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Combined use of the nearest vector selecting technique and low-frequency quarter-wave symmetric space vector PWM in control of three-phase multilevel voltage source inverter

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Abstract. The paper offers the problem's solution of the three-phase multilevel voltage source inverter (MLVSI) control with low dynamic losses in power semiconductor switches. The proposed technique is based on a single approach that operates with values of the integer and fractional parts of the relative values of the reference delta voltages, and it allows combining the use of the space-vector control, choosing the nearest vector (SVC), with the low-frequency space-vector pulse-width modulation with the quarter-wave symmetry of the output voltages (QWS-SVPWM). Depending on the impedance kind of a generalized load, the integrated voltage harmonics factor (weighted THD) of the appropriate order is being minimized by applying one of the two mentioned modulation methods (SVC or QWS-SVPWM) for a current range of voltage amplitude modulation index. Like its components, the combined algorithm is suitable for MLVSI of any circuit configuration and any number of equal output voltage levels.

1. Introduction

In addition to existing for several decades circuitry solutions of multilevel converters [1], this concept brings new and yet more effective results, leading, as expected, to the wide industrial application [2], including applications in photovoltaics power conversions with their grid integration [3]. Modified control techniques are constantly being developed to improve performances of both existing and new converters, and in particular inverters [4].

With an increase in the number of levels of multilevel voltage source inverters (MLVSI), that is precipitated by the necessary increase in their output voltages and powers, the high quality of the voltages and currents is achievable without increasing the frequency of power switches' switching. The possible solutions are both the space vector control with the choice of the nearest vector (SVC) [5], and the low-frequency space vector pulse-width modulation (SVPWM).

The combined use of two alternatives, SVC and SVPWM, is possible within the framework of a single approach, operating with the values of the integer and fractional parts of the samples of the relative values of the reference delta voltages and realizing the two-component formation of the two output delta voltages. This is exactly how the SVPWM with quarter-wave output voltage symmetry (QWS-SVPWM) [6] and the new algorithm for realizing the choice of the closest vector without storing data in lookup tables [7] were implemented separately.

This paper describes both the main principles of the combined use of the techniques and the approach to the specified MLVSI output variable THD minimizing by means of consideration of the integrated voltage harmonics factor [8] (weighted THD) of the appropriate order, depending on the



impedance kind of a generalized load. The ranges of using SVC and QWS-SVPWM modulation methods are revealed for the cases of the generalized load equivalent circuit, having numbers of the orders from zero to three (by the order of a corresponding differential equation and the relevant to it weighted THD).

2. Space vector's oblique-angled coordinates and barycentric coordinates on triangle of nearest vectors

Despite the space vector PWM algorithms, using the orthogonal decomposition has been well-established ones [9], the techniques operating in the context of oblique-angled coordinates' systems have proved to be a more simple and fast [10].

The so-called g- and h-coordinates in [10] have turned out to be the coordinates of the delta voltages relative values u_{ab}^* and u_{bc}^* . Here and after

$$u_{xy}^* = u_{xy} / U_d, \quad (1)$$

where U_d is the direct voltage of the unit (base) level, i.e. the delta voltages' step value.

So, one of the available and convenient representation for a complex-plane form $\dot{u}$ of some voltage space vector (VSV) is [11]

$$\dot{u} = \frac{2}{3} U_d \cdot \left(u_{ab}^* + u_{bc}^* \cdot e^{j\frac{\pi}{3}} \right). \quad (2)$$

MLVSI voltage space vectors are shown in figure 1. Figure 1,a demonstrates the vectors' diagram in the oblique-angled 60-degree coordinates with the space vectors designated as (u_{ab}^*, u_{bc}^*) .

If some VSV is hitting a point R inside a certain triangle of the nearest vectors, its decomposition can be presented in accordance with figure 1,b. Thanks to a property of barycentric coordinates on the upward-pointing and downward-pointing elementary triangles, the VSV $\vec{OR}$ is described, respectively, as follows [11]:

$$O_{\Delta} \vec{R} = b_{ab} \cdot O_{\Delta} \vec{D} + b_{bc} \cdot O_{\Delta} \vec{E} + b_{ca} \cdot O_{\Delta} \vec{F} \quad \text{and} \quad \dot{u} = \vec{OR} = b_{ab} \cdot \vec{OD} + b_{bc} \cdot \vec{OE} + b_{ca} \cdot \vec{OF}, \quad (3)$$

$$O_{\nabla} \vec{R} = b_{ba} \cdot O_{\nabla} \vec{E} + b_{cb} \cdot O_{\nabla} \vec{D} + b_{ac} \cdot O_{\nabla} \vec{G} \quad \text{and} \quad \dot{u} = \vec{OR} = b_{ba} \cdot \vec{OE} + b_{cb} \cdot \vec{OD} + b_{ac} \cdot \vec{OG}, \quad (4)$$

