Harmonic Elimination PWM Technique for Multilevel Inverter

Shatha K. Baqir,* Saad A. Hasan**

Received on: 28/11/2004
Accepted on: 9/6/2005

Abstract

The multilevel voltage source inverters are recently applied in many applications such as ac power supplies and ac motor drives. The concept of multilevel voltage source inverter is explained. The concept of the harmonic elimination pulse-width-modulation (HEPWM) technique based on Walsh function is presented. By using HEPWM technique, specific low order harmonics can be eliminated, a total harmonic content is minimized and the output voltage THD can be improved. A procedure to achieve the appropriate switching angles of HEPWM is proposed. The harmonic amplitudes of the multilevel inverter output voltage can be expressed as functions of switching angles. Thus, solving linear algebraic equations instead of solving nonlinear transcendental equations optimizes the switching angles.

The HEPWM technique is implemented on a multilevel inverter using cascaded-inverter with separated dc sources. Simulation results are presented using ORCAD package version 9.1.

تقنية تضمين عرض النبضة بحذف التوافقيات لعاكس متعدد المستويات

الخلاصة

استخدمت عواكس مصدرالفولتية متعددة المستويات حديثا في اغلب التطبيقات كاستخدامها في مصادر القدرة المتناوبة وفي مسوقات محركات التيار المتناوب. في هذا البحث تم عرض مفهوم عاكس مصدر الفولتية متعدد المستويات وتقنية تضمين عرض النبضة لحذف التوافقيات باستخدام دالة والش.

العاكس المتتالي متعدد المستويات المستخدم في هذا البحث يعمل بتقنية تضمين عرض النبضة لحذف توافقيات الفولتية الخارجة ذات الرتب الواطئة أو تقليل المحتوى التوافقي الكلي باستخدام دالة والش، يمكن مع هذه التقنية التعبير عن قيم التوافقيات للفولتية الخارجة للعاكس كدوال لزوايا التشغيل. لذلك يمكن تحسين حساب قيم زوايا التشغيل بحل معادلات جبرية بدلا من حل المعادلات Transcendental. نفذت تقنية HEPWM لعاكس متتالي متعدد المستويات مع مصادر منفصلة وكما مبين في النتائج باستخدام (ORCAD version 9.1).

Introduction

A. Background

The development of electric and hybrid-electric vehicles will offer many new opportunities and challenges to the power electronics industry, especially in the development of the main traction motor drive [1].Many current and future designs will incorporate the use of induction motors as the primary source for traction in electric vehicles.

Designs for heavy duty trucks and many military combat vehicles that have large electric drives will require advanced power electronic

* Dept., of Technical Education, UOT. , Baghdad-IRAQ.
**Dept., of Electrical Eng., UOT., Baghdad-IRAQ.

https://doi.org/10.30684/etj.24.10A.4
2412-0758/University of Technology-Iraq, Baghdad, Iraq

inverters to meet the high power demands required of them. Development of electric drive trains for these large vehicles will result in increased fuel efficiency, lower emissions, and likely better vehicle performance (acceleration and braking).

Multilevel inverters are uniquely suited for this application because of the high VA ratings possible with these inverters [2-3]. Where generated ac voltage is available as from an alternator or ac generator, a back-to-back diode-clamped converter can convert this source to variable frequency ac voltage for the driven motor. For electric vehicles (EVs),a cascaded full-bridges inverter can be used to drive the traction motor from a set of batteries or fuel cells.

Some traditional 2-level high-frequency pulse width modulation (PWM) inverters for drive systems can have problems associated with their high voltage change rates (dv/dt), which produces a common mode voltage across the motor windings. High frequency switching can exacerbate the problem because of the numerous times this common mode voltage is impressed upon the motor each cycle. PWM controlled inverters also requires an amount of heat removal because of the additional switching losses [4-5].

Multilevel inverters solve these problems because their individual devices have a much lower (dv/dt) per switching, and they operate at high efficiencies because they can switch at a lower frequency than PWM-controlled inverters.

B. Multilevel Inverters (MLIs)

The multilevel voltage source inverters unique structure allows them to reach high voltages with low harmonics without the use of transformers or series connected synchronized switching devices. The general function of the multilevel inverter is to synchronize a desired voltage from several levels of dc voltages. For this reason, multilevel voltage source inverter is recently applied in many industrial applications such as ac power supplies, static VAR compensatory, drives for electric motors, etc [3].

As the number of levels increases, the synchronized output waveform has more steps, which produces a staircase wave that approaches a desired waveform. Also, as more steps are added to the waveform, the harmonic distortion of the output wave decreases, approaching zero as the number of levels increases. As the number of levels increases, the voltage that can be spanned by summing multiple voltage levels also increases. The structure of the multilevel inverter is such that no voltage sharing problem is encountered by the series-connected devices [2].

Three types of multilevel inverters have been proposed by researches thus far: the diode-clamped inverter, the flying-capacitor inverter, and the cascade inverter with separated dc sources [6].

Three, four, and five level inverter drive systems which have used some form of multilevel PWM as a means to control the switching of the inverter section have been investigated in the literature [7-8]. Multilevel PWM has lower (dv/dt) than that experienced in some two-level PWM drives because switching is between several smaller voltage level. However, switching losses and voltage total harmonic distortion (THD) are still relatively high for these proposed schemes; the output voltage (THD) was reported to be (19.7%) for a four-level PWM inverter [9] and (5%) for a 11-level PWM inverter without the use of any filtering components[3].

