Abstract

The paper proposes some ways to reduce switching losses in inverters that supply the induction motor drives. Modulation systems with theoretical explanation and simulation results are analyzed. The benefits and drawbacks of the six-step, pulse-width, continuous and discontinuous space vector control techniques are described. The new approach can be effectively applied in the open-loop and direct torque controlled drives.

Keywords

Modulation, power electronics, electric drive, simulation.

1. Introduction

Today, alternating current (ac) motor drives are employed in different industrial areas with a wide power range [1]. In ac drive applications it is most desirable to use the sinusoidal waveform. However, the output of practical voltage source inverters (VSI) and current source inverters (CSI) inevitably has a certain amount of harmonic content. In practice, harmonic content can be reduced to a low value by a proper modulation strategy.

Several methods have been developed for this purpose, including simple six-step modulation, high-frequency pulse-width modulation (PWM) and progressive space vector modulation (SVM). The modulation method governs the voltage and current harmonics, torque ripple, acoustic noise emitted from the motor as well as electromagnetic interference.

Modulation techniques were originally restricted to the synthesis of stepped waveforms [2]. This strategy can be interpreted as a quantization process in which the reference sinusoidal waveform is approximated by the seven discrete voltage levels available at direct current (dc) supply. Implementation of this type of modulation is relatively simple and does not require high switching speed, which makes it suitable for converters. However, the six-step algorithm is the source of significant harmonic distortion as its waveforms are the low frequency rectangles, therefore the six-step inverters have found occasional use in motor drives.

Far more popular, especially for the converters with faster switching devices, are the PWM techniques. PWM is valuable for drive performance in respect to voltage and current harmonics, torque ripple, acoustic noise emitted from an induction motor and also electromagnetic interference.

Today, SVM is the most sophisticated method for generating a fundamental sine wave providing a higher voltage to the motor and lower switching losses as compared to both the six-step and PWM methods. This advanced computation-intensive control approach is possibly the best among all the power converter control techniques in the field of variable drive applications. Several SVM algorithms have been reported in [3]-[5]. In voltage source inverters they are based on the output voltage vectors and in the current source inverters on the output current vectors.

Both PWM and SVM suffer from high switching losses in inverter switches occurring because of high commutation frequencies. In an inverter operation above 100 Hz, the commutation losses with PWM switching become unacceptably high.

The objective of this study was to analyze and compare different modulation techniques from the point of view of their harmonic distortion and switching losses. This paper reviews the modulation systems of motor drive inverters. Advantages and disadvantages of different modulation methods are illustrated in the steady-state and dynamic modes of the inverter performance in ac drive applications. A new technique of reduced switching within the modulating interval is proposed.

The proposed approach can be effectively applied in drives working in the open-loop control mode. At the same time it is important for the vector and direct torque controlled drives.

2. Control Methods of Induction Motors

Contemporary electrical engineering industry disposes of several methods of induction motor control.

The torque of an induction machine is a result of the interaction of the effective flux and an active component of the stator current which generates the rotor electromagnetic force (EMF) and together with the rotor current participates in the torque burning.
At the same time, a reactive component of the stator current establishes the effective flux linkage which is approximately equal to the time integral of the impressed voltage. The machine works properly if the magnitude of the flux linkage is kept constant, which denotes a circular trajectory of the flux. Therefore, in the induction motor both of the current components are bounded together so that the alternation of one of them changes the other. The frequency control, slip control, mutual current / frequency, voltage / frequency, and flux / frequency controls known as a whole as the scalar controls are used along with the more sophisticated vector and direct torque controls.

The voltage/frequency control (VFC) is commonly applied today to control the speed of asynchronous motor drives having inferior dynamic quality. To adjust the speed, the supply frequency must be changed. But as the frequency falls, the flux rises and the magnetizing current increases as well, causing additional heating of the motor. Therefore, the current, slip, or EMF should be changed along with the frequency. Among these control “handles”, the frequency adjustment is by far the most critical as the small variations in frequency produce significant changes in current and torque.

