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Study of the Influence of Control System Parameters on the Quality of the Input Current of a Three-Phase Power Factor Corrector

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The paper suggests a control system of a three-phase power factor corrector. The study of the control system operation is carried out and the expressions for calculating the permissible values of error amplifier factors are obtained. The influence of the error amplifier parameters on phase current quality is investigated. The dependence of total harmonic distortion input current on a combination of error amplifier parameters is obtained at a given value of power factor. The conditions under which the total harmonic distortion input current has the minimum value are found out. This article is of interest to power electronics engineers, who are aimed at developing a three-phase power factor corrector.

K e y w o r d s : *three-phase power factor corrector, control system, error amplifier parameters, total harmonic distortion*

Nowadays the criterion of phase current consumption quality is the power factor which should tend to unity and total harmonic distortion which value should tend to zero. Upon that the total harmonic distortion (THD) is determined by:

$$THD40=\left(\sqrt{\sum_{n=2}^{40}\left(\frac{I_n}{I_1}\right)^2}\right)100\%, \tag{1}$$

where I_1 is the RMS value of the first harmonic consuming phase current; I_n is the RMS value of the n-th harmonic consuming phase current; n=40 is the largest harmonic component of consuming phase current which is considered when calculating the total harmonic distortion.

In order to ensure the required quality, the function of phase current consumption correction is used in three-phase AC converters. These converters are called power factor correctors (PFC). In response to specified function the consuming phase current of PFC has the power factor maximum possible value and it is in phase to the associated phase voltage. It should be noted that the choice of error amplifier factor of the three-phase PFC control system has a considerable impact on consuming phase current quality.

Control system of three-phase PFC. Using the results of studies [1–18], we selected a three-phase PFC power circuit with low power losses on power semiconductor devices. The developed mathematical models and block scheme of the control system of the chosen power circuit in standard operating mode as well as unbalanced bipolar load were described in detail at [19–21]. The control system structure of considered three-phase PFC can be found in Fig. 1.

The represented control system of PFC works in the following way.

The u_a, u_b and u_c signals from $BU1$–$BU3$ voltage sensors, and also the i_a, i_b and i_c signals from current $BI1$–$BI3$ sensors come to the $A1$ and $A2$ blocks, which complete Park–Gorev forward conversion and generate the u_d, u_q and i_d, i_q signals. Then the above-mentioned signals follow to the $A3$ block, which calculates currently instantaneous values of p and q powers consumed from the three-phase circuit. At the same time the signals from $BU4$ and $BU5$ sensors come to the $A8$ block which calculates u_{DC} and Δu_{DC} voltages.

The $i_{\text{H}1}$ signal from $BI4$, and u_{DC}, u_d voltage signals and given voltage value come to the $A9$ block, which calculates the preset real power p^*:

$$p^*=\frac{3}{4}\frac{u_d^2}{R_L}-\sqrt{\left(\frac{3}{4}\frac{u_d^2}{R_L}\right)^2-\frac{3}{2}\frac{u_d^2u_{DC}^{*2}}{R_LR}}. \tag{2}$$

Upon that the p_e power error signal, which equals p и p^* signal difference, the q, u_d, u_q, u_{DC} and Δu_{DC} signals

