_{k1} & a'_{k2} & \dots & a'_{kk} & | & a_k\Psi_n & & & \\ - & - & - & - & + & - & - & - & - \\ & & & a_{k+1}\Psi_n & | & a''_{11} & a''_{12} & \dots & a''_{1k} \\ & & & \vdots & | & & & & \\ & & & a_{k+m}\Psi_n & | & a''_{m1} & a''_{12} & \dots & a''_{mm} \end{vmatrix} \begin{vmatrix} x_1 \\ x_2 \\ \vdots \\ x_k \\ x_{k+1} \\ \vdots \\ x_{k+m} \end{vmatrix} + \begin{vmatrix} b'_1 e \\ b'_2 e \\ \vdots \\ b'_k e \\ 0 \\ \vdots \\ 0 \end{vmatrix} \tag{5.2.14}$$

From matrix equation (5.2.14) the method of obtaining of resulting system matrixes from continuous matrixes of subsystems F_1 and F_2 becomes clear. Variable coefficients appear in additional right column of F_1 coefficient system matrix $\mathbf{A}'$ and in left additional column of F_2 coefficient system matrix $\mathbf{A}''$. In convoluted form of submatrixes the equation (5.2.14) would look like

$$\begin{vmatrix} \dot{\mathbf{x}}_k \\ --- \\ \dot{\mathbf{x}}_m \end{vmatrix} = \begin{vmatrix} \mathbf{A}' & | & \mathbf{A}_k\Psi_n \\ - \; - & + & - \; - \\ \mathbf{A}_m\Psi_n & | & \mathbf{A}'' \end{vmatrix} + \begin{vmatrix} \mathbf{Bu} \\ --- \\ 0 \end{vmatrix}. \tag{5.2.15}$$

It is evident that coefficients of submatrixes $\mathbf{A}_k$ and $\mathbf{A_m}$ are defined by reverse values of correspondent coefficients of derivatives $\dot{\mathbf{x}}_k, \dot{\mathbf{x}}_m$ before their normalization (reduction to unity), i.e. by inductances L_k and capacitances C_k; the correspondent members of state equations are $L_k \dfrac{di_k}{dt}$ and $C_k \dfrac{du_k}{dt}$. If the converter is presented not as quadruple but as six-pole element (3-phase systems at the input and output of the converter) or octopole (3-phase systems with zero wire) or in general case as $2p$-pole the matrixes $\mathbf{A}_k$ and $\mathbf{A}_m$ will contain the number of columns equal to the number of auxiliary variables at the converter input and output correspondingly.

The same method is also suitable for obtaining converter state equations if it contains not only valves but also other elements such as RC-circuits, current-limiting or paralleling reactors and so on.

In general case the equation look like

$$\dot{\mathbf{x}} - \mathbf{A}(t)\mathbf{x} = \mathbf{B}(t)\mathbf{u} \,, \tag{5.2.16}$$

i.e. we obtained a system of differential equations with variable periodical discontinuous coefficients of system matrixes $\mathbf{A}(t)$ and $\mathbf{B}(t)$.

As a result we conclude that algebraization of system of differential equations (5.2.16) is very difficult in general case because the result depends on character of variation of $\mathbf{A}(t)$ matrix coefficients, i.e. on the converter type and its commutating function type that is defined by the control strategy. The application of considered algebraization procedure to the concrete converter devices allows to obtain closed analytical equations for rms values of first harmonics of state variables, i.e. for smooth components of electromagnetic noises. The calculation of rms values of total of high-frequency components of processes is considered in the next chapter as a part of development of direct calculation methods in sections 5.1 and 5.2.

5.3.* DEVELOPMENT OF DIRECT CALCULATION METHODS OF POWER FACTORS FOR VALVE CONVERTERS IN STATE SPACE

Differential equations if the state space that define mathematical models of the system such as SIMO, MIMU could be effectively used in building of two new versions of ADE method – ADESS1 for calculation of rms values of system reaction for nonsinusoidal actions and ADESS2 for calculation of rms values reaction higher harmonics for nonsinusoidal actions. These methods are alternative to ADE1 and ADE2 ones that are used for SISO and MISO systems (refer to section 1.5).

The procedure of algebraization in ADESS1 method will be introduced as in previous section – for the 2-nd order system and further generalization for n-th order system. The structure of differential equations here is similar to one (5.2.3)

$$\begin{aligned} &\dot{x}_1 - a_{11}x_1 - a_{12}x_2 = b_1u_1, \\ &\dot{x}_2 - a_{21}x_1 - a_{22}x_2 = b_2u_2. \end{aligned} \tag{5.3.1}$$

First step is converting the differential equations to integral ones by single integration:

$$x_1 - a_{11}\bar{x}_1 - a_{12}\bar{x}_2 = b_1\bar{u}_1,$$
$$x_2 - a_{21}\bar{x}_1 - a_{22}\bar{x}_2 = b_2\bar{u}_2. \tag{5.3.2}$$

Second step towards algebraization is the conversion of integral equations (5.2.3) to algebraic ones by squaring and averaging within a period of action, the result is:

$$X_1^2 + a_{11}^2\bar{X}_1^2 + a_{12}^2\bar{X}_2^2 - 2a_{12}(x_1,\bar{x}_2) + 2a_{11}a_{12}(\bar{x}_1,\bar{x}_2) = b_1^2\bar{U}_1^2,$$
$$X_2^2 + a_{21}^2\bar{X}_1^2 + a_{22}^2\bar{X}_2^2 - 2a_{21}(\bar{x}_1,x_2) + 2a_{21}a_{22}(\bar{x}_1,\bar{x}_2) = b_2^2\bar{U}_2^2. \tag{5.3.3}$$

The following labeling for scalar multiplications is used:

$$(x_1,\bar{x}_2) = \frac{1}{T}\int_0^T x_1\bar{x}_2\,dt, \qquad (\bar{x}_1,x_2) = \frac{1}{T}\int_0^T \bar{x}_1 x_2\,dt,$$
$$(\bar{x}_1,\bar{x}_2) = \frac{1}{T}\int_0^T \bar{x}_1\bar{x}_2\,dt, \qquad (x_1,\bar{x}_1) = (x_2,\bar{x}_2) = 0. \tag{5.3.4}$$

We obtain two algebraic equations (5.3.3) and six unknowns: $X_1, X_2, \bar{X}_1, \bar{X}_2, (\bar{x}_1,\bar{x}_2)$ и $(x_1,\bar{x}_2) = -(\bar{x}_1,x_2)$. Last equality could be proven by integration of scalar multiplication by parts.

Third step is redefining of obtained system composed of two algebraic equations by four algebraic correlations for unknowns to make the equation system compatible. With the framework of first level of assumption of ADE method N = 1 [21] we still rely that integrals of instant value of nonsinusoidal periodic functions could be substituted by their first harmonics because the integration process makes all k-th harmonics k – times weaker, i.e. this action is a kind of filtering of higher harmonic components. As a result we have

$$\bar{x}_1 \approx \bar{x}_{1(1)} = -\frac{\sqrt{2}}{\omega}X_{1(1)}\cos(\omega t - \varphi_{1(1)}), \qquad \bar{X}_{1(1)} \cong \frac{X_{1(1)}}{\omega},$$
$$\bar{x}_2 \approx \bar{x}_{2(1)} = -\frac{\sqrt{2}}{\omega}X_{2(1)}\cos(\omega t - \varphi_{2(1)}), \qquad \bar{X}_{2(1)} \cong \frac{X_{2(1)}}{\omega}.$$

As a result of calculation of (5.3.4) considering (5.3.5) we obtain

$$(x_1,\bar{x}_2) = -\frac{1}{T}\int_0^T \sqrt{2}X_{1(1)}\sin(\omega t-\varphi_{1(1)})\sqrt{2}\frac{X_{2(1)}}{\omega}\cos(\omega t-\varphi_{2(1)})dt =$$

$$= -\frac{X_{1(1)}X_{2(1)}}{\omega}\cos(\varphi_{2(1)}-\varphi_{1(1)}),$$

$$(\bar{x}_1,\bar{x}_2) = \frac{1}{T}\int_0^T \sqrt{2}\frac{X_{1(1)}}{\omega}\cos(\omega t-\varphi_{1(1)})\sqrt{2}\frac{X_{2(1)}}{\omega}\cos(\omega t-\varphi_{2(1)})dt =$$

$$= \frac{X_{1(1)}X_{2(1)}}{\omega^2}\cos(\varphi_{2(1)}-\varphi_{1(1)}).$$

(5.3.6)

Fourth step is solving of algebraic equation system (5.3.3) considering additional algebraic correlations (5.3.5) and (5.3.6), as a result from (5.3.3) we obtain

$$X_1^2 = b_1^2\bar{U}_1^2 - a_{11}^2\frac{X_{1(1)}}{\omega^2} - \frac{a_{12}^2}{\omega^2}X_{2(1)}^2 -$$

$$-2\frac{\left[a_{12}+\frac{a_{11}a_{12}}{\omega}\right]}{\omega}X_{1(1)}X_{2(1)}\cos(\varphi_{2(1)}-\varphi_{1(1)}),$$

$$X_2^2 = b_2^2\bar{U}_2^2 - \frac{a_{21}^2}{\omega^2}X_{1(1)}^2 - \frac{a_{22}^2}{\omega^2}X_{2(1)}^2 + \tag{5.3.7}$$

$$+2\frac{\left[a_{21}+\frac{a_{21}a_{22}}{\omega}\right]}{\omega}X_{1(1)}X_{2(1)}\cos(\varphi_{2(1)}-\varphi_{1(1)}).$$

right away.

Amplitude values of first harmonics of variables are calculated via formulas (5.2.9).

Due to the some awkwardness of (5.3.7) for rms value of the reactions for nonsinusoidal actions for SIMU, MIMU mathematical models the method ASESS2 appears to be mo rational than ADESS1, because

ADESS2 calculates rms value of overall value of higher harmonics of the reaction. Then the rms value of overall nonsinusoidal reaction of the system could evidently be found using formulas of calculation for first harmonic component

$$X_j = \sqrt{X_{j(1)}^2 + X_{h.h}^2}\,. \tag{5.3.8}$$

The procedure of algebraization of differential equations by ADESS2 method for total of reaction higher harmonic component has some specific features. First step is obtaining of integral equation again for totality of higher harmonics of state variables from correspondent differential equations analogous to (5.3.1):

$$\begin{aligned} &\dot{x}_{1h.h} - a_{11}x_{1h.h} - a_{12}x_{2h.h} = b_1 u_{1h.h}, \\ &\dot{x}_{2h.h} - a_{21}x_{1h.h} - a_{22}x_{2h.h} = b_2 u_{2h.h}, \end{aligned} \tag{5.3.9}$$

that produces

$$\begin{aligned} &x_{1h.h} - a_{11}\bar{x}_{1h.h} - a_{12}\bar{x}_{2h.h} = b_1 \bar{u}_{1h.h}, \\ &x_{2h.h} - a_{21}\bar{x}_{1h.h} - a_{22}\bar{x}_{2h.h} = b_2 \bar{u}_{2h.h}. \end{aligned} \tag{5.3.10}$$

Second step is making the system of two integral equations with four unknowns $x_{1h.h}, x_{2h.h}, \bar{x}_{1h.h}, \bar{x}_{2h.h}$ compatible (5.3.10) by redefining it with two additional correlations obtained from (5.3.5) within framework of first level of assumption of ADE method

$$\bar{x}_{1h.h} \approx 0, \qquad \bar{x}_{2h.h} \approx 0. \tag{5.3.11}$$

As a result the integral equations (5.3.10) comes to simplest combined system

$$x_{1h.h} = b_1 \bar{u}_{1h.h}, \qquad x_{2h.h} = b_2 \bar{u}_{2h.h}. \tag{5.3.12}$$

Third step is algebraization of integral equations (5.3.12) by squaring them and averaging within period of action that immediately gives a solution for totality of higher harmonics of the action

$$\begin{aligned} X_{1h.h} &= b_1 \bar{U}_{1h.h} = \frac{b_1 U_{1(1)}}{\omega}\sqrt{1 + \left(\bar{K}_{h1}\right)^2}, \\ X_{2h.h} &= b_2 \bar{U}_{2h.h} = \frac{b_2 U_{2(1)}}{\omega}\sqrt{1 + \left(\bar{K}_{h}\right)^2}. \end{aligned} \tag{5.3.13}$$

Here $\bar{K}_{\mathrm{h}1}, \bar{K}_{\mathrm{h}2}$ - integral harmonic coefficients of the action (nonsinusoidal current or voltage source).

The elementary nature of resulting formulas in ADESS2 method within first level of assumption frameworks makes obtaining of common output formulas for the s e c o n d l e v e l o f a s s u m m p t i o n $N = 2$ well-grounded. The principle of their obtaining follows below.

First step is obtaining of second system of integral equations for higher harmonics of variables by single integration of first system of integral equations (5.3.10):

$$\begin{aligned} &\bar{x}_{1h.h} - a_{11}\bar{\bar{x}}_{1h.h} - a_{12}\bar{\bar{x}}_{2h.h} = b_1\bar{\bar{u}}_{1h.h}, \\ &\bar{x}_{2h.h} - a_{21}\bar{\bar{x}}_{1h.h} - a_{22}\bar{\bar{x}}_{2h.h} = b_2\bar{\bar{u}}_{2h.h} \end{aligned} \tag{5.3.14}$$

and its combination with first system (5.3.10) that results in

$$\begin{aligned} &x_{1h.h} - a_{11}\bar{x}_{1h.h} - a_{12}\bar{x}_{2h.h} = b_1\bar{u}_{1h.h}, \\ &x_{2h.h} - a_{21}\bar{x}_{1h.h} - a_{22}\bar{x}_{2h.h} = b_2\bar{u}_{2h.h}, \\ &\bar{x}_{1h.h} - a_{11}\bar{\bar{x}}_{1h.h} - a_{12}\bar{\bar{x}}_{2h.h} = b_1\bar{\bar{u}}_{1h.h}, \\ &\bar{x}_{2h.h} - a_{21}\bar{\bar{x}}_{1h.h} - a_{22}\bar{\bar{x}}_{2h.h} = b_2\bar{\bar{u}}_{2h.h}. \end{aligned} \tag{5.3.15}$$

Then obtain a system of four equations with six unknowns $x_{1h.h}, x_{2h.h}, \bar{x}_{1h.h}, \bar{x}_{2h.h}, \bar{\bar{x}}_{1h.h}, \bar{\bar{x}}_{2h.h}$.

Second step is making this system compatible by its redefining by two correlations that following from the second level of the assumption $N = 2$ of ADE method; this means

$$\bar{\bar{x}}_{1h.h} \approx 0, \qquad \bar{\bar{x}}_{2h.h} \approx 0 . \tag{5.3.16}$$

Then the system (5.3.15) is being simplified to two independent equations:

$$\begin{aligned} &x_{1h.h} = b_1\bar{u}_{1h.h} + a_{11}b_1\bar{\bar{u}}_{1h.h} + a_{12}b_2\bar{\bar{u}}_{2h.h}, \\ &x_{2h.h} = b_2\bar{u}_{2h.h} + a_{21}b_1\bar{\bar{u}}_{1h.h} + a_{22}b_2\bar{\bar{u}}_{2h.h}. \end{aligned} \tag{5.3.17}$$

Third step is algebraization of integral equations (5.3.17) by squaring and averaging them. The result is immediate solution for rms values of totality of higher harmonics of reaction

$$X_{1h.h}^2 = b_1^2 \overline{U}_{1h.h}^2 + a_{11}^2 b_1^2 \overline{\overline{U}}_{1h.h}^2 + a_{12}^2 b_2^2 \overline{\overline{U}}_{2h.h}^2 +$$
$$+2a_{12} b_2 \left(\overline{u}_{1h.h}, \overline{\overline{u}}_{2h.h}\right) + 2a_{11} a_{12} b_1 b_2 \left(\overline{\overline{u}}_{1h.h}, \overline{\overline{u}}_{2h.h}\right),$$
$$X_{2h.h}^2 = b_2^2 \overline{U}_{2h.h}^2 + a_{21}^2 b_1^2 \overline{\overline{U}}_{1h.h}^2 + a_{22}^2 b_2^2 \overline{\overline{U}}_{2h.h}^2 + \tag{5.3.18}$$
$$+2a_{21} b_1 \left(\overline{\overline{u}}_{1h.h}, \overline{u}_{2h.h}\right) + 2a_{21} a_{22} b_1 b_2 \left(\overline{\overline{u}}_{1h.h}, \overline{\overline{u}}_{2h.h}\right).$$

At the second level of the assumption in ADESS2 method it is necessary to know not only action integral harmonic coefficients of 1-st and 2-nd order $\overline{K}_{\text{h1}}, \overline{\overline{K}}_{\text{h2}}$, but also additional cross integral coefficient of the action of 1-2 order that is defined as

$$\overline{K}_{\text{h12}}^{(1-2)} = \frac{\omega^3}{U_{1(1)} U_{2(1)}} \left(\overline{u}_{1h.h}, \overline{\overline{u}}_{2h.h}\right) = \frac{\omega^3}{U_{1(1)} U_{2(1)}} \left(\overline{\overline{u}}_{1h.h}, \overline{u}_{2h.h}\right). \tag{5.3.19}$$

If the system has only one action source (single-phase circuits) the formulas (5.3.18) becomes extremely simple:

$$X_{1h.h}^2 = \left(b_1 \frac{U_{1(1)}}{\omega} \overline{K}_{\text{h1}} \right)^2 + \left(a_{11} b_1 \frac{U_{1(1)}}{\omega^2} \overline{\overline{K}}_{\text{h1}} \right)^2,$$
$$X_{2h.h} = a_{21} b_1 \frac{U_{1(1)}}{\omega^2} \overline{\overline{K}}_{\text{h1}}. \tag{5.3.20}$$

It is easy to prove that the system of 3-rd order with single action source the formulas for second level of assumption for rms values of higher harmonic components of the variables for $X_{1h.h}, X_{2h.h}$ will be the same, but for third variable we can write

$$X_{3h.h} = a_{13} b_1 \frac{U_{1(1)}}{\omega^2} \overline{\overline{K}}_{\text{h1}}. \tag{5.3.21}$$

The difference between this equation and equation for $X_{2h.h}$ lies only in substitution of coefficient a_{21} by a_{31}; this fact lets generalize the results for the system of n-th order (SIMO model):

$$X_{nh.h}^2 = a_{n1} b_1 \frac{U_{1(1)}}{\omega^2} \overline{\overline{K}}_{\text{h1}}, \tag{5.3.22}$$

where changes in number n from 2 to n (order of the system) produces formulas for correspondent variables; the formulae for first variable ($X_{1h.h}$) remains invariable as in (5.3.20).

Thus, the common method of algebraization of differential equation system in state space shape was built; this method allow to obtain closed analytical equations (engineering formulas) to calculate rms values of nonsinusoidal reactions in circuit with nonsinusoidal actions (ADESS1 method) or only for calculation of rms values of summary of high frequency components of nonsinusoidal reactions (ADESS2 method). In second case the correlations is much simpler; this fact allow deriving a formulae the system of any order.

In case of mathematical model of the system of differential equations of state space with variable coefficients such as (5.2.16) the algebraization of such equations becomes rather complicated due to the singularity of scalar multiplications of (5.3.4) type because of discontinuous coefficients. Here the further development of direct methods is necessary; one of the possible approaches to that using projecting method of Galerkin could be found in [19].

5.4. DIRECT CALCULATION METHODS OF POWER FACTORS FOR 3-PHASE CIRCUITS WITH POWER CONVERTERS

Any dissymmetry whether of phase element parameters of multiphase converter or of controls or load results in dissymmetry of electromagnetical processes in phases relative to first harmonic components and high frequency components. That is why we are going to briefly consider the procedure of algebraization of differential equations for direct calculation methods for calculation for first harmonic components within ADE(1) method frameworks and further for higher harmonic components with ADE2 method frameworks.

Calculation for first harmonic component by ADE(1) method. The calculation of dissymmetrical modes in linear multiphase circuits with sinusoidal voltage sources in most cases done by method of symmetrical components. At that difficult calculation task is decomposed into series (three or more) simple tasks of calculations for individual components (forward, reverse and zero). This provides an opportunity to consider the difference in system parameters for method components;

first of all it is necessary in calculation of circuits with rotating multiphase electrical machines. Common solution is found by composition method of individual components (the superposition principle).

An alternative method to solve the same task could also be build on the basis of direct calculation methods. At that there is no necessity to decompose multiphase system of supply voltages into symmetrical components and thrice-repeated calculation for each component.

Firstly we consider common case of 3-phase 4-wire circuit; it is easy to further consider 3-wire case by setting the parameters of zero wire infinite. Analytical model of the circuit is shown in Fig. 5.4.1 for active-inductive nonsymmetrical load and arbitrary dissymmetry of 3-phase e.m.f. system (for phase or amplitude).

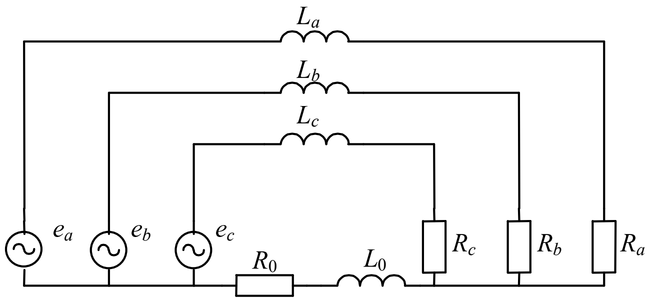


Fig. 5.4.1

Differential equations for this circuit look like

$$
\begin{aligned}
&L_a \frac{di_{a(1)}}{dt} + R_a i_{a(1)} + L_0 \frac{di_{0(1)}}{dt} + R_0 i_{0(1)} = e_{a(1)}\,, \\
&L_b \frac{di_{b(1)}}{dt} + R_b i_{b(1)} + L_0 \frac{di_{0(1)}}{dt} + R_0 i_{0(1)} = e_{b(1)}\,, \\
&L_c \frac{di_{c(1)}}{dt} + R_c i_{c(1)} + L_0 \frac{di_{0(1)}}{dt} + R_0 i_{0(1)} = e_{c(1)}\,, \\
&i_{a(1)} + i_{b(1)} + i_{c(1)} = i_{0(1)}\,.
\end{aligned}
\tag{5.4.1}
$$

By excluding current i_0 using last equality and transferring to matrix form we obtain

$$\begin{vmatrix} L_a+L_0 & L_0 & L_0 \\ L_0 & L_b+L_0 & L_0 \\ L_0 & L_0 & L_c+L_0 \end{vmatrix} \cdot \begin{vmatrix} \dfrac{di_{a(1)}}{dt} \\ \dfrac{di_{b(1)}}{dt} \\ \dfrac{di_{c(1)}}{dt} \end{vmatrix} +$$

$$+\begin{vmatrix} R_a+R_0 & R_0 & R_0 \\ R_0 & R_b+R_0 & R_0 \\ R_0 & R_0 & R_c+R_0 \end{vmatrix} \cdot \begin{vmatrix} i_{a(1)} \\ i_{b(1)} \\ i_{c(1)} \end{vmatrix} = \begin{vmatrix} e_{a(1)} \\ e_{b(1)} \\ e_{c(1)} \end{vmatrix}. \qquad (5.4.2)$$

By introducing $p = d/dt$ for differentiation operator, the equation (5.4.2) changes to:

$$\begin{vmatrix} p(L_a+L_0)+R_a+R_0 & pL_0+R_0 & pL_0+R_0 \\ pL_0+R_0 & p(L_b+L_0)+R_b+R_0 & pL_0+R_0 \\ pL_0+R_0 & pL_0+R_0 & p(L_c+L_0)+R_c+R_0 \end{vmatrix} \times$$

$$\times \begin{vmatrix} i_{a(1)} \\ i_{b(1)} \\ i_{c(1)} \end{vmatrix} = \begin{vmatrix} e_{a(1)} \\ e_{b(1)} \\ e_{c(1)} \end{vmatrix}. \qquad (5.4.3)$$

To apply algebraization procedure (from ADE1 method) we obtain from (5.4.3) differential equations (of 3-rd order here) for each of phase currents i_a, i_b, i_c using formulae

$$\Delta(p)\, i_{km(1)} = \sum_{k=1}^{n} \Delta_{km}(p)\, e_{m(1)}, \qquad m = a,b,c\,, \qquad (5.4.4)$$

where $\Delta(p)$ -differential operator, obtained from determinant of coefficient matrix in (5.4.3):

$$\Delta(p)=\begin{vmatrix} p(L_a+L_0)+R_a+R_0 & pL_0+R_0 & pL_0+R_0 \\ pL_0+R_0 & p(L_b+L_0)+R_b+R_0 & pL_0+R_0 \\ pL_0+R_0 & pL_0+R_0 & p(L_b+L_0)+R_b+R_0 \end{vmatrix}, \quad (5.4.5)$$

and $\Delta_{km}(p)$ – differential operator, defined by algebraic adjunct, i.e. by determinant of $\Delta(p)$ with crossed out k-th row and m-th column, multiplied by $(-1)^{k+m}$.