This paper proposes a multilevel inverter control scheme where devices are switched only at the fundamental frequency and the inverter output line voltage (THD) is general less than (2) percent without the use of any filtering components.

Cascade Inverter with Separated DC Sources

A cascaded multilevel inverter consists of a series of full-bridge (single-phase) inverter units. The general function of this multilevel inverter is to synthesize a desired voltage from several separate dc sources (SDCSs), which may be obtained from batteries, fuel cells, or solar cells [10]. Fig.(1) shows a single-phase structure of a cascade inverter with SDCSs. Each SDCS is connected to a single-phase full-bridge inverter. Each inverter level can generate three different voltage outputs, +Vdc, 0, and –Vdc, by connecting the dc source to the ac output side by different combinations of the four switches, S_1, S_2, S_3, and S_4. To obtain +Vdc, switches S_1 and S_4 are turned on. Turning on switches S_2 and S_3 yields –Vdc. By turning on S_1 and S_2 or S_3 and S_4, the output voltage is 0.

The ac outputs of each the different level full - bridge inverters are connected in series such that the synthesized voltage waveform is the sum of the inverter outputs. The number of output phase voltage levels in a cascade inverter is defined by m=2S+1, where S is the number of dc sources. The phase voltage $V_{an}=V_1+V_2+\text{---}V_s$ [11].

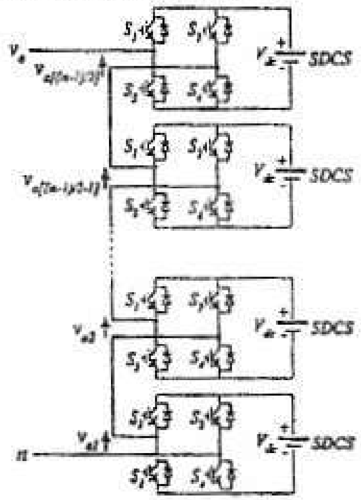

Fig. (1) Single-phase structure of a multilevel cascade full-bridges inverter.

Harmonic Elimination PWM Technique

The Walsh function method of harmonic elimination has been proposed to simplify the process of solving nonlinear harmonic elimination equations on-line and to make the technique more flexible and interactive one [12].

Walsh function for an ordered set of rectangular waveforms takes only two amplitude values, +1 and –1, over one normalized frequency period. The Walsh functions form a complete orthogonal set; hence, Walsh functions can be used to represent signals in the same way as the Fourier series. Since the sinusoidal waveform has quarter-symmetry, it is assumed that the PWM output waveform f (t) have unit amplitude and quarter-wave symmetry.

Using Fourier series expansion the output waveform f (t) can be expressed as follows:

$$f(t)=\sum_{k=1}^{k}A_{2k-1}\sin[(2k-1)\omega_o t]..(1)$$

where

$$A_{2k-1}=\frac{2}{T}\int_0^T f(t)\sin[(2k-1)\omega_o t]dt$$

$$=\frac{8}{T}\int_0^{\pi/2} f(t)\sin[(2k-1)\omega_o t]d(\omega t).....(2)$$

Walsh functions which have only two amplitudes +1 and -1 over one normalized time period can also be used to express the PWM output waveform f(t) as follows:

$$f(t)=\sum_{n=1}^{N}W_{4n-3}\,WAL(4n-3,t).....(3)$$

Where

$$W_{4n-3}=\int_0^1 f(t)\,WAL(4n-3,t)\,dt.....(4)$$

The W_{4n-3} is the Walsh coefficient of the inverter output voltage. Only the (4n-3) order of Walsh functions are used because of quarter-wave symmetry of the inverter output voltage.

By replacing f(t) in eq.(2) with eq.(3) the equation becomes

$$A_{2k-1}=\sum_{n=1}^{N}W_{4n-3}\left[\frac{8}{\pi}\int_0^{\pi/2}WAL(4n-3,\omega_o t)\sin[(2k-1)\omega_o t]d\omega t\right]$$

$$=\sum_{n=1}^{N}B_{2k-1,4n-3}W_{4n-3}\quad (5)$$

where: $B_{2k-1,4n-3}$

$$=\frac{8}{\pi}\int_0^{\pi/2}WAL(4n-3,\omega_o t)\sin[(2k-1)\omega_o t]d\omega t\ ....(6)$$

$B_{2k-1,4n-3}$ is the (2k-1) harmonic coefficient of the (2n-3) Walsh function . The values of $B_{2k-1,4n-3}$ coefficient can be calculated directly from the equation derived Seimens and Kitai [13].

From eq.(5)

$A_1=B_{1,1}W_1+B_{1,5}W_5+B_{1,9}W_9+\ldots$
$+B_{1,4N-3}W_{4N-3}$

$A_3=B_{3,1}W_1+B_{3,5}W_5+B_{3,9}W_9+\ldots$
$+B_{3,4N-3}W_{4N-3}$

. = . . .

. = . . .