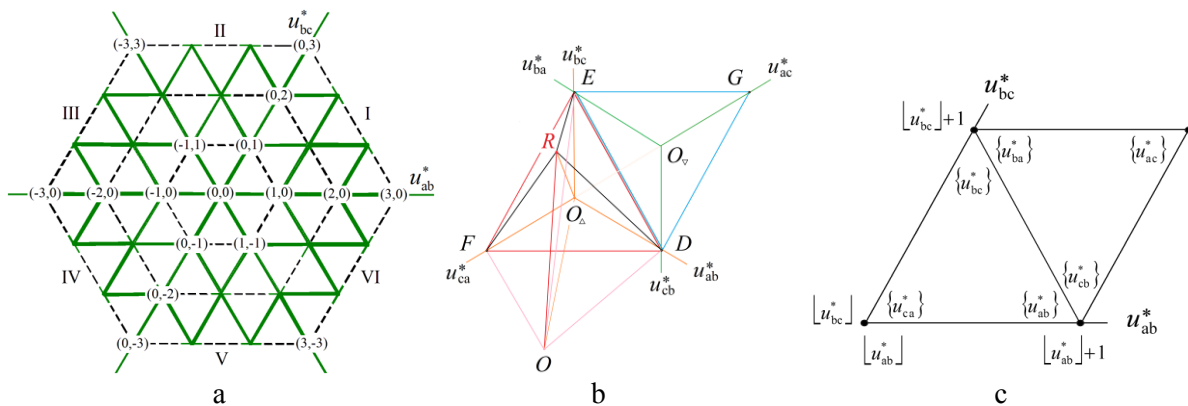


Figure 1. Voltage space vectors of three-phase multilevel voltage source inverters: a – vectors' diagram in 60-degree coordinates; b – elementary triangles of nearest vectors with six delta voltages' axes; c – parts of delta voltages' relative values as coordinates of specified voltage space vector.

where point O is the zero vector $(0,0)$ position, points O_Δ and O_∇ are the barycentres of triangles ΔDEF and ∇EDG , b_{xy} and b_{yx} are VSV barycentric coordinates on these two triangles, respectively,

$$b_{xy} = u_{xy}^* - \lfloor u_{xy}^* \rfloor = \{u_{xy}^*\}, \quad b_{yx} = u_{yx}^* - \lfloor u_{yx}^* \rfloor = \{u_{yx}^*\}, \quad (5)$$

here $\lfloor x \rfloor$ and $\{x\}$ are the integer and fractional parts of x ,

$$b_{ab} + b_{bc} + b_{ca} = 1, \quad b_{ba} + b_{cb} + b_{ac} = 1. \quad (6)$$

Thus, as can be seen in figure 1,c, the integer parts of the two delta voltages' relative values are the oblique-angled coordinates of the base vector of the three vectors which are the nearest vectors to the specified VSV, and the fractional parts of the three delta voltages' relative values are the barycentric coordinates of the specified VSV on these vectors' triangle. The three direct delta voltages should be used if the VSV's endpoint is in some upward-pointing elementary triangle, otherwise, the three inverse delta voltages should be considered (see figure 1,b).

The same quantities serve as affine coordinates on the upward-pointing and downward-pointing elementary triangles in following VSV decompositions, respectively:

$$\dot{u} = \dot{OR} = \dot{OF} + \dot{FR} = \dot{OF} + b_{ab} \cdot \dot{FD} + b_{bc} \cdot \dot{FE}, \quad (7)$$

$$\dot{OF} = \frac{2}{3}U_d \cdot \left(\lfloor u_{ab}^* \rfloor + \lfloor u_{bc}^* \rfloor \cdot e^{j\frac{\pi}{3}} \right), \quad \dot{FR} = \frac{2}{3}U_d \cdot \left(\{u_{ab}^*\} + \{u_{bc}^*\} \cdot e^{j\frac{\pi}{3}} \right); \quad (8)$$

$$\dot{u} = \dot{OR} = \dot{OG} + \dot{GR} = \dot{OG} + b_{ba} \cdot \dot{GE} + b_{cb} \cdot \dot{GD}, \quad (9)$$

$$\dot{OG} = \frac{2}{3}U_d \cdot \left(-\lfloor u_{ba}^* \rfloor + \lfloor u_{cb}^* \rfloor \cdot e^{-j\frac{2\pi}{3}} \right), \quad \dot{GR} = \frac{2}{3}U_d \cdot \left(-\{u_{ba}^*\} + \{u_{cb}^*\} \cdot e^{-j\frac{2\pi}{3}} \right). \quad (10)$$

3. Space vector pulse-width modulation with quarter-wave output voltage symmetry, new nearest vector selecting technique in space vector control and two-component formation of two output delta voltages

The executed output voltages here mean the voltages with waveforms, that are generated at MLVSI output by switching available space vectors as vertices of vectors' diagram in figure 1,a. They also can be treated as the coordinates of the MLVSI executed voltage space vector (EVSU), the latter is understood in the same sense.