Obtained differential equation for currents by solution using (5.4.4) looks like

$$\begin{aligned} & a_{3m}p^3 i_{m(1)} + a_{2m}p^2 i_{m(1)} + a_{1m}p i_{m(1)} + a_{0m} i_{m(1)} = \\ & = b_{2a}p^2 e_{a(1)} + b_{1a}p e_{a(1)} + b_{0a}e_{a(1)} + b_{2b}p^2 e_{b(1)} + b_{1b}p e_{b(1)} + \qquad (5.4.6) \\ & + b_{0b}e_{b(1)} + b_{2c}p^2 e_{c(1)} + b_{1c}p e_{c(1)} + b_{0c}e_{c(1)}\,. \end{aligned}$$

After thrice-repeated integration the differential equation (5.4.6) turns to integral one:

$$\begin{aligned} & a_{3m} i_{m(1)} + a_{2m}\bar{i}_{m(1)} + a_{1m}\bar{i}^{(2)}_{m(1)} + a_{0m}\bar{i}^{(3)}_{m(1)} = \\ & = b_{2a}\bar{e}_{a(1)} + b_{1a}\bar{e}^{(2)}_{a(1)} + b_{0a}\bar{e}^{(3)}_{a(1)} + b_{2b}\bar{e}_{b(1)} + \qquad (5.4.7) \\ & + b_{1b}\bar{e}^{(2)}_{b(1)} + b_{0b}\bar{e}^{(3)}_{b(1)} + b_{2c}\bar{e}_{c(1)} + b_{1c}\bar{e}^{(2)}_{c(1)} + b_{0c}\bar{e}^{(3)}_{c(1)}\,. \end{aligned}$$

Coefficients $a_{3m}, a_{2m}, a_{1m}, a_{0m}$ are expressed via parameters of electrical circuit $L_0, L_a, L_b, L_c, R_0, R_a, R_b, R_c$ after exposing of determinant $\Delta(p)$ by (5.4.5); coefficients b_{2m}, b_{1m}, b_{0m} are also expressed via parameters of electrical circuit after exposing of determinant $\Delta_{km}(p)$.

After squaring if (5.4.7) and averaging it within supply voltage period by solution formulae of ADE1 method we obtain an expression for calculation of average value of sinusoidal current of m-th phase of current

$$I_{m(1)} : I^2_{m(1)}\left[a^2_{3m} + \frac{a^2_{2m} - 2a_{3m}a_{1m}}{\omega^2} + \frac{a^2_{1m} - 2a_{2m}a_{0m}}{\omega^4} + \frac{a^2_{0m}}{\omega^6}\right] =$$

$$= \sum_{m=a,b,c} \frac{b^2_{2m}}{\omega^2}E^2_{m(1)} + \sum_{m=a,b,c} \frac{b^2_{1m} - 2b_{2m}b_{0m}}{\omega^4}E^2_{m(1)} + \sum_{m=a,b,c} \frac{b^2_{0m}}{\omega^6}E^2_{m(1)} +$$

$$+2\sum_{\substack{k_1,k_2=0,1,2\\ m_1,m_2=a,b,c\\ m_1\neq m_2}} b_{km_1}b_{km_2}\left(\overline{e}_{m_1(1)}^{(3-k_1)},\overline{e}_{m_2(1)}^{(3-k_2)}\right). \tag{5.4.8}$$

Here the integrals of cross products are denoted as

$$\left(\overline{e}_{m_1(1)}^{(3-k_1)},\ \overline{e}_{m_2(1)}^{(3-k_2)}\right)=\frac{1}{2\pi}\int_0^{2\pi}\overline{e}_{m_1(1)}^{(3-k_1)}\ \overline{e}_{m_2(1)}^{(3-k_2)}d\vartheta, \tag{5.4.9}$$

they are calculated easily for sinusoidal source voltages shape.

For example, scalar product such as (5.4.9) for amplitude phase voltages $e_{a(1)}$ and $e_{b(1)}$ dissymmetry, and phase deviation of phase voltage e_b from symmetrical shift of 120° on $\Delta\varphi_b$ looks like:

$$\left(e_a,e_b\right)=\frac{1}{T}\int_0^T\sqrt{2}E_{a(1)}\sqrt{2}E_{b(1)}\sin\omega t\cdot\sin\left(\omega t-120^\circ-\Delta\varphi_B\right)dt=$$

$$=E_{a(1)}E_{b(1)}\cos\left(120^\circ-\Delta\varphi_B\right). \tag{5.4.10}$$

Thus, the equation (5.4.8) gives a solution for rms values of phase currents of 3-phase dissymmetrical circuit without voltage decomposition procedures into symmetrical components. But, the calculation of integrals of cross products phase voltages of the sources is required instead. But the solution here has closed analytical form; this feature is practically unavailable in method of symmetrical components because of awkwardness of algorithm makes obtaining of closed form impossible.

Calculation of higher harmonic components by ADE2 method. As it was already shown, the procedure of algebraization of differential equations for 3-phase circuit for higher harmonics is similar to the procedure of algebraization of differential equations for first harmonic considered above in section 1.5; one should only change index «(1)» of all variables in (5.4.1)-(5.4.7) to index that refer to higher harmonics - «h.h.». Then instead of (5.4.7) we will obtain following integral equation for higher harmonics of the variables:

$$a_{3m}\dot{i}_{mh.h} + a_{2m}\bar{i}_{mh.h} + a_{1m}\bar{i}^{(2)}_{mh.h} + a_{0m}\bar{i}^{(3)}_{mh.h} =$$

$$= b_{2a}\bar{e}_{a.h.h} + b_{1a}\bar{e}^{(2)}_{a.h.h} + b_{0a}\bar{e}^{(3)}_{a.h.h} + b_{2b}\bar{e}_{b.h.h} +$$

$$+ b_{1b}\bar{e}^{(2)}_{b.h.h} + b_{0b}\bar{e}^{(3)}_{b.h.h} + b_{2c}\bar{e}_{c.h.h} + b_{1c}\bar{e}^{(2)}_{c.h.h} + b_{0c}\bar{e}^{(3)}_{c.h.h}\,, \tag{5.4.11}$$

$$m = a, b, c.$$

Its algebraization within first level of assumption $N = 1$ of ADE method results in following equation for rms value of higher harmonic components of phase currents:

$$I^2_{mh.h} = \frac{1}{a^2_{3m}} \sum_{m=a,b,c} \frac{b^2_{2m}}{\omega^2} E^2_m \bar{K}^2_{\text{h}} + \sum_{m=a,b,c} \frac{b^2_{1m} - 2b_{2m}b_{0m}}{\omega^4} E^2_m \left(\bar{K}^{(2)}_{\text{h}}\right)^2 +$$

$$+ \sum_{m=a,b,c} \frac{b^2_{0m}}{\omega^4} E^2_m \left(\bar{K}^{(3)}_{\text{h.h}}\right)^2 + 2 \sum_{\substack{k_1,k_2=0,1,2\\ m_1,m_2=a,b,c\\ m_1\neq m_2}} b_{km_1} b_{km_2} \left(\bar{e}^{(3-k_1)}_{m_1} \bar{e}^{(3-k_2)}_{m_2}\right). \tag{5.4.12}$$

Here the scalar products in last member define cross integral coefficients of harmonic components of $m_1 m_2$ order similar to (5.4.9).

Thus, direct methods allow us to obtain analytical equations for power quality factors of nonsinusoidal, dissymmetrical electromagnetic processes in multiphase electrical circuits. But the formulas in these cases are difficult enough and require such mathematical software as Mathcad (or other) to obtain desired characteristics. The usage of these formulas is defensible in those cases when calculation of great number of characteristics for hundreds of operational modes is required, but utilization of such systems as Pspice, Mcap, Parus-Pargraph for direct modeling would require too much time and resources. Calculation of approach to steady-state operation even in a single point for difficult circuits could take several seconds or minutes.

5.5**. PRECISE SOLUTIONS FOR NORMS BY DIRECT METHODS

It would be shown below that for some electrical circuits by direct methods it is possible to obtain p r e c i s e s o l u t i o n s for rms values of nonsinusoidal reactions (currents) for general shape of nonsinusoidal action (voltage). This allows introducing such term as *modulus* *of generalized circuit resistance at nonsinusoidal processes* and aims to expand search for exact solutions in more general cases.

5.5.1. PARALLEL RL (RC)-CIRCUIT

Such equivalent circuit as in Fig. 5.5.1,*a* is usually as equivalent of load. The same circuit could be used for transformer with active load if stray inductances are negotiated.

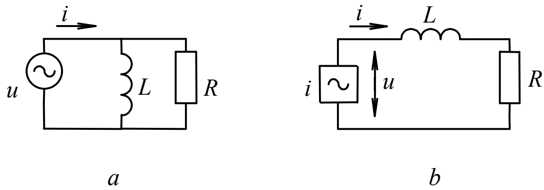


Fig. 5.5.1

To calculate rms source current value we utilize direct method ADE1. The differential equation for current is

$$i = \frac{u}{R} + \frac{1}{L}\int u dt = \frac{u}{R} + \frac{\overline{u}}{L}. \qquad (5.5.1)$$

Algebraization of obtained equation by ADE1 gives

$$I^2 = \frac{U^2}{R^2} + \left(\frac{\overline{U}}{L}\right)^2 = U^2\left\{\frac{1}{R^2} + \frac{U_{(1)}^2\left[1+\overline{K}_{\text{h}}^2\right]}{U^2\omega^2 L^2}\right\} =$$
$$= U^2\left\{\frac{1}{R^2} + \left(\frac{1}{\omega L}\right)^2 \frac{1+\overline{K}_{\text{h}}^2}{1+K_{\text{h}}^2}\right\}. \qquad (5.5.2)$$

An equation for modulus of generalized resistance of parallel RL - circuit for given voltage value follows from previous equation:

$$\left|Z_{RL}\right| = \frac{U}{I} = \sqrt{\left[\frac{1}{R^2} + \left(\frac{1}{\omega L}\right)^2 \frac{1+\overline{K}_{\text{h}}^2}{1+K_{\text{h}}^2}\right]^{-1}}. \qquad (5.5.3)$$

Similar conversions for parallel RC - circuit produce an equation for modulus of generalized resistance

$$\left|Z_{RC}\right| = \frac{U}{I} = \sqrt{\left[\frac{1}{R^2} + \left(\omega C\right)^2 \frac{1+\tilde{K}_{\text{h}}^2}{1+K_{\text{h}}^2}\right]^{-1}} . \tag{5.5.4}$$

Obtained equations for modules of generalized resistances in considered electrical circuits with nonsinusoidal voltage differ from modules of complex resistance of the same circuits with sinusoidal currents due to the presence of additional members of reactive resistance components of the branches that consider frequency spectrum of nonsinusoidal voltage (($\overline{K}_{\text{h}}, \tilde{K}_{\text{h}}$ together with $K_{\text{г}}$).)

5.5.2. SERIES RL (RC) - BRANCH

Such equivalent RL – circuit as in Fig. 5.5.1,*b* is used to equivalent the main for higher harmonics at the calculation of how valve converter affects the main; the converter in this case represented as power source with given current shape. Differential equation for the input voltage looks like

$$u = L\frac{di}{dt} + Ri . \tag{5.5.5}$$

Algebraization of this equation results in following equation for rms value of input voltage:

$$U^2 = L^2\tilde{I}^2 + R^2 I^2 = I^2\left[R^2 + \left(\omega L\right)^2 \frac{1+\tilde{K}_{\text{c.h}}^2}{1+K_{\text{c.h}}^2}\right]. \tag{5.5.6}$$

From (5.5.6) a formula for modulus of generalized resistance of series RL – circuit is obtained

$$\left|Z_{R-L}\right| = \frac{U}{I} = \sqrt{R^2 + \left(\omega L\right)^2 \frac{1+\tilde{K}_{\text{c.h}}^2}{1+K_{\text{c.h}}^2}} . \tag{5.5.7}$$

The same conversions for series RC-circuit gives an equation for modulus of generalized complex resistance

$$|Z_{R-C}| = \frac{U}{I} = \sqrt{R^2 + \left(\frac{1}{\omega C}\right)^2 \frac{1+\overline{K}_{\text{c.h}}^2}{1+K_{\text{c.h}}^2}}\,. \tag{5.5.8}$$

In contrast to parallel circuit with given voltage in series circuit with given current additional members at reactive resistances in (5.5.7) and (5.5.8) consider weighted frequency spectrum of nonsinusoidal current ($\tilde{K}_{\text{c.h}}$, $\overline{K}_{\text{c.h}}$, together with current harmonic coefficient $K_{\text{c.h}}$).

It is typical that such constructive attribute of linear electrical circuits with sinusoidal voltage s and currents as modulus of complex resistance has analogous *modulus of generalized (complex) resistance* in considered circuits of 1-st order with nonsinusoidal voltage and currents.

Second constructive attribute of electrical circuits with sinusoidal currents is vector diagrams that allow visual demonstration of qualitative correlation current and voltage vectors in electrical circuit. By in classical theory of electrical technique vector representation is possible only for sinusoidal voltages and currents. But such section of contemporary mathematics as functional analysis allows representing nonsinusoidal periodic functions as vectors by introducing such terms as vector and phase norms for scalar and vector multiplications [51]. In general case the norm of a function (that is considered in a vector in infinitely dimensional space) is defined such as:

$$X = \left[\frac{1}{T}\int_0^T |x|^p \, dt\right]^{\frac{1}{p}}, \tag{5.5.9}$$

for $p = 2$ the equation gives rms norm called in electrical technique root mean square value of the function; for $p = 1$ average norm by modulus and for $p = \infty$ - extreme function value (majorant norm).

Phase angle between two vectors (for $p = 2$) that represent periodical functions x and y is defined via their scalar product (x, y):

$$\cos\varphi = \frac{\frac{1}{T}\int_0^T xy\,dt}{XY} = \frac{(x,y)}{XY}\,. \tag{5.5.10}$$

Then for considered above series RL-circuit a vector diagram for norms of voltage U and current I would look the same as in case with circuit with sinusoidal voltage and current, but shift angle between vectors **U** and **I** accordingly to (5.5.10) will be calculated as

$$\cos\varphi = \frac{\frac{1}{T}\int_0^T uidt}{UI} = \frac{P}{UI} = \frac{I^2 R}{UI} = \frac{I}{U}R = \frac{R}{\left|Z_{R-L}\right|}, \qquad (5.5.11)$$

i.e. the angle is defined by known formula of *symbolic method* with correspondent replacement of complex resistance modulus to generalized complex resistance modulus.

5.5.3. PARALLEL RLC - CIRCUIT

Such circuit is used to equivalent RL – load (Fig. 5.5.2,a) at electrical circuit node where also presented compensation capacitors. Differential equation for current in the main looks like

$$i = \frac{u}{R} + \frac{1}{L}\int udt + C\frac{du}{dt}. \qquad (5.5.12)$$

After the algebraization the correlation for vector norms is obtained:

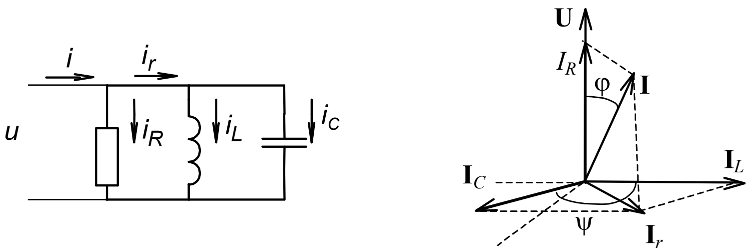
$$I^2 = U^2\left(\frac{1}{R^2} - 2\frac{C}{L}\right) + \left(\frac{1}{\omega L}\right)^2 \overline{U}^2 + (\omega C)^2 \overline{\overline{U}}^2. \qquad (5.5.13)$$

After transformations considering (5.5.2)-(5.5.4) we obtain

$$\left|Z_{RLC}\right| = \frac{U}{I} = \sqrt{\left[\left(\frac{1}{R^2} - 2\frac{C}{L}\right) + \left(\frac{1}{\omega L}\right)^2 \frac{1+\overline{K}_{\mathrm{h}}^2}{1+K_{\mathrm{h}}^2} + (\omega C)^2 \frac{1+\overline{\overline{K}}_{\mathrm{h}}^2}{1+K_{\mathrm{h}}^2}\right]^{-1}}.$$

(5.5.14)

It is typical that in complex resistance modulus there is not only correction of reactive resistances as in 1-st order circuit but also a correction of active part of full resistance. This result becomes obvious with help of vector diagram of considered circuit shown in Fig. 5.5.2,b.



a *b*

Fig. 5.5.2

Phase shift between current vectors $\mathbf{I}_L$ and $\mathbf{I}_C$ is not equal to 180°, but it is defined by (5.5.10) as:

$$\cos\varphi=\frac{\left(i_L,i_C\right)}{I_L I_C}=\frac{\left(\frac{1}{L}u,C\frac{du}{dt}\right)}{\frac{1}{L}\overline{U}C\overline{U}'}=-\frac{U^2}{\overline{U}\overline{U}'}=-\frac{1+K_h^2}{\sqrt{1+\overline{K}_h^2}\sqrt{1+K_h'^2}},\quad(5.5.15)$$

at that current vectors $\mathbf{I}_L$ и $\mathbf{I}_C$ lie on a plane that is perpendicular to voltage vector *u*, because correspondent scalar products are equal to zero:

$$(u,i_L)=0,\quad (u,i_C)=0.$$

Shift angle between current and voltage vectors at the input of the circuit is calculated as

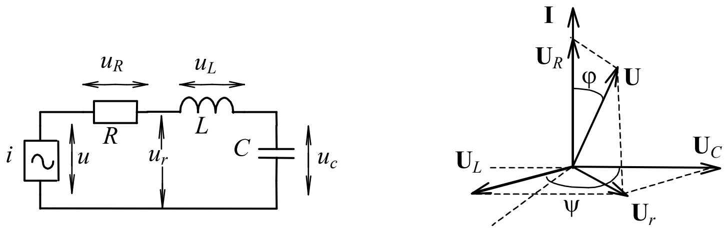
$$\cos\varphi=\frac{(u,i)}{UI}=\frac{(u,i_R)}{UI}=\frac{U^2}{UIR}=\frac{Z_{RLC}}{R}=\frac{\frac{1}{R}}{\frac{1}{Z_{RLC}}}.\qquad(5.5.16)$$

Thus, current and voltage vector of circuit with nonsinusoidal processes are locates not on the plane but in the space, here in 3-dimensional one [53]. The aspects of selection of different bases for such spaces are minutely described in [33].

5.5.4. SERIES RLC - CIRCUIT

Differential equation for input voltage of such circuit (Fig. 5.5.3, a) that is fed from current source *i*, looks like

$$u = iR + L\frac{di}{dt} + \frac{1}{C}\int idt\,. \qquad (5.5.17)$$



а *б*

Fig. 5.5.3

Its algebraization results in correlation for vector norms:

$$U^2 = I^2\left(R^2 - 2\frac{L}{C}\right) + L^2\tilde{I}^2 + \frac{1}{C^2}\bar{I}^2\,. \qquad (5.5.18)$$

After conversion for complex resistance modulus $|Z_{R-L-C}|$ of series *RLC*-circuit following equation is obtained

$$\left|Z_{R-L-C}\right| = \frac{U}{I} = \sqrt{\left(R^2 - 2\frac{L}{C}\right) + (\omega L)^2\frac{1+\tilde{K}_{\text{h}}^2}{1+K_{\text{h}}^2} + \left(\frac{1}{\omega C}\right)^2\frac{1+\bar{K}_{\text{h}}^2}{1+K_{\text{h}}^2}}\,. \qquad (5.5.19)$$

Vector diagram for series RLC – circuit is shown in Fig. 5.5.3,b. Inductance and capacitance voltage vectors $\mathbf{U}_L$ and $\mathbf{U}_C$ lie on the plane that is perpendicular input current vector; the angle between these vectors is defined from the correlation similar to (5.5.15):

$$\cos\varphi = \frac{(u_L, u_C)}{U_L U_C} = \frac{\left(L\frac{di}{dt}, \frac{1}{C}\bar{i}\right)}{U_L U_C} = -\frac{I^2}{\tilde{I}\bar{I}} = -\frac{1+K_{\text{c.h}}^2}{\sqrt{1+\bar{K}_{\text{c.h}}^2}\sqrt{1+\tilde{K}_{\text{c.h}}^2}}\,.$$

(5.5.20)

Shift angle between input voltage and current **U** and **I** is

$$\cos\varphi = \frac{(u, i)}{UI} = \frac{(u_R, i)}{UI} = \frac{I^2 R}{UI} = \frac{R}{|Z_{R-L-C}|}. \qquad (5.5.21)$$

Equations (5.5.20) as well as (5.5.21) similar to those for circuits with sinusoidal currents but considering correspondent correction of generalized complex resistance modules.

From formulas for generalized complex resistance modules one constructive generalization could be made by development of approaches from [54, 55]. Let's introduce four definitions of reactive resistance frequency (names are conditional yet):

–inductive resistance frequency when current is given accordingly to (5.5.7)

$$\omega_{LI} = \omega\sqrt{\frac{1+\left(\tilde{K}_{\text{c.h}}\right)^2}{1+K_{\text{c.h}}^2}}\,; \qquad (5.5.22)$$

–inductive resistance frequency when voltage is given accordingly to (5.5.3)

$$\omega_{LU} = \omega\sqrt{\frac{1+\left(K_{\text{h}}\right)^2}{1+\bar{K}_{\text{h}}^2}}\,; \qquad (5.5.23)$$

–capacitive resistance frequency when current is given accordingly to (5.5.8)

$$\omega_{CI} = \omega\sqrt{\frac{1+K_{\text{c.h}}^2}{1+\left(\bar{K}_{\text{c.h}}\right)^2}}\,; \qquad (5.5.24)$$

–capacitive resistance frequency when voltage is given accordingly to (5.5.4)

$$\omega_{CU} = \omega\sqrt{\frac{1+\left(\tilde{K}_{\text{h}}\right)^2}{1+K_{\text{h}}^2}}\,. \qquad (5.5.25)$$

This similarly to definition of reactive resistance for sinusoidal current to find equivalent reactive resistance for nonsinusoidal current, i.e. to introduce two reactances for inductance

$$X_{LI} = \omega_{LI} L, \qquad X_{LU} = \omega_{LU} L \qquad (5.5.26)$$

and two reactances for capacitance as well

$$X_{CI} = \frac{1}{\omega_{CI} C}, \qquad X_{CU} = \frac{1}{\omega_{CU} C}. \tag{5.5.27}$$

5.5.5. CIRCUITS OF *N*-TH ORDER

Exact value of solution norm for system of n-th order could be obtained if its differential equation looks like:

$$x = \sum_{m=0}^{n} a_m \frac{d^m y}{dt^m}, \tag{5.5.28}$$

where $x(y)$ – reaction (action).

Equation (5.5.28) could not be modeled by electrical circuit with passive elements but could be visually equivalented by the automatic regulation structure.

The algebraization of (5.5.28) results in following correlation for norms:

$$X^2 = \sum_{m=0}^{n} a_{mk}^2 (\tilde{Y}^{(m)})^2 = Y_2^2 \sum_{m=0}^{n} a_{mk}^2 \frac{1+\left(\tilde{K}_{\mathrm{h}}^{(m)}\right)^2}{1+K_{\mathrm{h}}^2}. \tag{5.5.29}$$

At that a_{mk}^2 are defined via coefficients a_m by (5.1.12).

From (5.5.29) for system transfer coefficient modulus for norm the next equation if following

$$K = \frac{X}{Y} = \sqrt{\sum_{m=0}^{n} a_{mk}^2 \frac{1+\left(\tilde{K}_{\mathrm{h}}^{(m)}\right)^2}{1+K_{\mathrm{h}}^2}}. \tag{5.5.30}$$

Thus, the presence of exact solutions in ADE1 method let us think that further development of this method is perspective. The question about direct calculation of reaction norm still remains unsolved in more general case with differential equation of circuit of n-th order when this equation unlike (5.5.28) contains derivatives of the reaction x in its left part. Successful solution of this task would be a creation of *generalized symbolic calculation method* for rms values of the variables in systems with nonsinusoidal voltage and currents. The value of such a result couldn't be overestimated.

5.6. DISCRETE MODELS OF POWER CONVERTERS

Extreme operating speed, control and accuracy resources of power electronics converters could only be realized by utilizing exact models of the converters for control processes that consider the discontinuity of the converter. Section 5.6.1 presents an approach to building discrete model of the converter on the basis of the equations in finite difference; section 5.6.2 introduce general approach to building the model of valve converter in small deviations mode as a pulse system model in terms of z- conversion.

5.6.1. COMPOSITION OF DIFFERENCE EQUATIONS AND THEIR SOLUTION FOR RECTIFIER

Let's begin consideration of difference equation method from its application to transient mode calculation at the rectifier output; rectifier is presented by simple equivalent circuit composed of e.m.f. source loaded by filter inductive resistance, active resistance (losses) and anti-e.m.f of load source connected in series.

The curves of rectified voltage and current in m-phase circuit of valve converter for phase regulation are presented in Fig. 5.6.1,a. Fig.5.6.1,b shows the ordinates of load current in moments of state changing. Changes of these ordinates from state to state define the law of variation of initial conditions for current in each state. Such a function that is defined for discrete equidistant moments of time

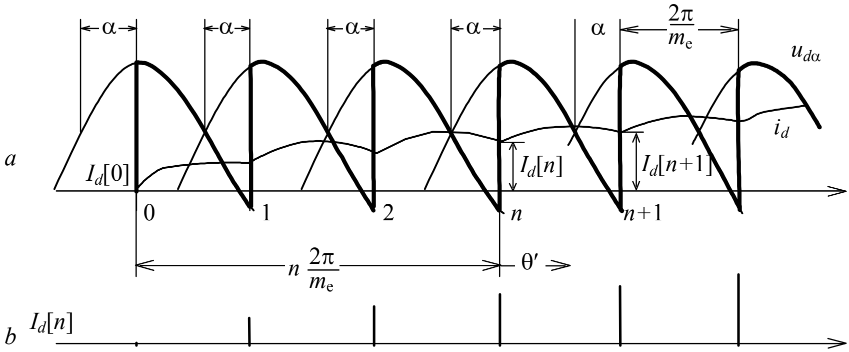


Fig. 5.6.1

$\left(\vartheta = n\dfrac{2\pi}{m}, \text{где } n = 0, 1, 2, 3, ...\right)$, is called lattice function; the system whose processes could be represented by lattice functions are called discrete or pulse functions [56].