$A_{2k-1}=B_{2k-1,1}W_1+B_{2k-1,5}W_5+\ldots$
$+B_{2k-1,4N-3}W_{4N-3}\quad (7)$

Equation (7) in matrix form is

$$\begin{bmatrix}A_1\\A_3\\ \vdots\\A_{2k-1}\end{bmatrix}=\begin{bmatrix}B_{1,1} & B_{1,5},B_{1,9}\ldots B_{1,4N-3}\\B_{3,1} & B_{3,5},B_{3,9}\ldots B_{3,4N-3}\\ \vdots & \vdots\\B_{2k-1,1} & B_{2k-1,5}\ldots B_{2k-1,4N-3}\end{bmatrix}\begin{bmatrix}W_1\\W_5\\ \vdots\\W_{4N-3}\end{bmatrix}\quad (8)$$

By initializing the firing angles α_k the Walsh function coefficients W_{4n-3} of the PWM output waveforms can be calculated from eq.(4). By sampling f(t) with M equidistant points, the integral in eq. (4) is replaced by summation

$$W_{4n-3}=\int_0^1 f(t)\,WAL(4n-3,t)\,dt$$

$$=\int_0^{1/M} f(t)\,WAL(4n-3,t)\,dt+$$

$$\cdots+\int_{1/M}^{2/M} f(t)\,WAL(4n-3,t)\,dt$$

$$+\int_{(M-1)/M}^{1} f(t)\,WAL(4n-3,t)\,dt$$

(9)

M is an integer power of two determined by the highest sequence-ordered component of the Walsh function [14].

The value of M is always to be equal to 4N, where N is given in eq. (3). Since WAL (4n-3, t) is replaced by WAL (4n-3,m). Thus eq. (9) becomes

$$W_{4n-3}=WAL(4n-3,0)\int_0^{1/M} f(t)\,dt$$

$$+ WAL(4n-3,1)\int_{1/M}^{2/M} f(t)dt + \dots\dots$$

$$+ WAL(4n-3,M-1)\int_{(M-1)/M}^{1} f(t)dt$$

$$= \sum_{m=0}^{(M-1)} W(4n-3,m)\left[\int_{m/M}^{(m+1)/M} f(t)dt\right]$$

$$= 4\sum_{m=0}^{(M/4)-1} W(4n-3,m)\left[\int_{m/M}^{(m+1)/M} f(t)dt\right] \quad (10)$$

Although the exact values of switching angles $\alpha_1, \alpha_3, .. \alpha_{2k-1}$ are unknown in each of them must be initialized in sampling intervals m_1, m2, ...m_k. So each switching angle α_k is constrained within the following range

$$\frac{m_k}{M} < \alpha_k < \frac{m_k + 1}{M},$$

for k=0, 1,..., k-1 (11)

W_{4n-3} can be expressed as a function of switching angles $\alpha_1, \alpha_3, .. \alpha_{2k-1}$ [15-16].

Thus

$$W_1 = C_{1,1}\alpha_1 + C_{1,2}\alpha_2 + \dots C_{1,k}\alpha_k + D_1$$

$$W_5 = C_{5,1}\alpha_1 + C_{5,2}\alpha_2 + \dots C_{5,k}\alpha_k + D_5$$

. .

. .

$$W_{4N-3} = C_{4N-3,1}\alpha_1 + C_{4N-3,2}\alpha_2 + \dots C_{4N-3k}\alpha_k + D_{4N-3} \quad (12)$$

where $C_{1,1}$, $C_{1,2}$, ... $C_{4N\text{-}3k}$ are the coefficients of the switching angles and D_1, D_5, ... D_{4N-3} are constants. The matrix form of eq. (12) is

$$\begin{bmatrix} W_1 \\ W_5 \\ \vdots \\ W_{4N-3} \end{bmatrix} = \begin{bmatrix} C_{1,1} & C_{1,2} & C_{1,3} & \cdots & C_{1,k} \\ C_{5,1} & C_{5,2} & C_{5,3} & \cdots & C_{5,k} \\ \vdots \\ C_{4N-3,1} & & C_{4N-3,2} & \cdots & C_{4N-3,k} \end{bmatrix} * \begin{bmatrix} \alpha_1 \\ \alpha_2 \\ \vdots \\ \alpha_k \end{bmatrix} + \begin{bmatrix} D_1 \\ D_5 \\ \vdots \\ D_{4N-3} \end{bmatrix} \quad (13)$$

From eq. (8) and (13)

$$[A] = [B][C][\alpha] + [B][D]$$
$$= [E][\alpha] + [F] \quad (14)$$

When k switching angles per quarter period, the degree of freedom in eq.(14) is k. It is desirable to use one degree of freedom control the fundamental amplitude, and (k-1) degrees of freedom are used to eliminate (k-1) unwanted harmonics. By setting the selected (k-1) harmonic amplitudes equal to zero eq.(14) becomes

$$A_1 = E_{1,1}\alpha_1 + E_{1,2}\alpha_2 + . + E_{1,k}\alpha_k + F_1$$

$$0 = E_{2,1}\alpha_1 + E_{2,2}\alpha_2 + \dots + E_{2,k}\alpha_k + F_2$$

. .

. .

$$0 = E_{k,1}\alpha_1 + E_{k,2}\alpha_2 + \dots + E_{k,k}\alpha_k + F_k \quad (15)$$

Eq.(15) can be used to analyze a quarter period of PWM waveform. By m_k sampling interval. The E and F can be computed and eq.(15) is solved to compute the switching angles with the desires fundamental amplitudes under the constraints of the initial conditions.

