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A Multi-Frequency PWM Scheme for Multi-Level Five-Phase Open-End Winding Drives

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***Abstract* – This paper presents a multi-frequency pulse width modulation (PWM) scheme for a multi-level five-phase open-end winding drive using the unified modulation technique. The open-end winding machine is supplied via a two-level voltage source inverter from each side of the windings. One inverter operates in ten-step mode while the other is operated under PWM. The multi-frequency PWM scheme is aimed to control the magnitude of the fundamental output and to cancel low-order harmonics produced by the inverter operating in ten-step mode. Simulation results show excellent performance resulting in multi-level output voltages with a low THD. Implementation of the scheme is straightforward, compared with the space-vector modulation based approach, since space vector transformation, sector identification, and look up tables are not required.**

***Index Terms* – dual-inverter supply, open-end winding machine, multi-frequency pulse width modulation, unified modulation technique.**

I. INTRODUCTION

Multi-level and multi-phase voltage source inverters (VSI) have been attracting increasing research interest recently due to their ability to overcome voltage and current limitations of power semiconductors and their inherent ability to tolerate faults [1]-[2]. There are several configurations of multi-level converters, the main ones being the neutral point clamped (NPC), the flying capacitor (FC) and cascaded converters [2-5].

Among the cascaded converters, the dual two-level inverter configuration has received growing attention due to its simple structure. The additional diodes used in the diode-clamped (NPC) VSI are not needed, leading to a saving in the overall number of components. Furthermore, the issue of proper capacitor voltage balancing does not exist if the supply is two-level at each winding side. Typically, three-phase VSIs are utilised. Application of such a dual-inverter supply enables drive operation with voltage waveform equivalent to the one obtainable with a three-level VSI in single-sided supply mode [5]. Such drives are being considered as alternative supply solutions in EVs/HEVs [6]-[7] and electric ship propulsion [8].

Due to their well known advantages [1], multi-phase drives have also been considered for similar applications as the multi-level drives. The only known examples where multi-phase drive systems in open-end winding configuration have been developed are [9-11] where an asymmetrical six-phase induction motor drive was considered in [9]-[10].

This paper proposes a new, and relatively simple, PWM algorithm for the five-phase open-end winding three-level drive supplied by two two-level five-phase inverters. The PWM algorithm is based on the decomposition of the there-level space vector decagon into a number of two-level decagons. A similar idea has been proposed in [12] for a three-phase NPC inverter and in [13] for a three-phase open-end winding drive. In the case of the five-phase open-end winding drive the situation is significantly more complicated since both 2-D planes have to be considered. It is shown that the proposed modulation strategy is capable of achieving the target fundamental while eliminating any low-order harmonic content in the output load phase voltage.

The paper begins with a review of the five-phase two-level drive characteristics, which is followed by a general description and mathematical model of the open-end winding topology, along with mapping of the space vectors into the 2-D planes. Next, the proposed modulation method is described. It is shown that due to the nature of the five-phase topology one of the inverters needs to operate with so called multi-frequency PWM. Finally, the performance of the proposed modulation method is verified via simulation.

II. GENERAL PROPERTIES OF TWO-LEVEL FIVE-PHASE DRIVES

Although the proposed modulation technique can be considered a carrier based PWM method the space-vector approach is an extremely useful tool when trying to understand the relationships which govern the performance of five-phase drives and the corresponding VSI modulation techniques. A five-phase machine can be modelled in two 2-D sub-spaces, so-called *α-β* and *x-y* sub-spaces [14]. It can be shown that only current harmonic components which map into the *α-β* sub-space develop useful torque and torque ripple, whereas those that map into the *x-y* sub-space do not contribute to the torque at all. A multi-phase machine with near-sinusoidal magneto-motive force distribution presents extremely low impedance to all non-flux/torque producing supply harmonics and it is therefore mandatory that the supply does not generate such harmonics. What this means is that the design of a five-phase PWM strategy must consider simultaneously both 2-D sub-spaces, where the reference voltage, assuming pure sinusoidal references, is in the first plane while reference in the other plane is zero. Two-level five-phase inverters can generate up to 2^5=32 voltage space vectors with corresponding components in the *α-β* and *x-y* sub-spaces. Active (non-zero) space vectors belong to three groups in accordance with their magnitudes - small, medium and large space vector groups. The magnitudes are identified