Dynamical behavior of such systems is described by difference equations that in most cases similar to differential equations that describe behavior of continuous systems. There are two well-known forms of difference equations. First one similar to differential equation of l-th order looks like:

$$b_l \Delta^l y[n] + b_{l-i} \Delta^{l-1} y[n] + ... + b_1 \Delta y[n] + b_0 y[n] = f[n], \tag{5.6.1}$$

where $\Delta y[n] = y[n+1] - y[n]$ – first order difference; $\Delta^2 y[n] = \Delta y[n+1] - \Delta y[n] = y[n+2] - 2y[n+1] + y[n]$ – second order difference and so on., i.e. the differences of lattice functions similar to derivatives of continuous functions; $b_l, b_{l-1}, ..., b_0$ - coefficients; $f[n]$- lattice action function.

Second type of difference equation that is obtained from the firs type by substitution of all the differences by lattice functions often appear to be more convenient than firs one, and it looks like

$$a_l y[n+l] + a_{l-i} y[n+l-1] + ... + a_1 y[n+1] + a_0 y[n] = f[n]. \tag{5.6.2}$$

Equation (5.6.2) could be treated as recurrent correlation that for given initial conditions $y[0], y[1], y[2], ..., y[l-1]$ (for difference l-th order difference equation there are l initial conditions) lets consequent finding of $y[l]$, $y[l+1], ..., y[n+l]$. Such a technique is often used for approximate integration of differential equation when it is substituted by difference equation of second type; the last one are further solved or used as recurrent correlation for consequent calculation of $I_d[1], I_d[2], ..., I_d[n]$.

Now let's start to compose difference equation that describes the behavior of lattice function of rectified current that defines initial conditions for it in each state. Let's compose differential equation for rectified current on the n-th state existence interval that is distanced from the time origin that is also a start of transition process for time interval equal to $n\frac{2\pi}{m}$. "Private" time within n-th state could be denoted as $\vartheta' = \vartheta - n\frac{2\pi}{m}$. Then considering the equivalent circuit of valve converter we will obtain ($R_{\text{fw}} = 0$, ΔU_0 consider in U_0)

$$X_d \frac{di_d}{d\vartheta'} + i_d R_d = \sqrt{2}U \cos\left(\vartheta' - \frac{\pi}{m} + \alpha\right) - U_0. \tag{5.6.3}$$

Solution of (5.6.3)

$$i_d = \frac{\sqrt{2}U}{Z_d} \cos\left(\vartheta' - \frac{\pi}{m} + \alpha - \varphi_d\right) - \frac{U_0}{R_d} + A_1 e^{-\frac{\vartheta'}{\omega\tau_d}}, \tag{5.6.4}$$

where

$$\omega\tau_d = \frac{X_d}{R_d}; \quad Z_d = \sqrt{R_d^2 + X_d^2}; \quad \varphi_d = \text{arctg}\frac{X_d}{R_d}. \tag{5.6.5}$$

Integration constant A_1 is defined from initial conditions for current i_d in considered state:

$$i_d = I_d[n] \quad \text{при} \quad \vartheta' = 0. \tag{5.6.6}$$

Then

$$A_1 = I_d[n] - \frac{\sqrt{2}U}{Z_d} \cos\left(-\frac{\pi}{m} + \alpha - \varphi_d\right) + \frac{U_0}{R_d}, \tag{5.6.7}$$

and by substituting (5.6.7) in (5.6.4), we obtain

$$i_d = \frac{\sqrt{2}U}{Z_d}\cos\left(\vartheta' - \frac{\pi}{m} + \alpha - \varphi_d\right) - \frac{U_0}{R_d} +$$
$$+ \left\{ I_d[n] + \frac{U_0}{R_d} - \frac{\sqrt{2}U}{Z_d}\cos\left(-\frac{\pi}{m} + \alpha - \varphi_d\right)\right\} e^{-\frac{\vartheta'}{\omega\tau_d}}. \tag{5.6.8}$$

Ia one add to (5.6.8)

$$i_d = I_d[n+1] \quad \text{at} \quad \vartheta' = \frac{2\pi}{m}, \tag{5.6.9}$$

i.e. current value at the end of the n-th state is the initial current value for current of n+1 – th state due to the discontinuity of the current curve, so (5.6.8) turns to difference equation that connects initial conditions for current and following states:

$$I_d[n+1] = \frac{\sqrt{2}U}{Z_d}\cos\left(\frac{\pi}{m} + \alpha - \varphi_d\right) - \frac{U_0}{R_d} +$$
$$+ \left\{ I_d[n] + \frac{U_0}{R_d} - \frac{\sqrt{2}U}{Z_d}\cos\left(\alpha - \frac{\pi}{m} - \varphi_d\right)\right\} e^{-\frac{2\pi}{m\omega\tau_d}}. \tag{5.6.10}$$

Lets transform difference equation (5.6.10) to standard form (5.6.2)

$$I_d[n+1] - e^{-\frac{2\pi}{m\omega\tau_d}} I_d[n] =$$
$$= \frac{\sqrt{2}U}{Z_d}\left[\cos\left(\frac{\pi}{m} + \alpha - \varphi_d\right) - \cos\left(\alpha - \frac{\pi}{m} - \varphi_d\right) e^{-\frac{2\pi}{m\omega\tau_d}}\right] -$$
$$- \frac{U_0}{R_d}\left(1 - e^{-\frac{2\pi}{m\omega\tau_d}}\right). \tag{5.6.11}$$

We obtained difference equation of 1-st order with constant coefficients and constant right part that could be denoted as K_1. The solution of difference equation could be done whether by classical or operator method by utilizing special discrete Laplace transformation or B

–transformation [56]. Within this transformation the representation and original of lattice functions are connected by following formulas

$$F^*(q) = D\{f[n]\} = \sum_{n=0}^{\infty} f[n]e^{-qn}, \tag{5.6.12}$$

$$f[n] = D^{-1}\{F^*(q)\} = \frac{1}{2\pi j}\int_{c-j\pi}^{c+j\pi} F^*(q)e^{qn}dq. \tag{5.6.13}$$

Let's define D-representations of some required for solution of (5.6.11) lattice functions:

1) *D*-representation of constant *K*

$$D\{K\} = \sum_{n=0}^{\infty} Ke^{-qn} = K\frac{1}{1-e^{-q}} = K\frac{e^q}{e^q-1} \tag{5.6.14}$$

considering that the sum forms infinite decreasing geometrical progression with denominator e^{-q} [56];

2) *D*- representation of lattice function

$$D\{e^{-\theta n}\} = \sum_{n=0}^{\infty} e^{-\theta n}e^{-qn} = \sum_{n=0}^{\infty} e^{-(\theta+q)n} = \frac{1}{e^{-(\theta+q)}-1} = \frac{e^q}{e^{-\theta}-e^q}; \tag{5.6.15}$$

3) *D*- representation of shifted lattice function

$$\begin{aligned} D\{I[n+1]\} &= \sum_{n=0}^{\infty} I[n+1]e^{-qn} = \sum_{m=1}^{\infty} I[m]e^{-q(m-1)} = \\ &= e^q\sum_{m=0}^{\infty} I[m]e^{-qm} - e^q I[0] = e^q[\{I[n]\} - I[0]]; \end{aligned} \tag{5.6.16}$$

4) *D*- representation of partial sum of lattice functions as shown in [56] is

$$D\left\{\sum_{m=0}^{n} I[m]\right\} = \frac{I(q)}{e^q-1}. \tag{5.6.17}$$

Note that together with *D*-representation of lattice functions defined by (5.6.12) vary popular is z –representation of lattice functions [57] that is connected with *D*-representation by equality

$$z = e^q .$$

By applying *D*-representation to equation (5.6.11) we obtain

$$e^q \left\{ I_d^*(q) - I_d[0] \right\} - e^{-\frac{2\pi}{m\omega\tau_d}} I_d^*(q) = K_1 \frac{e^q}{e^q - 1}, \qquad (5.6.18)$$

where I_d [0] –initial value of rectified current at the moment of the beginning of transition process (when valve converter starts operating the I [0] = 0 as it is shown in Fig. 5.6.1,a, but in general case I [0] ≠ 0, for example for transient process in operating valve converter from sudden change of α).

Solution for current representation from (5.6.18)

$$I_d^*(q) = K_1 \frac{e^q}{(e^q - 1)\left(e^q - e^{-\frac{2\pi}{\omega m\tau_d}}\right)} + I_d[0] \frac{e^q}{e^q - e^{-\frac{2\pi}{m\omega\tau_d}}}. \qquad (5.6.19)$$

By transferring to originals considering (5.6.51) and (5.6.17) we obtain

$$\begin{aligned} I_d[n] &= K_1 \sum_{n=0}^{n} e^{-\frac{2\pi}{\omega m\tau_d} n} + I_d[0] e^{-\frac{2\pi}{m\omega\tau_d}} = \\ &= \frac{K_1}{1 - e^{-\frac{2\pi}{m\omega\tau_d}}} \left(1 - e^{-\frac{2\pi}{\omega m\tau_d} n}\right) + I_d[0] e^{-\frac{2\pi}{m\omega\tau_d} n}. \end{aligned} \qquad (5.6.20)$$

First member in (5.6.20) that doesn't depend on *n,* gives steady state component (forced component); second and third components – are free components the decrease with increasing of *n*.

Sometimes the exponent of free component decreasing could be conveniently represented as:

$$e^{-\frac{2\pi}{m\omega\tau_d} n} = e^{-\frac{T}{m\tau_d} n} = e^{-\frac{n}{\overline{\tau}_d}}, \qquad (5.6.21)$$

where $\overline{\tau}_d = \dfrac{\tau_d}{T_1} = \dfrac{\tau m}{T}$ – relative time constant, expressed by quantization step T_1 of scheme processes.

Then time interval of forced mode establishment is defined by the number of quantization steps n, as usual

$$n \geq (3-4)\overline{\tau}_d. \tag{5.6.22}$$

We should also note that (5.6.20) give rectified current minorant; the minorant equation could be obtained by substitution in (5.6.20) n by ϑ / T_1.

As in investigation of continuous dynamical systems behavior by differential equation methods in time domain or by transfer functions on complex plane the investigation of discrete systems could also be performed by two ways: by utilizing deference equations in discrete time domain as it was done above or by utilizing pulse transfer functions for discrete systems. Such function is defined as relation of discrete representation of output variable to discrete representation of input action. Here the action is the firing angle α, that is introduced nonlinearly (in cosine argument) in the equation of output current (5.6.11).That is why it is necessary to linearize this equation for small control deviation mode $\Delta\alpha$.

Ia discrete rectifier model is used to control its rectified output then the most suitable model is not one for instant values but one average values between commutations. To do this it is necessary to integrate the equation (5.6.8) within n-th and n+1 time interval between commutations; this would result in difference equation such as (5.6.11) but with another right part:

$$I_d[n+1] - I_d[n]e^{-\frac{2\pi}{m\omega\tau_d}} = K_2.$$

Then one should linearize obtained difference equation relative to small deviation of firing angle α, and perform D-transformation of linearized difference equation. This would define pulse transfer function for current-control channel that would be look like [58]

$$W^*(q) = \frac{\Delta I_d(q)}{\Delta\alpha(q)} = -2\sin\frac{\pi}{m}\sin\frac{2\pi}{m}\frac{1-e^{-\frac{2\pi}{m\omega\tau_d}}}{e^q - e^{-\frac{2\pi}{m\omega\tau_d}}}\alpha[0]. \tag{5.6.23}$$

The knowledge of pulse transfer function of the rectifier allows synthesizing by known methods a current regulator, verifying closed loop system stability and evaluating the quality of transition processes of the control [58]. Next section consider general method of stability analysis for small pulse-width dc regulator; there is also an example of application of this method for 2-nd order system. The operation principle of pulse – width dc voltage regulator is described in chapter 7.

5.6.2**. THE MODEL OF DC PULSE-WIDTH CONVERTER AS A PULSE SYSTEM

To represent pulse-width converter by width-pulse model system (WPM) it is necessary to formalize the method of its obtaining. There is known method for one side pulse – width modulation [59]. Here presented the further development of this method for double-sided pulse width modulation [60]; the obtaining of double channel (two parallel channels with pulse elements) pulse model performed in formal way by representation of WPM as a relay system using Y.Z. Tsipkin methodology [61]. More general task requires formalization за calculation procedure that would be oriented on computer application because of difficult analysis of WPM that is represented by equations of order greater than two.

In general, the solution of named task ads up to performing of series of formalized stages: a) obtaining of pulse model for small deviations mode when WPM is considered as relay system; b) composition of characteristic equation with help of z - transformation for pulse system; c) performing of computer stability analysis.

Obtaining of pulse model. Generalized functional circuit of closed-loop WPM is presented in Fig. 5.6.2,a; the diagrams explaining its operation principle as well as operation principles of circuits obtained from this circuit (Fig. 5.6.2, b,c) are shown in Fig. 5.6.3. Fig. 5.6.2 uses following indications: S – current source; PWC – pulse width converter; L – load; R – regulator; $r(t)$ – reference signal.

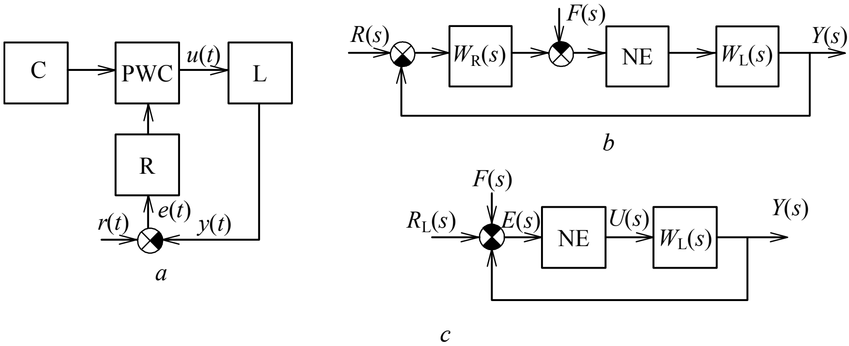


Fig. 5.6.2

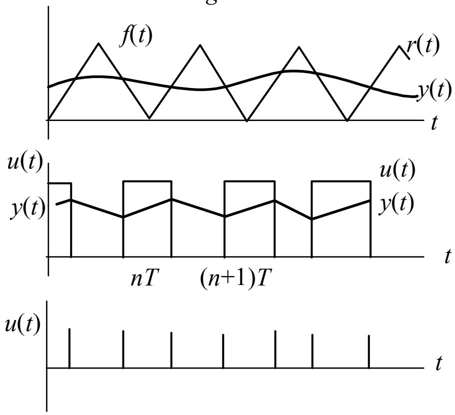


Fig. 5.6.3

Block diagram of such automation control system is shown in Fig. 5.6.2, b; after the conversions to standard type this circuit is presented in Fig. 5.6.2, c.

Here $W_r(s)$, $W_L(s), W_c(s)$ - correspondent transfer functions of regulator, load, and continuous part $W_c(s) = W_r(s)W_L(s)$, NE – relay nonlinear element that simulates PWC; $f(t)$ – sawtooth; $r_n(t)$ – normalized reference signal. The type of NE doesn't depend on PWC structure and type of PWM. In case of irreversible PWC and unipolar PWM the characteristic of NE is similar to one of two-position nonsymmetrical relay element without dead space.

In steady state at the relay element input following periodical signal is presented

$$e(t) = r_n(t) - y(t) - f(t), \tag{5.6.24}$$

on the output of the relay element a periodic square pulse sequence is formed (uni- or bipolar depending on the type of modulation).

The consideration of system stability "in small" ads up to the investigation of small deviations from steady state at the relay element output, that in [61] written as follows

$$\dot{u}(t) = \frac{k_p}{|\dot{e}-(T)|}\sum_{k=0}^{\infty}\delta(t-kT) + \frac{k_p}{|\dot{e}-(\gamma T)|}\sum_{k=0}^{\infty}\delta(t-(k+\gamma)T), \tag{5.6.25}$$

where k_r - parameter of relay element (pulse double amplitude); $\delta(t)$ - delta function; $kT, (k+\gamma)T$ –moments of relay element switching (k = 0, 1, 2, …); $|\dot{e}-(T)|$, $|\dot{e}-(\gamma T)|$ – absolute values of limits on the left when signal derivative switches defined by (5.6.24).

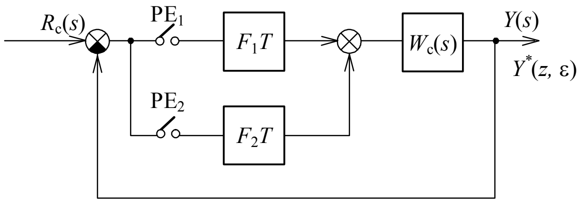
In converters technique terminology the value

$$F = \frac{1}{T}\frac{1}{|\dot{e}(t)|} = \frac{1}{T}\frac{1}{1+\dot{y}(t)}$$

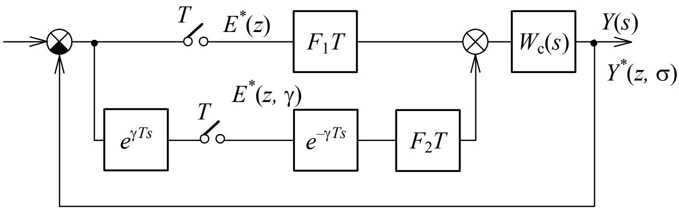
is called *pulse factor* a it is calculated constant input reference and saw-tooth voltage of unity amplitude [59.62.63]. Pulse factor defines the degree of deviation of control system transfer ratio due to instability (rippling) of error signal. It could be obtained from general equation accordingly to (5.6.24)

$$F = \frac{1}{T}\frac{1}{|\dot{e}(t)|} = \frac{1}{|\dot{r}_p(t) - \dot{y}(t) - \dot{f}(t)|}\frac{1}{T}. \tag{5.6.26}$$

Equation (5.6.25) corresponds to double channel pulse system (Fig. 5.6.4, a) with pulse elements (PE) that have operation period T and phase shift γT .



a



b

Fig. 5.6.4

Transfer ratios of the channels are

$$F_1T = \frac{k_p}{|\dot{e}-(T)|}, \quad F_2T = \frac{k_p}{|\dot{e}-(\gamma T)|}.$$

Thus, to obtain pulse model by formal way to analyze system stability the relay element could be substituted by parallel circuits with synchronous pulse elements in each circuit which number is equal to number of regulated (in time domain) switchings of relay element within a period. The moments of pulse elements switching clash with correspondent moments of relay element switching; partial transfer coefficients of each circuit are defined by values of generalized pulsation factor at the moments of relay element switchings. Note that the relay element could be of any type including ones with hysteresis. At that hysteresis value is absent in pulse model because the stability of oscillation of given shape is under the investigation.

To bring pulse system to a standard form with cophased and synchronous pulse elements the forestalling and delaying links are introduced into each channel [57]; phase shift of these links is defined by the shift of the operation moment of correspondent pulse element. Final circuit of pulse system in case of double sided unipolar PWM is presented in Fig. 5.6.4,b.

Composition of characteristic equation. Let's obtain the equation for transfer coefficient of each channel of pulse system for transfer function of linear part such as

$$W(s) = \frac{P(s)}{Q(s)} = \sum_{i=1}^{n} \frac{c_i}{s+p_i}, \quad c_i = \frac{P(-p_i)}{Q(-p_i)}, \quad p_i = \frac{1}{T_i}. \qquad (5.6.27)$$

For unity amplitude of sawtooth voltage and outputs of relay element the derivatives of signal (5.6.24) in moments of switching are defined by

$$\left|\dot{e}-(T)\right|=\frac{2}{T}+\sum_{i=1}^{n}\frac{c_i(e^{\gamma T_{pi}}-1)}{1-e^{-T_{pi}}}e^{-T_{pi}}, \tag{5.6.28}$$

$$\left|\dot{e}-(\gamma T)\right|=\frac{2}{T}+\sum_{i=1}^{n}\frac{c_i(e^{-\gamma T_{pi}}-e^{-T_{pi}})}{1-e^{-T_{pi}}}. \tag{5.6.29}$$

To simplify formulas we introduce next labeling

$$F_1T=\frac{1}{\left|\dot{e}-(T)\right|},\quad F_2T=\frac{1}{\left|\dot{e}-(\gamma T)\right|}.$$

The characteristic equation of the system looks like [64]

$$\begin{vmatrix}1+F_1TZ\{W(s)\} & F_2TZ\{e^{-\gamma Ts}W(s)\}\\ F_1TZ\{e^{\gamma Ts}W(s)\} & 1+F_2TZ\{W(s)\}\end{vmatrix}=0, \tag{5.6.30}$$

where

$$Z\left\{\sum_{i=1}^{n}\frac{c_i}{s+p_i}\right\}=\sum_{i=1}^{n}\frac{c_id_i}{z-d_i};\ Z\left\{e^{\gamma Ts}\sum_{i=1}^{n}\frac{c_i}{s+p_i}\right\}=\sum_{i=1}^{n}\frac{c_id_i^{\gamma}z}{z-d_i};$$

$$Z\left\{e^{-\gamma Ts}\sum_{i=1}^{n}\frac{c_i}{s+p_i}\right\}=\sum_{i=1}^{n}\frac{c_id_i^{1-\gamma}}{z-d_i};\ d_i=e^{-T_{p_i}}.$$

After transformation of (5.6.30) we obtain

$$1+(F_1+F_2)T\sum_{i=1}^{n}\frac{c_id_i}{z-d_i}+F_1F_2T^2\sum_{i=1}^{n}\frac{A_i}{z-d_i}=0 \tag{5.6.31}$$

or

$$\prod_{i=1}^{n}(z-d_i)+(F_1+F_2)T\sum_{i=1}^{n}c_id_i\prod_{\substack{j=1\\ j\neq i}}^{n}(z-d_i)+$$

$$+F_1F_2T^2\sum_{i=1}^{n}A_i\prod_{\substack{j=1\\ j\neq i}}^{n}(z-d_i)=0, \qquad (5.6.32)$$

where

$$A_i=-c_i^2d_i+2c_id_i\sum_{\substack{j=1\\ j\neq i}}^{n}\frac{c_jd_j}{d_i-d_j}-c_id_i^{1+\gamma}\sum_{\substack{j=1\\ j\neq i}}^{n}\frac{c_jd_j^{1-\gamma}}{d_i-d_j}-c_id_i^{2-\gamma}\sum_{\substack{j=1\\ j\neq i}}^{n}\frac{c_jd_j^{\gamma}}{d_i-d_j}.$$

The equation (5.6.32) turns to normal form by decreasing degrees. To have a possibility to use Rays criteria, that is convenient for computer analysis let's apply bilinear transformation. By substitution in (5.6.32):

$$z=\frac{1+\omega}{1-\omega}. \qquad (5.6.33)$$

After simple transformation

$$\prod_{i=1}^{n}\left(\omega+\frac{1-d_i}{1+d_i}\right)+(F_1+F_2)T\sum_{i=1}^{n}\frac{c_id_i}{1+d_i}(1-\omega)\prod_{\substack{j=1\\ j\neq i}}^{n}\left(\omega+\frac{1-d_i}{1+d_i}\right)+$$

$$+F_1F_2T^2\sum_{i=1}^{n}\frac{A_i(1-\omega)}{1+d_i}\prod_{\substack{j=1\\ j\neq i}}^{n}\left(\omega+\frac{1-d_i}{1+d_i}\right)=0. \qquad (5.6.34)$$

To simplify the equation

$$l_i=\frac{d_i-1}{d_i+1}. \qquad (5.6.35)$$

Let's turn the characteristic equation as

$$B_0\omega^n+B_1\omega^{n-1}+\ldots+B_n=0, \qquad (5.6.36)$$

where

$$B_0 = 1-(F_1+F_2)T\sum_{i=1}^{n}\frac{c_i d_i}{1+d_i} - F_1F_2T^2\sum_{i=1}^{n}\frac{A_i}{1+d_i};$$

$$B_1 = b_1 + (F_1+F_2)T\sum_{i=1}^{n}\frac{c_i d_i}{1+d_i}(1-b_1^i) + F_1F_2T^2\sum_{i=1}^{n}\frac{A_i}{1+d_i}(1-b_1^i);$$

$$B_k = b_k + (F_1+F_2)T\sum_{i=1}^{n}\frac{c_i d_i}{1+d_i}(b_{k-1}^i - b_k^i) + F_1F_2T^2\sum_{i=1}^{n}\frac{A_i}{1+d_i}(b_{k-1}^i - b_k^i);$$

$$B_n = b_n + (F_1+F_2)T\sum_{i=1}^{n}\frac{c_i d_i}{1+d_i}b_{n-1}^i + F_1F_2T^2\sum_{i=1}^{n}\frac{A_i}{1+d_i}b_{n-1}^i;$$

(b_k, b_k^i are expressed via elementary symmetrical functions)

$$\prod_{i=1}^{n}(\omega - l_i) = \omega^n + b_1\omega^{n-1} + \ldots + b_n;$$

$$b_k = \frac{(-1)^k}{k}\cdot\begin{vmatrix} s_1 & 1 & 0 & \ldots & 0 \\ s_2 & s_1 & 2 & \ldots & 0 \\ s_3 & s_2 & s_1 & \ldots & 0 \\ \ldots & \ldots & \ldots & \ldots & 0 \\ s_k & s_{k-1} & \ldots & \ldots & s_1 \end{vmatrix}, \quad k = 1,\ldots,n;$$

$$s_1 = \sum_{i=1}^{n} l_i; \qquad s_1 = \sum_{i=1}^{n} l_i^2; \qquad s_1 = \sum_{i=1}^{n} l_i^2; \quad \ldots;$$

i means that efficient $(\omega - l_i)$ is absent in the multiplication product.