If the reference voltage space vector (RVSU) of SVPWM is sampled inside a certain elementary triangle of the nearest vectors, called as modulating triangle, its sampled coordinates are being reproduced at the related PWM clock period (sampling period) in EVSU as average values of the output coordinates. Using the delta voltages as the reference voltages is especially convenient for SVPWM, since the duty-cycles of the three modulating vectors are exactly equal to the values of the above mentioned barycentric coordinates of the RVSU, i.e. the fractional parts of the relative values of the MLVSI reference delta voltages [11].

Thanks to (6) and other simple relationships for the rhombus in figure 1,c, transition from three to two VSV barycentric coordinates is easily accomplished. So, all the calculations for both kinds of the elementary triangles of the rhombus have been reduced to treating the only two direct delta voltages relative values instead of the values of six related delta voltages. In particular, direct using descriptions of the downward-pointing triangle, which are like (10), is no longer needed.

SVPWM with quarter-wave output voltage symmetry provides the absence of the even-order harmonics and the absence of the cosine component of the remaining harmonics in Fourier decomposition of the output voltage' waveform. As a result, we have reached considerable progress in reducing low-order higher harmonics contents. QWS-SVPWM and this supplementary advantage have

been achieved due to applying a new switching sequence algorithm with a three-segment switching sequence scheme [6].

The used here scheme of the nearest vector selecting space-vector control SVC [7] applies no special operations and generates the same output voltage waveforms as the traditional nearest vector selecting space-vector control [5]. But it utilizes the conventional delta voltages SVPWM attributes, namely the duty-cycles of the three vectors of the modulating triangle (i.e. of the three vectors, which are nearest to the reference voltage space vector RVSV). As have been shown above, these values are also the barycentric coordinates on the corresponding modulating triangle for the reference voltage space vector RVSV endpoint. So, the closer to unity the fractional value of certain relative delta voltage is, the closer to the related vertex the RVSV endpoint is. Moreover, the highest value of the three barycentric coordinates on the modulating triangle makes it possible to recognize the appropriate vertex as the closest to the reference point, and such a way one can choose the space vector nearest to the vector RVSV endpoint on each sampling period [7].

This approach to space vector control, which is choosing the nearest vector, needs no preliminary finding of anything coefficients and holding them in look-up tables. Similar to the presented SVPWM technique, it is suitable for any arbitrary MLVSI circuit with any arbitrary number of the equal feeding DC voltage levels. All the related logical functions and Matlab-Simulink implementation of the controller model have been offered in [7].

The two-component formation of two independent output delta voltages is another one base of the general technique for combining SVC and QWS-SVPWM modulation methods. Specifically, in accordance with the main principle of separated treating the integer and fractional parts of the samples of the relative values of the reference delta voltages, each of the two executed output delta voltages is generated as follows:

$$u_{EXEXy}^*(t) = \lfloor u_{sREFxy}^*(t) \rfloor + f_{sxy}(t), \quad (11)$$

where $u_{sREFxy}^*(t)$ is a respective relative delta voltage reference signal, which is sampled and held in each sampling period with a duration of T_s , and being a function of continuous model time, namely, we have two such reference signals, that are shifted to be corresponding to the middle point of the sampling period,

$$u_{sREFab}^*(t) = m_{aD} \sin((2\pi k(t) - \pi)/m_{sf} + 2\pi/3), \quad u_{sREFbc}^*(t) = m_{aD} \sin((2\pi k(t) - \pi)/m_{sf}), \quad (12)$$

m_{aD} is the delta voltage amplitude modulation index,

$$m_{aD} = \sqrt{3} m_{aY} = \sqrt{3} U/U_d = \sqrt{3} U^* = U_{\Delta m}^*, \quad m_{aY} = U/U_d = U^*, \quad (13)$$

m_{aY} is the phase voltage (wye voltage) amplitude modulation index, U and U^* are the value and relative value of RVSV magnitude, $U_{\Delta m}^*$ is the amplitude relative value of the reference delta voltages, $k(t)$ is the clock-cycle number function and it runs the integer values from 1 to m_{sf} within each MLVSI output voltage's period with a duration of T ,

$$k(t) = \lfloor (t/T - \lfloor t/T \rfloor) \cdot m_{sf} \rfloor + 1, \quad (14)$$

m_{sf} is the sampling frequency index, it is similar to the frequency modulation index in PWM,

$$m_{sf} = f_s/f = T/T_s, \quad (15)$$

here f_s and f are, respectively, the sampling frequency (sampling rate) and modulating (output) frequency;