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with indices s, m, and l and are given as $|\bar{v}_s| = 4/5\cos(2\pi/5)V_{dc}$, $|\bar{v}_m| = 2/5V_{dc}$, and $|\bar{v}_l| = 4/5\cos(\pi/5)V_{dc}$, respectively. Four active space vectors are required to generate sinusoidal voltages [1]. In order to provide zero average voltage in the x-y plane two neighbouring large and two medium space vectors are selected [14]. It is shown in [14] that the maximum peak value of the output fundamental phase-to-neutral voltage in the linear modulation region is $v_{max} = 0.525V_{dc}$, resulting in a maximum modulation index, M =1.05.

III. FIVE-PHASE OPEN-END WINDING TOPOLOGY

Fig. 1 illustrates the open-end winding structure, based on utilisation of two two-level five-phase VSIs. The two inverters are identified with indices 1 and 2. Inverter legs are denoted with capital letters, A,B,C,D,E and the negative rails of the two dc links are identified as $N1$ and $N2$. Machine phases are labelled as a,b,c,d,e. Phase voltage positive direction is with reference to the left inverter (inverter 1). Two isolated dc supplies are assumed so that the common mode voltage (CMV) v_{N1N2} is of non-zero value (the issue of CMV elimination is not addressed here). The resulting space vectors in dual-inverter supply mode will depend on the ratio of the two dc link voltages. The situation considered further on is the setting $V_{dc1} = V_{dc2} = V_{dc}/2$, which gives the equivalent of single-sided three-level supply. Using the notation of Fig. 1, phase voltages of the stator winding can be given as [11]:

$$\begin{aligned}
v_{as} &= v_{A1N1} + v_{N1N2} - v_{A2N2} \\
v_{bs} &= v_{B1N1} + v_{N1N2} - v_{B2N2} \\
v_{cs} &= v_{C1N1} + v_{N1N2} - v_{C2N2} \\
v_{ds} &= v_{D1N1} + v_{N1N2} - v_{D2N2} \\
v_{es} &= v_{E1N1} + v_{N1N2} - v_{E2N2}
\end{aligned} \tag{1}$$

Space vectors of phase voltages in the two planes are determined with;

$$\begin{aligned}
\bar{v}_{\alpha-\beta} &= (2/5)\left(v_a + \underline{a}v_b + \underline{a}^2 v_c + \underline{a}^3 v_d + \underline{a}^4 v_e\right) \\
\bar{v}_{x-y} &= (2/5)\left(v_a + \underline{a}^2 v_b + \underline{a}^4 v_c + \underline{a}^6 v_d + \underline{a}^8 v_e\right)
\end{aligned} \tag{2}$$

where $\underline{a} = \exp(j2\pi/5)$. Using (1) and (2), one gets

$$\begin{aligned}
\bar{v}_{\alpha-\beta} &= \bar{v}_{\alpha-\beta(A1B1C1D1E1)} - \bar{v}_{\alpha-\beta(A2B2C2D2E2)} \\
\bar{v}_{x-y} &= \bar{v}_{x-y(A1B1C1D1E1)} - \bar{v}_{x-y(A2B2C2D2E2)}
\end{aligned} \tag{3}$$

since $v_{N1N2}(1+\underline{a}+\underline{a}^2+\underline{a}^3+\underline{a}^4) = 0$. In (3) the two space vectors on the right-hand sides of the two equations are corresponding voltage space vectors of the two five-phase two-level VSIs. There are three vector lengths, large, medium and small: $0.32366V_{dc}$, $0.2V_{dc}$, $0.123V_{dc}$. Voltage space vectors, produced by the 1024 possible switching states, are illustrated in Fig. 2 for the α–β plane. There are 211 space vectors, 21 active phase voltage space vectors per 36 degrees sector. Space vector mapping into the x-y plane follows the pattern that exists for a five-phase VSI, the largest vectors of the first plane map into the smallest vectors of the second plane, and vice versa [11].