The characteristic equation for WPM stability analysis with integral regulator in found in the same manner. In steady state mode the value of constant input value is connected with the duration of the pulses by formula

$$r = \frac{P(0)}{Q(0)}\gamma. \tag{5.6.37}$$

The values of derivatives of the signals at the input of the relay element at the moments of switching are defined by following expressions

$$\left|\dot{x}-(T)\right| = \frac{2}{T} + \left[\gamma - \sum_{i=1}^{n} \frac{c_i}{p_i T_0} \frac{e^{-(1-\gamma)T_{pi}} - e^{-T_{pi}}}{1 - e^{-T_{pi}}}\right], \tag{5.6.38}$$

$$\left|\dot{x}-(\gamma T)\right| = \frac{2}{T} + \left[\sum_{i=1}^{n} \frac{c_i}{p_i T_0} \frac{e^{\gamma T_{pi}} - 1}{1 - e^{-T_{pi}}} - \gamma\right]. \tag{5.6.39}$$

Given transfer function of linear part looks like

$$W_n(s) = \frac{P(s)}{Q(s)T_0(s)} = \frac{c_0'}{s} + \sum_{i=1}^{n} \frac{c_i'}{s + p_i} = \sum_{i=1}^{n} \frac{c_i'}{s + p_i}, \tag{5.6.40}$$

where

$$c_0' = \frac{P(0)}{T_0 Q(0)}; \qquad c_i' = -\frac{c_i}{T_0 p_i}.$$

Further transformations to obtain characteristic equation are similar to those performed in (5.6.30)-(5.6.36). Characteristic equation for WPM with one sided modulation is obtained from general expressions by setting $F_2 = 0$ for leading edge modulation and $F_1 = 0$ for falling edge modulation. At that the similar equation for characteristic equation are obtained if raising edge modulation γ corresponds to falling edge modulation.

Stability analysis (example calculation). By utilizing obtained characteristic equation of WPM is possible to build stability thresholds in a domain of variable parameters with help of special software [64] but with the only difference that to reach the threshold to minimize calculation time one should not scan row of parameter table but one should perform perimeter rounding. In case of continuous parts of first or second orders the stability conditions are obtained in analytical form that is illustrated by the example given below.

Let's obtain stability conditions for PWC with continuous right part of second order:

$$W(s) = \frac{K}{(T_1 s + 1)(T_2 s + 1)}.$$

Coefficients F_1 and F_2 are defined from formulas (5.6.28) and (5.6.29):

$$F_1 = \left[2 + \frac{KT}{T_1 - T_2}\left(\frac{d_1^{\gamma} - d_1}{1 - d_1} - \frac{d_2^{\gamma} - d_2}{1 - d_2}\right)\right]^{-1},$$

$$F_2 = \left[2 + \frac{KT}{T_1 - T_2}\left(\frac{d_1^{1-\gamma} - d_1}{1 - d_1} - \frac{d_2^{1-\gamma} - d_2}{1 - d_2}\right)\right]^{-1}, \quad d_1 = e^{-\frac{T_1}{T}}, \quad d_2 = e^{-\frac{T_2}{T}}.$$

Characteristic equation of pulse model looks like

$$1 + (F_1 + F_2)\frac{KT(d_1 - d_2)z}{(T_1 - T_2)(z - d_1)(z - d_2)} -$$

$$-\frac{KT^2(d_1 - d_1^{1-\gamma}d_2^{\gamma} - d_1^{\gamma}d_2^{1-\gamma} + d_2)z}{(T_1 - T_2)^2(z - d_1)(z - d_2)}F_1F_2 = 0.$$

Let’s perform bilinear transform by (5.6.33). The characteristic equation will look like

$$A\omega^2 + B\omega + C = 0,$$

where

$$A = (1 + d_1)(1 + d_2) - (F_1 + F_2)\frac{KT(d_1 - d_2)}{(T_1 - T_2)} +$$

$$+\frac{K^2T^2}{(T_1 - T_2)^2}(d_1 - d_1^{1-\gamma}d_2^{\gamma} - d_1^{\gamma}d_2^{1-\gamma} + d_2)F_1F_2;$$

$$B = 2\left(1 - d_1d_2\right);$$

$$C = (1 - d_1)(1 - d_2) + (F_1 + F_2)\frac{KT(d_1 - d_2)}{(T_1 - T_2)} -$$

$$-\frac{K^2T^2F_1F_2}{(T_1 - T_2)^2}(d_1 - d_1^{1-\gamma}d_2^{\gamma} - d_1^{\gamma}d_2^{1-\gamma} + d_2).$$

For equation of second order the stability condition requires the coefficients to be greater than zero. Boundary gain coefficient could be defined from quadratic equation $A = 0$ for given values of $T_1/T, T_2/T, \gamma$. Conditions $B > 0$ and $C > 0$ are satisfied for all possible parameters.

Graphs for series of time constants values are shown in Fig. 5.6.5.

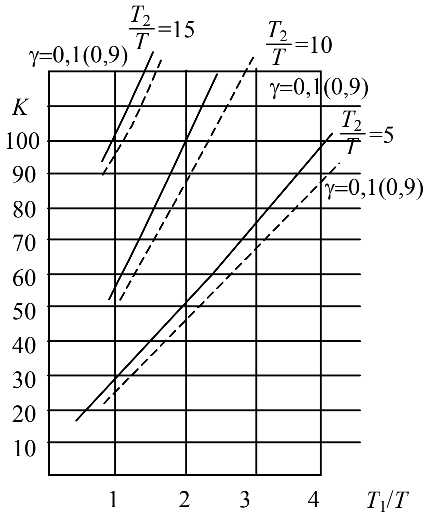


Fig. 5.6.5

In lots of important practice cases the rippling of output coordinate oa WPM are small comparing to the smooth component and they could be negotiated to simplify the calculations. In all the obtained equations pulse factor values are supposed to be equal to $1/2(F_1 = F_2 = 1/2)$. For example under the consideration correspondent close stability boundaries are drawn with dashed lines.

Thus, firstly on the basis of representation of pulse –width system by relay system the procedure of obtaining of pulse WPM model is formalized; this is necessary to investigate stability "in small" for any PWM type and continuous part of any order with simple poles. The equation for pulse factor of valve converter is also generalized. Secondly, the general expression of characteristic equation for closed-loop pulse system is obtained via its parameters is obtained; that equation by bilinear transformation driven to a form that is convenient for numerical calculation of stability thresholds by standard methodology using PC. For the system of second order the thresholds are obtained in analytical form. A method of simplified definition of stability thresholds by negotiating of WPM output coordinate rippling is proposed.

QUASTIONS

1. Which method of power factors calculation could be called direct ones?
2. What does producing equation sets?
3. How to build a solution for norm in ADE1 method for second level of assumption?
4. What methods one should apply to obtain closed expressions for integral harmonic coefficients $\overline{K}_{\mathrm{h}}^{(q)}$?
5. What do the last two letters stands for in abbreviation ADESS?

6. What are the advantages of ADESS method comparing to ADE one?
7. For what form of differential equations it is possible to achieve exact solutions in ADE and ADESS methods?
8. What are the advantages of calculation by ADESS(1) method comparing to symbolic method?
9. Which equations are called difference ones?
10. What is the pulse factor in power electronics?

EXERCISES

1. Build a solution for second level of assumption in ADE method for second order circuit.
2. Find an error of first level of assumption of ADE method for series RL – circuit for typical action voltage.
3. Build a solution for second level of assumption by ADESS1 method for second order circuit.
4. Define the influence of 3-phase voltage nonsymmetry on degree of nonsymmetry of currents in 3-phase symmetrical load of first order.
5. Define the influence of 3-phase 1-st order load nonsymmetry on the nonsymmetry of currents of symmetrical 3-phase voltage.
6. Define by ADESS(1) method the equations of resistances of source voltage influence upon the state space currents in the 2-nd order system.
7. Derive a difference equation for current of ideal rectifier with zero valve and RL – load.
8. Derive a difference equation for average values of rectified current of ideal rectifier.
9. Linearize the difference equation obtained in exercise 8 relative to firing angle α.
10. Build pulse system model using linearized equation (refer to exercise 9).

Chapter 6

ELECTROMAGNETIC COMPATIBILITY OF POWER ELECTRONIC CONVERTERS

Закон не должен быть пужало из тряпицы,
На кои, наконец, уже садятся птицы.

Там буйную вольность законы теснят.

А. С. Пушкин

This chapter is dedicated to the problem of electromagnetic compatibility concerning power electronic converters and this is quite new reference material on this topic. After the definition of the general problem content (part 6.1), the problems of quality of electrical energy in mains and autonomous power systems (parts 6.2 and 6.5), a problem of noise immunity (part 6.3) and noise emission (part 6.4) for power electronic devices are concerned. The material is based on analytical review of all-Union State Standard literature and our scientific proposals that should be added to the mentioned standards. In the sixth part there is review of power calculation theory for nonsinusoidal currents and voltages. This is really actual for power electronics systems, but hardly concerned in textbooks on the theory of electrotechnology.

6.1. THE CONTENT OF ELECTROMAGNETIC COMPATIBILITY PROBLEM

The problems of electromagnetic compatibility of power converters with technical surrounding and biosphere are the parts of the present-day ecological problems because they are concerned with electromagnetic “pollution” of environment.

Originally, the problem of electromagnetic compatibility appeared in radio engineering as a problem of broadcast surrounding pollution. For that area the electromagnetic compatibility is an ability of devices to work simultaneously in conditions of casual noises presence not generating unallowed noises for other devices.

Thus, the most important thing there is informational point of view of compatibility between useful signal and noise, i.e. their co-existence without useful signal distortion or information loss. So this aspect is mainly dealt with problem of inducted noises.

Two ways of electromagnetic noises transmission are evident: *inductive* (through electromagnetic field induction) and *conductive* (through the wires), although these two ways are interconnected and so we can speak only about domination of one of them concerning operation of power or informational units.

Later, this problem appeared to be actual for power engineering as a problem of pollution of the mains connected conductively (through the wires) with popular valve power converters and other nonlinear loads, which are the sources of higher harmonics and subharmonics (i.e. harmonics whose frequency is lower than frequency of the main's voltage).

Although, each word in word-combination "electromagnetic compatibility" is usually correctly understood, the whole sense of the phrase sometimes isn't clear enough despite the fact of its similarity to the phrase "human compatibility in the any group with limited resources". *Electromagnetic compatibility of electrical devices* means the compatibility of the hardware to operate properly despite the presence of unpremeditated conductive (from the main) and inductive (from the environment) electromagnetic noises without generation objectionable interferences to the main and\or environment.

Now we can **see three groups of electromagnetic compatibility (EMC) problems**:

- firstly, the quality of the electrical energy of the main and the influence of the power converter on the main [44, 65];
- secondly, resistance (noise-immunity), mainly of the control circuit, to the conductive and inductive noises, or general noise immunity;
- thirdly, the noise emission of the valve converters to the environment.

The content of electromagnetic compatibility problem is depicted in Fig 6.1.1. The first part of the problem of quality of the electrical energy is divided into three conditional subproblems:

– definition of the set of electrical energy quality factors and their normalization (numerical value setting). This set provides us an opportunity to calculate the losses caused by electrical energy of low quality;

– definition of ways to register the negative counter-effect of consumer's current (especially transitional and nonlinear) on the quality of electrical energy in the mains;

– definition of ways to detect the consumers that negative influence on the quality of electrical energy of the main.

The second part of EMC (electromagnetic compatibility) problem is the *noise-immunity aspect* of converters' control circuits. This aspect is divided into *conductive noise-immunity problem* and *inductive noise-immunity problem* like a part of the whole noise-immunity problem of electrical, electronic and radioelectronic systems, depending on the place of noise origin (i.e. main or the air). Conductive noises are also divided into special groups and each group has its own standards and noise-proof tests for hardware:

– nanosecond pulse noises [66];
– microsecond high-power pulse noises [67];
– dynamical voltage fluctuations (sags, interruptions, spikes) [68];
– decaying oscillations (single or recurring) [69];
– conductive noises induced by electromagnetic field of radio frequency [70];
– contact electrostatic discharge (conditionally) [71].

Inductive noises are also divided into special types relating to those the special tests on noise-immunity for hardware are provided:

– air electrostatic discharges (conditional) [71];
– electromagnetic field of radio frequency [72];
– magnetic field of industrial frequency [73];
– pulse magnetic field [74];

For each group of electromagnetic noises there is its own state standard (the references are indicated above). Besides that, there are the standards for special hardware that should be tested for noise-immunity accordingly to special sets of conductive and inductive noises. These sets may differ a bit from standards on certain types of electromagnetic noises. There are some special standards on noise-immunity for power electronics systems:

– for hardware used in industrial areas [75];

– for hardware used in residential and commercial zones, and also in industrial zones with low power consuming [76];

– for information technology hardware [77];

– for lightning hardware [78];

– for electrical hardware for measurements, control or laboratory using [79];

– for electromotor drive systems with variable speed [80].

Together with the problem of inductive noise immunity for hardware systems it is suitable to concern here the maximum allowable levels of electromagnetic exposures on people. These levels are regulated not only by State Standards for industrial enterprises [81], [82], but also by the sanitary codes and regulations [83] and they hadn't been provided in technical literature before.

The third part of the whole EMC problem is *noise emission* of technical units. The emitted noises could be again conductive and inductive. The *conductive noises* are characterized with levels of voltage and current induced by industrial radio noises [84, 85], and *inductive noises* – with electromagnetic field level [86, 87]. The terminology [88] and the procedure of certification tests on electromagnetic compatibility are approved by standards [89].

The basic principles of State Standards on the enumerated problems of EMC are considered below. Despite the fact that the first standard on electrical energy quality in Russia was approved in 1967, the most of the mentioned standards on noise-immunity and noise emission were introduced in 2000-2007 yrs. They are created on the base of the standards of International Electrotechnical Committee (IEC) and European committee (СИСПР) in the middle 90-ties.

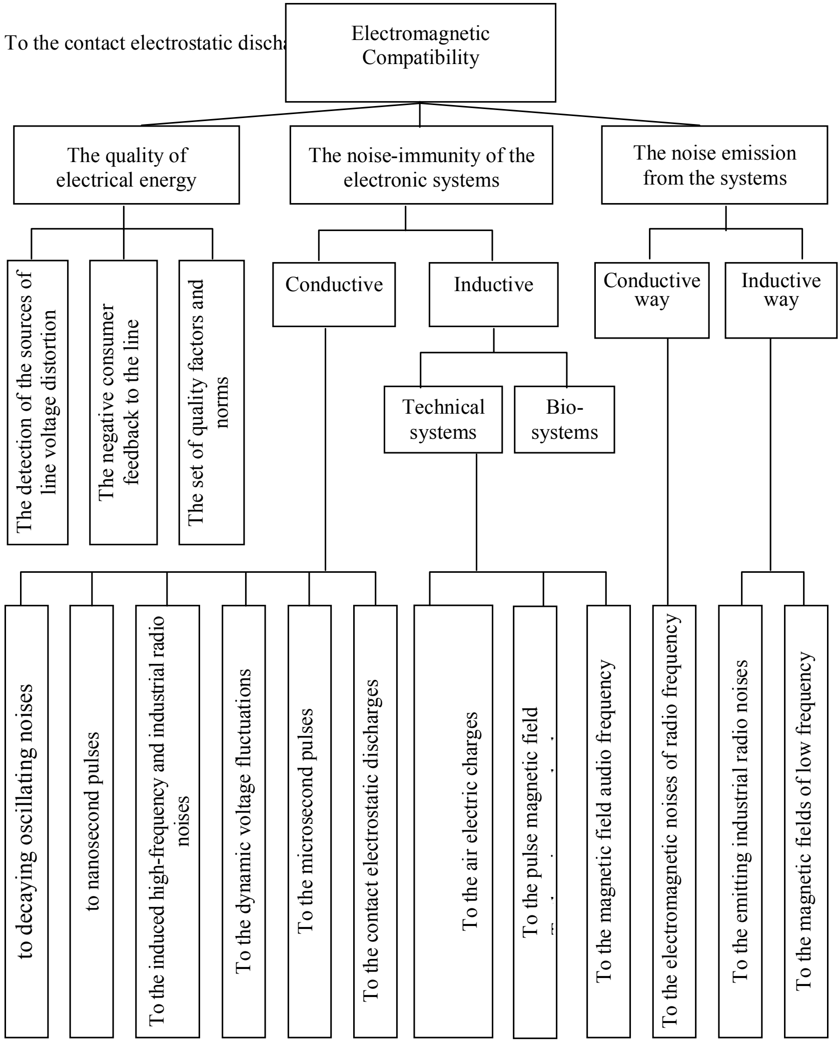

To the contact electrostatic disch
Electromagnetic Compatibility
The quality of electrical energy
The noise-immunity of the electronic systems
The noise emission from the systems
The detection of the sources of line voltage distortion
The negative consumer feedback to the line
The set of quality factors and norms
Conductive
Inductive
Technical systems
Bio-systems
Conductive way
Inductive way
to decaying oscillating noises
to nanosecond pulses
To the induced high-frequency and industrial radio noises
To the dynamic voltage fluctuations
To the microsecond pulses
To the contact electrostatic discharges
To the air electric charges
To the pulse magnetic field
To the magnetic field audio frequency
To the electromagnetic noises of radio frequency
To the emitting industrial radio noises
To the magnetic fields of low frequency

Fig. 6.1.1

6.2. THE QUALITY OF THE ELECTRICAL ENERGY IN THE COMMON MAINS

6.2.1. THE SYSTEM OF ELECTRICAL ENERGY QUALITY FACTORS AND THEIR NORMS

The electrical energy quality factors and their numerical norms for the common power lines of general power supply for 3-phase and 1-phase AC of 50 Hz frequency are defined by GOST 13109-97 for common connection points (CCP), i.e. for points to where electrical consumers or local power lines are connected.

The norms of electrical energy quality that are set for conductive electromagnetic noises are necessary for all operation modes of power systems except the modes caused by:

– extraordinary weather impacts and the acts of God (i.e. hurricanes, flood, earthquake);
– unforeseeable consequence caused by the acts of the party that are neither an power saving organization nor power consumer (i.e. fire, explosion, military operations and so on);
– conditions that are regulated by the state department and due to the elimination of the consequences caused by extraordinary whether impacts or unforeseeable consequence.

There are two kinds of electrical energy quality norms – normal and maximum allowable. The factors of electrical energy quality characterizing its disturbances are to be standardized. They are:

1. *Voltage deviation δU.* Voltage deviation is the difference between nominal rms phase voltage and steady (more than 1 min) rms phase voltage as shown in Fig. 6.2.1 [44]. Normal and maximum allowable values of voltage deviation δU_u are ±5 and ±10 % of nominal phase voltage:

$$\delta U_{\text{у}} = \frac{U_{(1)} - U_{\text{ном}}}{U_{\text{ном}}} \cdot 100 \approx \frac{U_y - U_{\text{ном}}}{U_{\text{ном}}} \cdot 100, \qquad (6.2.1)$$

where $U_{(1)}$ – rms phase voltage of 1st harmonic component.

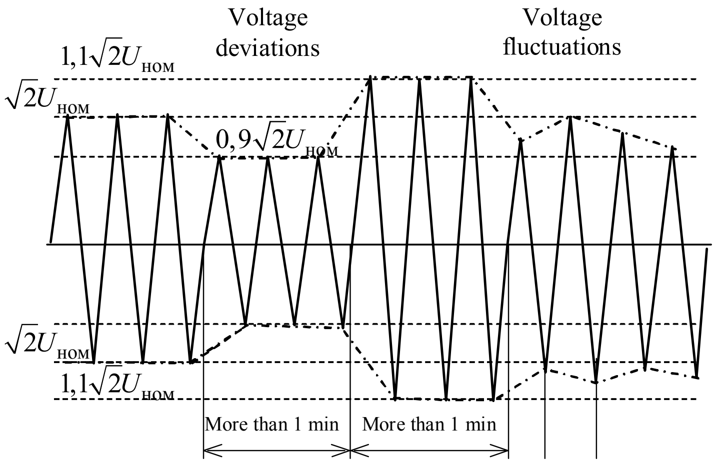


Fig. 6.2.1

2. *Voltage fluctuations* that are characterized by the *swing of voltage fluctuation* δU_t and *flicker dose* P_t. To determine these factors the idea of the envelope of root mean square meanings of the voltage is used. This envelope is multigraded function of time that is made of rms voltage meanings discretely determined at the each half period of the main frequency voltage. The voltage swing is determined as the difference between adjacent levels of the envelope and it is also characterized by the frequency of voltage changing $F_{\delta U_t}$ or by the interval between voltage changing $\Delta t_i, t_{i+1}$. Maximum allowable meanings of the voltage swing δU_t at the PCC (points of common connection) when the envelope is a meander depending on the frequency are shown in Fig. 6.2.2. The curve 1 is responsible for all the consumers except those who use incandescent lamps in the apartments where the significant eyestrain is necessary (for them more strict limitations are determined by the curve 2).

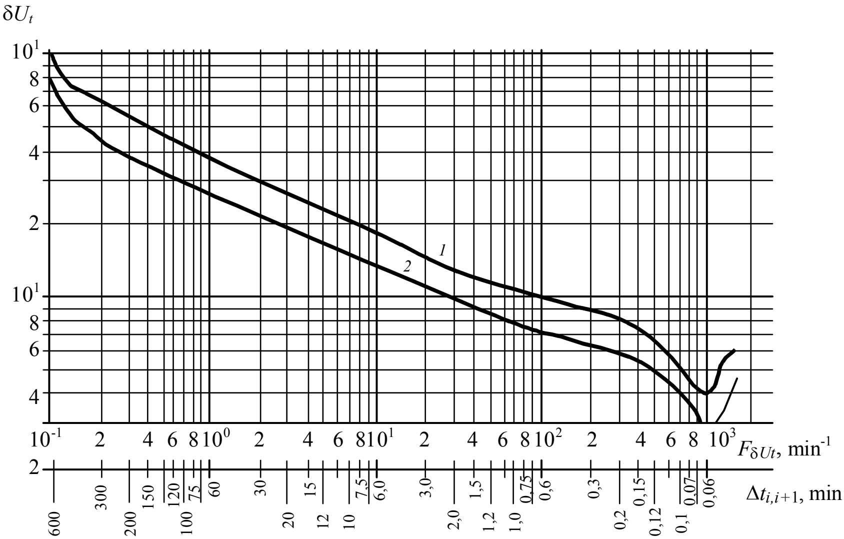


Fig. 6.2.3

The second characteristic is a flicker dose. The Flicker is determined as a people's subjective perception of luminous flux fluctuation of the candlelight caused by voltage fluctuations in the main to which the lightings are connected.

The Flicker dose is a degree of people's receptivity to the flicker influence during the fixed interval of time. To determine flicker dose we ought to calculate the time of flicker perception t_f (in seconds) for each curve of relative voltage fluctuations:

$$t_f = 2{,}3\left(Fd_{\max}\right)^{3/2}, \tag{6.2.2}$$

where $d_{\max}$ – maximum voltage deviation from nominal voltage in percents; F – the reduction coefficient that depends on the shape of voltage waveforms, that are shown in [44]. So, we can calculate the flicker dose using (6.2.3) for common time interval T_p of 10 min.

$$P_{st} = \left(\sum_f \frac{t_f}{T_p}\right)^{1/3{,}2} \tag{6.2.3}$$

Prolonged flicker dose is calculated for the time interval of 2 hours.

Maximum allowable meaning of short-term flicker dose at PCC for 0.38 KW mains is 1.38 and of long-term flicker dose is 1.0. For the places with incandescent lamps where significant eyestrain takes place these doses consequently are 1.0 and 0.74.

3. *Voltage non-harmonicity.* The voltage *non-harmonicity* is determined by:

– *coefficient of distortion of voltage harmonicity K_U;*

– *coefficient of the n^{th} harmonic component of the voltage $K_{U(n)}$.*

First coefficient is determined as a ratio of rms voltage of high harmonics to the rms voltage of the 1^{st} harmonic. The second one is determined as the ratio of the rms meaning of the n^{th} harmonic component to the rms meaning of the 1^{st} harmonic component.

The normal and maximum allowable meanings (in percents) of the voltage harmonicity distortion coefficient K_{Un} at the PCC for the mains with different nominal voltages are shown in Table 6.2.1.

Table 6.2.1

The meanings of the harmonicity distortion coefficient

Normal meanings of K_U at U_{nom}, кV				Maximum allowable meanings K_U at U_{nom}, кV			
0,38	6...20	35	110...330	0,38	6...20	35	110...330
8,0	5,0	4,0	2,0	12,0	8,0	6,0	3,0

Normal meanings of the n^{th} harmonic component $K_{U(n)}$ at the PCC for the mains of different nominal voltages are shown in Table 6.2.2. The maximum allowable meanings of $K_{U(n)}$ are 1.5 times greater than those shown in the table.

