DIGITAL SCALAR PULSE WIDTH MODULATION METHODS FOR VOLTAGE SOURCE INVERTER FED AC DRIVE

M.NAYEEMUDDIN[1], DR.C.SRINIVASA RAO[2]
[1]Assistant Professor, E.E.E Department, G.Pulla Reddy Engineering College (Autonomous), Kurnool, Andhra Pradesh, India
[2]Professor, EEE Department, G.Pullaiah College of Engineering & Technology, Kurnool, Andhra Pradesh, India

ABSTRACT

As there is increase in the variable AC supply in the industrial drives resulting need for a controllable AC device, and is obtained from an inverter using various controlling techniques. A technique known as space vector PWM (SVPWM) which is computationally intensive & efficient method. It is proved that SVPWM will reduce total harmonic distortion than conventional sinusoidal PWM. The proposed digital scalar PWM (DSPWM) gathers the characteristics of simplicity of implementation found in the regular sampling with the flexibility of manipulation of switching patterns in the SVPWM. This paper shows how to make DSPWM equivalent to the SVPWM without losing the simplicity of implementation. In the proposed method, the introduction of the "distribution ratio" allows the development of a systematic approach of either conventional or any modified vector strategies with changing the modulator scheme. This corresponds to generate any attractive non sinusoidal modulating signals in the carrier based modulation techniques. Simulated results demonstrate the validity of the proposed methods.

Keywords *- CSVPWM, DSPWM, NSMS, V/f control.*

I. INTRODUCTION

Major improvements in modern industrial processes over the past 50 years can be largely attributed to advancements in variable speed drives. Prior to the 1950's most factories used DC motors because 3-phase induction motors could only be operated at one frequency. Now thanks to advancements in power electronic devices, reliable and cost effective control of induction motor. The increase in the use of induction motors was largely attributed to major industries converting existing diesel engines to run off electricity.

The area of AC motor control has continued to expand because induction motors are excellent candidates for use in electrical traction. The classical sine-triangle modulation, or natural sampling modulation (NSPWM), compares a high frequency triangular carrier with three reference signals, known as modulating signals, to create gating pulses for the switches of the power converter.

However Space vector based PWM (conventional) results in better dc bus utilization [1-5]. Apart from this the current ripple in steady state operation and also harmonic distortion will be reduced when conventional SVPWM is employed.

The alternative "digital scalar pulse-width modulation" (DSPWM) technique imposes, to the pole voltage of an inverter leg, an average value that corresponds to each reference phase within the sampling interval. Such strategy is of simpler implementation than the SVPWM technique, reducing the effort of calculation. The technique introduced in has a similar treatment by using the concept of reallocation of the "effective time." This "effective time" is in fact the sum of the times of application of the active vectors. The pulse-widths are ordered and the sum of times is appropriately moved within the sampling period.

This paper establishes a correlation among the SVPWM and DSPWM techniques. It also shows how to make the DSPWM strategy equivalent to the SVPWM technique without losing its simplicity of implementation. The introduction of the "distribution ratio" in this technique, allows the development of a systematic approach for implementation of either conventional or any modified vector strategies without changing the modulator scheme. This corresponds to generate any attractive NSMS in the carrier-based modulation techniques. Simulated results demonstrate the validity of the proposed methods.

II. SPACE VECTOR PULSE WIDTH MODULATION:

In recent years voltage source inverter (VSI) is widely used to generate variable voltage, variable frequency, 3-phase ac required for variable speed ac drive. The ac voltage is defined by two characters, namely amplitude and frequency. Hence it is essential to attain control over both of these quantities. Pulse width modulation (PWM) controls the average output voltage over a sufficiently small period called sampling period, by producing pulse of variable duty-cycle [6-7].