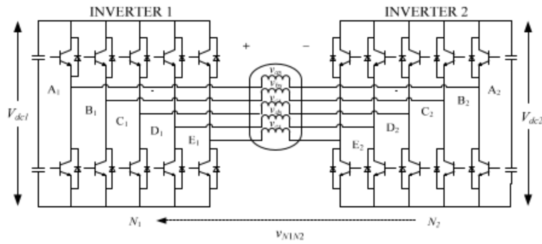

Fig. 1. Dual-inverter fed five-phase open-end winding topology.

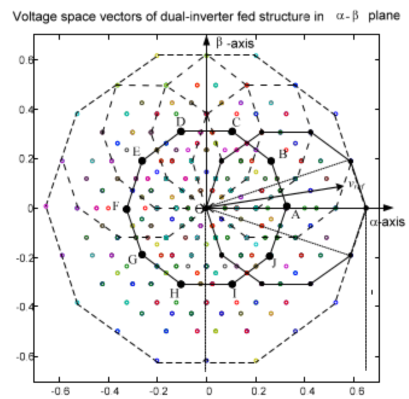

Fig. 2. Voltage space vectors of dual-sided structure decomposed into smaller decagons.

IV. PRINCIPLE OF THE PROPOSED ALGORITHM

The design of a suitable modulation method for the dual-fed structure represents a considerable challenge due to the high number of switching states. An attempt is made here, by at first decomposing the voltage vectors into smaller decagons equal to two-level structure, as shown in Fig. 2. There are 10 small decagons in the outer ring (the centre points are labelled as A, B, C, D, E, F, G, H, I, J) and one decagon in the centre (centred at point O). The centre decagon comprises of vectors which can be activated if one inverter is used up to half of the achievable maximum voltage with the other one short-circuited. As a consequence, the converter is in two-level mode of operation and the PWM is based on four active and zero vector application, as discussed in section II and [14]. As can be seen in Fig. 2, the origins of the outer decagons are located on the outer vectors of the inner decagon, denoted by the black dots in Fig 2. In the case of the three-phase topology [12] operation in the outer region (outer hexagon) is achieved when one inverter operates with a single voltage space vector applied (the nearest one to the reference) and the second inverter is modulated using the standard three-phase two-level space vector modulation (SVM) technique. Let the applied vector for the one inverter be $\underline{v}_i$ and let the reference be $\underline{v}^*$. Here $\underline{v}_i$ is the vector produced by the inverter that is the nearest to the reference, and $\underline{v}^*$ exceeds, in magnitude, maximum voltage realisable with one inverter. The reference for the other inverter is then set as:

$$\underline{v}^{**} = -(\underline{v}^{*} - \underline{v}_i) \tag{4}$$

In other words, when the magnitude of the reference voltage exceeds the maximum value obtainable with one inverter, one inverter is operated in six-step mode while the second inverter is modulated in the standard way.

It is well known that operation of a five-phase inverter in ten-step mode without a controllable dc link leads to uncontrollable fundamental output voltage magnitude and unwanted low order harmonics, which for the leg voltage can be expressed as a Fourier series as follows:

$$v(t) = \frac{1}{\pi} V_{dc} [\sin \omega t + \frac{1}{3} \sin 3\omega t + \frac{1}{5} \sin 5\omega t + \frac{1}{7} \sin 7\omega t ...] \tag{5}$$

In a five-phase system, harmonics of the order $10k\pm1$ (k = 0, 1, 2, 3...) map into the torque/flux producing subspace, α-β, while harmonics of the order $10k\pm3$ map into the x-y subspace. They do not produce any useful torque/flux and simply lead to large unwanted loss-producing currents. The large currents are a consequence of the relatively small impedance presented in x-y plane. $10k\pm5$ are zero-sequence components. This leads to the requirement that the second inverter must be able to not only control the fundamental but also eliminate the unwanted low order harmonics, which are produced by applying only the large vector in α–β from one inverter. This causes unwanted harmonics in both planes since a large vector has a corresponding non-zero value in the second, x-y plane [14]. In order to achieve this objective, the second inverter modulation scheme will need to operate in both the α-β and the x-y planes. Such PWM modulators are known as multi-frequency PWM modulators and were first developed in order to control multiphase multi-motor drive systems [15].