$f_{sxy}(t)$ is a pulse function, possessing the values 0 and 1, and exactly a difference in formation of this function in SVC and QWS-SVPWM defines an overall difference in generated output voltages.

While in QWS-SVPWM technique a sector's number, a type of a modulating vectors' triangle and values of the barycentric coordinates of the RVSV specifies positions and durations of the high-potential pulses on each clock cycle period [6], SVC sets the only level per the sampling period in accordance with a maximum barycentric coordinate on a related triangle [7]. In contrast to PWM, SVC at some enough sampling frequency value generates the same voltages' waveforms as for "instantaneous SVC" with a vanishingly small sampling period, and the switching frequency of power switches in MLVSI is therefore absolutely independent of the sampling frequency value.

Thus, to have in the offered generalized approach a low level of the switching losses in semiconductor switching devices (IGBT, MOSFET, etc), QWS-SVPWM should be applied at enough low values of the sampling frequency and switches' commutation frequency, while the SVC, as a priori a low-frequency method with regard to MLVSI power circuits, should be used with enough high sampling frequency value.

4. Integrated voltage harmonics factors of inverter voltage waveform as indicators of power quality of a generalized load

The simplified equivalent circuits of the MLVSI for the cases of four various kinds of its load are shown in figure 2, where $u_{ag}(t)$ and $u_{an}(t)$ are the phase-to-ground ("g") and phase-to-neutral ("n") generated voltages of the phase "a", respectively; z_a is the phase "a" impedance, L_a and R_a are the prospective phase "a" load inductance and resistance, L_f and C_f are the elements of phase "a" MLVSI output LC filter; i_a is the phase "a" load current.

The MLVSI load current's THD, designated here as the factor of current's harmonics K_{hia} , can be calculated as follows:

$$K_{hia} = I_{a(hh)} / I_{a(1)}, \quad (16)$$

where $I_{a(1)}$ and $I_{a(hh)}$ are the RMS values of the current's fundamental harmonic and higher harmonics component, respectively.

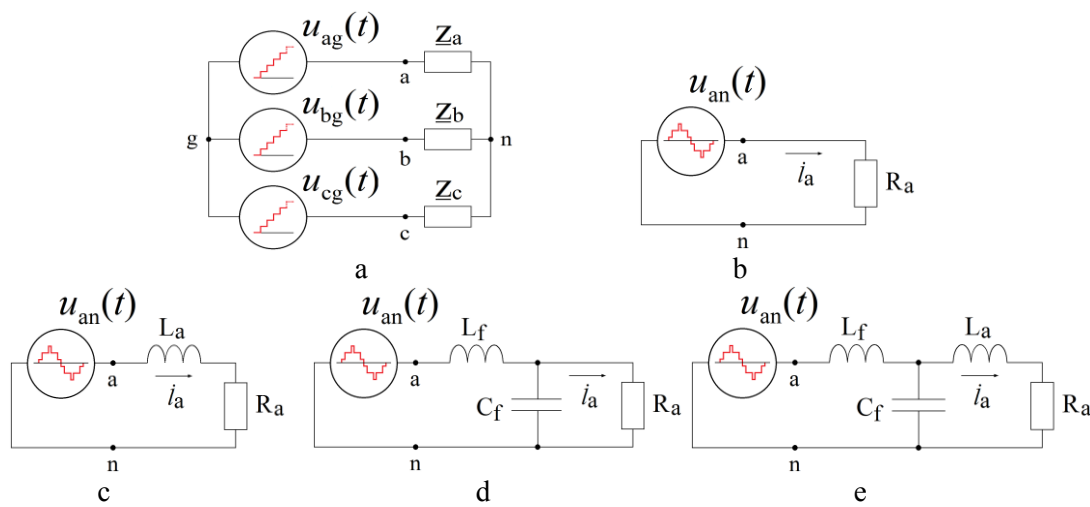


Figure 2. Simplified equivalent circuits of three-phase multilevel voltage source inverter and its load: a – generalized three-phase three-wire circuit; b – per-phase R load circuit; c – per-phase RL load circuit; d – per-phase R load circuit with inverter output LC filter; e – per-phase RL load circuit with inverter output LC filter.