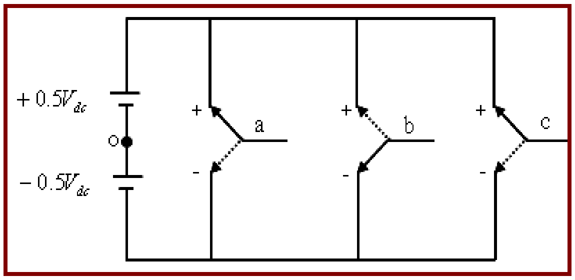

Fig 1 Three phase voltage source inverter

In fig 3.3 the a-phase and c- phase are shown connected to positive dc bus, while b- phase is connected to the negative dc bus. This state can be designed as(+-+)or 101. Thus every phase of 3-phase VSI can be connected either to the positive or the negative dc bus. Hence, 3-phase together can have 2^3 or 8 combinations of switching stages. To designate the switching state code numbers 0 to 7 are used and these switching states --- (V_0) and +++ (V_7), all three poles are connected to the same dc bus, effectively shorting the induction motor and resulting in no transfer of power between the dc bus and induction motor. These two states are called 'Zero voltage vectors' or Zero states'. In case of the other switching states, power gets transferred between the dc bus and induction motor. These states (1 to 6) are called 'active voltage vectors' or 'active states'.

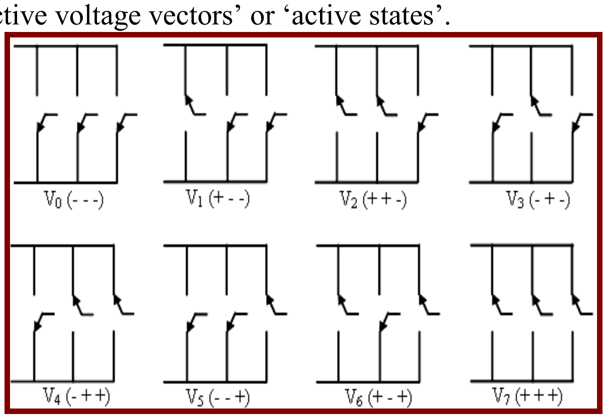

Fig. 2 possible switching states of the inverter

Figure 1 shows the voltage vectors produced by a three-phase, two-level inverter, with six active voltage vectors dividing the space vector plane into six sectors. In CSVPWM algorithm, the desired reference voltage vector is generated by time averaging the suitable discrete voltage vectors in every sub cycle or sampling time period T_s. For example, the reference vector in sector I, as shown in figure 3, is generated by applying the active state 1, active state 2 and the zero states 0 and 7 for durations T1,T2 and TZ respectively [1].

For a given reference voltage vector and Ts, the duration T1, T2 and TZ are unique. The expressions for the various active state time durations and zero states time duration in the first sector can be given as in formulae(1)-(3).

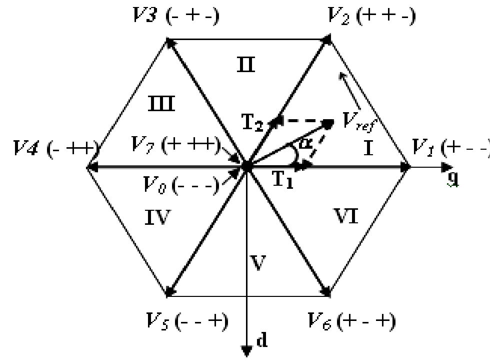

Fig.3: Voltage space vectors for SVPWM

$$T_1 = \frac{2\sqrt{3}}{\pi} M_i \left(\sin(60^o - \alpha)\right) T_s \qquad (1)$$

$$T_2 = \frac{2\sqrt{3}}{\pi} M_i (\sin \alpha) T_s \qquad (2)$$

$$T_Z = T_s - T_1 - T_2 \qquad (3)$$

$$where\ M_i = \frac{\pi V_{ref}}{2V_{dc}}$$

$$V_k = \frac{2}{3} V_{dc} e^{j(k-1)\frac{\pi}{3}} \qquad \text{where } k = 1,2,\ldots,6$$

Different switching sequences can be obtained by using conventional space vector approach as given in (1)-(3), which uses the reference voltage vector and angle information. This basic CSVPWM is the base for further developments of SVPWM algorithms. Many researchers are striving hard and continuing their efforts to improve drive performance with the advancement in technological developments.

In the view of harmonic analysis of various PWM algorithms, simple expressions have been derived in [8-9] by using the notion of current ripple. But the analysis is carried with the conventional space vector approach which is complex due to the calculation of sector and angle information. To overcome this complexity a unified PWM algorithm has been developed in [10-12] using the concept of effective times. Different simplified approaches for the real time implementation have been discussed in [13].