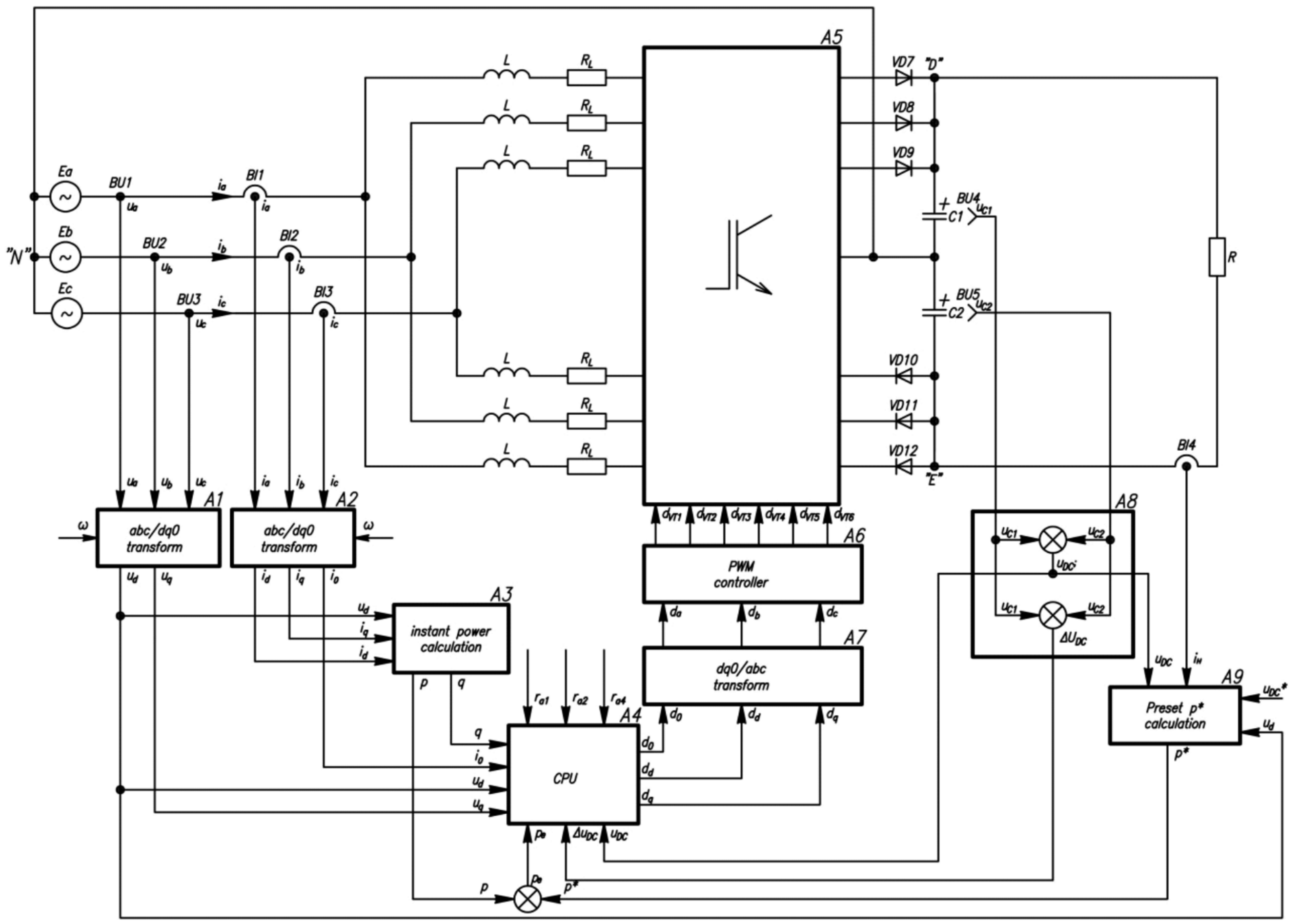

Fig. 1. The control system structure of considered three-phase PFC

and r_{a1}, r_{a2} and r_{a3} coefficients come to the $A4$ block calculating d_d, d_q, d_0 switching functions.

Then the d_d, d_q, d_0 signals come to the $A7$ block completing the reverse vector Park–Gorev conversion, and calculates the d_a, d_b and d_c signals. The obtained d_a, d_b and d_c signals come to the $A6$ block which generates the PWM control signals for the power transistors of the $A5$ block of the PFC.

As a result, the considered control system receiving the signals from the BU–$BU5$ voltage sensors, the $BI1$–$BI4$ current sensors and the given u^*_{DC} voltage value and circular frequency calculates the p real power, the preset p^* real power, and p_e power error signal. Then using the r_{a1}, r_{a2} and r_{a3} coefficients, it calculates the d_d, d_q, d_0 switching functions, and after the backward vector Park–Gorev conversion generates a switching of power transistors of the PFC where the p_e power error signal tends to zero.