The switching angles can be expressed as first-order polynomial of fundamental amplitude A_1 so:

$$\alpha_k = p_k A_1 + q_k \quad \text{(degree)}$$
$$= (p_k A_1 + q_k)/ 360\, F_1 \text{ (sec)} \quad (16)$$

where A_1 and F_1 indicate the fundamental frequency and fundamental amplitude, p_k and q_k are polynomial coefficients for the switching angle α_k .

The coefficients p_k and q_k are determined by off-line via the Walsh function harmonic method and straight- line curve-fitting method [12]. Table (1) in appendix (a) shows the switching angles as linear functions of fundamental amplitude in the first quarter-period. Fig.(2-a) shows the linear relationship between switching angles and fundamental amplitude. The output waveform of an 33-level cascaded inverter is shown in Fig.(2-b).

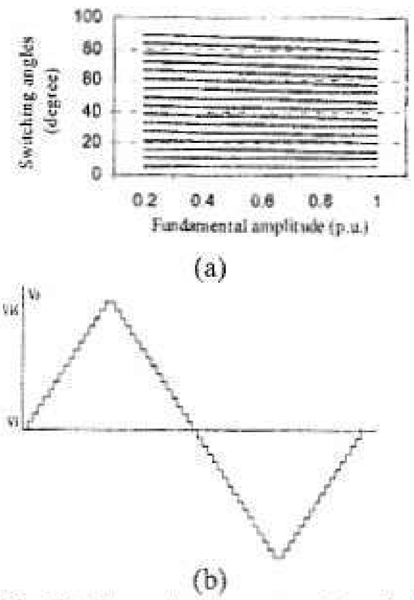

Fig.(2) Cascaded inverter (a) solution trajectories (b) output waveform.

Simulation Results

To design a simulation system for cascaded inverter with separate dc sources (batteries) and used to drive an induction motor, ORCAD program is employed as a simulation tool. A piecewise linear (PWL) voltage source is used to represent a full-bridge inverter. The schematic of 33-level HEPWM inverter is illustrated in appendix (A). Sixteen 12 PWL voltage sources are used to generated (136 V_{rms}-400Hz) output phase voltage. Thus, the total number of PWL for three-phase system is (48). A (10Ω) resistor and a (2.5 mH) inductor, which connected in series, are used as the load per phase in inverter circuit. The simulated fundamental amplitude (A_1=1.00, and 0.4). The simulation results will be shown in Appendix (B).

Evaluation of HEPWM to Eliminate Harmonics

There are four performances used in this paper to evaluate of HEPWM to eliminate harmonics are defined as follows.

Harmonic Factor (HF)

The harmonic factor (of the nth harmonic), which is a measure of individual harmonic contribution, is defined as

$$HF_n = \sqrt{\left[\left(\frac{V_n}{V_1}\right)^2 - 1\right]} \qquad (17)$$

where

V_1 fundamental voltage (rms)

V_n harmonic voltage (rms)

n order of harmonics.

Harmonic Loss Factor (HLF)

The harmonic loss factor, which is proportional to total rms harmonic current, is defined as

$$HLF = \frac{1}{V_1}\sqrt{\sum_{n=5}^{n=125}\left[\frac{V_n}{n}\right]^2} \qquad (18)$$

Total Harmonic Distortion Factor (THD)

The total harmonic, which is a measure of closeness in shape between a waveform and its fundamental voltage, is defined as

$$THD = \frac{1}{V_1}\sqrt{\sum_{n=5}^{n=125}(V_n)^2} \qquad (19)$$

Distortion Factor (DF)

The distortion factor indicates the amount of harmonic distortion that remains in a particular waveform after the harmonics of that waveform have been subjected to a second-order attenuation (i.e., divided by n^2). Thus DF is a measure of effectiveness in reducing unwanted harmonics without having to specify the values of a second-order load filter and is defined as

$$DF = \frac{1}{V_1}\sqrt{\sum_{n=5}^{n=125}\left[\frac{V_n}{n^2}\right]^2} \qquad (20)$$

[17].

It is further noted here that the various performance factors HF, HLF, THD and DF shown in appendix(C) have been plotted versus the fundamental amplitude A_1. Fig.(6) illustrate that performances factors decrease progressively when the fundamental voltage amplitude A_1 is increased.

The voltage harmonic spectra for HEPWM of operation are given in appendix (D), with number of harmonics to be eliminated (15). For example, eliminating of (15) harmonics requires (16)-switching angles per quarter cycle of the output voltage waveform. This figure shows that, increasing of switching angles will cause increase in the number of low order eliminated harmonics, which causes to push more harmonic energy into high frequency regions; therefore low frequency harmonics are well attenuated .

Conclusions

1. A multilevel cascade inverter with separate dc source has been proposed because it can use the batteries or fuel cell to synthesize a sinusoidal output voltage waveform to drive an induction motor.
2. This system has higher efficiency because the minimum switching frequency.
3. No voltage sharing problems exit for series connected devices unlike traditional inverters.
4. The HEPWM technique has been used to minimize the harmonic content of the inverter output voltage. In this technique the fundamental amplitude can be expressed by a set of linear equations instead of non-linear equations by using linear functions such as (WALSH) function to obtain the switching angles on-line resulting in eliminating the unwanted harmonics, to get nearly sinusoidal output be small compared with off- line solution, and fewer computations are required for the hardware implementat-ion.
5. The high number of harmonics to eliminate is (15). It is very practical to eliminate the harmonics up to the (47^{st}) so that the lowest harmonic component, which may appear in the inverter output waveform, is the (49^{rd}). Then, we find that it not needed to use of any filtering components.