Modulation algorithm:

Step 1: read three phase reference voltages (V_a, V_b, V_c)

Step 2: obtain three –phase to two phase transformation (a,b,c $\rightarrow$ d,q)

Step 3: calculate obsolete value of V_d, V_q & tan $^{-1}(\frac{V_d}{V_q})$

Step 4: identify the sector in which the reference voltage vector lies.

Step 5: select the switching vectors corresponding to the identified sector.

Step 6: the switching times are calculated depending on the output voltage vector magnitude

III. DIGITAL SCALAR MODULATION:

It is possible to impose, to each leg voltage pole of the inverter, an average voltage corresponding to its reference phase voltage during the sampling interval. Modified voltage references and can be defined from the three-phase sinusoidal reference voltages as follows:

$$V^1_{ref_j} = V_{ref_j} + V_h \quad j=a,b,c \qquad (4)$$

Where V_h=zero -sequence component

In the analysis that follows, voltages V_{refi} , and are considered as constants in the interval T_s . As a first step lets make the average values of these voltages equal to the average values of the midpoint voltages V_{ao}, V_{bo} & V_{co} This leads to

$$\frac{1}{T_s}\int_0^{T_s} V^1_{ref_j}(t)dt = \frac{1}{T_s}\int_0^{T_s} V_{jo}(t)dt \qquad (5)$$

and consequently to

$$V^1_{ref_j} = \left[\frac{V_{dc}}{2}\times T_j - \frac{V_{dc}}{2}\times(T_s - T_j)\right]\times\frac{1}{T_s} \qquad (6)$$

Since $V^1_{ref_j}$ are assumed to be constant over T_s. From expression (6) it is possible to calculate the time intervals, that is

$$T_j = \left(\frac{V^1_{ref_j}}{V_{dc}} + \frac{1}{2}\right)\times T_s \qquad (7)$$

In equation (7)T_1,T_2,and T_3 are the time intervals in which switches q_1, $q_2$and $q_3$are closed. respectively, Although the average voltages obtained by this approach refer to the point 0 , it can be shown that the average phase voltages equal the reference voltages respectively, for symmetrical loads. Using the conservative three-phase to two-phase transformation, the equation (7) can be expressed in terms of d-q components as

$$T_1 = \left(\sqrt{\frac{2}{3}} * \frac{V_d}{V_{dc}} + \frac{1}{2}\right) * T_s$$

$$T_2 = \left(\frac{-1}{\sqrt{6}} * \frac{V_d - \sqrt{3}V_q}{V_{dc}} + \frac{1}{2}\right) * T_s$$

$$T_3 = \left(\frac{-1}{\sqrt{6}} * \frac{V_d + \sqrt{3}V_q}{V_{dc}} + \frac{1}{2}\right) * T$$

- (8), (9), (10)

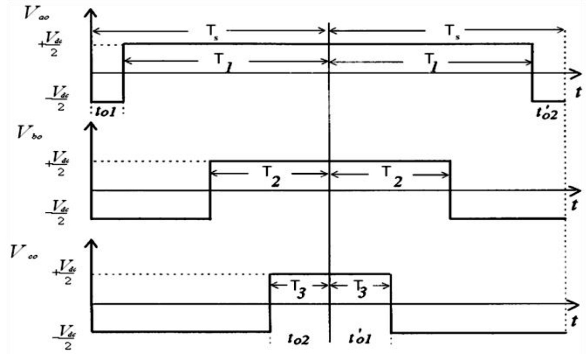

Fig.4 three phase voltages generated with DSPWM

The time intervals $t_1$and t_2 corresponding to the vectors V_1&V_2 (Sector I), respectively, are given by

$$t_1 = T_1 - T_2$$
$$t_2 = T_2 - T_3 \qquad (11), (12)$$

From expressions (3) and (4), these time intervals can be calculated, for k=1 and l=2 as

$$t_1 = \frac{1}{\sqrt{2}} * \frac{T_s}{V_{dc}} * \left(\sqrt{3} * V_d - V_q\right)$$
$$t_2 = \sqrt{2} * \frac{T_s}{V_{dc}} * V_q \qquad (13), (14)$$

Tests that associate vector and scalar approaches are synthesized in below Table.