T a b l e 6.2.2

Normal meanings of K_U

Odd harmonics not divisible by 3 at $U_{nom,\,k}$V					Odd harmonics divisible by 3 * at U_{nom}, kV					Even harmonics at U_{nom}, kV				
n	0,38	6–20	35	110–330	*n*	0,38	6–20	35	110–330	*n*	0,38	6–20	35	110–330
5	6,0	4,0	3,0	1,5	3	5,0	3,0	3,0	1,5	2	2,0	1,5	1,0	0,5
7	5,0	3,0	2,5	1,0	9	1,5	1,0	1,0	0,4	4	1,0	0,7	0,5	0,3
11	3,5	2,0	2,0	1,0	15	0,3	0,3	0,3	0,2	6	0,5	0,3	0,3	0,2
13	3,0	2,0	1,5	0,7	21	0,2	0,2	0,2	0,2	8	0,5	0,3	0,3	0,2
17	2,0	1,5	1,0	0,5	>21	0,2	0,2	0,2	0,2	10	0,5	0,3	0,3	0,2
19	1,5	1,0	1,0	0,4	–	–	–	–	–	12	0,2	0,2	0,2	0,2
23	1,5	1,0	1,0	0,4	–	–	–	–	–	>12	0,2	0,2	0,2	0,2
25	1,5	1,0	1,0	0,4	–	–	–	–	–	–	–	–	–	–
>25	0,2+1,3x 25/*n*	0,2+1,3x 25/*n*	0,2+1,3x 25/*n*	0,2+1,3x 25/*n*	–	–	–	–	–	–	–	–	–	–
					–	–	–	–	–	–	–	–	–	–
					–	–	–	–	–	–	–	–	–	–

Comments:

1) *n* – number of harmonic component of voltage;

2) * –normal meanings for *n* equal to 3 and 9 for single phase mains, for the 3 – phase 3 - wire mains these meanings should be reduced to a half .

4. *The non-symmetry of the Voltage.* This kind of the three-phase voltage distortion is characterized by *coefficients of non-symmetry*:

– *for inverted sequence* K_{2U};
– *for zero sequence (string)* K_{0U}.

These coefficients are the ratios of voltage of inverted sequence or zero sequence to the voltage of the direct sequence. The components of the voltages mentioned above are to be determined either via the *method of symmetrical components* or through the standard formulas that link measured phase-to-phase voltages.

Normal and maximum allowable meanings of the coefficients of voltage non-symmetry are normalized for the equal levels of 2 and 4 % at the nominal voltage of 0.38 kV. At the same time, the norms for zero sequence are relevant to the PCC of 4 – wire mains.

5. *Frequency deviation.* This factor is determined as the difference between current value of the ac voltage frequency and its nominal value. Normal and maximum allowable meanings of frequency deviation factor are correspondingly ± 0,2 and ± 0,4 Hz.

6. *Voltage sag*. The voltage sag is a sudden decrease of the voltage at the power point lower than $0{,}9U_{\text{nom}}$, and its further recovering during the time interval of 10ms to dozens of seconds (Fig. 6.2.3) [44].

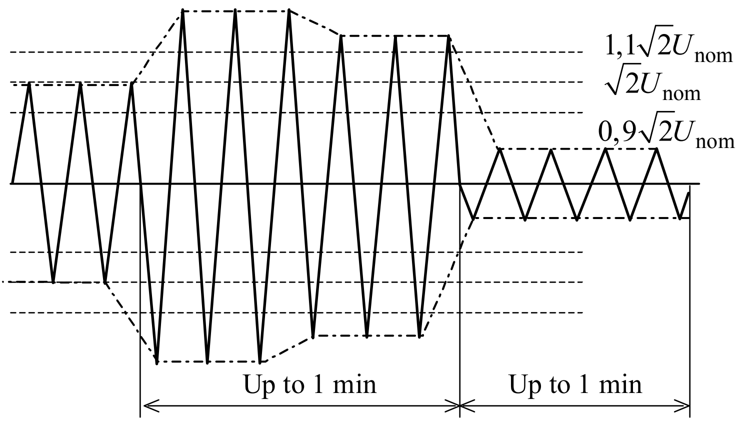


Fig. 6.2.3

The difference between voltage sag and voltage deviation is their duration and changing in amplitude. Duration of the voltage sag is much shorter, but amplitude of the voltage decreases many times more than during the deviation process. The parameter to normalize here is the duration of the sag, the maximum allowable value of this parameter for the 20 kV main is 30 seconds. The duration of the sag that would be corrected automatically at any main's connection point is defined by the tolerance parameter of the relay protection system.

There are some data in the appendix of the GOST 13109-97 that characterize the values of the sags, their duration and frequency of their appearance in Russian 6…10kV mains and in EC mains. The correspondence of the sags characteristics to the characteristics of ARS (Automatic Reserve Systems) are given in Table 6.2.3.

Table 6.2.3

The characteristics of the voltage sags

Depth of the sag, %	Interval portion, %, at the sag duration, sec				Total, %
	0,2	0,5...0,7	1,5...3,0	3,0...30	
10...35	–	–	18	–	18
35...99	38	3	8	–	49
100	26	–	–	7	33
Sum total	64	3	26	7	100

On average, every consumer of described mains suffers sags 12 times a year (energy supply side data).

7. *Voltage pulse.* The voltage pulse is abrupt voltage change at the power point for several milliseconds that is followed by fast reestablishment. There are no norms for voltage pulses, but there are some statistic data for lightning and commutation impulses that could appear in the main. This data are derived from GOST 13109-97. The lightning impulses have half-sine waveform with amplitude of 100kV (2000kV max for some mains) and 5…10 microseconds duration for 6….10 kV mains. Commutation pulses are characterized with 1000…5000 microseconds duration (at 0.5 of amplitude level) and amplitude of 25..40kV in the 6…10kV mains.

8. *Voltage spikes.* The voltage spike is the situation when voltage at the power point increases more than $1{,}1U_{\mathrm{nom}}$ for more than 10 ms. This happens at the commutation process or when the short circuit occurs (see Fig. 6.2.3). The parameter is s*pike time factor* $K_{\mathrm{sp}U}$ that is ratio of the spike envelope to the amplitude of nominal voltage. Voltage spike is also characterized by its duration $\Delta t_{\mathrm{sp}U}$. The values of the voltage spike factors for power utility are shown in Table6.2.4.

Table 6.2.4
Voltage spike factors

Duration of the spike Δt_{spU}, s	up to 1	up to 20	up to 60
Spike time factor K_{spU}	1,47	1,31	1,15

On average there are approximately 30 voltage spikes at the PCC during the year. But if the neutral wire of the 3-phase system with dead grounded neutral (up to 1 kV) is damaged, temporary phase-to-ground overvoltages occur. At the same time if the load phase non-symmetry is significant, the K_{spU} could reach $\sqrt{3}$ meaning and Δt_{spU} could be up to several hours. Such event could make severe damages to the consumer electronics connected to the main. The protection system could use automatic load disconnector (but it could respond to the voltage pulses or short spikes) or power line conditioner (see Chapter 11).

To determine whether the values of the energy quality factors correspond to their norms, excluding voltage sag, voltage pulses, voltage spikes, the time interval of minimum 24 hours is set. Values of the voltage non-harmonicity factor, the n-th harmonic factor, the factors of non-symmetry for zero and inverted sequences mustn't overcome their maximum allowable meanings. The same about their meanings measured with 95% probability for the same observation period, they mustn't overcome their normal values. The same requirements for steady state voltage and frequency deviation are set considering their sign.

Total duration of power quality factors measurement (i.e. voltage pulses, short overvoltages) excluding voltage sag duration is recommended to be 7 days period. The power quality control at PCC is provided by power supplying organization. Power qualitu measurements are to have place:

– for steady state voltage deviation – two times a year or more;
– for other factors – one time during two years or more on conditions that power circuit and its elements are invariable, load changing is insignificant.

Consumers that aggravating electrical energy quality ought to provide control measures at their own power points near the PCC, and also at the front ends of the loads that generate conductive electromagnetic noises. This should take place periodically and the period is to be coordinated with power supplying organization.

The inaccuracy of power quality factors measurements is defined in absolute and relative forms and shown in Table6.2.5.

Table 6.2.5

The inaccuracy of power quality factors measurements

Innacuracy of measurements, %	δU_y	δU_t	P_{st}	P_{Lt}	K_U	$K_{U(n)}$	K_{2U}	K_{0U}	Δf
Absolute	±0,5	–	–	–	–	±0,05 at the $K_{U(n)} < 1,0$	±0,3	±0,5	±0,03
Relative	–	±8	±5	±5	±10	±5 at the $K_{U(n)} \geq 1,0$	–	–	–

Power quality measurements are to be provided only by accredited laboratories that have special license and complex measurement equipment such as «Omsk», Eris-KE, PPKE, Resurs VF-2.

However, the lack of power quality factors set is evident, first of all concerning the voltage non-harmonicity. From the direct calculation methods (section 1.6, chapter 5) we have that load current quality at the non-sinusoidal voltage is defined not by standardized *voltage non-harmonicity factor* K_U, but with set of integral harmonic factors. It is easy to make certain that current non-harmonicity factor (harmonic coefficient) of induction motor stator, if machine model for higher harmonics is sum of the windings leakage inductances, is:

$$K_I = \overline{K}_U K_\mathrm{r}$$

(K_r – the factor of induction motor startup current).

The non-harmonicity factor (harmonic coefficient) of capacitor is

$$K_I = \breve{K}_U \, .$$

The non-harmonicity factor (harmonic coefficient) of inductance current is

$$K_I = \overline{K}_U \, .$$

So, knowing our new voltage quality factors $\overline{K}_U$, $\breve{K}_U$, it is possible to predict current quality of typical customer's loads. Because almost all the loads could be represented as a model of active resistance and inductance, the knowledge of only two additional power quality factors appears to be enough.

In that way, addition of the integral and differential harmonic factors $\bar{K}_{г}, \tilde{K}_{г}$, defining the consumer's prejudice due to the non-harmonicity, to the power quality standard [44] seems to be rational.

6.2.2. GENERAL EVALUATION OF POWER CONVERTERS' CONDUCTIVE INFLUENCE TO THE MAIN

Here we consider only one type of negative backward influence of the power converter to the main known as main's voltage distortion due to the converter's non-sinusoidal input current.

The degree of voltage distortion due to the non-linear consumer can be calculated if the main's equivalent circuit and its frequency characteristic as well as non-linear consumer's input current spectrum are known. That is why international power quality standards (MEC, IEEE) contain norms for consumption current up to 40th harmonic component.

Accordingly to the Russian classification standard on harmonic emission for units with phase current up to 16 amps:

– class A – symmetrical 3 – phase units that are not corresponded to the B C , or D classes;

– class B – portable electric tools;

– class C – lighting units including adjustment devices;

– class D – technical units with rectangular multigraded input current and active power not less than 600W.

Maximum allowable values of class A current harmonics are shown in Table 6.2.6

Table 6.2.6

n	3	5	7	9	11	13	$15<n<39$ odd	2	4	6	$8<n<40$
I_n, A	2,3	1,4	0,77	0,4	0,33	0,21	$0{,}15\cdot 15/n$	1,08	0,43	0,3	$0{,}23\cdot 8/n$

For class B consumers these values should be increased 1.5 times. For class C consumers with more than 25W power maximum allowable harmonics values are shown in Table 6.2.7.

Table 6.2.7

n	2	3	5	7	9	11 < n < 39 (odd)
% from $I_{(1)}$	2	30χ	10	7	5	3

Here χ is the consumer's power factor. For current consumer of this class with power less than 25W maximum allowable current harmonics values are shown in Table 6.2.8.

Table 6.2.8

n	2	3	5	7	9	11 < n < 39 (odd)
$I_{(n)}/W$, mA/W	3,4	1,9	1,0	0,5	0,35	3,85/n
$I_{(n)}$ max, A	2,3	1,14	0,77	0,4	0,33	As the A class

For class D consumer's harmonic components at the load power more than 75W mustn't overcome current values of previous table, and if the load power is less than 75W norms aren't set. After 4 years the 75W barrier would be reduced to 50W.

Thus, knowledge (control) of consumer's current harmonic composition allows us to calculate easily harmonic composition of main's voltage that occurs due to the backward influence of nonlinear consumer. At the same time, low-voltage mains network (up to 1000V) is usually simulated for higher harmonics (up to the 40^{th} i.e. 2kHz) using inductive reactance. In this case we could possibly use direct voltage distortion degree calculation method with is alternative to the spectral method.

6.2.3. DETERMINATION OF CONTRIBUTION OF NONLINEAR CONSUMERS TO THE VOLTAGE DISTORTION AT THE POINTS OF COMMON CONNECTION

In the common mains there could be several consumers that have nonsinusoidal input currents connected to the PCC. In this case there is necessity to determine their partial contribution into the overall voltage distortion at the PCC to made proper decisions (organizational or penal). There are no direct indications about that in the electrical energy standards.

Main difficulty in the voltage quality control at the node of energy system is the detection of the parts of higher harmonics in overall voltage distortion that are introduced by individual nonlinear load connected to the node.

In the power industry the is a method to define the part of distortion introduced by individual load by consequent power quality factor measurement when the load is connected L_{con}, and disconnected L_{discon}, and further calculation of partial contribution of the distortion at the PCC by formulae

$$b = 1 - \frac{L_{con}}{L_{discon}}. \qquad (6.2.10)$$

This method has several disadvantages:

- result is not always correct (as it would be shown below);
- it is necessary to disconnect load during the measurements, but this is not always possible.

There is also a method of definition of partial load contribution into the voltage quality distortion using active and reactive secondary distorting load powers. In this method partial contribution is defined by picking out of anomalistic component of voltage, defining of active and reactive powers of all the loads connected to the system none, and by defining partial contribution factor (PCF) accordingly to the proposed formulae by the sum portion of active and reactive secondary powers of given load relative to one generated by all distorting loads [90].

The limitation of this method is error of load partial contribution definition into the voltage quality distortion connected with two factors:

- firstly, not all the factors that influence on the voltage quality are taken into consideration, because secondary distortion powers caused by interaction of different current and voltage harmonics aren't considered in the secondary power balance.

- secondly, voltage quality changes at the energy system node is evaluated in GOST13109-97 by voltage quality factors, not power ones that depend not only on voltage quality but also on current quality. This fact causes the truncation error. Besides, partial contribution of EMF into the voltage quality isn't considered in this method.

Here follows a description of the method that doesn't have above mentioned limitations.

Fig. 6.2.4 shows the circuit of energy system node with single power source, and Fig. 6.2.5 shows energy system node with several electrical energy sources. There are EMF sources (e_1, e_2, ... e_n, in general case the are n sources), equivalent inductances with correspondent sources (L_1, L_2, ... L_n), current sources (i_1, i_2, ... i_k, in general case the number of partial current sources is k) that equivalent k nonlinear loads.

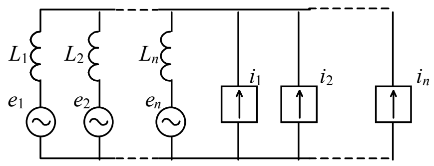


Fig. 6.2.5

Voltage distortions at the node are connected with its anomalistic components, that are caused by whether voltage higher harmonics presence or by reverse or zero voltage sequences in multiphase voltage systems. Any anomalistic components will be further denoted with «a» index.

Let's begin the analysis with the single EMF source case and further generalization to the n sources case. Accordingly to the figure 6.2.4 for anomalistic component in the typical energy system node picked out with known methods it is possible to derive following differential equation:

$$u_{\mathrm{a}} = e_{\mathrm{a}} - L\frac{di_{1\mathrm{a}}}{dt} - L\frac{di_{2\mathrm{a}}}{dt}\,. \tag{6.2.11}$$

After the algebraization of differential equation for anomalistic components by raising differential equation of anomalistic components of type (6.2.11) to the second power and its further integration on the period accordingly to the ADE2 method, we will have an algebraic equation for rms values of *anomalistic (distorting) components of voltages and currents:*

$$U_{\mathrm{a}}^2 = \frac{1}{T}\int_0^T e_{\mathrm{a}}^2 dt + \frac{L^2}{T}\int_0^T \left(\frac{di_{1\mathrm{a}}}{dt}\right)^2 dt + \frac{L^2}{T}\int_0^T \left(\frac{di_{2\mathrm{a}}}{dt}\right)^2 dt +$$

$$+2L^2\frac{1}{T}\int_0^T \frac{di_{1\mathrm{a}}}{dt}\frac{di_{2\mathrm{a}}}{dt}dt - 2L\frac{1}{T}\int_0^T e_{\mathrm{a}}\frac{d(i_{1\mathrm{a}}+i_{2\mathrm{a}})}{dt}dt = \mathrm{OPC}(e_{\mathrm{a}}) + \mathrm{OPC}(i_{1\mathrm{a}}) +$$

$$+\mathrm{OPC}(i_{2\mathrm{a}}) + \mathrm{OPC}(i_{1\mathrm{a}}, i_{2\mathrm{a}}) + \mathrm{MPC}(e_{\mathrm{a}}, i_{1\mathrm{a}} + i_{2\mathrm{a}}). \tag{6.2.12}$$

Here OPC – own partial contribution, of the correspondent branch connected to the node, into the overall voltage quality change, MPC – mutual partial contribution of two branches connected to the node, into the overall voltage change at the node.

Physical sense of OPC factor is the determination of node voltage degradation that would be caused on the condition of absence of other voltage distortion sources (i.e. nonlinear loads, EMF sources with distorted waveform and so on).

Physical sense of MPC of two branches connected to the node is the determination of node voltage quality change (improvement or degradation) that is caused by summarizing of anomalistic voltage components of the branches.

An equation for relative values of own (OPC*) and mutual (MPC*) partial contributions of branches into resulting node voltage quality change could be obtained by division of both parts of the equation (6.2.12) by U_a^2:

$$1=\frac{1}{U_{\mathrm{a}}^2}\frac{1}{T}\int_0^T e_{\mathrm{a}}^2 dt+\frac{1}{U_{\mathrm{a}}^2}\frac{L^2}{T}\int_0^T\left(\frac{di_{1\mathrm{a}}}{dt}\right)^2 dt+\frac{1}{U_{\mathrm{a}}^2}\frac{L^2}{T}\int_0^T\left(\frac{di_{2\mathrm{a}}}{dt}\right)^2 dt+$$

$$+\frac{2L^2}{U_{\mathrm{a}}^2}\frac{1}{T}\int_0^T\frac{di_{1\mathrm{a}}}{dt}\frac{di_{2\mathrm{a}}}{dt}dt-\frac{2L}{U_{\mathrm{a}}^2}\frac{1}{T}\int_0^T e_{\mathrm{a}}\frac{d(i_{1\mathrm{a}}+i_{2\mathrm{a}})}{dt}dt=\mathrm{OPC}^*(e_{\mathrm{a}})+\mathrm{OPC}^*(i_{1\mathrm{a}})+$$

$$+\mathrm{OPC}^*(i_{2\mathrm{a}})+\mathrm{MPC}^*(i_{1\mathrm{a}},i_{2\mathrm{a}})+\mathrm{MPC}^*(e_{\mathrm{a}},(i_{1\mathrm{a}}+i_{2\mathrm{a}})). \quad (6.2.13)$$

An equation for correspondent coefficient that characterizes individual types of anomalistic components could be obtained by division of equation (6.2.12) by rms value of first harmonic component of direct voltage sequence. At that, if the anomalistic voltage is the voltage of higher harmonics, the voltage nonsinusoidality coefficients are obtained (general, own and mutual ones); if the anomalistic voltage is the reverse sequence voltage then the coefficient of reverse sequence are obtained; and finally, if the anomalistic voltage is the voltage of zero sequence of nonsymmetrical 3-phase system then the coefficients of zero sequence are obtained accordingly to that as they defined by GOST13109-97.

For the common case shown in Fig. 6.2.5 for *n* branches containing EMF, the equation to determine own and mutual partial contributions of branches into the overall voltage quality change could be obtained by generalization of the equation (6.2.13):

$$U_{\mathrm{a}}^2=\sum_{n=1}^{n}\frac{1}{T}\int_0^T e_{\mathrm{a}(n)}^2dt+\sum_{k=1}^{k}\frac{L^2}{T}\int_0^T\left(\frac{di_{\mathrm{a}(k)}}{dt}\right)^2dt+2\sum_{\substack{l,m\\ l\neq m}}^{n}\frac{1}{T}\int_0^T e_{\mathrm{a}(l)}e_{\mathrm{a}(m)}dt-$$

$$-2L\sum_{n,k}\frac{1}{T}\int_0^T e_{\mathrm{a}(n)}\frac{di_{\mathrm{a}(k)}}{dt}dt=\sum_n \mathrm{OPC}(e_{\mathrm{a}(n)})+\sum_k \mathrm{OPC}(i_{\mathrm{a}(k)})+$$

$$+\sum_{\substack{l,m\\ l\neq m}}^{n}\mathrm{MPC}(e_{\mathrm{a}(l)},e_{\mathrm{a}(m)})+\sum_{n,k}\mathrm{MPC}(e_{\mathrm{a}(n)},i_{\mathrm{a}(k)}). \qquad (6.2.14)$$

From the equations (6.2.11) – (6.2.14) it is clear that resulting node voltage distortions are defined not only by own characteristics (spectrums) of all energy objects connected to the node, but also by mutual characteristics. Mutual characteristics of the energy objects define whether there would be improvement or degradation of power quality at the node because these coefficients could have positive or negative sign in contrast to OPC that always have positive value. That the reason why the method of definition of partial contribution by disconnection of some consumers that doesn't consider this fact is not correct.

Coefficients of partial contribution of each nonlinear consumer into the overall voltage distortion at the PCC could be defined on the basis of equation (6.2.13).

The coefficient of partial contribution of electrical system of the overall voltage distortion at the PCC is defined by the equation where own partial distribution of the system as well as its mutual contribution in the interaction with consumers is considered:

$$K(e_{\mathrm{a}})=\mathrm{OPC}^*(e_{\mathrm{a}})+\mathrm{MPC}^*(e_{\mathrm{a}},(i_{1\mathrm{a}}+i_{2\mathrm{a}}))\frac{\mathrm{OPC}^*(e_{\mathrm{a}})}{\mathrm{OPC}^*(e_{\mathrm{a}})+\mathrm{OPC}^*(i_{1\mathrm{a}})+\mathrm{OPC}^*(i_{2\mathrm{a}})}$$

(6.2.15)

The coefficient of partial contribution of the first nonlinear consumer is defined in the same manner by considering own contribution and its mutual contribution in the interaction process with other electrical energy consumers:

$$K(i_{1a}) = \mathrm{OPC}^*(i_{1a}) + \mathrm{MPC}^*(i_{1a}, i_{2a}) \frac{\mathrm{OPC}^*(i_{1a})}{\mathrm{OPC}^*(i_{1a}) + OPC^*(i_{2a})} +$$
$$+\mathrm{MPC}^*(e_a, (i_{1a} + i_{2a})) - \frac{\mathrm{OPC}^*(i_{1a})}{\mathrm{OPC}^*(e_a) + \mathrm{OPC}^*(i_{1a}) + \mathrm{OPC}^*(i_{2a})}. \quad (6.2.16)$$

The coefficient of partial distribution of the second nonlinear consumer is defined in the same manner

$$K(i_{2a}) = \mathrm{OPC}^*(i_{2a}) + \mathrm{MPC}^*(i_{1a}, i_{2a}) \frac{\mathrm{OPC}^*(i_{2a})}{\mathrm{OPC}^*(i_{1a}) + \mathrm{OPC}^*(i_{2a})} +$$
$$+\mathrm{MPC}^*(e_a, (i_{1a} + i_{2a})) - \frac{\mathrm{OPC}^*(i_{2a})}{\mathrm{OPC}^*(e_a) + \mathrm{OPC}^*(i_{1a}) + \mathrm{OPC}^*(i_{2a})} \quad (6.2.17)$$

It is evident that these coefficients are connected by the common equation

$$K(e_a) + K(i_{1a}) + K(i_{2a}) = 1, \quad (6.2.18)$$

This fact allows calculating (measure) only two of them; the third one could be found using this equation.

Thus, the method of load partial contribution definition of load and energy system into the voltage quality change of the system PCC, that generalizes and evolves the method that us proposed together with M.V. Gnatenko and V.I. Popov, also allows to consider not only contribution of the loads but also contribution of energy system into the voltage quality at the PCC. This fact allows to detect voltage quality distortion sources more adequately in comparison with today's methods [44].

6.3. NOISE-IMMUNITY OF ELECTROTECHNICAL AND ELECTRONICAL SYSTEMS WITH POWER ELECTRONICS CONVERTERS

Electromagnetic noise-immunity is the capability of technical device to continue the operation with given quality under the action of external noises with regulated parameters without special noise protection devices, except ones that appurtenant to the device operation principle or building principle of the device [88]. At that, the *electromagnetic noise* is the electromagnetic phenomenon or process that degrade or could possibly degrade the quality of operation of the device.

Due to the huge variety of technical systems and multiplicity of possible kinds of electromagnetic noises in, different requirements to the predictable reaction of technical system on the noises, all the typical situations are classified into the finite number of typical situations.

First of all, the “painful points” of technical system are unified. These “painful points” are also called ports; they are the border or outpost between technical system and external electromagnetic environment (clamp, cutoff point, terminal and so on). Fig. 6.3.1 shows unified technical system with five types of ports.

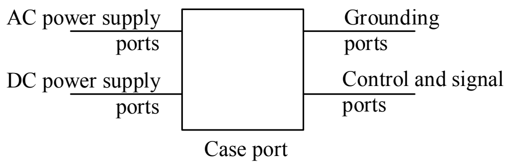


Fig. 6.3.1

Secondly, different types of c o n d u c t i v e electromagnetic noises are defined, they are:

- nanosecond pulse noises;
- microsecond high-power pulse noises;
- dynamical voltage changes (voltage hollows, interruptions, spikes);
- noises induced by electromagnetic field of radio frequency (150 kHz...80 MHz);
- contact electrostatic discharges;
- oscillating decay noises.

Moreover, i n d u c t i v e electromagnetic noises are also defined, they are:

- air electrostatic discharges;
- audio frequency magnetic fields;
- pulse magnetic fields;
- electromagnetic fields of radio frequency (80...1000 MHz).

Thirdly, there are five degrees of test levels of system noise-immunity depending on the value (power) of electromagnetic noises, created at the test. This is to classify operation conditions of technical systems.