Instead of considering the full three-phase equivalent circuit (figure 2,a), for some three-phase three-wire circuit with symmetrically-controlled MLVSI and some balanced load one of the simplified per-phase equivalent circuits may be studied.

The trivial case is presented in figure 2, b for cases when a generalized load, including some output circuit of the MLVSI itself, can be treated as the active resistance. The respective equation is a degenerate one and lead to THD simple and obvious relation:

$$R_a i_a = u_{an}, K_{hia} = K_{hu} = \overline{K}_{hu}^{(0)}, \quad (17)$$

where K_{hu} is THD of MLVSI phase-to-neutral voltage u_{an} , also treated as the zero-order integrated voltage harmonics factor $\overline{K}_{hu}^{(0)}$ (here and after index “a” is omitted from subscript of such the factors).

The simplest of the more realistic MLVSI phase load circuits in figure 2,c is described by the first order differential equation and reduced integral equation derived from it:

$$L_a \frac{di_a}{dt} + R_a i_a = u_{an}, i_a + \frac{R_a}{L_a} \overline{i}_a^{(1)} = \frac{1}{L_a} \overline{u}_{an}^{(1)}, \quad (18)$$

here and after $\overline{x}^{(n)}$ designates the result of the taking of n-fold indefinite integral of some instantaneous AC variable x (can have any subscript).

To assess the load current's higher harmonics component and THD the method, developed by Prof. G.S. Zinoviev (NSTU, Novosibirsk), namely the differential equations' algebraization (ADE) method [8] is used here. The ADE method is intended for direct obtaining expressions of RMS voltages and currents values in a closed analytical form from source differential equations without their solving, and ADE is asymptotically approximate method.

After the ADE procedure applying to (18), written for the higher harmonics component $i_{a(hh)}$, the phase “a” load current i_a THD value can be defined as asymptotic quantity, being approximately calculated as the sum of the finite series with alternating members' signs:

$$K_{hia} \approx \left[\left(\left(R_a / L_a \right)^2 + \omega^2 \right) \cdot \sum_{n=1}^{N_f} \left((-1)^{n-1} \cdot \left(\frac{(R_a / L_a)^{n-1}}{\omega^n} \cdot \overline{K}_{hu}^{(n)} \right)^2 \right) \right]^{1/2}, \quad (19)$$

where N_f is the approximation level, related to the “filter hypothesis” [8], [12], namely $\overline{i}_{a(hh)}^{(n \geq N_f)} \equiv 0$, $\overline{K}_{hu}^{(n)}$ is the n-order weighted THD (WTHD) coefficient of the MLVSI output voltage $u_{an}(t)$, which ADE method's author called n-order integrated voltage harmonics factor (IHF).

So, the first approximation $N_f = 1$ corresponds to

$$K_{hia} \approx \overline{K}_{hu}^{(1)} \cdot \left[1 + \left(R_a / (\omega L_a) \right)^2 \right]^{1/2}. \quad (20)$$

Since the load current's THD, with only one exception (see (17)), does not correlate directly with the output voltage's THD, the latter cannot be a criterion for completely sufficient quality comparative evaluation of some voltage source output voltage. It shows only a quantity (“weight”) of higher harmonics in relation to a fundamental one and does not talk about the distribution of harmonics in a frequency range of a voltage spectrum, in particular, about their proximity to the voltage fundamental component, which is important, for instance, in some designing process, which is taking into account filtering properties of both load itself and specially installed circuits. The criteria that take into account both the “weight” of high-frequency harmonics and their position in the spectrum are the IHF factors, i.e. WTHD, of various orders. They make weighted (by the harmonic order number) summation of harmonics, as a matter of fact, modeling the effect of the action of the amplitude-frequency

characteristic of the idealized electric circuit of the corresponding order. Such the ideal filter reduces harmonics' magnitudes by the number of times equal to the harmonic order number raised to the filter order power. That is exactly what we can see in the following equation for the n-order integrated voltage harmonics factor [8]:

$$\overline{K}_{hu}^{(n)} = WTHD^{(n)} = \frac{\overline{U}_{(hh)}^{(n)}}{\overline{U}_{(1)}^{(n)}} = \frac{\overline{U}_{(hh)}^{(n)} \cdot \omega^n}{U_{(1)}} = \left[\sum_{k=2}^{\infty} \left(\frac{U_{(k)}}{k^n \cdot U_{(1)}} \right)^2 \right]^{1/2}, \quad (21)$$

where for estimated voltage $u_{an}(t)$ values $U_{(k)}$, $\overline{U}_{(1)}^{(n)}$, $\overline{U}_{(hh)}^{(n)}$ are the RMS values of the k-order harmonic component $u_{an(k)}$, and of the n times integrated values of the fundamental harmonic $\overline{u}_{an(1)}^{(n)}$ and the higher harmonics component $\overline{u}_{an(hh)}^{(n)}$.