References

[1] Miller, J.M. and Gale, A.R., "Hybrid-electric Vehicle Success

Will Depend On Low Cost, Efficient Power Electronics Systems", Power Conversion & Intelligent Motion (PCIM), Nov., 1997.

[2] Tolbert, L.M., Peng, F.Z., Adams, D.J. and Mckeever, J.W., "Multilevel Inverters For Large Automotive Electric Drives", All Electric Vehicle Second International Conf., June 8-12, 1997.

[3] Tolbert, L.M., Peng, F.Z. and Habetler, G.,"Multilevel Converters For Large Electric Drives", IEEE Trans. Ind. Applicat., Vol.35, No.1, Jan. /Feb., 1999.

[4] Bell, S. and Sung, J.,"Will your Motor Insulation Survive A New Adjustable Frequency Drive?" IEEE Trans. Ind. Applicat., Vol.33, No.5, Sep., 1997.

[5] Bonnett, A.H.,"A Comparison Between Insulation Systems Available for PWM-inverter-fed motors", IEEE Trans. Ind. Applicat., Vol.33, No.5, Sep., 1997.

[6] Tolbert, L.M., Peng, F.Z. and Habetler, G., "Multilevel Inverters For Electric Vehicle Applications", All Electric Vehicle second international conf., Oct.22-23, 1998.

[7] Steinke, J.K., "Control Strategy For A Three Phase Ac Traction Drive With Three Levels GTO PWM inverter", PESC, 1988.

[8] Menzies, R.W., Steimer, P. and Steinke, J.K.,"Five-level GTO Inverters For Large Induction Motor Drives", IEEE Trans. Ind. Applicat., Vol.30, No.4, July, 1994.

[9] Sinha, G. and Lipo, T.A., "A Four Level Inverter System for Drive Applications", Conf. Rec., 1996.

[10] Rashid, M.H., "Power Electronic: Circuits, Devices And Applications", Prentice-Hall, India, 2004.

[11] Peng, F.Z., Lai, J.S. Mckeever, J.W. and Coevering J.V., "A Multilevel Voltage-Source Inverter With Separate Dc Sources For Static VAR Generation", IEEE Trans. Ind. Applicat., Vol.32, Sept. /oct., 1996.

[12] Liang, T.J., Connell ,M.O. and Hoft,G. "Inverter Harmonic Reduction Using Walsh Function Harmonic Elimination Method", IEEE Trans. Power Elec., Vol.12, No.6, Nov., 1997.

[13] Siemens, K.H. and Kital, R., "A Nonrrcursive Equation For The Fourier Transform Of The Walsh Function", IEEE Trans., Elec., Compat., Vol. EMC-15, No.2, 1973.

[14] Chen, B.D and Sun,Y.Y, "Waveform Synthesis Via Inverse Walsh Transform", Int. J. Elec., Vol. 48, No.3, 1980.

[15] Asumadu, J.A. and Hoft, R.G., "Microprocessor-based Sinusoidal Waveform Synthesis Using Walsh And Related Orthogonal Functions", IEEE Trans. Power Elec., Vo.14, No.2, Nov., 1989.

[16] Swift, F. and Kamberis, A., "A new Walsh Domain Technique Of Harmonic Elimination And Voltage Control In PWM Inverters", IEEE Trans.Power Elec., Vo.8, No.2, Nov., 1993.

[17] Krein, P.T., "Elements Of Power Electronics", Oxford University, New York, 1998.

Appendix (A)

Table (1) The switching angles (in degrees) as linear functions of fundamental amplitude

Switching angles (in degrees)	The linear algebraic equation
α_1	$-0.5488A_1 + 5.6517$
α_2	$-0.8582A_1 + 11.1909$
α_3	$-1.7633A_1 + 16.9314$
α_4	$-1.5543A_1 + 22.3165$
α_5	$-2.7541A1 + 28.1630$
α_6	$-2.3713A_1 + 33.5305$
α_7	$-3.7300A_1 + 39.4324$
α_8	$-3.1799A_1 + 44.7839$
α_9	$-4.4979A_1 + 50.6610$
α_{10}	$-3.8498A_1 + 56.0344$
α_{11}	$-5.1594A_1 + 61.9069$
α_{12}	$-4.4522A_1 + 67.3360$
α_{13}	$-5.5610A_1 + 73.1413$
α_{14}	$-4.9529A_1 + 78.7006$
α_{15}	$-5.6306A_1 + 84.6396$
α_{16}	$-6.1158A_1 + 90.9664$