Modulation	
Vector	Scalar
Sector I - (V_1 $-$ V_2)	$\tau_1 > \tau_2 > \tau_3$
Sector II - (V_2 $-$ V_3)	$\tau_2 > \tau_1 > \tau_3$
Sector III - (V_3 $-$ V_4)	$\tau_2 > \tau_3 > \tau_1$
Sector IV - (V_4 $-$ V_5)	$\tau_3 > \tau_2 > \tau_1$
Sector V - (V_5 $-$ V_6)	$\tau_3 > \tau_1 > \tau_2$
Sector VI - (V_6 $-$ V_1)	$\tau_1 > \tau_3 > \tau_2$

A similar result is obtained when t_1 and t_2 are calculated from (13) and (14), with T_1 and T_2 given by (8)–(10). This fact demonstrates that the inverter controlled by either SVPWM or DSPWM methods provides the same average voltage as applied to the load [13].

The expressions in (13) and (14) may be generalized by ordering the values computed using equations (8)–(10), i.e., by introducing the concept of maximum, T_{max}, minimum , T_{min} , and intermediate, pulse-widths values. In the case of Fig. 4.4, the ordering is$T_{max} = T_1$, $T_{min} = T_3$ and $T_i = T_2$. From this, it can determined that

$$t_k = T_{max} - T_i$$
$$t_l = t_i - T_{min} \qquad (15), (16)$$

From (15), (16) one can also obtain the following expression for the free-wheeling time interval

$$t_0 = T_s - T_{max} + T_{min} \quad (17)$$

Considering the first cycle of Fig. 4 there are two freewheeling time intervals $t_{01} = T_s - T_1$ and $t_{02} = T_3$, at the beginning and at the end of the interval, respectively. The free-wheeling time intervals of the DSPWM are always determined by

$$t_{01} = T_s - T_{max}$$
$$t_{02} = T_{min} \quad (18), (19)$$

In general, these free-wheeling time intervals are different. A complete equivalence between SVPWM and DSPWM is achieved only if $t_{01} = t_{oi}$ and $t_{02} = t_{0f}$. To achieve such condition, a time interval T_h is added to each of T_1 , T_2 andT_3 . This action is equivalent to compute (8)–(10) with

$$V_h = T_{h*} V_{dc}/T_s \quad -(20)$$

The term is modified to be $T^1_{max} = T_{max} + T_h$, when V_h is added to the reference voltages. By using (22), one can obtain T_h as given by

$$T_h = T_S - t_{01} - T_{max} \quad (21)$$

Similarly $T^1_{min} = T_{min} + T_h$ is modified to be . By using (18), one can find out that T_h is

$$T_h = t_{02} - T_{min} \quad (22)$$

Since the time intervals T_h computed by (21) and (26) must be equal, it can be written that

$$t_{01} + t_{02} = T_s - T_{max} + T_{min} \quad (23)$$

Taking into account the fact that $t_{01} + t_{02} = t_0 = t_{oi} + t_{of}$, the equivalence is completely established. The most usual case corresponds to $t_{0i} = t_{0f} = t_0/2$ and uses (21) or (22) to calculate T_h.

3.1 NON-SINUSOIDAL WAVEFORMS:

It has been seen from Fig 4 that t_1 and t_2 do not change if a zero-sequence component, V_h, is added to each reference signal within the sampling interval. In contrast, the values t_{oi} of and t_{of} change to t^1_{01} and t^1_{02}. This change in the distribution of the null-vectors within the sampling interval, which can be represented in terms of the parameter μ.The total time interval for application of the zero-sequence components, as a function of μ, is given

$$t01 = \mu(T_S - T_{max} + T_{min})$$
$$t02 = (1-\mu) \times (T_S - T_{max} + T_{min}) \quad -(24), (25)$$

Therefore, one can say that the distribution of the null-vectors in the DSPWM plays the same role as the non sinusoidal modulating signals in the carrier based modulation technique, within the same sampling interval. This is beneficial since it is already known that the use of certain NSMS permits to increase the switching frequency of the inverter. The use of NSMS also improves the performance of the modulator in terms of the total harmonic distortion when the modulation depth increases [14].