Adjusting the induction motor speed by changing the supply frequency along with the current control is called a current-frequency control (CFC). Unlike the VFC, this approach requires the implementation of CSIs instead VSCs. The CSI is less sensitive to the parameter instability, however, its bulky inductor is of high weight and low speed of response. Therefore, these inverters are designed often by introducing the current feedback into the traditional dc link VSI.

A field-oriented control (FOC) became the industry standard for high dynamic asynchronous drives the performance of which is close to that of dc motor drives. It was one of the most important innovations in ac motor drives which opened the door to researchers aiming for ever enhanced control performance. For this, the controller needs to obtain accurate information about the rotor speed or the air gap flux from the proper sensors. To arrange the control which instantly aligns the stator current orthogonally to the rotor flux, a Park transformer is used to convert the motor currents and voltages acquired by sensors into the rectangle frame. Further, using the motor model the rotor flux linkage is calculated. Thus, the closed-loop system fixes the rotor torque. As the resulting torque is proportional to the current at the constant flux, an induction motor is adjusted quite similar to a dc motor.

Unlike the FOC, which provides the stator current control method, a direct torque control (DTC) method works directly with the stator flux and torque without having a need for inner loops with current controllers. In this technique, the inverter switches are directly adjusted using a look-up table in order to control the stator flux and torque. An important point of the DTC is the correct selection of the stator voltage vector in order to maintain flux and torque within the limits of two hysteresis bands. For this purpose, the control system should be able to generate multiple voltage vectors, which implies the use of the SVM.

3. Analysis of Six-Step Modulation

An induction motor is mostly supplied from the three-phase bridge inverter shown in Fig. 1. To develop the control algorithms, a simplified switching model shown schematically in Fig. 2, a is initially used. Further, the enhanced model of the three-leg inverter built on IGBTs VT1 and VT4, VT2 and VT5, and VT3 and VT6 with the freewheeling diodes is used. The switches are examined in eight different combinations designated by the binary variables 100, 110, 010, 011, 001, 101, 111, and 000, which indicate whether the switch is in the top (1) or bottom (0) position, thus defining all the possible switching states shown in Fig. 2, b. During the modulating period 1/ω = 1/(2f_m), phase voltages sequentially change their values depending on which the IGBTs switch on and off. Here, ω is an angular motor speed, f_m = 1/ T_m the modulating frequency, T_m a modulation period, and U_d the dc supply voltage.

The modulating period is divided into six sectors. Referring to the circuit in Fig. 2, in the six-step mode of operation the states of the switches and the phase, neutral, line-to-neutral, and line-to-line voltages have the waveforms plotted in Fig. 3. Here, the switching of the three inverter legs supplied by the dc voltage U_d is the phase-shifted by 120°. Each phase L1, L2, L3 is under the current during half a modulating

![Fig. 1](image1.png)

![Fig. 2](image2.png)
the voltages of phases (line-to-dc neutral) times the inverter switching frequency. Therefore, the load and dc neutral points remain in series with the remaining two parallel-connected positive pole to the negative one being sequentially specific phase is alternately switched from the period and open during another half-period. A

The rms values of the six-step line-to-neutral and line-to-line voltages do not depend on the load but rather on the load current pulsations and instability with additional energy losses especially when the frequency is low.

When a reactive load is connected to an inverter, it becomes necessary to take into account reverse-connected diodes that carry return current (Fig. 1). The presence of the diodes identifies the circuit as a voltage source inverter rather than a current source inverter for which return diodes are unnecessary.

The currents sketched by the dotted lines in Fig. 3 undergo an exponential increase of value and at each switching there is an abrupt change of their slopes. A motor impedance value is speed related, thus there are magnetically induced voltages in the windings. Hence, the fundamental component of the current lags the voltage by about \( \pi / 3 \) of phase angle, which is typical of low-speed induction motor operations.