The proposed scheme is realised by utilizing two separate two-level modulators. Fig. 3 shows the voltage vectors that are used by the modulators. Inverter 1, which operates in ten-step mode, applies a single vector in one switching period. The vectors are mapped in the points of the outer ring of decagon (labelled as *A*, *B*, *C*, *D*, *E*, *F*, *G*, *H*, *I*, *J* in Fig. 2); these correspond to the large vectors in a two-level modulator (Fig. 3(a)). The modulator for inverter 2, which operates in PWM mode, is able to choose from all 32 voltage vectors (small, large, medium and zero) and select suitable 4 active and one zero vectors in-order to control the fundamental and cancel unwanted harmonics in the output. The multi-frequency PWM is employed for the higher modulation indices ($0.525 < M \leq 1.05$). In the lower modulation index region ($M \leq 0.525$) only one inverter operates, which reverts the system into single-sided two-level SVM operation mode. Fig. 4 shows the voltage reference waveform (trace (c)) of the multi-frequency modulator for modulation index of 1.05. It is obtained by subtracting the ten-step mode inverter phase voltage output (trace (b)) from the total voltage reference (trace (a)) and inverting the signal. It can be noticed that the resulting reference of the PWM modulator is multi-frequency. The non-sinusoidal reference is difficult to achieve using the space-vector modulation technique. The control method developed in the paper adapts the unified PWM technique

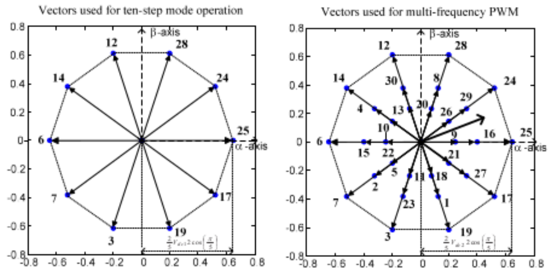

Fig. 3. The voltage space vectors employed for ten-step modulator (a) and the multi-frequency modulator (b).

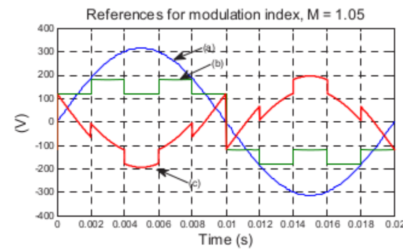

Fig. 4. Generation of the voltage references for multi-frequency modulator.

[16] to function as a multi-frequency PWM modulator, which is capable of controlling the fundamental, eliminating unwanted low-order harmonics and maintaining the maximum dc link utilisation. The technique is well-established for three-phase systems and has been recently extended to the single sided two-level five-phase system, as reported in [17]. Implementation of the unified method is advantageous as the computational load is greatly reduced owing to the removal of the space vector transformation, sector identification and switching state look-up table.

V. Unified Voltage Modulation Technique for Multi-Frequency Output Generation

In the conventional SVM scheme, calculation of dwell time is based on the magnitude of the voltage reference space vector and its relative angle with respect to the reference axis. Space vector transformation, sector identification and information regarding the selected vectors are required to perform the task. In the unified PWM technique the dwell time is determined directly from instantaneous values of the phase voltage reference. This is so since the effective time evaluated in the conventional SVM is just the difference value between two application times corresponding to the phase voltage [16], [17]. The basic principle of the method is shown in Figs. 5 and 6.