Different types of NSMS are obtained depending on how the null-vectors are distributed within the sampling interval, that is, as a function of the parameter μ. Fig. 5 shows one cycle of some of the well-known distorted modulating signals for different values of the μ parameter. Fig. 4.6(a) shows waveform for constant μ and Fig. 4.6(b) shows waveforms for variable . It should be noted that when one of the phases is clamped the number of commutations per cycle of the power switches is reduced.

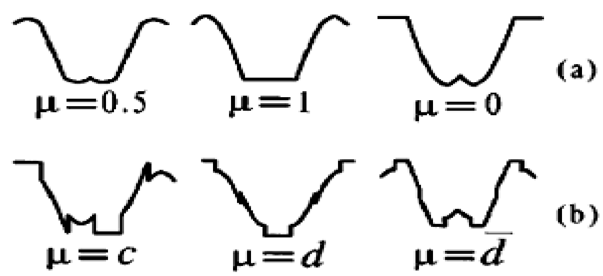

Fig 5 non sinusoidal modulating signals (a): constant μ, (b) variable μ

3.2 IMPLEMENTING NSMS:

The distribution ratio μ can be defined in terms of logic signals originated from the comparison among the calculated pulse-widths in the DSPWM technique [14].

The six intervals 1 to 6 define the ordered Voltage Segments (OVS) which are ordered in terms of their magnitude. It can be noticed that each OVS has a characteristic switching pattern. As shown in that figure, a_1,a_2 and a_3 signals are obtained from the following comparisons.

$$a_1 = 1 \ \ if T_1 >= T_2 else a_1 = 0$$
$$a_2 = 1 \ \ if T_2 >= T_3 else a_2 = 0$$
$$a_3 = 1 \ \ if T_3 >= T_1 else a_3 = 0$$

From these signals one can obtain $c = a_1 \oplus a_2 \oplus a_3$

Where $\oplus$ is the exclusive OR operator.

Signals b_j result from comparisons as given below

$$if\ T_1 >= \frac{T_s}{2}, b_1 = 1 \ \ else\ b_1 = 0$$
$$if\ T_2 >= \frac{T_s}{2}, b_2 = 1 \ \ else\ b_2 = 0$$
$$if\ T_3 >= \frac{T_s}{2}, b_3 = 1 \ \ else\ b_3 = 0$$

Where b_j is the sign of the modulating waveform. From the signal $d = b_1 \oplus b_2 \oplus b_3$ is obtained. In Fig. 4.6(b) three discontinuous NSMS are presented, of which the logic signals that determine μ are also indicated. For μ=c the phases that correspond to subscripts max and min are clamped. For μ=d only phases that correspond to the maximum absolute value are clamped. For μ=d^* only phases with medium absolute value are clamped.

The introduction of NSMS in the DSPWM technique is, then, direct: the parameter μ is defined by the non sinusoidal modulating signals wished and t_{01} or t_{02} are calculated by (24) and (25).

IV. RESULTS

The DSPWM algorithm was validated by simulation tests for three-phase voltage source inverter fed Induction Motor drive with the following rating.
400 Volts, 4 kW, 4-pole, 1470 RPM, 50 Hz, rated torque = 30 N-m, R_s =1.57 ohm, R_r = 1.21 ohm
L_m = 0.165 H,L_s = 0.17 H, L_r = 0.17 H,J = 0.089 Kg - m^2
The DSPWM mainly generates different types of modulating signals by varying distribution ratio values. The distribution ratio can be a constant or varying value. The figure 6 to figure11 show the wave forms of modulating signal, pulses, line voltage and stator current, line for various values of distribution ratio and figure 12 to figure 17 shows total harmonic distortion of stator current of induction motor drive.