The study of error amplifier factors. In the article [21] we got some equations for the calculation of switching functions in $dq0$-coordinates:

$$\begin{cases} \bar{d}_d = \dfrac{2}{u_d u_{DC}}\left(u_d^2 + r_{a1}(p-p^*) - \dfrac{2}{3}L\omega q\right); \\ \bar{d}_q = \dfrac{2}{u_d u_{DC}}\left(u_d u_q + r_{a2} q - \dfrac{2}{3}L\omega p\right); \\ \bar{d}_0 = \dfrac{2}{u_d u_{DC}}(u_d u_0 + r_{a3} z). \end{cases} \tag{3}$$

We are proposed to present the aforementioned switching functions using two components:

$$\begin{cases} \bar{d}_d = f_{d.tr} + f_{d.st}; \\ \bar{d}_q = f_{q.tr} + f_{q.st}; \\ \bar{d}_0 = f_{0tr} + f_{0st}, \end{cases} \tag{4}$$

where $f_{d.tr}$, $f_{q.tr}$, f_{0tr} are alternating components of switching functions; $f_{d.st}$, $f_{q.st}$, f_{0st} are steady-state components of $\bar{d}_d$, $\bar{d}_q$, $\bar{d}_0$ functions.

Taking into account (4), we obtained the following formulas from (3):

$$f_{d.st} = \frac{2}{u_d u_{DC}}\left(u_d^2 - \frac{2}{3}R_L p^*) - \frac{2}{3}L\omega q\right);$$

$$f_{q.st} = \frac{2}{u_d u_{DC}}\left(u_d u_q + \frac{2}{3}L\omega p\right);$$

$$f_{0st} = \frac{2u_0}{u_{DC}};$$

$$f_{d.tr} = \frac{2}{u_d u_{DC}} r_{a1}(p - p^*);$$

$$f_{q.tr} = \frac{2}{u_d u_{DC}} r_{a2} q;$$

$$f_{0tr} = \frac{2}{u_d u_{DC}} r_{a3} z.$$

The values of steady-state components of switching functions can be calculated according to the suggested steady-state mode of the converter. Upon that the ripples of $f_{d.st}$ steady-state component can be neglected.

The analysis shows that the $f_{d.tr}$ ripple value depends directly on r_{a1} error amplifier factor value. For the operation of PWM *A*6 block it is necessary for the $\bar{d}_d$ switching function values to remain within the interval [0;1], and also for the geometrical mean value of the $\bar{d}_d$ and $\bar{d}_q$ steady-state components to be less than unity. From the first assumption we inferred the existence of r_{a1} maximum value:

$$f_{d.tr} \leqslant 1 - f_{d.st}. \tag{5}$$

It should be noted that $f_{d.tr}$ value contains ripples caused by the ripples of the consumed phase currents of the PFC. Upon that the mentioned switching function component has deviations during the transients. For example, when the load resistance changes, a small voltage jump occurs, that for some period of time tends to the u^*_{DC} present value. Upon that the duration of transient depends on r_{a1} value. As the r_{a1} coefficient increases, the duration of the transition decreases. Therefore, the coefficient r_{a1} must have the maximum possible value corresponding to (5).

Taking into account that the i_d is vector projection of the input phase currents and is approximately equal to the amplitude of phase current, we obtained the following expression for calculating the range of the real power ripples:

$$\Delta p = p - p^* = \frac{3}{4}u_d \Delta i_d = \frac{3}{8}\frac{u_d u_{DC}}{Lf_{VT}}. \tag{6}$$

As a result, from (5) and with regard to (6) we got the range of the r_{a1} error amplifier factor:

$$0 < r_{a1} \leqslant \frac{4}{3}Lf_{VT}(1 - f_{d.st}). \tag{7}$$

Upon that we determined the maximum r_{a1} coefficient value, where *A*6 PWM block ensures operation taking into account (7):

$$r_{a1\max} = \frac{4}{3}Lf_{VT}(1 - f_{d.st}). \tag{8}$$

From the second condition of the *A*6 block correct operation and taking into account the symmetry of the input phase voltages, the range of the r_{a2} coefficient was estimated. In order for the amplitude of the switching functions in *abc*-coordinates to be less than unity, the following condition must be completed:

$$\sqrt{\bar{d}_d^2 + \bar{d}_q^2 + \bar{d}_0^2} = 1. \tag{9}$$

From (9) and if the r_{a1} is fixed, we defined the relation between the r_{a2} and r_{a3} coefficients. Disregarding the $\bar{d}_d$ switching function ripple and taking into account (4) we got the equation of the second degree:

$$f_{q.tr}^2 + 2f_{q.st}f_{q.tr} + f_{q.sr}^2 + \bar{d}_{d\max}^2 + \bar{d}_0^2 - 1 = 0.$$