4. Analysis of PWM Modulation

As opposed to the six-step modulation, the PWM method is gradually taking over the inverter market of control applications. This technique combines both the voltage and the frequency control within the inverter itself. The PWM circuit output represents the chain of constant magnitude pulses, the duration of which is modulated to obtain the necessary specific waveform under the desired modulating frequency \( f_m \) ranging from a few kilohertz in simple motor control systems up to several tens kilohertz. For the three-phase operation, the symmetrical triangle double-sided wave of the carrier frequency \( f_c \) is currently compared with the modulating wave, thus generating the control pulses.

Table of Harmonics of Six-Step Modulation

<table>
<thead>
<tr>
<th>Harmonic Order</th>
<th>THD ( U_{LN} )</th>
<th>THD ( U_{LL} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.47</td>
<td>0.78</td>
</tr>
<tr>
<td>5</td>
<td>0.09</td>
<td>0.156</td>
</tr>
<tr>
<td>7</td>
<td>0.07</td>
<td>0.111</td>
</tr>
<tr>
<td>11</td>
<td>0.04</td>
<td>0.071</td>
</tr>
<tr>
<td>13</td>
<td>0.03</td>
<td>0.06</td>
</tr>
<tr>
<td>17</td>
<td>0.02</td>
<td>0.04</td>
</tr>
</tbody>
</table>

To implement VFC, the output voltage must be adjusted in almost direct proportion to frequency. Commonly, voltage control is impossible under the six-step approach, and the need in a phase-controlled rectifier to adjust the dc voltage is an inherent weakness of this circuit. Another disadvantage of the six-step modulated inverters is that they suffer from low order harmonics of significant amplitudes due to the non-sinusoidal inverter voltage shape, which leads to the load current pulsations and instability with additional energy losses especially when the frequency is low.

The total harmonic distortions are as follows:

\[
THD_{LN} = \sqrt{\frac{U_{LN}^2}{U_{LN}^2}} = 0.31, \\
THD_{LL} = \sqrt{\frac{U_{LL}^2}{U_{LL}^2}} = 0.27
\]

The presence of the diodes identifies the circuit as a voltage source inverter rather than a current source inverter for which return diodes are unnecessary.
The rms value of the PWM line-to-neutral and line-to-line voltages are presented.

The set of the PWM waveforms is shown in Fig. 4. Here, the states of the top switches VT1, VT2 and VT3 of an inverter shown in Fig. 1 as well as the phase, line-to-neutral and line-to-line voltages are presented.

The rms value of the PWM line-to-neutral and line-to-line voltages with a modulation ratio $m$ are as follows:

$$U_{LN\text{ rms}} = \frac{\sqrt{2}}{\pi} m U_d = 0.399 m U_d, U_{LL\text{ rms}} = 0.7 m U_d$$

The fundamental sinusoidal components of the voltages obtained by the Fourier analysis of the PWM waveforms will be

$$U_{1LN\text{ max}} = \frac{mU_d}{2} = 0.5 m U_d, \quad U_{1LN\text{ min}} = 0.355 m U_d,$$

$$U_{1LL\text{ max}} = \frac{3mU_d}{2} = 0.87 m U_d, \quad U_{1LL\text{ min}} = 0.612 m U_d$$

The total harmonic distortions are as follows:

$$THD_{1LN} = 0.463, \quad THD_{1LL} = 0.485$$

For the balanced supply, the frequency ratio $m_f$ of the carrier and modulating frequencies should be an odd multiple of three. The carrier frequency is then a triplen of the modulating frequency so that the output waveform does not contain the carrier frequency or its harmonics.

Harmonic content of the PWM modulation depends on the frequency ratio $m_f$ and the modulation ratio $m$, as shown in Table 2 for the line-to-line rms voltage.