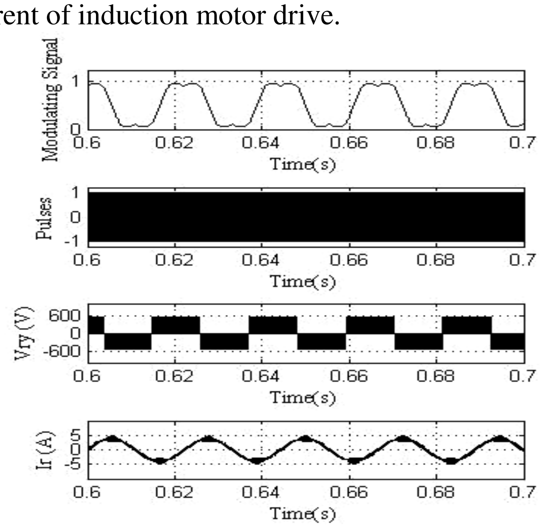

Fig. 6 for μ (distribution ratio)=0.5:Modulating Waveform, Pulses, Line voltage & Stator Current

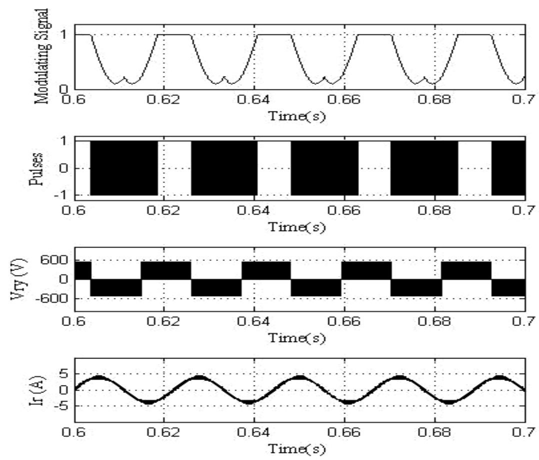

Fig. 7 for μ =0:Modulating Waveform, Pulses, Line voltage & Stator Current

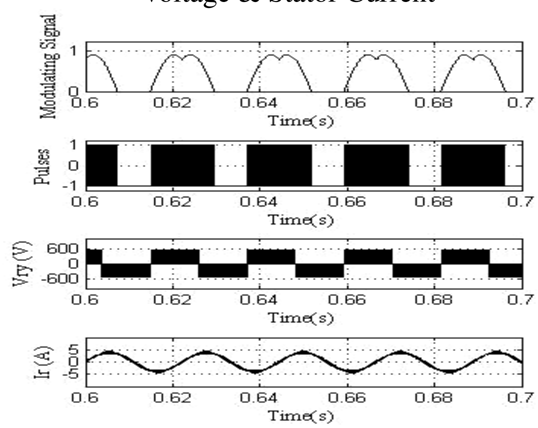

Fig. 8 for μ =1: Modulating Waveform, Pulses, Line voltage & Stator Current

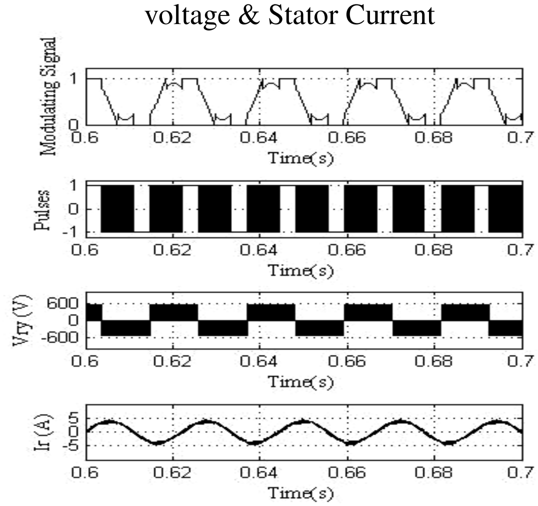

Fig. 9 for μ =c: Modulating Waveform, Pulses, Line voltage & Stator Current

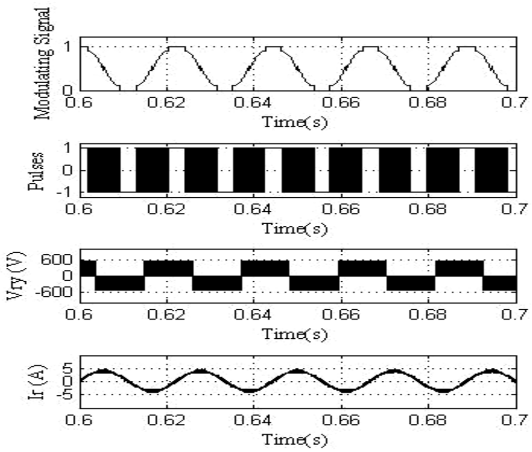

Fig. 10 for μ =d: Modulating Waveform, Pulses, Line voltage & Stator Current

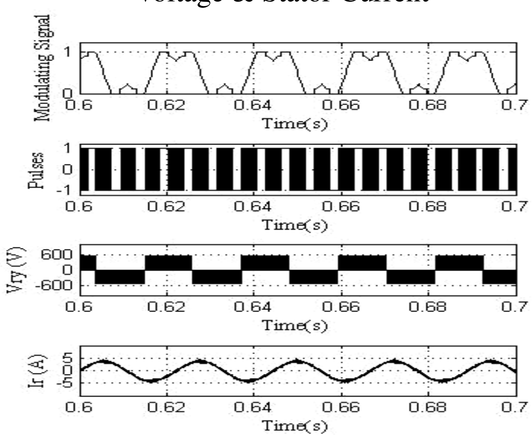

Fig. 11 for μ =~d: Modulating Waveform, Pulses, Line voltage & Stator Current

Stator Current Harmonic Comparison:

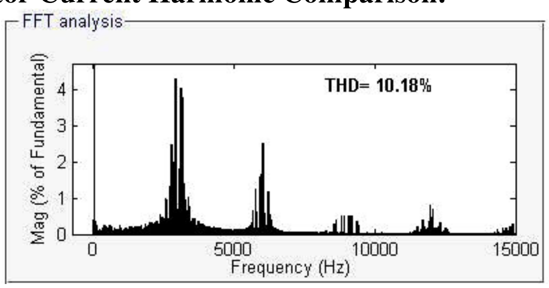

Fig. 12 for μ =0.5: harmonic spectrum of stator current

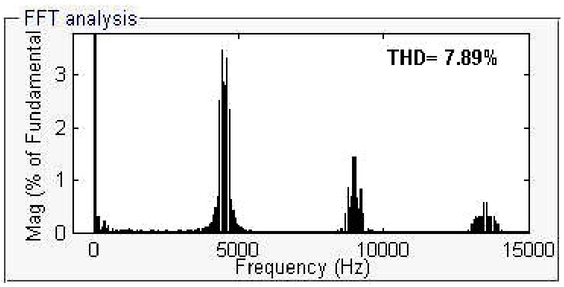

Fig. 13 for μ =0: harmonic spectrum of stator current

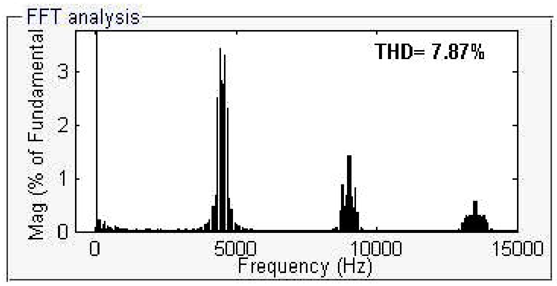

Fig. 14 for μ =1: harmonic spectrum of stator current

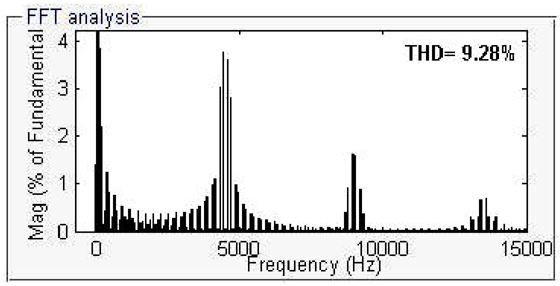

Fig. 15 for μ =c: harmonic spectrum of stator current

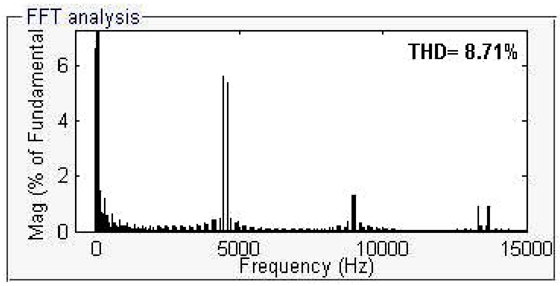

Fig. 16 for μ =d: harmonic spectrum of stator current

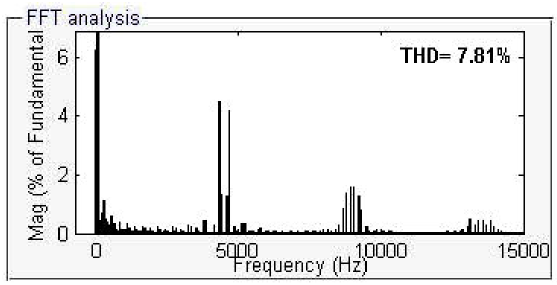

Fig. 17 for μ =~d: harmonic spectrum of stator current

V. CONCLUSION

It is possible to obtain the same results as those obtained with the space vector modulation by using a digital scalar modulation approach. Such equivalence was employed to propose a simple software algorithm to generate the space vector modulation from the scalar implementation. The proposed scheme provides a direct method to deal with non sinusoidal modulating waveforms. The proposed scheme was evaluated mathematically and tested via Matlab simulink.

REFERENCES

[1] A. Schönung and H. Stemmler, "Static frequency changers with sub harmonic control in conjunction with reversible variable speed ac drives," *Brown Boveri Rev.*, pp. 555–577, 1964.

[2] S. R. Bowes, "New sinusoidal pulse width-modulated inverter," *Proc. Inst. Elect. Eng.*, vol. 122, pp. 1279–1285, Nov. 1975.

[3] J. Holtz, "Pulse width modulation for electronic power conversion,"*Proc. IEEE*, vol. 82, pp. 1194–1214, Aug. 1994.

[4] G. Pfaff, A. Weschta, and A. F. Wick, "Design and experimental results of a brushless ac servo drive," *IEEE Trans. Ind. Applicat.*, vol. 20, pp. 814–821, Jul./Aug. 1984.