Fourthly, test results are classified on the basis of the electrical system operation criteria:

– A – normal operation;

– B – temporary operation quality worsening with further recovery without help of operator when noise disappears;

– C – temporary operation quality worsening that requires help of operator when noise disappears;

– Д – constant operation quality worsening or termination of operation without recovery due to the damage caused by the noise.

Owing to the evident complexity of electromagnetic compatibility problem analysis the norms for all types of noises are defined on the statistic basis. Accordingly to this basis not less than 80% of production-run equipment should correspond to these norms with 80% reliability.

To obtain the knowledge about these norms let's briefly consider their application to the all listed types of conductive and inductive electromagnetic noises that are used in noise-immunity tests.

Nanosecond pulse noises with 5ns edge with 150ns pulse duration of 0,25...4 kV are fed into the power supply ports and input-output signal ports depending on the level of test harshness ant the type of electrical system. Microsecond pulse noises of the same voltage band should have steep leading edge of about 1ms with pulse duration of 50ms. Immunity of the system to the dynamical supply voltage changes is tested using: a) voltage hollows (sags) that are 30% of nominal voltage with duration from 10 to 100 periods depending on the harshness of the test; b) voltage spikes, i.e. when voltage is 20% higher than nominal value during the same time interval; c) total voltage interruption for time interval from 1 to 25 periods. The norm of immunity to the conductive noises, inducted magnetic field of radio frequency of 150kHz to 80MHz is 120…140dB (relative to 1μV/m). The norm of immunity to the individual (2…3 periods) or periodic (duration from 2s) oscillating decay noises of 100 kHz and 1 MHz frequency, that are shown in Fig. 6.3.1, is 0.5…4kV when noise is fed "from port wire to ground", and 0.25…2kV when noise is fed "from ground to ground".

The norm of immunity to the magnetic field of industrial frequency is changed within band limit of 1100A/m for continuous magnetic field and within 300…1000 A/m for short term magnetic field of 1..3sec duration depending on the harshness of the test.

The norm of immunity to the pulse magnetic field is varied within 100…1000 A/m band when duration of the pulse is 16 μs and front edge is 6.4μs.

The norm of immunity to the electrostatic discharge that is contact to the case port is up to 4kV and for air discharge up to 8kV.

The norm of immunity to the test electromagnetic field strength is 1…10V/m within frequency band from 80…1000MHz, when field is 80% modulated with 1kHz frequency.

All possible technical systems are divided into aggregative groups [89]. For each group there is own standard that defines the set of typical electromagnetic noises for noise-immunity tests. For power electronics such technical systems are:

- measurement [79], control and laboratory equipment;
- electromotor drive systems with adjustable rotation speed [80];
- professional audio, video, audiovisual devices and control equipment for lightning devices for sight events [91];
- measuring relays and protection devices [92];
- uninterruptible power supplies [93].

Bioelectromagnetic compatibility. Inductive electromagnetic noises influence not only technical systems (that was the reason to develop state standards on electromagnetic compatibility of technical systems) but also biological ones, especially on humans.

Maximum allowable level of influence of electromagnetic field of 30 kHz…300GHz frequency on the peoples is set by sanitary code and regulations [83]. There also were two GOSTs before that norms [81,82]. Influence evaluation is performed by two parameters. They are: *radiant exposure* that is defined by the radiation intensity and by the time of exposure to the human being whose work or study is connected with necessity to stay in the areas near the radiation source; *electromagnetic exposure intensity* (for other people). At that, the intensity of electromagnetic radiation within 30 kHz …300 MHz frequency band is evaluated by electrostatic intensity (E, V/m) and electromagnetic intensity (H, A/m). In the frequency band 300 MHz...300 GHz the intensity is evaluated by energy-flux density values (W/m^2, μW/sm^2).

Radiant exposure created by the electromagnetic field is equal to the product of second degree of electromagnetic field intensity and time of the exposure to the human, and appears as $(W/m)^2 \cdot h$ and $(A/m)^2 \cdot h$. In case of pulse excitation the average value is used for evaluation purposes. Stated radiant exposure during the working day (shift) shouldn't overcome the limits shown in table 6.3.1.

Maximum allowable levels of electrical and magnetic components within frequency band 30 kHz...300 MHz depending on the duration of exposure are defined in table 6.3.2.

If the exposure lasts more than 0.08 h further increase if its intensity is prohibited.

Table 6.3.1

Radiant exposures

Frequency band	Maximum allowable radiant exposures		
	for electrical component, $(W/m)^2 \cdot h$	for magnetic component, $(A/m)^2 \cdot h$	for energy-flux density, $(\mu W/sm^2) \cdot h$
30 KHz...3 MHz	20 000,0	200,0	–
3...30 MHz	7000,0	undeveloped	–
30...50 MHz	800,0	0,72	–
50...300 MHz	800,0	undeveloped	–
300 МГц...300 GHz	–	–	200,0

Table 6.3.2

Maximum allowable field intensity levels

Duration of exposure, h	*E*, V/m			*H*, A/m	
	0,03...3 MHz	3...30 MHz	30...300 MHz	0,03...3 MHz	30...50 MHz
8,0 and more	50	30	10	5,0	0,3
1,0	141	84	28	14,2	0,85
0,08 and less	500	296	80	50,0	3,0

To measure the intensity of electromagnetic radiation the next field-intensity meters are recommended: PZ-22 (for frequency band 0,01...1000 MHz) and flux-intensity meters PZ-18, PZ-19, PZ-20 (for frequency band 0,3...39,65 GHz).

6.4. NOISE EMISSION OF POWER ELECTRONICS CONVERTERS

The sense of by-word "live and let the others to live" concluded in the conception of electromagnetic compatibility suggests the absence of generation (emission) of inadmissible electromagnetic noises from power electronic devices for other technical units. Here, as in the noise immunity issues, two ways of noise propagation are possible: conductive (to the power supply) and inductive (to the environment), but the direction is reverse.

Conductive noise emission. Noise emission standard up to 40th harmonic component [65] covers frequency band from 50 (60) Hz to 2(2,4) kHz. Frequency band from 3 to 148.5kHz is dedicated to the service signals transmission for energy supply organizations, power inspecting organizations and for power consumers. Correspondent standard [94] defines maximum allowed value of noise voltage for stated subbands within 134...116 dB limits (relative to 1μV), i.e. on the 0,5...3 V level that is calculated as logarithm of relative noise U_2 value with constant base- denominator U_1, here U_1 = 1 μV:

$$u = 20\lg\frac{U_2}{U_1} \text{ [dB]}. \tag{6.4.1}$$

Intensity levels of electrical and magnetic fields are defined in the same manner. Power level of the noise is calculated via the formulae:

$$P = 10\lg\frac{P_2}{P_1} \text{ [dB]} \tag{6.4.2}$$

for fixed base value of P_1.

Because of difference of power supplies parameters, for unification of consumer noise emission level measurements, there is special device for supply parameter modeling is used. This device is called artificial mains network. Principal circuits of standard artificial mains networks for 3...9 kHz frequency band are shown in Fig. 6.4.1 [94]. Artificial mains network (AMN) is connected between power consumer (PC) input and filter (F) output, which is connected to the power supply to filter own electromagnetic noises of the supply, see Fig. 6.4.2. Then noise measuring device (NMD) would measure noises that are emitted by the power consumer to the main.

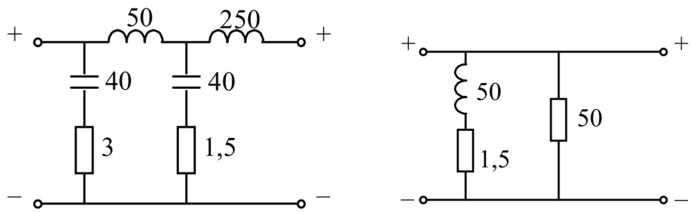


Fig. 6.4.1

Industrial radio noises (IRN)- electromagnetic noises in the radio frequency band (more than 150kHz), that are generated by electric and electronic devices without taking high-frequency radio transmitter sections into account. There are two common standards, that limit noise emission of technical devices what are used in living quarters, business areas and work areas with low power consumption [95], industrial areas [96].

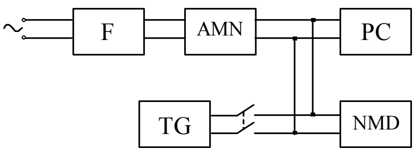


Fig. 6.4.2

Technical devise belongs to the industrial area if one of the next conditions is true:

1) industrial, medical, domestic high-frequency devices are presented;

2) frequent commutation of significant inductive or capacitive loads;

3) significant values of currents that are consumed by the equipment and therefore significant magnetic fields.

For living areas, business areas and industrial areas with low energy consumption maximum allowed levels of noise emission for 0,15...30MHz frequency band are set 84...74 dB (relative to 1μV) for quasi peak value and 74...64 dB for average voltage value. For noise current level these values are 40...30 dB (relative to 1 μA) and 30...20 dB correspondingly. For industrial areas these maximum allowed levels of noise emission are equal to 79...73 dB for quasi peak voltage value and 66...60 dB for average voltage value.

On the basis of these common standards special standards for concrete types of technical systems were developed. For example, there is a standard for conductive electromagnetic noises of stabilized dc current sources [97], that additionally norms noise emission within 10...150 kHz frequency band. Another standard is for industrial radio noises from domestic electric appliances, electric tools and similar devices, including regulation devices on the basis of semiconductor keys [98].

Inductive noise emission. Common noise emission standard from technical devices in industrial area [96] determines norms of inductive noise emission within 30...1000 MHz on 40...47 dB level (relative to 1 μV/m), that is measured 10 meters away from technical device. Tests are performed if people complain about presence of the noises in television and radio reception. To perform tests anechoic box and screened room are necessary. If there is no such places tests should be performed at the open area.

There are also special norms for maximum magnetic field intensity within 50 Hz...50 kHz frequency band for certain types of devices on the level 4...0,4 A/m for frequencies 50...500 Hz and 0,4...0,01 A/m for 0,5...50 kHz. Field intensity is measured at the certain remote point (approximately 1m from device).

Defining power electronics converters noise emission level.

Power electronics devices with their abrupt changes of currents and voltages of semiconductor devices during the commutations are the sources of intensive electromagnetic noises. Thus, when semiconductor diodes turns off due to the accumulated current carriers in the *p–n-* junction for some t_s time interval after forward current there is also a reverse current i_r, which could have significant amplitude (depending on forward current decrement rate at the turning off). Reverse current stops abruptly during time interval about 1 μs, this causes huge voltage

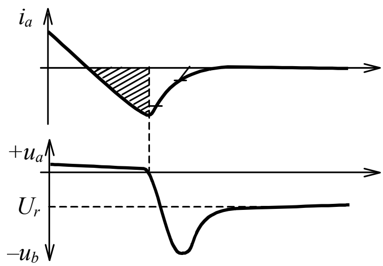


Fig. 6.4.3

with wide spectrum pulse on the diode due to the presence of inductance in the diode external chain, as it is shown in Fig. 6.4.3.

To limit this overvoltage pulse, and therefore noise emission from the diode, the diode is shunted with capacitance and resistor to limit the current – this is so called RC-chain or snubber.

On the contrary, thyristors generate noises at the turning on, that is accompanied by abrupt decrease of forward voltage and fast forward current increase that causes wideband noise spectrum. At the turning off thyristors have the same processes as diodes. Powerful transistors, such as IGBT, have turn on/off times as thyristor or less, especially FET – transistors, that is why they have broad noise spectrum.

To evaluate noise emission level a quasi peak electromagnetic noises U within 0,15...30 MHz frequency band were measured for 3-phase transistor bridge rectifier with resistive load [99]. Noise level in decibels relative to initial level of $1\mu V$ depending on the frequency is shown in Fig. 6.4.4. Curves *1* and *2* indicate the regime with $\alpha = 0$, nominal load and 1/5 of nominal load correspondingly. Curves *3*, *4* indicate the regime with $\alpha = 70°$, nominal load and 1/3 of nominal load correspondingly.

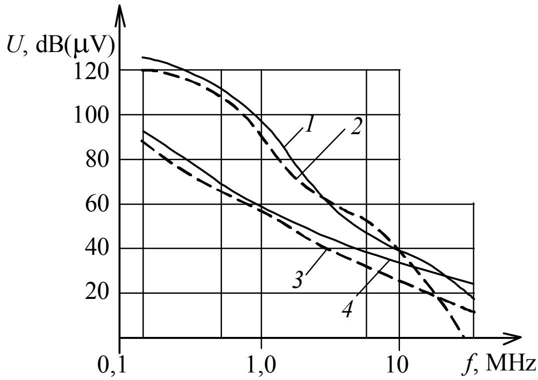


Fig. 6.4.4

Measurements confirm the hypothesis about domination of noises when thyristor turns on and there is abrupt voltage decrease that is aggravated when α increases. There is weak dependence of noises on the load current. Fig. 6.4.5 shows noise emission measurement results for the same frequency band for other types of converters. Curves *1*, *2* are for ac voltage regulator with active and active-inductive loads correspondingly. Curves *3*, *4* are for 3-phase half-wave rectifier with active and active-inductive loads. Noise emission level shows weak dependence on the load type. But noise level strongly depends on the rectifier topology. In the full-wave circuit (bridge) noises several times smaller than in half-wave one.

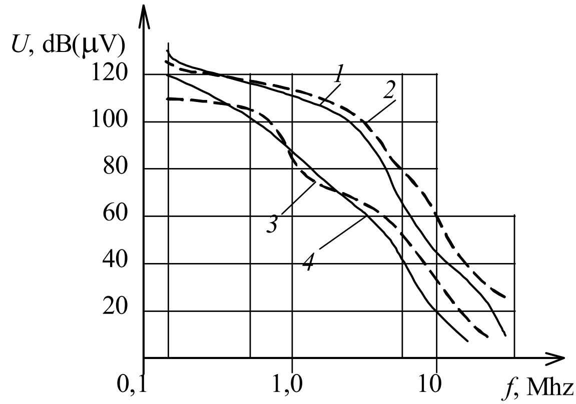


Fig. 6.4.5

It is evident that noise emission measurement for power electronics device is very difficult task because an anechoic box is required. This circumstance makes difficulties for necessary certification of devices that is required by state standards. That is why a simple calculation method for defining the level of electromagnetic noises on the basis of calculated voltage (current) spectrum with engineering accuracy is of utmost importance [99, 100].

For typical trapezoid form of pulses that modulate anode current with amplitude A, rise/fall time t^t, flat top duration t and period T, of idealized rectifier (see Fig. 6.4.6) the amplitude of n-th Fourier harmonic is

$$A_n = \frac{2A(t+t')}{T}\frac{\sin \pi f(t+t')}{\pi f(t+t')}\frac{\sin \pi f t'}{\pi f t'}, \tag{6.4.3}$$

where $f = n/T = nF$ - the frequency of n-th harmonic component relative to the pulse repetition rate F. Fig. 6.4.7 shows

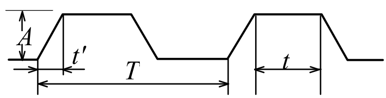


Fig. 6.4.6

harmonic spectrum (thin vertical lines), its envelope and also envelope of maximal (modulo) higher harmonics amplitude (dashed line). However, it is difficult to use these envelopes due to their nonlinearity and uncertainty of definition of wideband properties of signal frequency spectrum. It is more convenient to plot the envelope for maximal amplitudes in logarithmic scale.

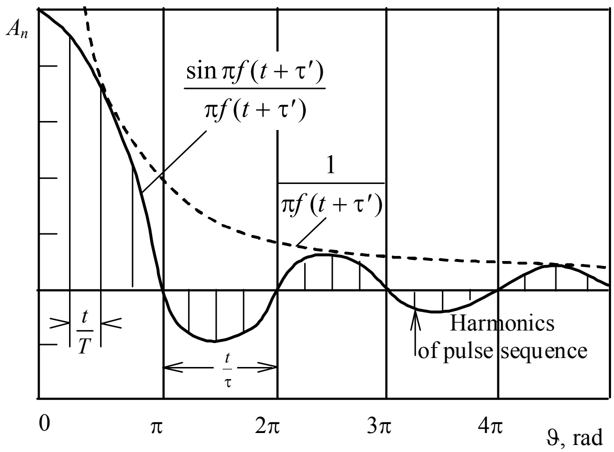


Fig. 6.4.7

For maximum simplification (piecewise linearization) of spectrum plot in logarithmic scale (Fig. 6.4.8) two additional transformations are made. Firstly, it is evaluation of noise level in microvolts relative to the frequency band, usually of 1MHz. The number of harmonic components that are contained in final frequency band (here $\Delta f = 1\text{Mhz}$) is equal to the width of this interval divided by frequency spacing (pulse repetition rate F). Total noise $V(\Delta f)$ in this frequency band is equal to the product of average spectral lines amplitude A_n and number of lines within this frequency band $(\Delta f / F)$.

By dividing total noise by bandwidth $\Delta f = 1\text{Mhz}$ we will have

$$\frac{V(\Delta f)}{\Delta f} = A_n \frac{\Delta f}{F} \frac{1}{\Delta f} = \frac{2A(t+t')}{TF} \frac{\Delta f}{\Delta f} \frac{\sin \pi f(t+t')}{\pi f(t+t')} \frac{\sin \pi f}{\pi f_1'} \left[\frac{\mu\text{V}}{\text{MHz}} \right]. \qquad (6.4.4)$$

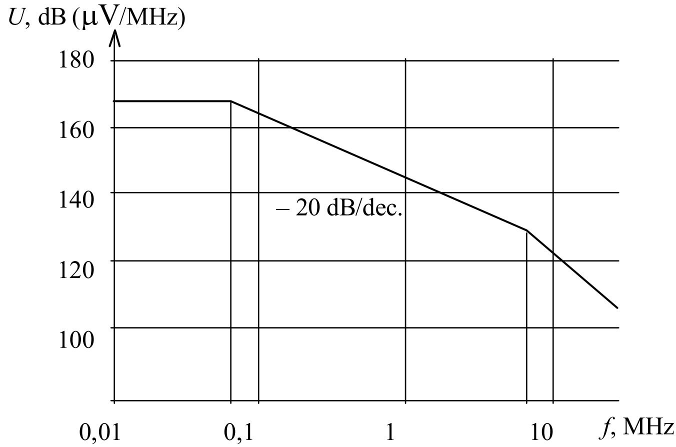


Fig. 6.4.8

Secondly, after taking the logarithm of this equation and its multiplication by the factor of 20 to transfer to decibel units, we will have:

$$V^{*} = 20\lg\left(\frac{V(\Delta f)}{\Delta f}\right) = 126 + 20\lg A(t+t') + \\ + 20\lg\frac{\sin \pi f(t+t')}{\pi f(t+t')} + 20\lg\frac{\sin \pi f t'}{\pi f_1'}, \quad (6.4.5)$$

then asymptotic piecewise-linear approximation is performed (6.4.5). Namely, for low frequency area when sinus of two last components could be substituted by their arguments, which will made this components equal to zero, instead of (6.4.5) we obtain

$$V_{1.f}^{*} = 126 + 20\lg A\left(t+t'\right). \quad (6.4.6)$$

Assume that the upper bound of low frequency area is the situation when an argument of sine of third component in (6.4.5) is unitary:

$$f_1 = \frac{1}{\pi(t+t')}. \quad (6.4.7)$$

Low frequency area of the wideband spectrum is presents as horizontal line in the logarithmic scale in Fig. 6.4.8. For medium frequencies area the equality $\sin \pi f t' = \pi f t'$ is still true due to the small value of t'. Therefore the last term in (6.4.5) is still negligibly small. But, for this area we can state that

$$\sin \pi f (t + t') = 1 .$$

This is prove by the fact that to define wideband spectral characteristic only the harmonics with maximum amplitude in each half-wave of common spectral function envelope are necessary.

Then, for medium frequencies approximation equation for (6.4.5) will be written as

$$V^{*}_{\mathrm{m.f}} = 116 + 20\lg A - 20\lg f \ . \qquad (6.4.8)$$

This second linear section begins from frequency corresponded to first fold (f_1) and has –20 dB/dec slope until the frequency of second fold *f2*. Upper bound for medium frequencies f_2 is defined in the same manner as f_1:

$$f_2 = \frac{1}{\pi t'} . \qquad (6.4.9)$$

For high frequencies area it is possible to state that $\sin \pi f t' = 1$. Then spectral function equation for high frequency area could be written as

$$V^{*}_{\mathrm{h.f}} = 106 + 20\lg\left(A/t'\right) - 40\lg f , \qquad (6.4.10)$$

i.e. linear slope of the plot here –40 dB/dec, and it starts from f_2 frequency.

Universal set of frequency characteristic of noises that are generated by trapezoid signals is shown in Fig. 6.4.9. Using this data it is possible to build local plot of noise spectral characteristics (as in Fig.6.4.8) for concrete pulse parameters A*, t, t'*, *T*.

Universal set of frequency characteristics of the truncated sine noise is shown in Fig. 6.4.10. There are similar plots for another pulse forms in [99, 100] – i.e. for sawtooth, normal distribution shape, sine module.

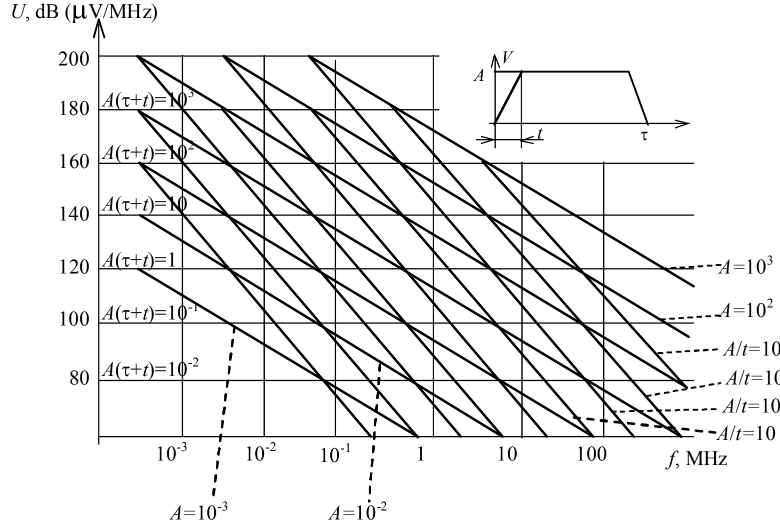


Fig. 6.4.9

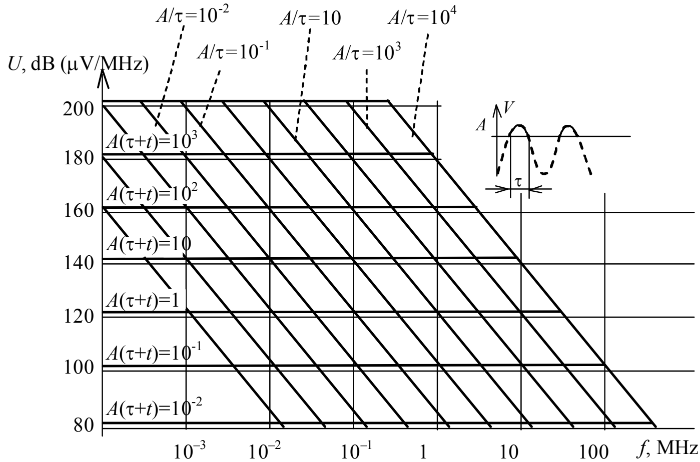


Fig. 6.4.10

Results that are obtained to calculate conductive noise emission could be used for inductive radio noises calculation. Evaluation of field that is radiated by long wire with known noise voltage spectrum one meter away from the wire is obtained by reducing level of conductive noise by 33dB (μV/MHz). In this case radiating field unit is dB (μV/m/MHz) for wideband emission. To obtain field value at other distances it is necessary to additionally subtract 20 lg (*d*), where *d* – distance measured in meters.

To calculate magnetic field level that is radiated by the wire coils with known current spectrum, it is necessary add to the current spectrum level next value:

$$H_{\Delta} = 20\lg\frac{NA}{4\pi d^3}, \qquad (6.4.11)$$

where N – number of coils in current circuit; A – coil area, m^2; d – distance to the current circuit, m.

Units to measure magnetic field in this case are dB (μA/MHz). Units to measure magnetic field power that is calculated with shown procedure are dBμA·coil/m/MHz.

Thus, costly noise emission measurement procedure for power electronics devices that requires expensive devices and special apartments could be substituted analytical engineering evaluation for known voltage and current waveforms.

6.5.* FEATURES OF THE ELECTRICAL ENERGY STANDARDS IN AUTONOMOUS POWER SUPPLY SYSTEMS

GOST 13109-97 for electrical energy quality in common mains doesn't apply to autonomous mains and power supply systems. For last named there are standardized requirements to electrical energy quality in valve converters [101], uninterruptible power supplies [102, 103], moving objects power supply systems: airplanes [104], sea crafts, river boat [105], automobiles [106, 107], electrical locomotives [108], and also in power supplies for communication devices [109, 110], computer techniques [111]. Let's consider additional functions for hardware-software complex power quality measurement that are caused by the standards for autonomous power supply systems made on the basis of valve converters.

ADDITIONAL POWER QUALITY FACTORS IN STANDARDS FOR POWER CONVERTERS

Common standard for all semiconductor converters of electrical energy [101] contains a set of additional power quality factors that are absent in standards for common mains [44].

Voltage amplitude modulation coefficient is defined as

$$k_{\text{mod}} = \frac{U_{\text{mod}}}{\sqrt{2}U_{\text{nom}}}, \tag{6.5.1}$$

where U_{mod} – half of the difference between minimal values of voltage amplitude envelopes;

$\sqrt{2U_{\text{nom}}}$ –nominal sine voltage amplitude.