In what follows it is assumed that the inverter 2 output is modulated. Regardless of the sector where the reference vector is, the application time for each phase voltage can be defined using:

$$\begin{aligned} v_{as}^{**} + v_{bs}^{**} + v_{cs}^{**} + v_{ds}^{**} + v_{es}^{**} &= 0 \\ T_{as} + T_{bs} + T_{cs} + T_{ds} + T_{es} &= 0 \end{aligned} \tag{6}$$

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$$T_{as} \equiv \frac{T_s}{V_{dc2}} v_{as}^{**},\ T_{bs} \equiv \frac{T_s}{V_{dc2}} v_{bs}^{**},\ T_{cs} \equiv \frac{T_s}{V_{dc2}} v_{cs}^{**},$$
$$T_{ds} \equiv \frac{T_s}{V_{dc2}} v_{ds}^{**},\ T_{es} \equiv \frac{T_s}{V_{dc2}} v_{es}^{**} \tag{7}$$

where T_s is the switching period. The application times, T_{as}, T_{bs}, T_{cs}, T_{ds} and T_{es} which are then named as "imaginary switching times" are proportional to the instantaneous phase voltage reference magnitude. The effective time, T_{eff} is defined as the difference between the maximum and minimum values among T_{as}, T_{bs}, T_{cs}, T_{ds} and T_{es}. The effective times are the time durations of the active vector application. The time duration of zero vector, T_{zero} is obtained by subtracting from T_s the effective time [16]:

$$\begin{aligned} &T_{eff} = T_{\max} - T_{\min} \\ &\Rightarrow T_{\max} = \max\{T_{as}, T_{bs}, T_{cs}, T_{ds}, T_{es}\} \\ &\Rightarrow T_{\min} = \min\{T_{as}, T_{bs}, T_{cs}, T_{ds}, T_{es}\} \\ &T_{zero} = T_s - T_{eff} \end{aligned} \tag{8}$$

In order to obtain a symmetrical switching pattern over switching period, the zero voltage has to be distributed equally. The offset time T_{offset} is required:

$$T_{offset} = (T_{zero}/2) - T_{\min} \tag{9}$$

The actual switching times for each inverter leg can be obtained by the time shifting operation as follows:

$$\begin{aligned} &T_{ga} = T_{as} + T_{offset},\ T_{gb} = T_{bs} + T_{offset} \\ &T_{gc} = T_{cs} + T_{offset},\ T_{gd} = T_{ds} + T_{offset} \\ &T_{ge} = T_{es} + T_{offset} \end{aligned} \tag{10}$$

VI. Simulation Verification

In order to verify the drive's performance a series of simulations were performed. The dc link voltage of each inverter is 300 V ($V_{dc1} = V_{dc2} = V_{dc}/2$), giving an effective dc link voltage of 600 V when M > 0.525. The switching frequency of the modulated inverter is 2 kHz. Ideal operation of the inverters is assumed, meaning that effects due to the dead-time and semiconductor voltage drops were neglected. Fig. 7 shows the load phase voltage and the inverter's leg voltage waveforms and spectra at the boundary between regions where the PWM modulated inverter reduces the fundamental voltage, created by the ten-step inverter, or adds to it (this takes place when M = 0.6366). The converter achieves the reference fundamental with minimum low order harmonics. The leg voltages of each inverter show that the harmonics created by the ten step inverter are counter balanced by the modulated inverter and the fundamental is overwhelmingly provided by the ten step modulator. This means that the modulated inverter is only compensating the unwanted harmonics. In order to examine the drives performance over the complete linear operating range simulations were performed with modulation index of 0.1 – 1.05 in 0.05 increments. The THD of the phase output voltage and the respective α and x-component is shown in Fig. 8. It can be seen that the THD remains low for the entire linear

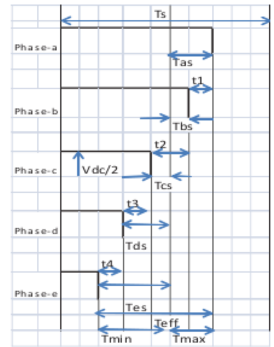

Fig. 5. Illustration of the imaginary switching times.