As for $f_{q.tr}$, we obtained the following solutions:

$$f_{q.tr_{1,2}} = -f_{q.st} \pm \sqrt{1 - \bar{d}_{d\max}^2 - \bar{d}_0^2}. \tag{10}$$

Please note, that the sign «–» in (10) reliably leads to the negative value of r_{a2} coefficient, that is contrary to its definition range. Substituting in (10) the equations (4), (9) and taking into account that if the symmetrical input phase voltage $u_0 = 0$, we obtained the equation which relates to both the r_{a2} and r_{a3} coefficients:

$$r_{a2} = -f_{q.st} + \frac{u_d u_{DC}}{2q}\sqrt{1 - \bar{d}_{d\max}^2 - \left(\frac{2}{u_d u_{DC}} r_{a3} z\right)^2}.$$

Taking into account the condition that the r_{a2} coefficient is real, from (10) we obtained the $r_{a3\max}$ maximum error amplifier factor value:

$$r_{a3\max} = \frac{u_d u_{DC}}{2z}\sqrt{1 - \bar{d}_{d\max}^2}. \tag{11}$$

Similarly, from the equation (11) we got the equation to calculate the $r_{a2\max}$ maximum error amplifier factor value:

$$r_{a2\max} = \frac{u_d u_{DC}}{2q}\sqrt{1 - \bar{d}_{d\max}^2} - f_{q.st}. \tag{12}$$

It should be noted that input phase currents are summarized in neutral conductor according to the Kirchhoff's first law. Upon that the sum of the basic