The ratio of voltage to frequency U/f, which is defined as rms voltage in volts and frequency in Hertz.

Ripple factor

$$k_{\text{rip}} = \frac{U_{\text{rip}}}{U_{\text{nom}}} \cdot 100\%, \tag{6.5.2}$$

where U_{rip} – maximum (amplitude) ac voltage component; U_{nom} – nominal dc voltage value.

Coefficient of efficiency and converter power factor are defined in section 1.2.

ADDITIONAL POWER QUALITY FACTORS IN POWER SUPPLY SYSTEMS OF AIRCRAFTS AND HELICOPTERS

Power quality standard for board supplies of aircrafts and helicopters contains an additional set of power quality factors that should be registered:

– *shearing angle between phase vectors* at normal and emergency operation (possible limits are 116...124°);
– *frequency drift rate* (not more than 2,5 Hz/sec);
– *frequency modulation ratio* (not more than 1%);
– *frequency components of main voltage frequency modulation envelope* with period within 0,01...10 sec range;

– *amplitude value coefficient* (amplitude ratio), that is limited by 1,4 ± 0,15 band;
– *voltage dc component* – not more than ±0,1V for nominal rms value of phase voltage of 115V;
– *voltage interruptions* – not more than 80ms.

The features of measurement of other power quality factors of aircraft and helicopter board systems are usually the broad bands of their deviation; this should be provided in linear monitoring mode. First of all this apply to ac voltage frequency that is equal to f_{nom} = 400 Hz in nominal mode and could deviate within 380...420 Hz band and within 360...440 Hz band in malfunction mode.

ADDITIONAL POWER QUALITY FACTORS IN STANDARDS FOR CRAFT BOARD STANDARDS

Due to the possible length of craft board mains power quality factors are standardize at power source taps and consumer taps separately. Factors that differ from ones for common mains are:

– dynamical voltage deviations at the consumer taps, limited within + 20 … – 30 %,
– *dynamical frequency deviation at the source and consumer taps* are limited within ±10 % value,
– *dynamical disconnection time for voltage (1.5sec) and for frequency (5sec)*, at the consumers taps,
– *linear (phase) voltage dissymmetry at the source and consumer taps* – not more than ±15 %

FEATURES OF STANDARDS FOR MOTOR TRANSPORT ELECTRICAL EQUIPMENT AND ITS ELECTROMAGNETIC COMPATIBILITY

Due to the limited power of automobile electrical sources (accumulators and generators) the standard sets different norms for voltages of sources (generators) and consumers. For the sources they are 7, 14, 28 V, for consumers 6, 12, 24 V.

Capability to operate for consumer should be provided in nominal mode (except starting mode) within voltage band of (90...125 %) U_{nom}. Nominal voltages for nominal parameters of current consumers are 6,7; 13,5 or 27 V. Radio noises from automobile electrical equipment are also standardized. Conductive radio noises in the electrical equipment main shouldn't be more than (relative to 1μV level)

$$L_U = 85 - 11{,}5\lg\frac{F}{0{,}15}\ \text{[dB]} \qquad (6.5.3)$$

within frequency band F 0,15...30 MHz but no more than 68dB within 30...110 MHz band.

Magnetic field intensity in the passenger compartment of automobile shouldn't be more than (relative to 1μV level)

$$L_E = 100 - 11{,}5\lg\frac{F}{0{,}15}\ \text{[dB]} \qquad (6.5.4)$$

within 0,15...30 MHz, but no more than 68 dB within 30...110MHz band.

FEATURES OF POWER QUALITY STANDARDS FOR POWER SUPPLIES OF COMMUNICATION EQUIPMENT

The standard [110] is now the standard for communication and informatization field, not GOST and regulates not only power quality but also nose immunity and noise emission. These were absent in previous GOST [109]. The standard has next features in definition of power quality factors:

– steady voltage deviations at the dc power supplies output taps are defined as 9,0...7,5 V for nominal voltage 48 and 12 V and for nominal voltage 60V;

– efficiency factor of ac-dc converters shouldn't be less than 0.8 for output power value up to 2kW and 0.9 for higher power values;

– voltage ripple for rms value odd harmonic components sum within frequency band from 20Hz to 150kHz shouldn't be more than 50mV;

– voltage unbalance coefficient and irregularity of currents consumed by individual phases are defined by GOST 29192-91 and GOST R513.174.1-2000, not by GOST 13709-97;

– voltage ripple for psophometric rms voltage value no more than 2mV:

$$U_{\text{r.psoph}} = \sqrt{\sum_{k=2}^{\infty} (pU_{(k)})^2} \,. \qquad (6.5.5)$$

Psophometric factor p is for considering different sensitivity of human ear to the harmonic components of different frequencies. Unitary value of this coefficient is background noise effect of 80Hz frequency, and relative intensity of effect of other frequencies are evaluated using acoustic effect coefficient *p*, that is defined by curve shown in Fig.6.5.1

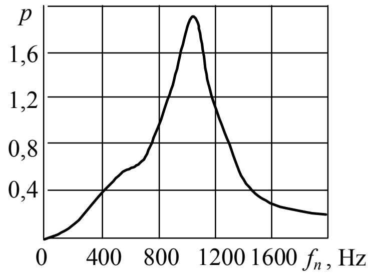


Fig. 6.5.1

It is appropriate mention here that analogous effect of different human eye sensibility to the luminous flux deviations is considered by flicker dose calculation (refer to section 6.2.1).

It is typical that railway power supply standard utilize psophometric normalization of current harmonics in pulling main due to the non-sinusoidal current, that is consumed by electrical locomotive. Its psophometric value shouldn't overcome 4A.

Thus, it is possible to conclude that:

1. Power quality standards for autonomous power systems contain lots of additional power quality factors, that are absent in power quality standard for common mains [44]. This fact requires significant increase of software-hardware power monitoring complex resources if this complex should monitor power quality of autonomous power supplies.

2. Technical terminology, that is used power quality standards was being developed during several decades and was contained in recommendations of USSR Academy of Sciences [112] and in electrical energy terminology and definitions standard [113]. Last standard is out of date relative to today's electrical technique with nonsinusoidal currents; it is also incomplete and erroneous for some definitions. This fact makes description and calculation of some power quality factors set more difficult.

3. Requirements specification for power electronics devices designing have to contain a section dedicated to their electromagnetic compatibility with mains and environment [114].

6.6.* PROBLEMS OF A POWER THEORY AT NONSINUSOIDAL VOLTAGES AND CURRENTS

6.6.1. APPROACH TO REACTIVE POWERS DETERMING

For electric mains with nonsinusoidal waveforms of voltages and currents that is typical for valve converters there are two very actual theoretic problems with a great practical meaning:

1) of a reactive power determining;
2) of a full power determining.

Nowadays there are monographs [21, 39, 47, 50] and a great number of articles in which different approaches to their solution are considered.

Here a conception of an approach to a solution of a reactive power problem is stated from the point of view of an electrical engineer-designer directed to a clarification of that and how we have to calculate in a different way at an energetic network with a nonsinusoidal current in contrast to a network where an electric energy transfer and consumption is at sinusoidal currents and voltages.

At the fig. 6.6.1 an outline equivalent circuit of an electric network (*a*) to which we can reduce the next typical cases of an autonomous and not autonomous power supply systems, at the fig. 6.6.1, *b* it's reductive diagram is represented.

1. e source emf and a current of an impassive load i_{load} are sinusoidal.
2. a source emf is sinusoidal, a current of a load current source is nonsinusoidal (valve converters and others).
3. a soure emf is nonsinusoidal (a valve converter), a current of a load current source is sinusoidal due to a presence of filters.
4. a source emf is nonsinusoidal, a current of alod current source i_{load} is also nonsinusoidal.

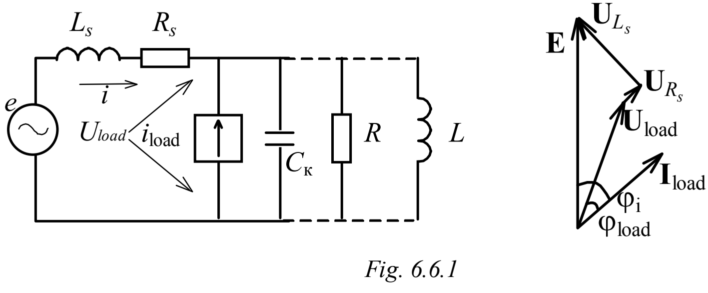


Fig. 6.6.1

Here L_s, R_s – channel parameters (an energy transfer line from a source to a consumer); $C_к$ – a parameter of a capacitate compensator of a reactive power at consumer clamps; L, R – a possible passive part of a linear load.

At first, let's enumerate that additional problems for an engineer-designer which are caused by a reactive power presence of a load in the first case, i.e. even at sinusoidal mains currents. A reactive power presence at a load Q_{load} that become here as φ_{load} shift of a current phase C_{load} relatively a load U_{load} voltage (then, relatively a source emf – a shift of a current φ_{s} source) leads to a necessity of calculation undesirable secondary manifestations ("a damage") due to Q_{load}, such as:

1) a current rms value increasing at e source to transfer the same active power to a load that entails a source full power increasing

$$\frac{I_{\text{s}}}{I_{\text{load.a}}} = \frac{1}{\cos\varphi_{\text{load}}} = f_1\left(Q_{\text{load}}\right), \quad \dot{I}_{\text{load}} = I_{\text{load.a}} + jI_{\text{load.r}}; \quad (6.6.1)$$

2) additional loss of an active power at transmission facilities R_s resistance

$$\Delta P_{R_s} = I_{\text{load.r}}^2 R_s = R_s \frac{Q_{\text{load}}^2}{U_{\text{load}}^2} \sin^2\varphi_{\text{load}} = f_2\left(Q_{\text{load}}\right); \qquad (6.6.2)$$

3) additional voltage loss at a line (at L_s, neglecting by R_s)

$$\Delta U_{L_s} = I_{\text{load.r}} \omega L_s = \omega L_s \frac{Q_{\text{load}}^2}{U_{\text{load}}^2} \sin^2\varphi_{\text{load}} = f_3\left(Q_{\text{load}}\right); \quad (6.6.3)$$

4) changing of a fundamental voltage at a load and it's rms value, correspondingly(neglecting by R_s)

$$U_{\text{load}} = \left[\frac{1}{T}\int_0^T \left(e - L_s \frac{di_{\text{load.r}}}{dt}\right)^2 dt\right]^{\frac{1}{2}} = f_4\left(Q_{\text{load}}\right); \qquad (6.6.4)$$

5) changing of a voltage phase at a load (a voltage of an autonomous mains) relatively a source voltage

$$\Delta\varphi = \varphi_{\text{s}} - \varphi_{\text{load}} = \operatorname{arctg}\frac{Q_{\text{load}} + Q_{L_S}}{P_{\text{load}}} - \operatorname{arctg}\frac{Q_{\text{load}}}{P_{\text{load}}} = f_5\left(Q_{\text{load}}\right); \quad (6.6.5)$$

6) costs for a setting (if it's necessary) of compensating (cosinusoidal) capacitors of a power

$$Q_{\text{comm}} = K_{\text{comm}} Q_{\text{load}} = f_6 \left(Q_{\text{load}} \right), \tag{6.6.6}$$

where $K_{\text{comm}} = Q_{\text{comm}} / Q_{\text{load}}$ – a compensation degree;

7) additional loss of an active power at compensator capacitances

$$\Delta P_{\text{comm}} = Q_{\text{comm}} \operatorname{tg} \delta_k = K_{\text{comm}} \operatorname{tg} \delta_{\kappa} Q_{\text{load}} = f_7 \left(Q_{\text{load}} \right). \tag{6.6.7}$$

From correlations (6.6.1) – (6.6.7) it can be seen that all secondary manifestations of a reactive power presence at a network are expressed through a value of this reactive power Q_{load}, i.e. all damages depend on the same calculated value – a load reactive power through their partial coefficients. That's why at calculation of networks with sinusoidal waveforms of a voltage and current it's sufficient to determine only a fundamental reactive power.

Now let's consider the same "damages" for the secondary typical case of a power supply system:

1) a current rms value increasing at a source (neglecting by L_s)

$$\frac{I}{I_a} = \frac{S}{P_{\text{load}}} = \frac{\sqrt{P_{\text{load}}^2 + Q^2}}{P_{\text{load}}} = f_{21}(Q), \tag{6.6.8}$$

where

$$Q = \sqrt{S^2 - P_{\text{load}}^2} = \sqrt{E^2 I^2 - P_{\text{load}}^2} \tag{6.6.9}$$

– all (reactive) power of a discrepancy between S and P;

2) additional loss of an active power at R_s resistance of transmission facilities

$$\Delta P_{R_s} = I_{\text{load}}^2 R_s = \frac{Q^2 + P_{\text{load}}^2}{E^2} R_s = f_{22}(Q); \tag{6.6.10}$$

3) a fundamental voltage decreasing at load clamps (at a distribution network) is determined on (6.6.4) through a fundamental reactive power;

4) a voltage rms value changing at a load due to a presence of L_s (at an absence of C_{κ}) is determined through a new value of a load voltage rms value

$$U_{\text{load}}^2 = \left[\frac{1}{T} \int_0^T \left(e - L_s \frac{di_{\text{load.r}}}{dt} \right)^2 dt \right]^{\frac{1}{2}} = f_{23}\left(Q_1, \bar{K}_{\text{c.h}} \right), Q_1 = \left(e, \frac{di}{dt} \right); \tag{6.6.11}$$

5) costs for setting up (if it's necessary) compensating capacitors estimated by an installed power that we determine as a production of rms values of this capacitor voltage and current

$$S_к = U_{\text{load}} I_к .$$

Rms values of pointed voltages and current we will determine by ADE1 method. A differential equation for a capacitor voltage based on the fig. 6.6.1, is the next

$$L_s C_к \frac{d^2 u}{dt^2} + L_s \frac{di_{\text{load}}}{dt} + u = e . \qquad (6.6.12)$$

Making an algebraization of the equation (6.6.12) we will write the next

$$U^2 = f_{24}(\bar{K}_{\text{v.h.}}, (\bar{e}^{(2)}, \bar{i}_{\text{load}})), \qquad (6.6.13)$$

where a scalar production

$$(\bar{e}^{(2)},\ \bar{i}_{\text{load}}) = \frac{1}{\omega^3} \sum_k \frac{U_k I_{\text{load}(k)}}{k^3} \sin \varphi_{\text{load}(k)} \equiv Q_2 \qquad (6.6.14)$$

has a dimension of a reactive power formed by a weighted (with $1/k^3$ coefficient) sum of separate harmonics reactive powers. Here $\varphi_{load.k}$ – a shifted degree on a phase between the same voltage harmonics at a capacitor and a load current.

To determine a current rms value through a compensating capacitor we obtain an equation relatively du/dt from (6.6.14) and after this equation algebraization we will have the next expression

$$I_k = (C\bar{U}^{(-1)}) = f(U, (\bar{e}, i_{\text{load}}), K_{\text{v.h.}}).$$

A scalar production which is the next

$$(\bar{e}, i_{\text{load}}) = \frac{1}{\omega} \sum_k \frac{U_k I_{\text{load}(k)}}{k} \sin \varphi_{\text{load}(k)} \equiv Q_3 \qquad (6.6.15)$$

again has a dimension of a reactive power but different from that at (6.6.14). From here it follows that using of some universal, general reactive power such as Budeanu power which is the next

$$Q_{\text{B}} = \sum_k Q_k = \sum_k U_k I_k \sin \varphi_k$$

isn’t represented as reasonable, especially that this power, as it's known, doesn't have a physics meaning and a calculated appliance [115].

So, in the case of a power supply system with a nonsinusoidal load current various partial “damages” from an inactive component of a load power are determined by various partial reactive powers of a system but not the same fundamental reactive power as it was in the case of a sinusoidal load current. At that to determine “a damage” from a nonsinusoidality of a load current it’s necessary to know an integral harmonic factor of the first order current (and higher orders – in the case of a line substitution by an electric network of a higher than the first order).

By similar calculations we can make sure that at two remaining typical cases of a power supply system (with a nonsinusoidal voltage source at an input) relatively “a damage” determining we have an identical situation (in the part of a quantity of calculated reactive powers). At that “a damage” of a nonsinusoidal voltage is characterized by a set of integral harmonic factors of a mains voltage.

Thus, at networks having nonsinusoidal energy processes various kinds of a damage connected with their waveform distortion are determined by different expressions having a structure of reactive powers formulas that can be called as partial reactive powers. At this a calculation of partial reactive powers at nonsinusoidal waveforms of energy processes at a network can be quickly made also by a direct method of calculation of energetic factors – an ADE method due to that it isn't required to find decisions of differential equations.

6.6.2. APPROACHES TO DETERMINE OF A FULL POWER AND IT’S COMPONENTS

At networks with sinusoidal waveforms of a voltage and current a determination of a reactive and active powers has an evident physical meaning. An active P power characterizes a part of an instantaneous power that can be converted into another types of an energy (a thermal, mechanic, chemical, electromagnetic). A reactive Q power characterizes an exchange energy between a source and a load that loads a network (makes loss in it) but can’t be used usefully. A vector summing of an active and reactive current (voltage) components at a network leads to a geometrical sum of an active and reactive powers at a resulting power that is called as an apparent or full S power:

$$S = \sqrt{P^2 + Q^2} = UI \; . \qquad (6.6.16)$$

The second form of calculation S through an rms value of a voltage and current of an electro-technical device allows to use a full power as a calculated value to characterize resource costs for a constructive realization of the given device. Conditionally we can suppose that a rms value of I current determines active power loss at a device from a through current flowing between an input and output, a rms value of U voltage – loss of an active power from a current flowing between input device clamps (isolation leakage currents, excitation currents of coils with steel cores) when there is an idling at an output. In other words, a full power characterizes sum costs (at a some scale) of resources (copper, steel, isolation) for a realization of a device.

A determination of a reactive and active power is significantly complicated when network energy processes are nonsinusoidal [115–125]. An appearing set of partial reactive powers was considered above. As a consequence a narrowness of a full power detremination through an active power and some partial reactive power became evident corresponding to the first equality at the expression (6.6.16). At the same time a proximity of a full power determination through a rms value of a voltage and current is seen. It connected with that for current higher harmonics at device conductors we have to take into account a correlation of a conductor resistance from a harmonic frequency due to a current displacement effect and a proximity effect. Often this correlation of r_f resistance from f frequency is approximated by the next expression

$$r_f = r_{50}\sqrt{\frac{f}{50}} = r_{50}\sqrt{k}\ , \qquad (6.6.17)$$

where r_{50} – a resistance at a frequency of 50 Hz; k – a number of a current I_k harmonic relatively a frequency of 50 Hz.

Now loss of an active power at a conductor ΔP_{Cu} will be determined not by a rms current value but by the next value

$$\Delta P_{\mathrm{Cu}} = \sum_{k=1}^{\infty} I_k^2 r_f = r_{50}\sum_{k=1}^{\infty} I_k^2\sqrt{k} = r_{50}\sum_{k=1}^{\infty}(I_k k^{1/4})^2\ , \qquad (6.6.18)$$

which depends on a current spectrum and can significantly surpass a current rms value [113].

It's interesting to notice that the last form of writing is an evidence of a possibility of it's interpretation as a derivative rms value square of a current of 1/4 degree at using terms of derivative and integral fractional degrees [127]. Correspondingly we can bring in *a current differential harmonic factor of ¼ order*

$$K_{\text{c.h.}}^{(1/4)} = \sqrt{\sum_{k=2}^{\infty}\left(\frac{I_k}{I_1}k^{1/4}\right)^2}\ . \qquad (6.6.19)$$

The more complicated is situation with a loss determination at a core magnetic path of reactors and transformers that are approximated by an expression

$$\Delta P_{\text{Fe}} = \rho_{\text{c}} V_{\text{c}} K_{\text{c}} f^{\alpha} B_m^{\beta}, \qquad (6.6.20)$$

where ρ_{c} – a specific weight of a core material, V_{c} – it's volume; K_{c} – a constant that depends on a core material; B_m – a limit induction.

Table 6.6.1

Values of magnetic path parameters

Material	B_m	μ	ρ_c kg/dm^3	K_c	α	β
Powdered Fe	2,1	4500	6	0,1–10	1,1	2
Si-steel	2,0	10000	7,65	$0{,}5\cdot10^{-3}$	1,7	1,9
Ni–Mo alloy	0,8	250	13	$5\cdot10^{-3}$	1,2	2,2
Mn–Zn ferrite	0,4	2000	4,8	$1{,}9\cdot10^{-3}$	1,24	2
Ni–Zn ferrite	0,3	400	4,8	$2{,}5\cdot10^{-3}$	1,6	2,3

Values of these parameters for various core materials are represented at the table 6.6.1 [128, 129].

Resulting loss at a core material are determined at a method of superposition (at a physical mechanism invariability of energy loss at various voltage frequencies) by calculating loss components from an each harmonic of an applied voltage and in the common case they can's be expressed through a rms value of an applied voltage. Resulting loss are also determined by some calculated value obtained by a some operator conversion of an applied voltage curve.

At a particular but an interesting for a practice case we can find this calculated value of a voltage if at the expression (6.6.20) for a simplicity set α = 1,5, β = 2 that is close to real values for Si-steel and Ni–Zn-ferrite. Then after some conversions the expression (6.6.20) can be led to the next look

$$\Delta P'_{\text{Fe}} = \rho_c V_c K_c f^{3/2} B_m^2 = K_1 U_1^2 \left[1 + \sum_{k=2}^{\infty} \left(\frac{U_k}{U_1 k^{1/4}} \right)^2 \right]. \qquad (6.6.21)$$

In this case core loss *are determined by a voltage integral harmonic factor of ¼ order.*

Loss at dielectrics (capacitors, isolators) are also nonlinearly depend on a frequency and value of a voltage and, for example, for cosinusoidal capacitors connected to improve a power factor are equal [130] if a coefficient of dielectric loss $\mathrm{tg}\,\delta$ doesn't depend on a frequency:

$$\Delta P_{di} = \omega C \,\mathrm{tg}\delta \sum_{k=1}^{\infty} k U_k^2 = \omega C \,\mathrm{tg}\delta U_1^2 \left[1 + \sum_{k=2}^{\infty} \left(\frac{U_k}{U_1} k^{\frac{1}{2}} \right)^2 \right] =$$

$$= \omega C \,\mathrm{tg}\,\delta \left[1 + \left(\overline{K}_h^{\left(\frac{1}{2}\right)} \right)^2 \right] U_1^2, \qquad (6.6.22)$$

where $\overline{K}_h^{\left(\frac{1}{2}\right)} = \sqrt{\sum_{k=2}^{\infty} \left(\frac{U_k}{U_1} k^{\frac{1}{2}} \right)^2}$ – *a voltage differential harmonic factor of ½ order* that also corresponds to a term of a fractional differentiation [127].

If dielectric loss coefficient $\mathrm{tg}\,\delta$ linearly depends on a frequency then loss at a capacitor dielectric will be equal to

$$\Delta P'_{di} = \omega C \,\mathrm{tg}\,\delta \sum_{k=1}^{\infty} k^2 U_k^2 = \omega C \,\mathrm{tg}\,\delta U_1^2 \left[1 + \sum_{k=2}^{\infty} \left(\frac{U_k}{U_1} k \right)^2 \right], \qquad (6.6.23)$$

i.e. they are determined by a voltage differential harmonic factor of the first order.

It's known that loss taking into account a current displacement from higher harmonics at induction motors can be approximately represented as [130]

$$\Delta P_{AD} = 2\Delta P_{\mathrm{c}} K_s^2 \sum_{k=1}^{\infty} \left(\frac{U_k}{U_1 k^{3/4}} \right)^2 = 2\Delta P_{\mathrm{c}} K_{\mathrm{s}}^2 \left[1 + \bar{K}_{\mathrm{h}}^{\left(\frac{3}{4}\right)} \right]^2 , \qquad (6.6.24)$$

where K_{s} – a starting current multiplicity; ΔP_{c} – loss at windings cooper at dc current; $\bar{K}_{\mathrm{h}}^{(3/4)}$ *– a voltage integral harmonic factor of ¾ order.*

Thus, an aforesaid demonstrates a possibility to connect resource costs for a creation of electro-technical devices operating at distorted waveforms of a voltage and current with rms values of a some function fron them in the common case. But this possibility also requires a further study and control. That's why nowadays for an electro-technical equipment operating at a public network of 50 (60) Hz with a practically sinusoidal voltage ($K_{\mathrm{h}} \leq 8$ % on [44]) and with known limitations on a spectrum of a current drawn from a mains [65], as before, they use a full power determination of an equipment through a production of rms values of a voltage and current.

Based on an accepted definition of a full power through a production of rms values of a voltage and current we can connect it by quality factors of a nonsinusoidal voltage and current through the next correlations:

$$S = UI = \sqrt{U_1^2 + U_{\mathrm{h.h}}^2}\sqrt{I_1^2 + I_{\mathrm{h.h}}^2} =$$

$$= \sqrt{U_1^2 I_1^2 + U_1^2 I_{\mathrm{h.h}}^2 + U_{\mathrm{h.h}}^2 I_1^2 + U_{\mathrm{h.h}}^2 I_{\mathrm{h.h}}^2} , \qquad (6.6.25)$$

dividing rms values of a voltage and current at a geometrical sum of rms values of a fundamental U_1, I_1 and higher $U_{\mathrm{h.h}}, I_{\mathrm{h.h}}$ harmonics, instantaneous values of which are orthogonal. Let's express a full power of nonsinusoidal energy processes through a full power of fundamental energy processes and a correcting multiplier considering processes nonsinusoidality degree:

$$S = U_1 I_1 \sqrt{1 + K_{\mathrm{c.h}}^2 + K_{\mathrm{h}}^2 + K_{\mathrm{h}}^2 K_{\mathrm{c.h}}^2} = \sqrt{P_1^2 + Q_1^2 + S_{\mathrm{d}}^2} , \quad (6.6.26)$$

where $K_{\mathrm{h}}, K_{\mathrm{c.h}}$ – a voltage and current harmonic factors, correspondingly;

$$S_{\text{d}} = \sqrt{\left(U_1 I_{\text{h.h}}\right)^2 + \left(U_{\text{h.h}} I_1\right)^2 + \left(U_{\text{h.h}} I_{\text{h.h}}\right)^2} =$$

$$= S_1 \sqrt{K_{\text{c.h}}^2 + K_{\text{h}}^2 + \left(K_{\text{h}} K_{\text{c.h}}\right)^2} \qquad (6.6.27)$$

– a distortion power; P_1, Q_1 – an active and reactive powers on a fundamental.