**TABLE 2. HARMONICS OF PWM MODULATION**

<table>
<thead>
<tr>
<th>Harmonic</th>
<th>$m = 0.2$</th>
<th>$m = 0.4$</th>
<th>$m = 0.6$</th>
<th>$m = 0.8$</th>
<th>$m = 1.0$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$m_f \pm 2$</td>
<td>0.012</td>
<td>0.245</td>
<td>0.367</td>
<td>0.490</td>
<td>0.612</td>
</tr>
<tr>
<td>$m_f \pm 4$</td>
<td>0.010</td>
<td>0.037</td>
<td>0.080</td>
<td>0.135</td>
<td>0.195</td>
</tr>
<tr>
<td>$2m_f \pm 1$</td>
<td>0.005</td>
<td>0.011</td>
<td>0.017</td>
<td>0.020</td>
<td>0.023</td>
</tr>
<tr>
<td>$2m_f \pm 5$</td>
<td>0.008</td>
<td>0.020</td>
<td>0.029</td>
<td>0.032</td>
<td>0.034</td>
</tr>
<tr>
<td>$3m_f \pm 2$</td>
<td>0.027</td>
<td>0.085</td>
<td>0.124</td>
<td>0.108</td>
<td>0.084</td>
</tr>
<tr>
<td>$3m_f \pm 4$</td>
<td>0.007</td>
<td>0.029</td>
<td>0.064</td>
<td>0.096</td>
<td>0.120</td>
</tr>
<tr>
<td>$4m_f \pm 1$</td>
<td>0.100</td>
<td>0.096</td>
<td>0.005</td>
<td>0.064</td>
<td>0.042</td>
</tr>
</tbody>
</table>

Hence, PWM inverters have several problems in terms of motor drives. They produce 16-18% lower voltage magnitudes than the six-step inverters, thus the supply dc voltage cannot be fully utilized. As the sinusoidal PWM prevents the full fundamental voltage from reaching its maximum value (27-32% lower than the six-step inverter), operation at a reduced voltage becomes costly, implying higher currents for a given power requirement. This duty has an impact on the choice of switching devices used in inverters. The THD factor is also actually lower for the six-step inverter: while it is 0.463-0.485 for the PWM waveforms, it is only 0.27-0.31 for the six-step system.

Nevertheless, the main advantage of PWM is the fact that with high switching frequency, the harmonic frequencies of the voltage are very much higher. When $m_f > 9$, the harmonic magnitudes are independent of $m_f$ and the lowest significant harmonic number is 11 whereas the six-step modulation is sensitive to the 5-7 harmonics.

The induction motor torque responds largely to the fundamental frequency component. As the motor winding circuits have an inductive character, the PWM current contributions due to the harmonic voltages are small, therefore the resultant motor currents are fairly close to the sine wave.

In contrast, the harmonics of the PWM voltage are often more significant than those of the consequent motor current. The eddy-current and hysteresis iron loss variation with flux and frequency is often greater than the copper losses in the windings. Hence, it is common practice that the PWM driven motors are derated by 5 to 10%. The presence of harmonics in the waveforms increases the voltage rms values while leaving the supply power unchanged. Therefore, harmonics do not lead to the net delivery of energy to the motor, yet they increase the losses in the system.

**5. Analysis of Space Vector Modulation**

The essence of SVM lies in the denial of the simultaneous commutation of all the inverter IGBTs in favour of the commutation between some preliminarily selected states. Additionally, as distinct from the significant computing time required to calculate the PWM signals in real time, the calculation process in SVM is simplified, therefore better performance can be obtained.

To proceed from the six-step modulation to the SVM, each of the binary variables 100, 110, 010, 011, 001, 101, 111, and 000, is associated with a particular space vector $U_{abc}$. The states of which correspond to the definite space position of the resultant voltage vector. Figure 5, a. illustrates the switching states and appropriate voltage vectors of the three-phase inverter shown in Fig.1. The desired voltages at the output of the inverter are represented...
by an equivalent space vector $u^*$ rotated in the counterclockwise direction at an angular frequency $\omega$ referenced for the steady state operating conditions. The magnitude of the space vector is related to the magnitude of the output voltage and the modulating time this vector takes to complete one revolution is the same as the fundamental time period of the output voltage.