[5] H.W. Van der Broeck, H.-C. Skudelny, and G. V. Stanke, "Analysis and realization of a pulse width modulator based on voltage space vectors," *IEEE Trans. Ind. Applicat.*, vol. 24, pp. 142–150, Jan./Feb. 1988.

[6] L. Abraham and R. Blumel, "Optimization of three phase pattern by variable zero sequence component," in *Proc. Conf. Rec. EPE*, 1991, pp.169–174.

[7] M. Depenbrock, "Pulse-width control of a 3-phase inverter with non sinusoidal phase voltages," in *Proc. Conf. Rec. IAS*, 1977, pp. 399–403.

[8] C. B. Jacobina, A. M. N. Lima, and E. R. C. da Silva, "PWM space vector based on digital scalar modulation," in *Proc. Conf. Rec. PESC*, 1997, pp.100–105.

[9] L. Abraham and R. Blumel, "Optimization of three phase pulse pattern by variable zero sequence component," in *Proc. Conf. Rec. EPE*, 1991,pp. 272–277.

[10] P. G. Handley and J. T. Boys, "Practical real-time pwm modulators: An assessment," *Proc. Inst. Elect. Eng. B*, vol. 139, pp. 96–102, Mar. 1992.

[11] H. W. van der Broeck, "Analysis of the harmonics in voltage fed inverter drives caused by pwm schemes with discontinuous switching operation," in *Proc. Conf. Rec. EPE*, 1991, pp. 3261–3266.

[12] R. N. C. Alves, A. M. N. Lima, E. R. C. da Silva, and C. B. Jacobina, "A new approach to the problem of synthesising non sinusoidal waveforms for analog and digital implementations of space vector pwm strategies,"in *Proc. Conf. Rec. COBEP-Brazil*, 1991, pp. 228–233.

[13] V. Blasko, "A hybrid pwm strategy combining modified space vector and triangle comparison methods," in *Proc. Conf. Rec. PESC*, 1996, pp.1872–1878.

[14] Cursino Brandão Jacobina, Antonio Marcus Nogueira Lima, Edison Roberto Cabral da Silva, "Digital Scalar Pulse-Width Modulation: A Simple Approach to Introduce Non-Sinusoidal Modulating Waveforms", IEEE Transactions on Power Electronics, Vol. 16, No. 3, May 2001, pp.351–359.