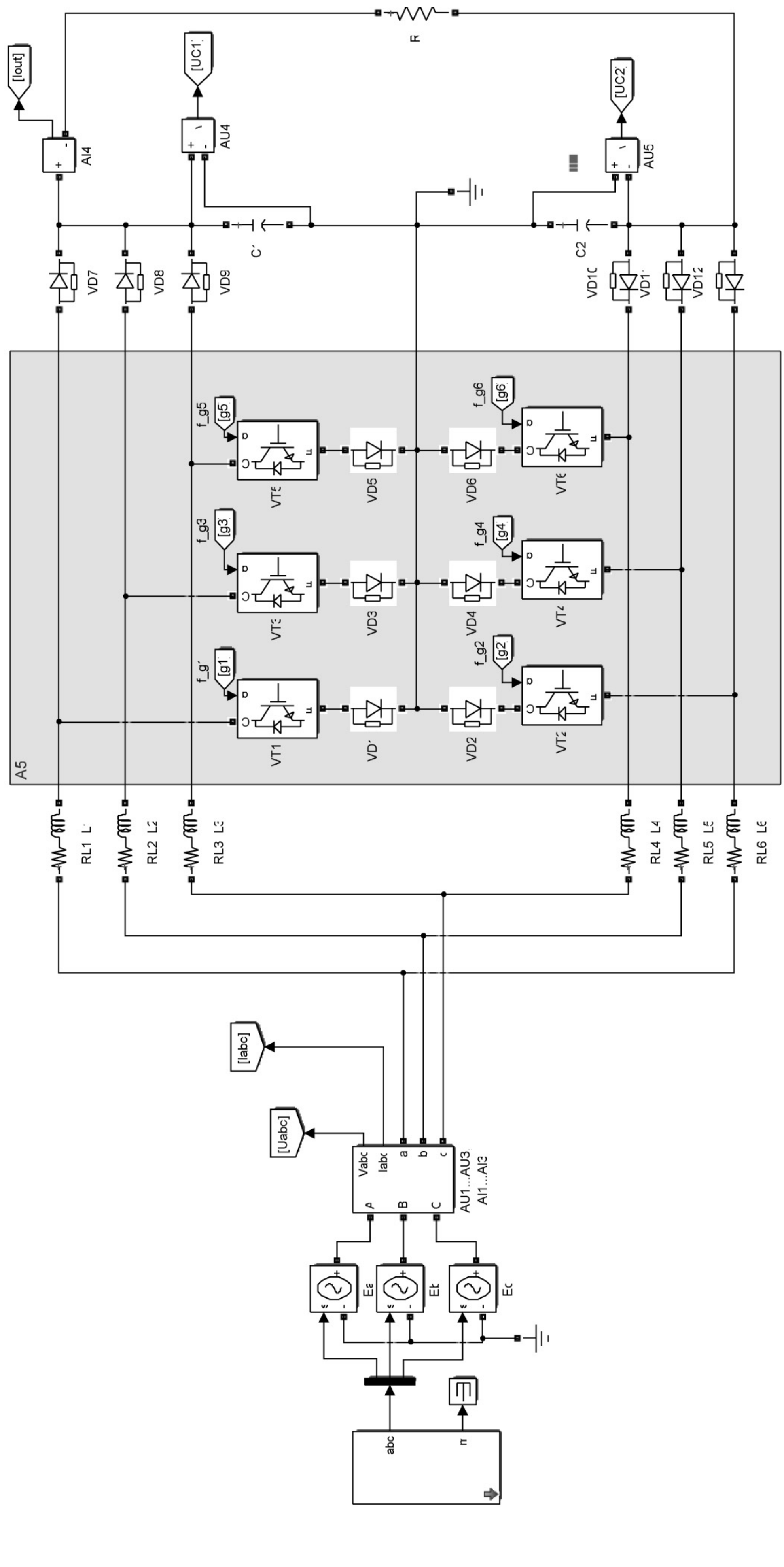

Fig. 2. The designed simulation computer model: E_a, E_b and E_c are EMF of the input three-phase power supply; *L*1–*L6* are input reactors; *VD*1–*VD*12 are power diodes; *VT*1–*VT*6 are power transistors; *C*1 and *C*2 are output capacitors; *R* is PFC load

harmonic components of the input phase currents in symmetrical mode equals zero. Consequently, the phase currents ripples generate current in the neutral conductor. So, the value of the power z due to the current in the neutral conductor is defined as:

$$z = \frac{3}{32}\frac{u_d u_{DC}}{Lf_{VT}}. \qquad (13)$$

The q reactive power value is obtained as:

$$q = \frac{1}{PF}\sqrt{p^2 - PF^2(p^2 - z^2)}. \qquad (14)$$

With fixed r_{a1} coefficient and using the equations (11)–(14) it is possible to create a great number of (r_{a2}, r_{a3}) couple coefficients.

The analysis of the $f_{d.st}$, $f_{q.st}$ and f_{0st} components shows that r_{a2} and r_{a3} coefficients can vary within wider limits than r_{a1} without affecting the *A*6 block operation of the PWM.

It should be noted that reduction of the reactive component that is consumed from the power network occurs with the increase in r_{a2} coefficient, but at the same time the total harmonic distortion input current is rising. The analysis shows that if r_{a3} coefficient is fixed, there is a r_{a2} value when THD40 has the minimum value.

On the other hand, when the r_{a3} increases, the power component decreases due to the current in the i_0 neutral conductor while the reactive component of the input power increases. Upon that if a substantial decrease of the load resistance happens, the longstanding transient occurs, during which the difference of average voltages on the output capacitors tends to take a minimum value for a long time. Moreover, the duration of the described transient has the minimum value at a r_{a3} coefficient equal to zero.

Based on the above it can be concluded that the determination of the r_{a1}, r_{a2} and r_{a3} error amplifier factors is a challenging task. Upon that the different combinations of r_{a1}, r_{a2} and r_{a3} combinations severely affect the PFC operating mode and the input phase current quality. Considering that it is difficult to obtain an analytical solution, it is reasonable to study the r_{a1}, r_{a2} and r_{a3} coefficients influence on the electric process using the simulation computer model (SCM) of the considered PFC basing on required quality criteria of the input phase current.