The vector set includes six active voltage space vectors $U_1$ to $U_6$ that correspond to the switching states 100, 110, 010, 011, 001, 101, and two zero voltage space vectors $U_0$, $U_7$ keeping with 111 and 000. On the plane shown in Fig. 5, $a$, active vectors are situated 60° apart, segmenting the plane by equal sectors 1 to 6. Voltage vectors $U_1$, $U_2$, $U_3$ are oriented along the axes of $L_1$, $L_2$, and $L_3$ phases. Supply voltage $U_d$ specifies the amplitude of the space vectors. The demanded reference vector $u^*$ is determined by its module and phase:

$$u^* \leq \frac{U_d}{\sqrt{3}}, \quad \theta^* \leq \frac{\pi}{3}.$$  

In SVM, vector $u^*$ is treated through adequate timing of adjacent non-zero and zero space vectors. It is composed by a switching sequence comprising the neighbor space vectors $U_{i1}$...$U_{6}$, while filling up the rest of the time interval with zero vectors $U_0$ or $U_7$ during the voltage alteration. As a result, the end of the vector $u^*$ travels along the hexagon or stops. Thus, the maximum modulation ratio is expressed by

$$m_{\text{max}} = u^*_{\text{max}}.$$  

The modulating period $2\pi$ involves six sectors, each including a number of fixed sampling intervals $T_c$. To move between the neighbor vectors $U_i$ and $U_{i+1}$, the switching sequence of pulses $U_j$, $U_{j+1}$, and $U_{0}$ has to be generated in each sampling interval, the time durations of whose are consequently $t_i$, $t_{i+1}$, and $t_0$ that is

$$u^* = (t_i U_i + t_{i+1} U_{i+1} + t_0 U_{0}) / T_c,$$

where $U_i$ is the magnitude of the first space vector; $U_{i+1}$ is the magnitude of the space vector valid in the next $T_c$ interval; $U_{0}$ = 0; $t_i$ and $t_{i+1}$ are the time durations for the two adjacent vectors that are to be computed in each $T_c$ in real time or beforehand with keeping in memory; $t_0$ is the zero vector duration. The solution for $t_i$, $t_{i+1}$ and $t_0$ results in

$$t_i = \frac{\sqrt{3} u^*}{2U_d} T_c \sin \left( \frac{\pi}{3} - \theta^* \right), \quad t_{i+1} = \frac{\sqrt{3} u^*}{2U_d} T_c \sin \theta^*,$$

$$t_0 = T_c - t_i - t_{i+1}, \quad \theta^* = T_c N,$$

where $N$ is a number of the current sampling period counted from the sector starting point. While travelling between the adjacent sectors, the computations repeat regularly basing on the value of the reference voltage and speed at the beginning of each sampling period. In this way, the reference is updated at every sampling interval $T_c$. In fact, this technique produces an average of three voltage space vectors $U_{i}$, $U_{i+1}$, and $U_{0}$ $(U_7)$ over a sampling interval $T_c$.

From Fig. 5 it is obvious that the possibility of reaching the line-to-line voltage as high as $U_d$:

$$m_{\text{max}} = \frac{\sqrt{3}}{2}, \quad U_{i+1,\text{max}} = \frac{2mU_d}{3} \leq 0.7U_d, \quad U_{i+1,\text{min}} = U_d$$

whereas PWM gives less at $m = 1$. The better modulation depth at the same level of harmonic distortions as compared to the PWM is an important advantage of the SVM algorithm. Due to the higher line-to-line voltage amplitude, the torque generated by the motor is higher, resulting in the better dynamic response of the motor. With the increased output voltage, the user can design the motor control system with reduced current rating, keeping the same power capacity. It helps to decrease inherent conduction loss of the inverter.