An operating group on nonsinusoidal modes of an international IEEE institute recommends a similar decomposition of a full power for practical calculations and measurements [121].

One else problem occurs at a full power determining for a three-phase mains. At present time the next variants of it's calculation are spread:

а) an arithmetical full power

$$S_A = S_a + S_b + S_c; \qquad (6.6.28)$$

б) a vector full power

$$S = \sqrt{(P_a^2 + P_b^2 + P_c^2) + (Q_a^2 + Q_b^2 + Q_c^2)}; \qquad (6.6.29)$$

в) a system full power

$$S = \sqrt{2} U_{\text{e}} I_{\text{e}}, \qquad (6.6.30)$$

where

$$I_e = \sqrt{\frac{I_a^2 + I_b^2 + I_c^2 + \rho I_0^2}{3}} \qquad (6.6.31)$$

– an equivalent current, $\rho = r_{\text{load}} / r$ – a correlation between a neutral wire resistance and a phase wire resistance;

– an equivalent voltage at a three-wire system

$$U_{\text{e}} = \sqrt{\frac{U_{ab}^2 + U_{bc}^2 + U_{ca}^2}{3}}; \qquad (6.6.32)$$

– an equivalent voltage at a four-wire system (with a null wire)

$$U_{\text{e}} = \sqrt{\frac{U_{an}^2 + U_{bn}^2 + U_{cn}^2 + U_{ab}^2 + U_{bc}^2 + U_{ca}^2}{4}}. \qquad (6.6.33)$$

A questioning of a fifty regional electric energy consumers of USA and Canada carried out by a working group on nonsinusoidal situations of an International organization IEEE [120] showed that by a determination of an arithmetical and vector full powers 22 % of questioned are used, only 4 % use a determination of a full power on an expression (6.6.30), also 6 % of questioned use to determine a full power of a phase “instrumentally” based it’s definition through average on a module U_{av} voltage and I_{av} phase current:

$$S_a = 1{,}11^2 U_{\text{av}} I_{\text{av}} . \tag{6.6.34}$$

Some users work with some of determinations (measurements) of a full power but 48 % of questioned didn’t answer at inquiry at all.

Thus, nowadays there is no an adopted way of determination (measurement) of a full power at a three-phase network. If for the case of a sinusoidal symmetrical three-phase network all ways give the same result then at the case of unbalanced, asymmetrical three-phase networks various ways of a full power determining will give different results at a measurement procedure on their basis. It is connected with various physical suppositions used at derivation of full power formulas.

An arithmetical full power characterizes a maximal active power that can be obtained from a mains at given rms values of it’s phase voltages and currents at connection wires.

A system full power characterizes maximal active power transferred from a source to a receiver at the same thermal influence to wires and isolation.

A carried out research of a correlation of active power normalized loss at a three-phase network from a square of a normalized full power determined on different formulas showed a linear correlation of these loss from only a system full power [124].

In the case of nonsinusoidal waveforms of voltages and currents of a three-phase network determinations of a full power will be more plural depending on it’s criteria and a kind of an electro-technical device for which it’s determined.

6.6.3. DECOMPOSITION WAYS OF AN ELECTIC NETWORK INSTANTANEOUS POWER

A basing of using an instantaneous power and a coordinate conversion. A full power decomposition considered above on it’s components (6.6.26) was based on integral values (as a rule, on rms values) corresponding components of voltages and currents. Using of these components is appropriate at carrying out calculations of electrical modes, choice of network real elements, measurement of electric energy characteristics. An averaging operation at a function rms value determining in a period carries in a some delay at an obtaining of the result. That's why using of current and voltage integral values to control electric modes (a stabilization, regulation on a program) is complicated if a high regulation dynamic is required. In these cases to control full power components it's more effective to use their definitions through corresponding components of an instantaneous power obtained (observed) at a real time mode.

Below there are considered *decompositions of an instantaneous power at three-phase networks* that are applied, first of all, at control systems of full power inactive components compensators (chapter 11, 13) and frequency converters for an electric drive.

A feature of a three-phase system of u_a, u_b, u_c voltages and i_a, i_b, i_c currents at a three-wire system (without a null wire) is an interdependency of phase variables since they are connected by equations

$$u_a + u_b + u_c = 0\,, \tag{6.6.35a}$$

$$i_a + i_b + i_c = 0\,. \tag{6.6.35б}$$

In other words, only two from three variables are independent and a value of the third variable is determined according to a bond equation (6.6.35). This specific of a three-phase system makes difficult a forming and regulating of phase variables at three-phase Power Electronics devices on an input (self-commutated inverters, direst frequency converters, ac voltage regulators, compensators of full power inactive components), although it can be used positively at a row of algorithms of three-phase converters control [21] when only two converter phases are actively controlled at any time instant.

T o obtain an independent control of converters output variables usually a constrained three-phase system of variables is converted to a two-phase system having independent variables for a separate (not constrained) regulation of active and reactive components of voltages, currents and powers.

The next two-phase systems are distinguished: α, β-system (a stationary frame of reference); *d*, *q*-system (a frame of reference synchronously rotating); *x*, *y*-system (a frame of reference rotating with an arbitrary velocity) and not constrained three-phase systems from which usually a system of symmetrical components is used.

α, β-axes conversion. Conversion formulas of three-phase system phase variables to α, β-two-phase system are the next

$$\begin{vmatrix} u_\alpha \\ u_\beta \end{vmatrix} = \sqrt{\frac{2}{3}} \begin{vmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} \end{vmatrix} \begin{vmatrix} u_a \\ u_b \\ u_c \end{vmatrix}, \tag{6.6.36}$$

$$\begin{vmatrix} i_\alpha \\ i_\beta \end{vmatrix} = \sqrt{\frac{2}{3}} \begin{vmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} \end{vmatrix} \begin{vmatrix} i_a \\ i_b \\ i_c \end{vmatrix} \tag{6.6.37}$$

or at a generalized matrix form

$$\mathbf{x}_\alpha = K_{a\alpha} \mathbf{M}_{a\alpha} \mathbf{y}_a, \tag{6.6.38}$$

where $\mathbf{x}_\alpha = (x_\alpha, x_\beta)$ at x equal to u or i, – a vector of α, β-conversion output; $\mathbf{y}_a = (y_a, y_b, y_c)$ at y equal to u or i, – a vector of α, β-conversion input; $\mathbf{M}_{a\alpha}$ – a conversion matrix from three-phase variables to two-phase α, β-variables; $K_{a\alpha}$ – a factor set a kind of invariance at a conversion (usually an invariance of a vector norm, power and others).

An instantaneous value of a real power at a frame of reference (α, β) are determined as a sum of scalar productions of similar voltage and current components and resolved at a constant $\bar{p}$ and variable $\tilde{p}$ components:

$$p = u_\alpha i_\alpha + u_\beta i_\beta = \bar{p} + \tilde{p}. \tag{6.6.39}$$

Also an instantaneous imaginary power [122] is brought in through a sum of vector productions of dissimilar voltage and current components which module also cab de divided at a constant $\bar{q}$ and variable $\tilde{q}$ components:

$$q = u_\alpha i_\beta + u_\beta i_\alpha = \bar{q} + \tilde{q}. \tag{6.6.40}$$

Two these scalar equations can be joint into one matrix equation

$$\begin{vmatrix} p \\ q \end{vmatrix} = \begin{vmatrix} u_\alpha & u_\beta \\ -u_\beta & u_\alpha \end{vmatrix} \begin{vmatrix} i_\alpha \\ i_\beta \end{vmatrix}. \tag{6.6.41}$$

Another form of this equation is possible

$$\begin{vmatrix} p \\ q \end{vmatrix} = \begin{vmatrix} i_\alpha & i_\beta \\ i_\beta & -i_\alpha \end{vmatrix} \begin{vmatrix} u_\alpha \\ u_\beta \end{vmatrix}. \qquad (6.6.42)$$

If a formula (6.6.41) is appropriate to analyze three-phase systems having voltage sources then a formula (6.6.42) is appropriate to analyze three-phase systems having current sources (current inverters, frequency converters at a current source mode).

To control by converters it's necessary to have back correlations, i.e. correlations of currents and voltages from instantaneous powers. From a formula (6.6.41) it can be obtained

$$\begin{vmatrix} i_\alpha \\ i_\beta \end{vmatrix} = \begin{vmatrix} u_\alpha & u_\beta \\ -u_\beta & u_\alpha \end{vmatrix}^{-1} \begin{vmatrix} p \\ q \end{vmatrix} = \begin{vmatrix} u_\alpha & u_\beta \\ -u_\beta & u_\alpha \end{vmatrix}^{-1} \begin{vmatrix} p \\ 0 \end{vmatrix} + \begin{vmatrix} u_\alpha & u_\beta \\ -u_\beta & u_\alpha \end{vmatrix}^{-1} \begin{vmatrix} 0 \\ q \end{vmatrix} = \begin{vmatrix} i_{\alpha p} \\ i_{\beta p} \end{vmatrix} + \begin{vmatrix} i_{\alpha q} \\ i_{\beta q} \end{vmatrix},$$

(6.6.43)

and from a formula (6.6.42) analogically

$$\begin{vmatrix} u_\alpha \\ u_\beta \end{vmatrix} = \begin{vmatrix} i_\alpha & i_\beta \\ i_\beta & -i_\alpha \end{vmatrix}^{-1} \begin{vmatrix} p \\ q \end{vmatrix} = \begin{vmatrix} i_\alpha & i_\beta \\ i_\beta & -i_\alpha \end{vmatrix}^{-1} \begin{vmatrix} p \\ 0 \end{vmatrix} + \begin{vmatrix} i_\alpha & i_\beta \\ i_\beta & -i_\alpha \end{vmatrix}^{-1} \begin{vmatrix} 0 \\ q \end{vmatrix} = \begin{vmatrix} u_{\alpha p} \\ u_{\beta p} \end{vmatrix} + \begin{vmatrix} u_{\alpha q} \\ u_{\beta q} \end{vmatrix}.$$

(6.6.44)

The last terms of an equalities at two these last formulas are a conditional decomposition of α, β-components of a two-phase system to a real and imaginary components that is appropriate for construction of active filter regulators as it will be shown at the section 13.4.3.

So, from an equation (6.6.43) follows the next decomposition of a current α-component taking into account (6.6.39) and (6.6.40):

$$i_\alpha = \frac{u_\alpha}{u_\alpha^2 + u_\beta^2}\bar{p} + \frac{u_\alpha}{u_\alpha^2 + u_\beta^2}\tilde{p} + \frac{u_\beta}{u_\alpha^2 + u_\beta^2}\bar{q} + \frac{u_\beta}{u_\alpha^2 + u_\beta^2}\tilde{q}. \qquad (6.6.45)$$

The first item here is an instantaneous value of an active fundamental current, the third item – an instantaneous value of a reactive fundamental current, a sum of the second and fourth items – an instantaneous value of current higher harmonics, subharmonics and harmonics of an inverted sequence of phases, i.e. an instantaneous value of an anomalistic current component (section 6.2.3). Such decomposition of a load current allows to selectively exclude from it undesirable inactive components by adding the same but anti-phase current components (by an active filter) to a mains (section 11.2).

Thus, a considered decomposition of an instantaneous power waveform is appropriate to use to construct regulators of separate components or all inactive component of an instantaneous power at current or voltage active filters but it's difficult to apply is difficultly applied to construct reactive power definitions since an introduced instantaneous imaginary power doesn't have an evident physical meaning.

Conversion to another coordinates. Equally with α, β-conversion a conversion to *d*, *q*-coordinates of a two-phase system is used rotating with a synchronous velocity. This conversion is appropriate because at it sinusoidal phase variables are converted into constant components at new coordinates that are proportional to an active and reactive (sinusoidal and cosinusoidal) components of an initial sinusoid. For such signals it's easier to make regulators according to methods of a conventional automatic control theory.

It's more convenient to convert to *d*, *q*-coordinates from α, β-coordinates by the next matrix conversion $\mathbf{M}_{\alpha d}$:

$$\begin{vmatrix} i_d \\ i_q \end{vmatrix} = \begin{vmatrix} \cos\omega t & \sin\omega t \\ -\sin\omega t & \cos\omega t \end{vmatrix} \begin{vmatrix} i_\alpha \\ i_\beta \end{vmatrix}. \qquad (6.6.46)$$

Applying the same conversion to voltages then we can determine *p*, *q*-components of an instantaneous power to designations identical to (6.6.41) and (6.6.42) then through them to determine *d, q*-components of voltages and currents at a form similar to (6.6.43) and (6.6.44) to a changing of α, β to *d, q* indexes, correspondingly. $\cos\omega t$ and $\sin\omega t$ reference signals have to be generated synchronously relatively a mains voltage frequency.

An inverted conversion from *d, q*-components to phase variables is realized the next way:

$$\begin{vmatrix} i_a \\ i_b \\ i_c \end{vmatrix} = \sqrt{\frac{2}{3}} \begin{vmatrix} 1 & 0 \\ -\frac{1}{2} & \frac{\sqrt{3}}{2} \\ -\frac{1}{2} & -\frac{\sqrt{3}}{2} \end{vmatrix} \begin{vmatrix} \cos\omega t & \sin\omega t \\ -\sin\omega t & \cos\omega t \end{vmatrix}^{-1} \begin{vmatrix} i_d \\ i_q \end{vmatrix}. \qquad (6.6.47)$$

If a dominating type of a low-quality (abnormality) of network energetic processes is an asymmetry then a decomposition of currents and voltages into symmetrical components is rational. The method of symmetrical components is designed conformably to sinusoidal functions that's why when there is a lot of harmonics at a nonsinusoidal function spectrum it's necessary to apply the method for an each three-phase harmonic system that requires great calculating resources. Indeed, with a widening of Digital Signal Processors (DSP) application at a control a pointed feature of symmetrical components method stops to be determining.

A conversion from sinusoidal phase variables of a three-phase system at representing them at a vector form $\dot{I}_a, \dot{I}_b, \dot{I}_c$ to vectors of symmetrical components $\dot{I}_+, \dot{I}_-, \dot{I}_0$ is realized the next way:

$$\begin{vmatrix} \dot{I}_+ \\ \dot{I}_- \\ \dot{I}_0 \end{vmatrix} = \frac{1}{3} \begin{vmatrix} 1 & a & a^2 \\ 1 & a^2 & a \\ 1 & 1 & 1 \end{vmatrix} \begin{vmatrix} \dot{I}_a \\ \dot{I}_b \\ \dot{I}_c \end{vmatrix} \tag{6.6.48}$$

$\left(a = e^{j2\pi/3} - rotation\ operator\right)$. A back conversion is the next

$$\begin{vmatrix} \dot{I}_a \\ \dot{I}_b \\ \dot{I}_c \end{vmatrix} = \begin{vmatrix} 1 & 1 & 1 \\ a^2 & a & 1 \\ a & a^2 & 1 \end{vmatrix} \begin{vmatrix} \dot{I}_+ \\ \dot{I}_- \\ \dot{I}_0 \end{vmatrix}. \tag{6.6.49}$$

In some cases the more convenient is a well-known obtaining of active and reactive components values of direct and back sequences currents through α, β-components by the next conversion

$$\begin{vmatrix} i_{+\mathrm{a}} \\ i_{+\mathrm{p}} \end{vmatrix} = \begin{vmatrix} \sin\omega t & -\cos\omega t \\ -\cos\omega t & -\sin\omega t \end{vmatrix} \begin{vmatrix} i_\alpha \\ i_\beta \end{vmatrix}, \tag{6.6.50}$$

$$\begin{vmatrix} i_{-\mathrm{a}} \\ i_{-\mathrm{p}} \end{vmatrix} = \begin{vmatrix} \sin\omega t & \cos\omega t \\ \cos\omega t & -\sin\omega t \end{vmatrix} \begin{vmatrix} i_\alpha \\ i_\beta \end{vmatrix}. \tag{6.6.51}$$

If i_α and i_β – purely harmonic functions then obtained values of active and reactive components of a direct (+) and back (–) sequences are constant values. If i_α and i_β are nonsinusoidal functions then new variables obtained after a conversion will contain constant components and pulsations. The last can be separated averaging new obtained variables in a reference signal period $(T = 2\pi / \omega)$.

Instantaneous values of active and reactive components of a direct and back sequences are obtained by a multiplication of detailed constant components at corresponding unity reference signals.

Decomposing in such a way three-phase system voltages then we can determine an instantaneous power of a three-phase system and to decompose it, in one’s part, at characteristic components conditioned by a presence of a direct, back and null sequences at currents and voltages. Because of a nonorthogonality of direct and back sequences components at an instantaneous power a component from their interaction appears. This component doesn’t have an evident physical meaning. That’s why such decomposition is especially appropriate at applying for cases with a sinusoidal current or a nonsinusoidal current at a sinusoidal voltage.

There is known a determination of components of an instantaneous power *p, q*-theory through phase variables without converting to two-phase coordinates.

In the common case for a three-phase four-wire electric system instantaneous space vectors of a voltage and current are bought in at a three-dimensional orthogonal system representing *a, b* and *c* phases:

$$\mathbf{u} = \begin{bmatrix} u_a \\ u_b \\ u_c \end{bmatrix} \qquad \mathbf{i} = \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix}. \tag{6.6.52}$$

At this space a scalar and vector productions of pointed vectors are determined:

$$p = \mathbf{u} \cdot \mathbf{i} = u_a i_a + u_b i_b + u_c i_c,$$

$$\mathbf{q} = \mathbf{u} \times \mathbf{i} = \begin{bmatrix} q_a \\ q_b \\ q_c \end{bmatrix} = \begin{bmatrix} \begin{vmatrix} u_b & u_c \\ i_b & i_c \end{vmatrix} \\ \begin{vmatrix} u_c & u_a \\ i_c & i_a \end{vmatrix} \\ \begin{vmatrix} u_a & u_b \\ i_a & i_b \end{vmatrix} \end{bmatrix}. \tag{6.6.53}$$

Here a scalar p represents an instantaneous active power, **q** vector – an instantaneous reactive (all inactive) power of a three-phase network. A value (norm) of a vector is equal:

$$q = \sqrt{q_a^2 + q_b^2 + q_c^2}\ . \tag{6.6.54}$$

Through these variables an instantaneous active current vector is determined analogically to p, q-theory but through phase variables

$$\mathbf{i}_p = \begin{bmatrix} i_{ap} \\ i_{bp} \\ i_{cp} \end{bmatrix} = \frac{p}{\mathbf{u}\cdot\mathbf{u}}\mathbf{u}\,, \tag{6.6.55}$$

an instantaneous reactive current vector

$$\mathbf{i}_q = \begin{bmatrix} i_{aq} \\ i_{bq} \\ i_{cq} \end{bmatrix} = \frac{\mathbf{q}}{\mathbf{u}\cdot\mathbf{u}}\mathbf{u}\,, \tag{6.6.56}$$

and also an instantaneous full power s and an instantaneous power factor

$$s = ui, \qquad \chi = \frac{p}{s}\,, \tag{6.6.57}$$

where $u = \|u\| = \sqrt{u_a^2 + u_b^2 + u_c^2}$ and $i = \|i\| = \sqrt{i_a^2 + i_b^2 + i_c^2}$ – instantaneous variables (norms) of three-phase voltage and current vectors.

Thus, at the considered summarizing of p,q-theory an instantaneous active power p characterizes a power transferred to a load (an active and reactive), an instantaneous reactive power q – a power interchange between phases without an active power transfer since $\mathbf{u}\cdot\mathbf{i}_q = 0$. This power compensation by an active filter is possible on the same technology that of components of an instantaneous imaginary power at p,q-theory. Leading in an instantaneous power factor value allows to estimate a compensation degree not only in a static but a dynamic mode.

C o m m o n c o n c l u s i o n on problems of a power theory at nonsunusoidal voltages and currents networks is reduced to the next.

First of all, a power theory development is, in general, on two directions: a) an integral determination of a full power and it's components necessary for calculations; b) a decomposition of an instantaneous power on components that is necessary to solve tasks of an active filtering (a compensation) of inactive full power components. Because of that a problem is multi-fold and mathematic models of nonsinusoidal energy processes are several-dimensional solutions usually are multiple-choice that provides a great material for a study, comparing alternatives and their further development.

FINAL REMARKS

1. An attempt of a complex consideration of all electromagnetic compatibility aspects was at a law about an electromagnetic compatibility that was accepted by a State Duma at the ending of 1999 but still deflected by the president of the Russian Federation.

2. A necessity of an electric energy quality certifying is determined by accepted laws.

3. A certifying sign of an equipment on an electromagnetic compatibility that have to be marked on an equipment case and proved by a corresponding certificate is determined by European Community Countries.

4. A standard at an electric energy quality of Russia (all-Union State Standard 13109-97) doesn't contain norms of a consumer current quality that a long time ago are set into International (МЭК, IEEE) and European (CENELEC) standards. But in 1999, 2000 in Russia about 40 standards on an electromagnetic compatibility are additionally accepted, 25 from them are set first from 2001, 2002. Finally, at 2002 a separate standard on current harmonic components emission of technical devices that draw a phase current no more than 16 A is set.

5. Nowadays requirements on an electromagnetic compatibility have to be included in a requirements specification to design new technical systems and contain a nomenclature of EMC parameters, their numerical values and methods of measurements.

6. Exacting requirements on a noise-emission of current harmonics drawn from a mains by one-phase technical devices to 3,5 kWA and by three-phase technical devices with a power to 10,5 kWA dictate a necessity of designing and applying of new technical solutions of valve converters with an improved electromagnetic compatibility with a mains (sections 3.11, 13.4) and a creation of a new segment of an industry of power electronic devices – a production of inactive full power components compensators (chapter 11).

7. The further development of a theory of electromagnetic compatibility various aspects begining from a power theory at nonsinusoidal processes to a theory of a noise-emission and noise-immunity of power electronics devices. Also a complex solution of a question about a terminology standardization in the field of nonsinusoidal electromagnetic processes and an electromagnetic compatibility is required.

QUESTIONS

1.1. From which components an electromagnetic compatibility at electrical engineering is summed?

1.2. Which factors of an electric energy quality are determined by a power supply side?

1.3. Which factors of an electric energy quality are determined by an energy consuming organization?

1.4. By which parameter a limit power of valve converter connected to a mains is determined?

2.1. By which influences a conductive noise-immunity of an electrical-technical equipment is determined?

2.2. By which influences an inductive noise-immunity of an electrical-technical equipment is determined?

2.3. By which kind of noise an inductive noise-emission from systems is determined?

2.4. By which kind of noise a conductive noise-emission from systems is determined?

2.5. By which way it's possible to determine a real culprit of a mains voltage waveform distortion among electric energy consumers?

2.6. By which norms electric energy quality factors at autonomous power supply systems are determined?

2.7. How to explain a presence of partial reactive powers set at nonsinusoidal energy processes?

2.8. How to explain a presence of some ways of a three-phase system full power determination?

2.9. Which definition of the full power of an asymmetrical three-phase system corresponds to a criteria of a maximal active power transfer at the same thermal loss effect?

2.10. Which definition corresponds to a system full power of a three-phase mains?

2.11*. Which parameters of some consumers does their partial back influence on a mains depend on?

2.12.What are the ways of an electromagnetic compatibility improvement of valve converters with a mains

EXERCISES

1. To an ac voltage source containing 300 V of a fundamental amplitude (50 Hz), 20 V of a fifth harmonic amplitude and 15 V of a seventh harmonic amplitude of a voltage and with it's own inductive reactance of 2 Ω a resistance R = 40 Ω, inductance 0,1 H and a one-phase bridge rectification scheme having a current I_d = 10 A and γ = 20° at a variable α are connected in parallel. Determine a distortion factor of a mains voltage sinusoidality at an angle α angle representing a rectifier as a current source of the given waveform.

2. To a mains voltage source (exercise 1) the same rectifier is connected in parallel relatively a compensating capacitor at the input. Determine local contributions of a source and consumer into a common mains voltage distortion.

3. To a mains voltage source (exercise 1) without higher harmonics two controlled one-phase bridge rectifiers are connected with the next currents I_d = 10 A and I_d = 15 A and γ = 20° at both schemes. Determine local contributions of consumers into a common mains voltage distortion.

4*. Construct a 24-phase rectifier scheme, determine it's input current waveform and a current differential harmonic factor at a commutation angle γ function, supposing that a current changing law is linear at a commutation interval.

5. Determine reactive powers of capacitors operating at the network with a nonsinusoidal voltage and a the network with a nonsinusoidal current.

6. Determine loss at an active resistance 1 Ω at a symmetrical triangle current flowing of 1 A amplitude without taking into account a current forcing effect and taking into account a current forcing effect.

7. Determine a difference degree of a determination on different formulas of a full power of a three-wire symmetrical three-phase voltage system from an asymmetry degree of an active-inductive load on phases.

8*. Determine a difference degree of a determination on different formulas of a full power of a four-wire symmetrical three-phase voltage system from an asymmetry degree of an active-inductive load.

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