The Fig. 2 shows the designed SCM of the control system of the three-phase PFC.

The results of simulation computer. Using designed SCM at specified PF coefficient, the influence of r_{a1}, r_{a2} and r_{a3} coefficients on input phase current quality was studied. We calculated the r_{a3} coefficient when the r_{a1} coefficient changes discretely from zero to the maximum value ($r_{a1\max}$) calculated from (8), and the r_{a2} coefficient changes discretely from zero to the maximum value ($r_{a2\max}$) calculated from (12). Then we obtained the THD 40 and PF values for the found r_{a3}, using the developed SCM. The r_{a3} coefficient functions vs r_{a2} value at different r_{a1} coefficient quantity are shown on Fig.3.

The results of r_{a2} and r_{a3} are shown in the Table at a given r_{a1} value when $THD40_{\min} = f(r_{a1}, r_{a2}, r_{a3})$ takes the minimum value. The study of the computer modeling results represented in Table allows to come to the conclusion that there is a unique r_{a1} coefficient value where *THD*40 has the minimum quantity, and it is close to the maximum accepted value calculated with the use of (8). Accordingly, it is reasonable to make a search of the r_{a2} и r_{a3} coefficients (where *THD*40 will have the minimum value) in proximity of the r_{a1} maximum accepted coefficient.

r_{a1}	r_{a2}	r_{a3}	$THD40_{\min} = f(r_{a1}, r_{a2}, r_{a3})$	PF
$0,100^{*}_{ra1\max} = 0,10$	10,0	3,60	2,56	0,9994
$0,30^{*}_{ra1\max} = 0,30$	8,93	3,21	1,30	0,9994
$0,50^{*}_{ra1\max} = 0,50$	7,62	2,74	0,81	0,9994
$0,70^{*}_{ra1\max} = 0,69$	5,96	2,14	0,59	0,9993
$0,90^{*}_{ra1\max} = 0,89$	2,95	2,40	0,47	0,9992
$0,95^{*}_{ra1\max} = 0,93$	1,72	2,16	0,51	0,9988
$0,99^{*}_{ra1\max} = 0,98$	2,15	1,60	1,20	0,9961

We assumed that the surface area $THD40 = f(r_{a2}, r_{a3})$ has only one r_{a2} and r_{a3} value pair where the THD40 derivation will be equal to zero. If this assumption is correct, then such r_{a2} and r_{a3} value pair will be easy to obtain using well-known mathematical optimization methods.

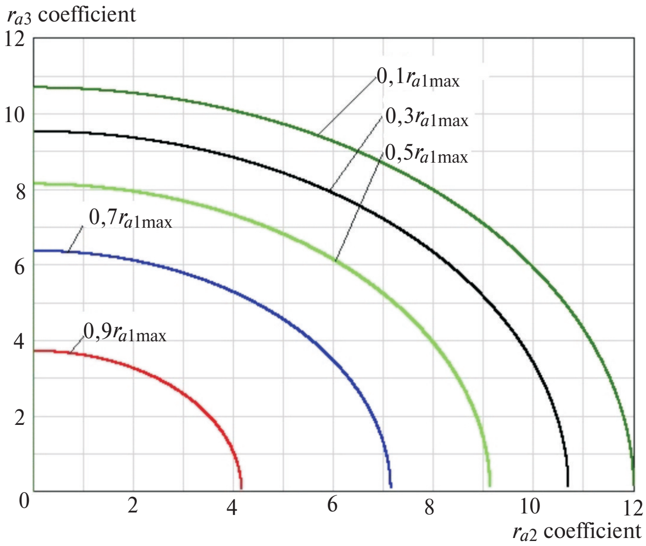

Fig. 3. The functions r_{a3} of error amplifier factor vs r_{a2} value at different r_{a1} coefficient value of control system