6. Comparison of Continuous and Discontinuous SVM Techniques

Knowing the values of $t_i$, $t_{i+1}$ and $t_0$ the switching patterns for the consecutive periods of $T_c$ can be constructed. There are many SVM schemes for the three-phase converters that contribute to different levels of performance. All of them produce the same switching time for the active vectors. The only difference between them is the choice of the zero vectors $U_0$ and $U_7$ and the sequence in which the vectors are applied within the sampling cycle $T_c$. Therefore, each scheme is suitable for a different operating condition.

Several switching rules for the SVM developers have been proposed in [6], [7], [8]. They require the circular trajectory of the reference voltage vector, only one switching per the state transition, and no more than three commutations in a sampling period. These rules aim in limiting the number of switching actions that would lead towards reduction in switching losses. Besides, symmetry property could be maintained in waveforms to achieve lower harmonic distortion.

In continuous space vector modes, both $U_0$ and $U_7$ are used in the same sampling cycle, as shown in Fig. 6, $a$, for the first sector. Here, $VT_1$, $VT_2$ and $VT_3$ are assumed to be the states of the top switches of a converter shown in Fig. 1. In this strategy, the total zero voltage vector time $t_0$ is divided equally between $U_0$ (111) and $U_7$ (000). Also, in this method, $t_0$ is distributed symmetrically at the start and end of
the sampling period. Opposite to the six-step modulation with the switching sequence of the inverter legs $U_1 \rightarrow U_2 \rightarrow U_3 \rightarrow U_4 \rightarrow U_5 \rightarrow U_6 \rightarrow \ldots$, within the modulating periods, in the continuous SVM the switching sequence of the upper legs of the inverter is given by $U_0 \rightarrow U_1 \rightarrow U_{101} \rightarrow U_7 \rightarrow U_8 \rightarrow U_1 \rightarrow U_0$ in two sampling periods. The scheme has three switch-ons and three switch-offs within $T_c$.

As far as Fig. 6 represents the first sector only, Table 3 regulates the vector in other sectors.

![Fig. 6](image)

<table>
<thead>
<tr>
<th>Sector</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
</tr>
</thead>
<tbody>
<tr>
<td>$t_0$</td>
<td>$t_i$</td>
<td>$t_0$</td>
<td>$t_i$</td>
<td>$t_0$</td>
<td>$t_i$</td>
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<tbody>
<tr>
<td>$U_0$</td>
<td>$U_1$</td>
<td>$U_{101}$</td>
<td>$U_7$</td>
<td>$U_8$</td>
<td>$U_1$</td>
<td>$U_0$</td>
</tr>
</tbody>
</table>

The waveforms for the continuous SVM are practically the same as for PWM shown in Fig. 4.

An idea of the discontinuous modulation is based on the assumption that one phase is clamped by 60° to the lower or upper level of the dc voltage [9]. When the reference vector is in Sector 1 of Fig. 5, $a$, the phase $L1$ voltage reaches the maximum as compared to other phases. Switch VT1 of the phase $L1$ keeps in conduction all the time and phases $L2$ and $L3$ modulate, as shown in Fig. 6, $b$. This means that only one state vector can be used in this 60° period of time. By extending this conclusion to a complete sampling period, a distribution of zero vectors in the hexagon can be obtained where 111 and 000 vectors are used alternatively in the sectors (Fig. 6, $c$).

As opposed to the continuous modulation, it gives only one zero state per sampling period, hence only one vector, $U_0$ or $U_7$, is used in the switching cycle. Thus, one of the main advantages of the discontinuous modulation is the reduction of the number of switching processes of VT1 – VT6 in a period $T_c$ from 6 to 4, providing a 33% decrease in the switching losses.

The set of waveforms for the discontinuous SVM is shown in Fig. 7. Referring to Fig. 7 it can be noted that the center pulses in the SVM waveform are broader than the pulses where the fundamental voltage passes through zero.