The MLVSI phase load circuit with a resistive load and some inverter output LC filter (figure 2,d) produces the second order reduced integral equation:

$$i_a + \frac{1}{R_a C_f} \overline{i}_a^{(1)} + \frac{1}{L_f C_f} \overline{i}_a^{(2)} = \frac{1}{L_f C_f R_a} \overline{u}_{an}^{(2)}, \quad (22)$$

Here the first level approximation current's THD estimate is

$$K_{hia} \approx \overline{K}_{hu}^{(2)} \cdot \left[1 + \left(\frac{1}{(R_a C_f)^2} - \frac{2}{L_f C_f} \right) \frac{1}{\omega^2} + \frac{1}{(L_f C_f)^2 \omega^4} \right]^{1/2}, \quad (23)$$

At last, the MLVSI phase RL-load circuit with LC filter (figure 2,e) leads to the following third order reduced integral equation and current's THD first level estimate:

$$i_a + a_1 \cdot \overline{i}_a^{(1)} + a_2 \cdot \overline{i}_a^{(2)} + a_3 \cdot \overline{i}_a^{(3)} = b_3 \cdot \overline{u}_{an}^{(3)}, \quad (24)$$

$$a_1 = \frac{R_a}{L_a}, \quad a_2 = \frac{1}{C_f} \left(\frac{1}{L_f} + \frac{1}{L_a} \right), \quad a_3 = \frac{R_a}{L_f C_f L_a}, \quad b_3 = \frac{1}{L_f C_f L_a},$$

$$K_{hia} \approx \overline{K}_{hu}^{(3)} \cdot \left[1 + \frac{a_1^2 - 2a_2}{\omega^2} + \frac{a_2^2 - 2a_1 a_3}{\omega^4} + \frac{a_3^2}{\omega^6} \right]^{1/2}. \quad (25)$$

As can be seen, each estimate of load current's THD correlates to the voltage IHF of the order (from zero to three), equal to the order of a differential equation corresponding to the considered equivalent circuit of a generalized load.

The similar THD assessing can be performed for any element of any equivalent circuit regarding current in this element or voltage across it. In particular, equations for voltage u_r across the resistance R_a in circuits of figures 2,b...e are, respectively, as follows:

$$u_r = u_{an}; \quad u_r + \frac{R_a}{L_a} \overline{u}_r^{(1)} = \frac{R_a}{L_a} \overline{u}_{an}^{(1)}; \quad u_r + \frac{1}{R_a C_f} \overline{u}_r^{(1)} + \frac{1}{L_f C_f} \overline{u}_r^{(2)} = \frac{1}{L_f C_f} \overline{u}_{an}^{(2)};$$

$$u_r + a_1 \cdot \overline{u}_r^{(1)} + a_2 \cdot \overline{u}_r^{(2)} + a_3 \cdot \overline{u}_r^{(3)} = a_3 \cdot \overline{u}_{an}^{(3)}. \quad (26)$$

Their compositions repeat the compositions of equations (17), (18), (22) and (24) with the difference in coefficients of the variables only. Thus, one can obtain first level estimates of voltage u_r

THD, similar to formulae (17), (20), (23) and (25), and check that they are proportional to $\overline{K}_{hu}^{(n)}$ of the appropriate orders. The approach to the required MLVSI output variable THD minimizing should be therefore based on consideration and minimization of the integrated voltage harmonics factor (weighted THD) of the appropriate order, depending on the impedance kind of a generalized load.

5. Results of computer simulation and delineation of modes and ranges of amplitude modulation index between two techniques

Regarding the choice of QWS-SVPWM frequency modulation index (here the sampling frequency index), we should abide by the principle of reasonable sufficiency in the sampling frequency increasing and the related improvement of MLVSI output voltages' higher harmonics content because of the higher frequency of power switches' commutations, the higher related power losses in them and the lower efficiency of MLVSI. So, the quarter-wave symmetric space vector PWM must remain a low-frequency and have a comparable with SVC technique number of switches' commutations per output voltages' period.

We shall confine ourselves here to considering the case of the QWS-SVPWM sampling frequency index value of 30 ($m_{sf} = 30$), i.e. for output voltages' frequency value of 50 Hz the sampling frequency value is 1500 Hz.