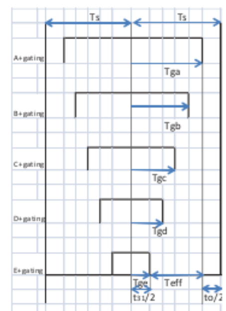

Fig. 6. Actual switching pulse pattern.

operating region. The performance is significantly improved when compared to the single-sided two-level drive in line with the higher number of phase voltage levels available particularly in the higher modulation index.

Next, the phase variable model of the five-phase machine is used and the drive is operated in open-loop V/f mode. The voltage reference profile is such that the supply frequency of the machine is ramped from zero to 50 Hz in 0.5s. At the operating frequency of 50 Hz the maximum modulation index is reached (M = 1.05). Voltage boost is not applied. The acceleration transient is presented in Figs. 9 and 10 for 25 Hz (M = 0.525) and 50 Hz (M = 1.05), respectively along with the steady-state load phase voltage waveform and spectrum. It can be seen in Fig. 9 that no voltage reference is applied to inverter 2 when M = 0.525 and so the drive operates in two-level mode. For this reason the steady-state phase voltage contains only nine levels. However, the spectrum is much improved when compared with its single sided two-level equivalent due to the 50% reduction in the effective dc link voltage. Fig. 10 shows that, as soon as M > 0.525 (at t = 0.25 s) inverter 1 operates in ten-step mode, while inverter 2 operates in multi-frequency PWM mode. At t = 0.5 s the reference frequency is reached and the modulation index is maximum, M = 1.05. The drive is operating in multi-level mode and the steady-state phase voltage now comprises 15 levels. The torque ripple is mininal and the stator current is sinusiodal without any low order harmonics. The load phase voltage spectra indicate absence of unwanted low-order

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harmonics and the target fundamental voltages have been met in both cases.

VII. Conclusion

A multi-frequency PWM scheme for multi-level dual-inverter fed open-end winding drives has been successfully developed based on the unified PWM technique. When $M <$ 0.525, the converter operates in two-level mode utilising a single inverter. When the reference fundamental exceeds the capabilities of a single inverter (M>0.525), one inverter is in ten-step mode and the other is multi-frequency modulated. As

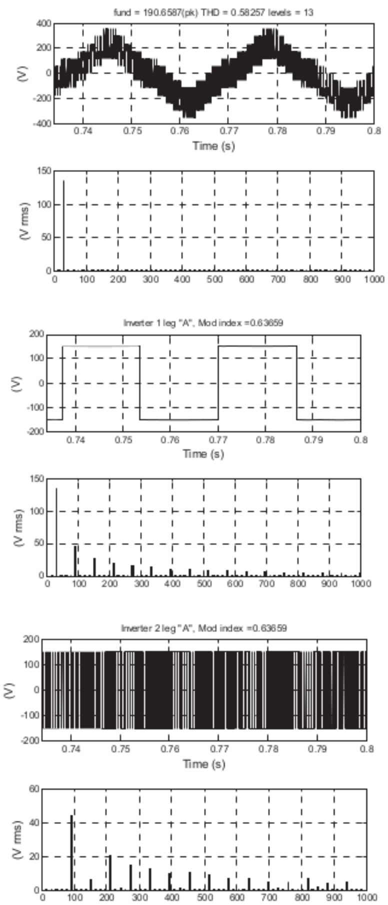

Fig. 7. Stator phase and inverter leg voltage waveforms and spectra for M = 0.636, f = 30.31Hz.

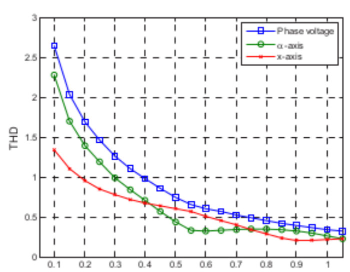

Fig. 8. Total harmonic distortion (THD) of the stator phase voltage and the α-axis and x-axis components.