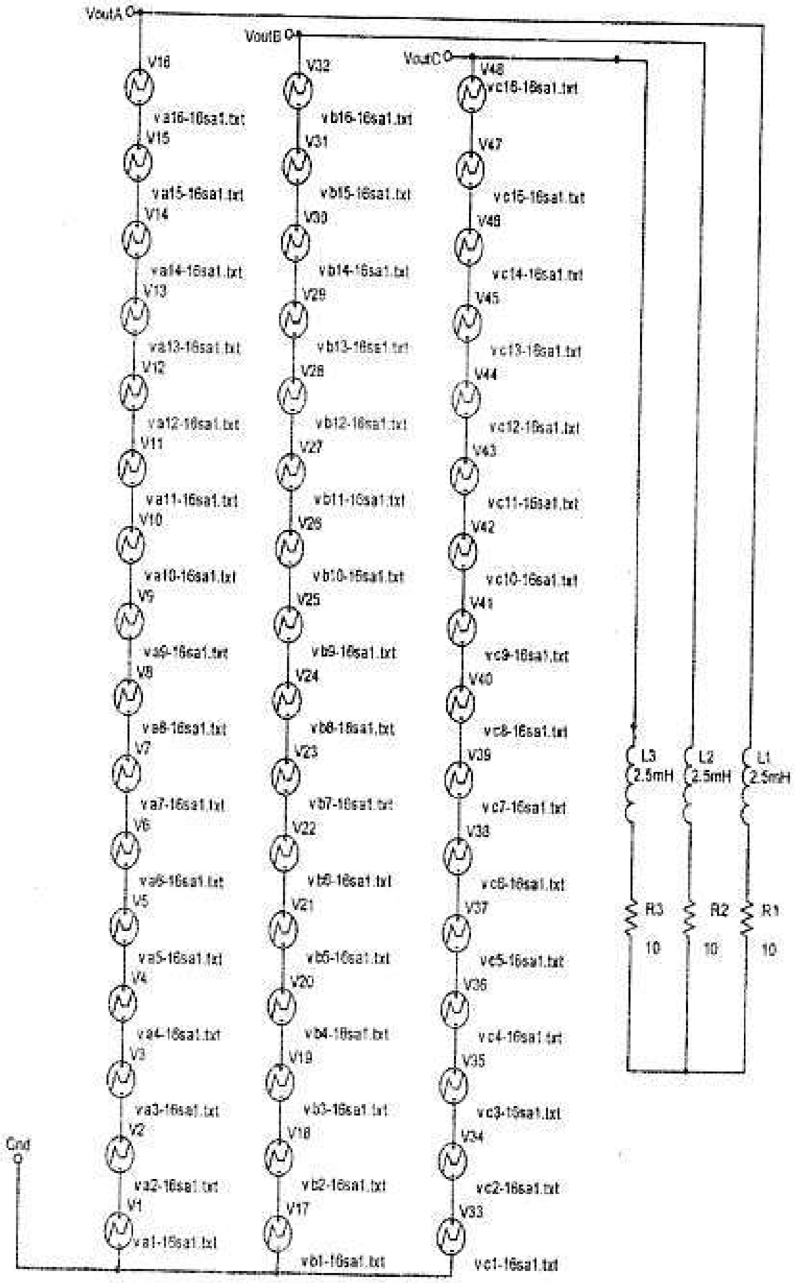

Fig.(3) A schematic of a three-phase 33-level cascade inverter

Appendix (B)