SVC simulation corresponds here to a continuous survey of the reference sinusoidal signals of the delta voltages. The same resulting voltage waveforms and the related values of the weighted THD (IHF) take place at enough high sampling frequency index value (likely, above 500). However, as have been mentioned above, transitions between available space vectors and changes in voltages values are comparatively rare.

The computer simulation for the two techniques was performed in LabVIEW, where the controllers' models have been implemented. The resulting waveforms of the delta and phase MLVSI output voltages for SVC and QWS-SVPWM at $m_{ay} = 3.14$ are shown in figure 3.

The curves of the dependences of the integrated voltage harmonics factors on the phase-voltage amplitude modulation index m_{ay} are presented in figure 4. They make it possible to delineate modes and to reveal ranges of using SVC and QWS-SVPWM techniques.

It should be noted that the sub-range $m_{ay} = U^* < 1/3$ is SVC zero space vector proximity zone, i.e. the dead band of SVC, therefore this is the sub-range of QWS-SVPWM technique, without distinction as to the kinds of the generalized load.

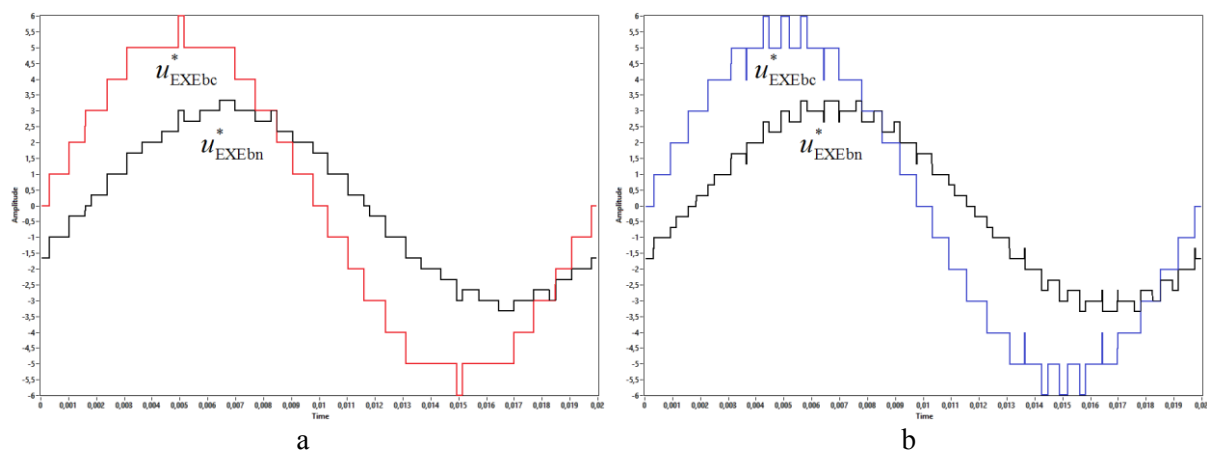


Figure 3. Examples of LabVIEW-simulated delta and phase voltages' waveforms of two techniques: a – space vector control executed voltages idealized signals (pu); b – quarter-wave symmetric space vector pulse-width modulation executed voltages idealized signals (pu).