7. A Method to Reduce Switching Losses

The permissible harmonic content in the voltage is dependent on the type of load on the ac side. Clearly, under the influence of electromagnetic processes occurring in an inverter-motor system, the character of the load voltage transients has no direct equivalence with the reference signals. The degree of such discrepancy depends on the motor winding inductions and resistances as well as on the mode of the motor operation, thus resulting in additional voltage distortion and reduced use of supply power.

In this context, it may be stated that the waveforms of the ac voltage and the ac current are normally different. While the load is an ac motor the harmonic content in the current is considerably less than the harmonic content in the voltage applied to the motor. This is because the motor inductance itself acts as a filter element, considerably attenuating the higher-order harmonic currents [10].

The current vector trajectory is a continuous function during each time interval for which the voltage is constant. Since the voltage space vector changes its discrete positions at each 60 degrees, the current space vector trajectory results close to hexagonal as presented in Fig. 5, $b$. In PWM and SVM modulation modes the current space vector moves smoothly [3].

The phase current can be derived by the integral of the phase voltage equation:

$$ U_L = L \frac{dI_L}{dt} = \frac{T_m}{3L} \frac{T_m}{6} U_d $$

The load currents in Sector I of Fig. 5, $a$, can be expressed from Fig. 3 as follows:
When the conduction delays become very narrow, there is insufficient time to reverse bias the off-going power device. This is the reason why the protective intervals and the inherent delays of the switching devices must also be taken into account. Particularly, it concerns the differences between turn-on and turn-off times, which can cause considerable distortion of the converter characteristics at low output voltage and frequency. Hence, if voltage across this device increases in a positive sense too soon, the device will again conduct and high losses will occur. Therefore, if the modulation scheme calls for a delay width below this minimum time, this delay should be omitted altogether.

An electromagnetic time constant of the motor drive \( T_e = \frac{L}{R} \) may be calculated as given in [11]. Parameters of the asynchronous electric drive having the stabilized rotor flux linkage are as follows:

\[
R = R_x + R_L, L = \sigma L_t, \quad \sigma = 1 - k_1k_2, \quad k_1 = \frac{L_1}{L_2}, \quad k_2 = \frac{L_2}{L_1},
\]

\[
L_{t2} = \frac{3}{2} \frac{X_{t2}}{314}, \quad L_{t1} = \frac{X_{t1}}{314} + L_{t2}, \quad L_2 = \frac{X_2}{314} + L_{t2},
\]

where index 1 designates the stator parameters, and index 2 indicates the rotor parameters, and \( X_1, X_2, X_{t1}, R_1, R_2 \) are the rated values from the motor data sheet.

An improved modulating algorithm is executed as follows. Each calculated time interval \( t_p \), \( t_{on1} \), and \( t_{off} \) is compared with an electromagnetic time constant \( T_e \) of the drive circuit. While such interval is less than \( T_e \), it is eliminated and its value is saved in the controller memory before the next sampling. Within the next sampling this value is added to the new calculated time interval and the sum is compared with \( T_e \) again. When the summing time overcomes \( T_e \), it is used in the commutation process.

The waveforms built using the proposed algorithm are shown in Fig. 8. As these waveforms show, switching is moved here to the side sectors of the device conducting periods resulting in the reduced switching losses. The harmonics typical for the six-step inverter are presented also with the modulation and the motor current and torque do not sense these holes.

8. Conclusion

Three main modulation techniques were compared in the paper: simple six-step modulation, high-frequency PWM, and SVM. An impact of the modulation method on the voltage and current harmonics, torque ripple, acoustic noise emitted from the motor and electromagnetic interference was studied. An SVM advantages in decreasing the current ripples and switching losses have been illustrated in the steady-state and dynamic modes of the converter performance in ac drive applications. The discussed category of the modulation techniques can be effectively applied in drives working in the open loop VFC modes as well as in the vector controlled DTC drives.

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References