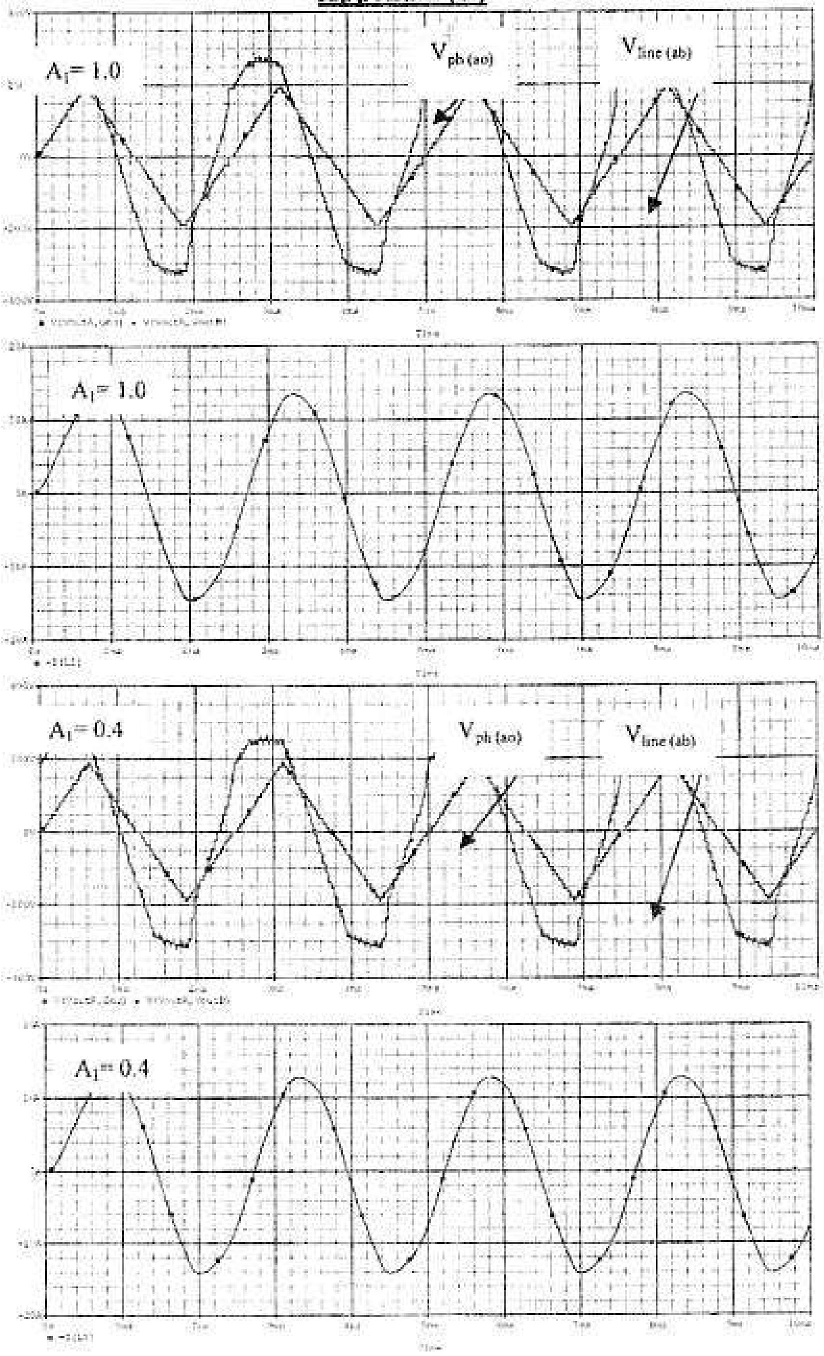

Fig.(4) Phase voltage, line voltage, and load current

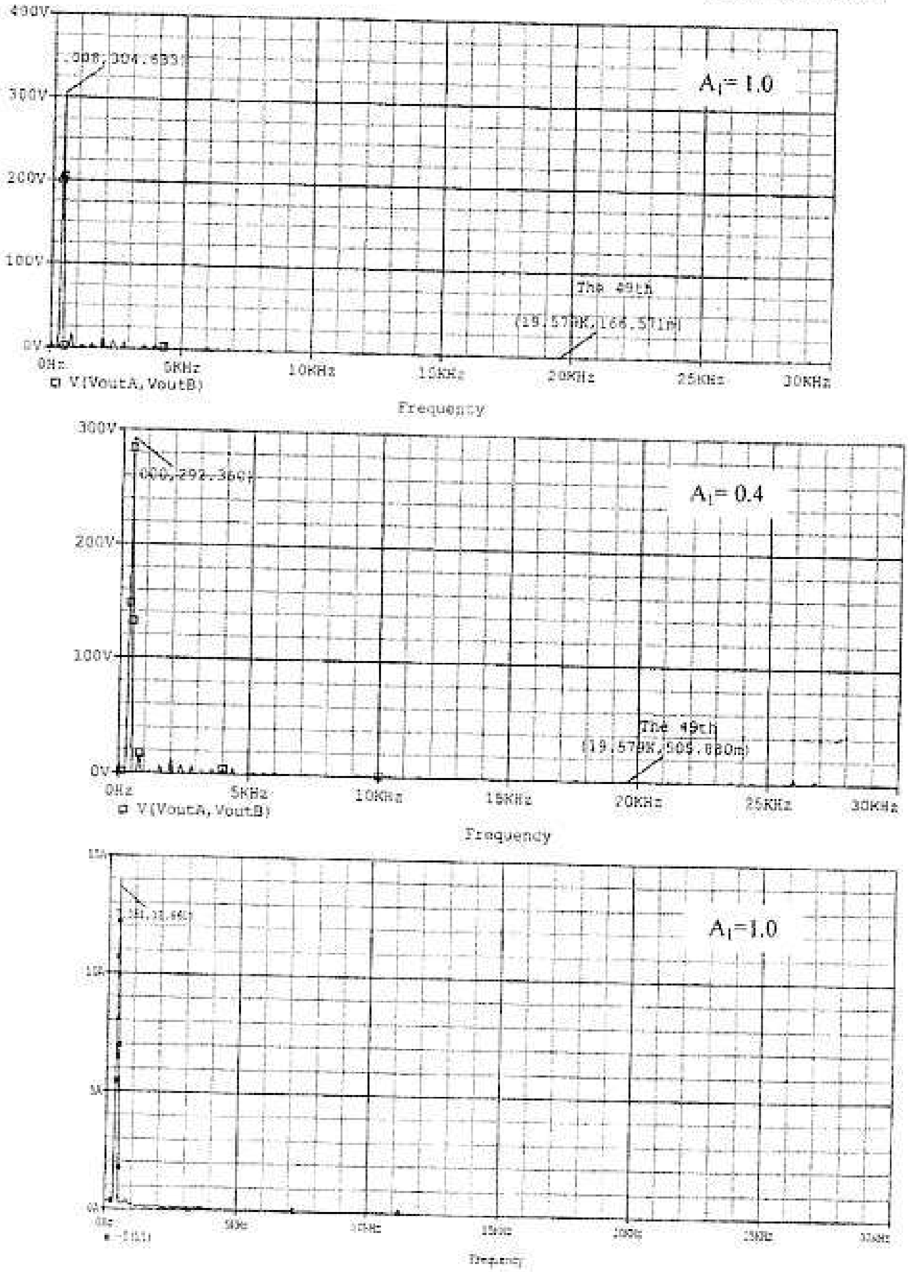

Fig.(5) Frequency spectrum of line voltage and load current in detail

Appendix (C)