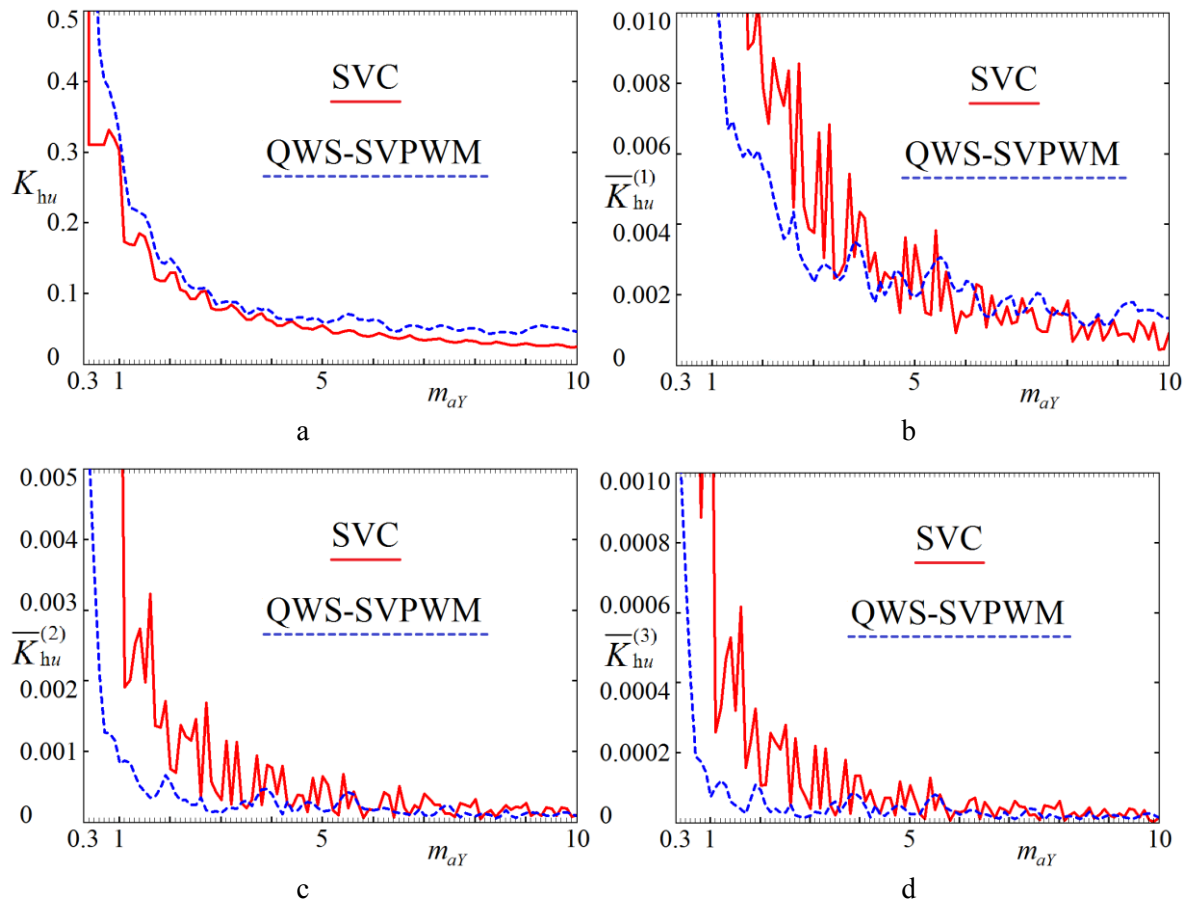


Figure 4. Integrated voltage harmonics factors (IHF, weighted THD) versus phase voltage amplitude modulation index: a – zero-order IHF (THD); b – first-order IHF; c – second-order IHF; d – third-order IHF.

The delineation of the main range $m_{a\gamma} > 1/3$ is as follows. If the generalized load for some being examined MLVSI application can be treated as a pure resistive circuit, the SVC is being employed, as having lower THD values in accordance with figure 4,a.

As can be seen from figure 4,c and d, for the both of the considered circuits with inverter output LC filter (figure 2,d and e), QWS-SVPWM with $m_{sf} = 30$ should be applied in the full range of the phase voltage amplitude modulation index as having an advantage in IHF of the second and third orders. This advantage is clear at low values of the amplitude modulation index. With the increase of output voltage magnitude, some competition from SVC takes place, but the related sub-ranges are too meager to be taken into account.

At last, some generalized load, representable in the form of a resistive inductive series circuit (figure 2,c), must of necessity combine both the considered control algorithms to minimize values of the voltage IHF of the first order, see figure 4,b. This case is very common, it occurs, for instance when some transformerless-by-output MLVSI circuit is loaded by some three-phase symmetrical squirrel-cage induction motor without any filter, and its equivalent per-phase circuit can be simplified and accordingly reduced to this low-order RL-form. Here SVC is losing the competition against QWS-SVPWM under low phase-voltage amplitude modulation index values, namely, when $m_{a\gamma} \leq 5.1$.

Thus, the offered concept can easily be implemented with a guaranteed positive effect on the quality of voltages, generated by an arbitrary circuit of a three-phase multilevel inverter loaded by a specified generalized consumer circuit.

6. Conclusion

The combined control algorithm for a three-phase multilevel voltage source inverter with any circuit configuration and any number of equal output voltage levels is developed to improve the quality of its output power, namely THD of some specified output variable. The approach is oriented to minimization of an integrated voltage harmonics factor (a weighted THD) with an order, dependent on the kind of an equivalent circuit of a generalized load. The algorithm employs the space-vector control, which is choosing the nearest vector to the reference one, and the recently offered space vector pulse-width modulation technique with quarter-wave output voltage symmetry to keep low-frequency nature of power switches' commutations and low level of related losses in them. The combined algorithm is based on processing values of the integer and fractional parts of the samples of the relative values of the reference delta voltages and realizing the two-component formation of the two output delta voltages. The obtained curves of the dependences of the integrated voltage harmonics factors on the phase-voltage amplitude modulation index made it possible to delineate modes and to reveal ranges of applying of the two alternatives.

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