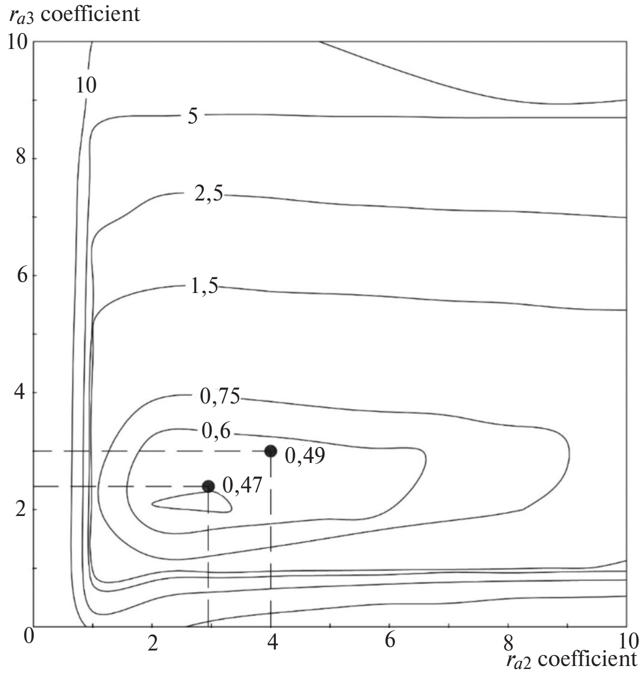

Fig. 4. The functions ra3 of error amplifier factors vs *THD*40 total harmonic distortion input current

To confirm the mentioned assumption, we carried out analysis of surface area $THD40 = f(r_{a2}, r_{a3})$ with the use of brute-force search method (exhaustive search) at $r_{a1} = 0{,}9 r_{a1\,max}\, r_{a2} = 0...10{,}0$, and $r_{a3} = 0...10{,}0$.

Using computer simulations, we obtained the *THD*40 lines (Fig. 4) which correspond to the 0.4; 0.6; 0.75; 1.5; 2.5; 5.0 and 10.0%. From Fig. 4 it is obvious that the given dependences have only one local (global) the THD40 minimum. Upon that the difference between obtained THD40 minimum values in Table and in Fig. 3 is due to the assumptions at the conclusion (14) and (13). Generally, in absolute magnitude the given difference has a small value and it is accepted in engineering calculations.

The obtained results confirm the fact that there is only one local (global) minimum of THD40 for a given PF. This factor at the given *PF* simplifies the possibilities of forming an algorithm for searching triple r_{a1}, r_{a2} and r_{a3} coefficients where *THD*40 takes the minimum value.

Conclusion. 1. The determination of the r_{a1}, r_{a2} and r_{a3} error amplifier factors of control system is a challenging task. Upon that the different combinations of r_{a1}, r_{a2} and r_{a3} severely affect the three-phase PFC operating mode and the input phase current quality.

2. It is proved that there is only one r_{a1} coefficient value where *THD*40 has its minimum, and it is close to the maximum accepted value.

3. It is acknowledged that at a given PF there is only one local (global) *THD*40 minimum. This factor at a given *PF* value simplifies the possibilities of forming an algorithm for searching triple r_{a1}, r_{a2} and r_{a3} coefficients where THD40 takes the minimum value.

The results received could be of interest for those developers of three-phase PFCs who are challenging to design converters with high powers factor.

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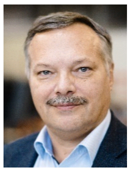

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Исследование влияния параметров системы управления на качество входного тока трехфазного корректора мощности

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Предложена система управления трехфазным корректором коэффициента мощности. Проведено исследование рассматриваемой системы управления и получены выражения для расчета допустимых значений коэффициентов усилителей ошибок. Исследовано влияние коэффициентов усилителей ошибок системы управления на качество потребляемого фазного тока устройства. При заданном значении коэффициента мощности получены зависимости коэффициента гармонических составляющих входного фазного тока трехфазного корректора мощности от сочетания коэффициентов усилителей ошибок. Показаны условия, при которых коэффициент гармонических составляющих входного фазного тока принимает минимальное значение. Статья представляет интерес для инженеров силовой электроники, занимающихся вопросами разработки трехфазного корректора коэффициента мощности и проектированием устройств, требующих повышенного качества электрической энергии.

Ключевые слова: *трехфазный корректор мощности, система управления, коэффициент усилителей ошибки, коэффициент гармонических составляющих*

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