A_1	HF%	HLF%	THD%	DF%
0.2	5.2137	7.4285e-02	2.0797	1.1144e-03
0.4	4.9742	7.1386e-02	1.6378	1.0736e-03
0.6	4.6602	6.2576e-02	1.6227	0.9493e-03
0.8	4.0979	5.2898e-02	1.3221	0.7557e-03
1.0	3.5283	5.0482e-02	1.0478	0.7209e-03

Table (2) the value of performance factors (HF, HLF, THD, DF)

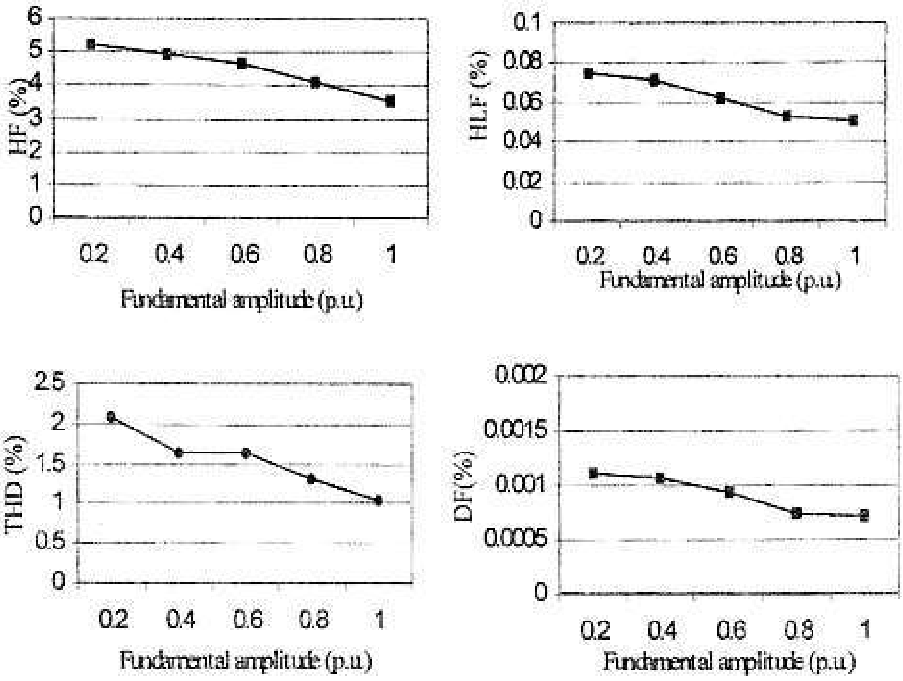

Fig.(6) Performance factors as function of the fundamental amplitude in (p.u.) of 16-level HEPWM

Appendix (D)

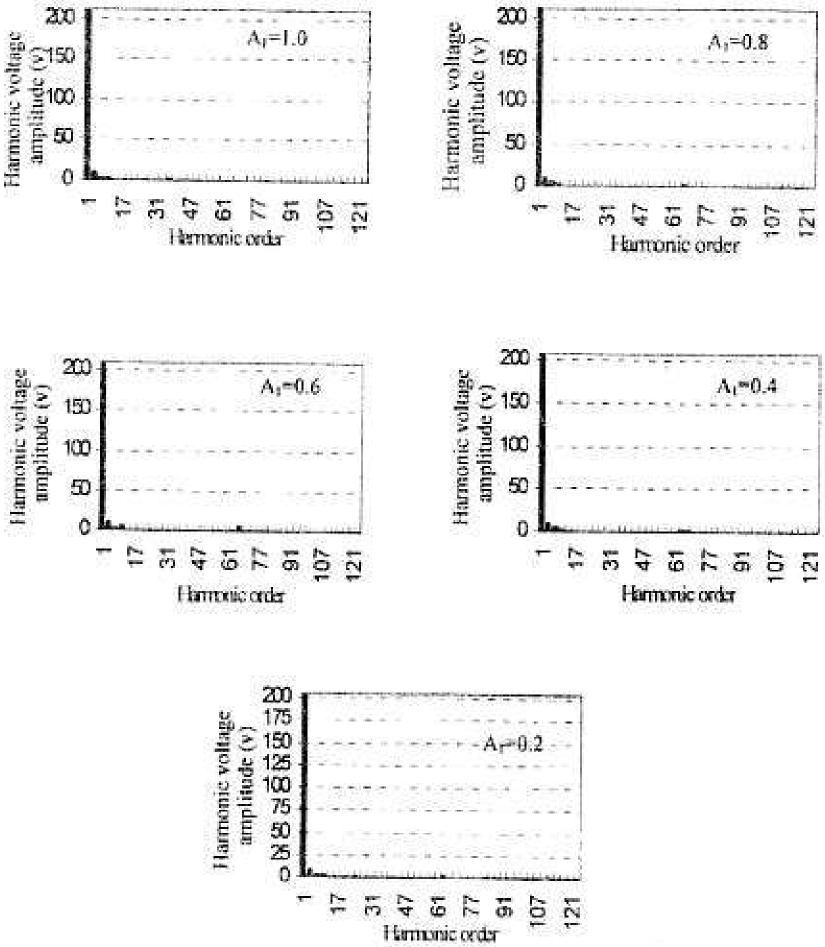

Fig. (7) Harmonic line voltage amplitude (v) as function of the harmonic order of 16-level HEPWM