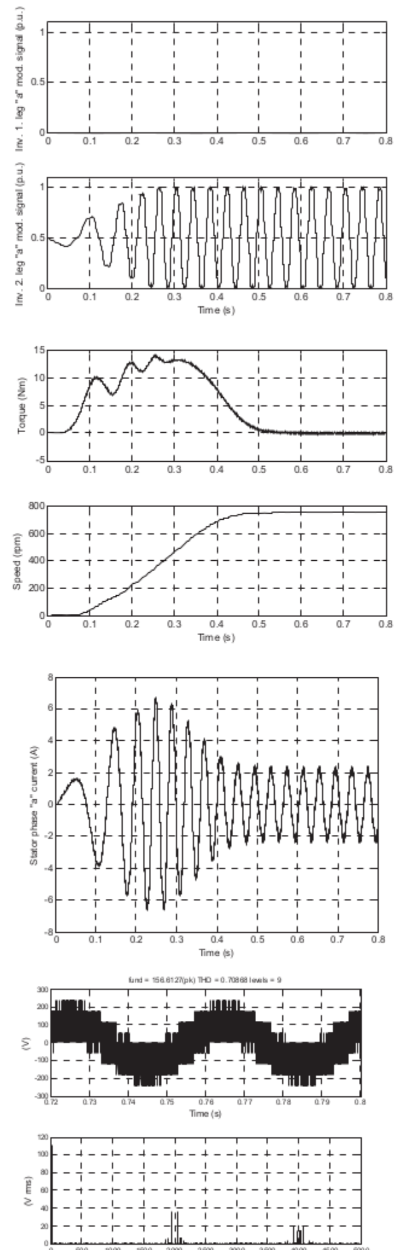

Fig. 9. Motor acceleration under V/f control (ref. 25 Hz, M = 0.525): inverter modulation signals, speed, torque, stator current and steady-state stator phase voltage waveform with spectrum.

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a result the converter is operated in multilevel mode. The multi-frequency modulated inverter is used to control the fundamental voltage and eliminate any low order harmonics created by the ten-step mode inverter. The method has been verified by simulation.

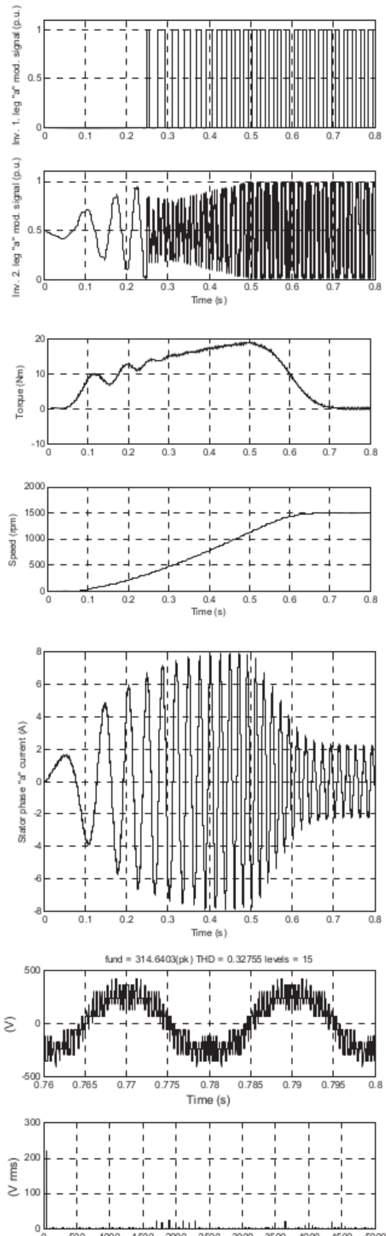

Fig. 10. Motor acceleration under *V/f* control (ref. 50Hz, $M = 1.05$): inverter modulation signals, speed, torque, stator current and steady-state stator phase voltage waveform with spectrum.

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