EMPLOYING ADVANCED SCHEME OF A DIGITAL PULSEWIDTH MODULATION FOR DATA TECHNIQUES [+]

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Abstract:

Advanced pulsewidth modulation (PWM) techniques for inverters show the high-efficiency ac drives with low torque and speed pulsations and with reduced losses. A digital control scheme for a thyristor inverter is presented, where firing patterns satisfying given requirements like harmonic elimination or loss minimization have been calculated and stored in Programmable Read-Only Memory (PROM). Moreover, such a control permits a continuous variation of output voltage over the modulation range for full inverter utilization. Evaluation using a 22 KW induction motor drive verified that the control provides a clean frequency spectrum over the operating range of the drive (0-100 Hz).

استخدام متطور لبرمجيات عرض النبضة الترددية لتقنيات البيانات

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المستخلص:

ان تقنية عرض النبضة الترددية تعطي كفاءة عالية من التجهيز المتناوب بعزم وسرعة منخفضة مع تسريع النبضات وتقليل المفاقيد المصاحبة للاخراج .ان مخطط سيطرة العاكس الرقمي الموضح في البحث يعطي افضلية في ازالة التوافقيات او تقليل الخسائر ويكون من خلال حساب وخزن المعلومات في ذاكرة القراءة المبرمجة (PROM)، اضافة الى ذلك أن هذه السيطرة تسمح باستمرارية تغير فولتية الاخراج ضمن المدى القصوى لعمل العاكس.عند استخدام(22KW) لتجهيز المحركات الحثية يحقق سيطرة مجهزة ذات تردد وطيف نقي على مدى تشغيل المجهز (0- 100 Hz).

Introduction:

The great advantages of induction motors compared with dc machines and the existence of fast switching semiconductor drives (thyristors, power transistors, and gate turn off's (GTO's) are the bases for the enormous interest in the development of variable-speed ac drives.

The performance of these pulsewidth modulations (PWM) inverter fed drives is directly influenced by the choice of the modulation strategy used in control circuits.
The aim of PWM is to reproduce a reference waveform-in this case a sine wave- as a composition of pulses of variable width and constant amplitude. The modulator must supply

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the switching information for the inverter in order to obtain a power output with the lowest possible harmonic distortion. In all PWM techniques, this distortion depends directly on the switching-to-output frequency ratio. For typical low-power applications, using high-frequency transistor inverters, this ratio is always high enough, and the harmonic problem becomes insigniflement even with simple PWM strategies. However, in the medium and high power range, the maximum switching frequency of thyristor inverter (fig.1) is restricted by the turn-off time and commutation losses of the devices to 300-1000 Hz.

Regarding this limitation, the well-known subharmonic PWM method [1] with free-running carrier (where a triangle carrier wave is compared with three reference sinewaves) has suitable results only at low frequencies of the fundamental. In order to avoid unwanted subharmonic components of the inverter output voltage, which increase at lower carrier to reference frequency ratio; it is necessary at higher frequencies to synchronize the reference and carrier waveforms, which is difficult to realize, with conventional analog circuitry. Also, when the modulation index (peak value of the fundamental voltage to half dc link voltage ratio) approaches unity, a distortion in the modulated wave must be accepted in order to guarantee the minimum turn-off time of the thyristors.

authors [2],[3],[4] have proposed modulation strategies with in order to achieve a better utilization of a dc power supply and inverter, but problems like voltage discontinuities in V/f characteristic for operation with constant flux and stage synchronization arise, diminishing the performance and dynamic of the drive.

In this paper a digital PWM modulator is presented, where sophisticated modulation techniques like harmonic elimination [5] and loss minimization [6] can be easily implemented. The inheret synchronization and flexibility of the digital circuitry prevents the occurrence of unwanted subharmonics and voltage discontinuities over the whole range of the modulation index. Therefore, a “switching pattern” is computed by assigning switching information to each angle (α) and magnitude stored in the memory. In the pattern table the number of commutations per phase and period of the fundamental is variable and is matched to the modulation index; thus a better utilization of the inverter and dc power supply and a better performance of the drive are obtained. [7].

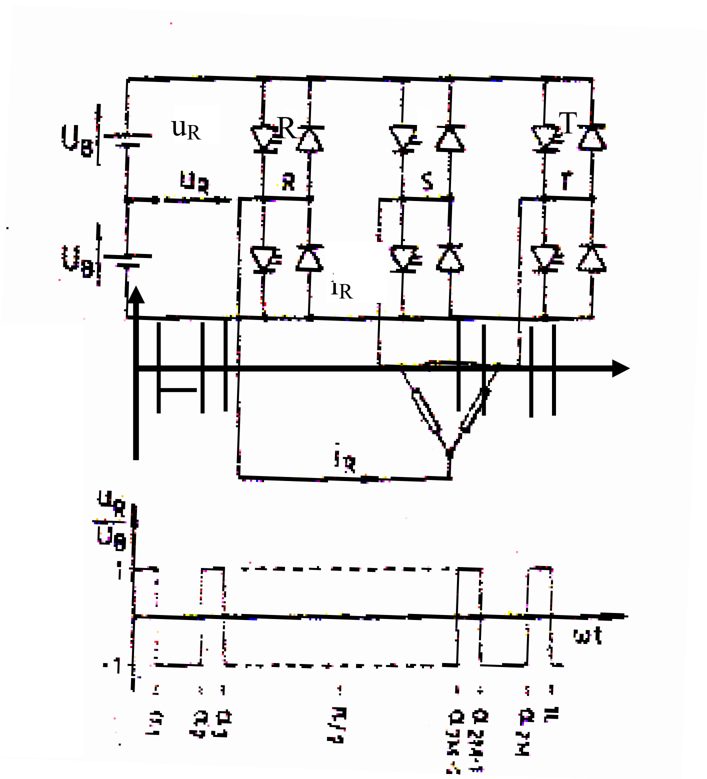

Fig.1 basic configuration of a three-phase inverter and its voltage waveform.
Computation of the Switching Pattern. [1].

One possibility of reducing harmonic distortion is to eliminate partain specific harmonics. It is possible to suppress one harmonic component by each commutation per quarter-period. The waveform in fig.1 can be described by a Fourier series as: [2].

$$U_{Ro}(\alpha)=\frac{U_R(\alpha)}{U_B}=\frac{R_o}{2}\sum_{n=1}^{\infty}[a_n\sin(n\alpha)+b_n\cos(n\alpha)] \qquad (1)$$

Where: $U_{RO}(\alpha)$ = phase voltage when ($0\leq\alpha\leq 120$),and (-phase voltage when ($180\leq\alpha\leq 300$)),$U_R(\alpha)$=voltage between phase R and center tap of dc link voltage with different value of α, U_B = dc voltage source with using center tap dc link.

By assuming quarter, and half-wave symmetry of the wave, b_n= 0 for all n and a_n= 0 for n even. The amplitude of the harmonic components is computed as follows [5]:

$$a_n=\frac{4}{n\pi}[1+2\sum_{k=1}^{n}(-1)^k\cos(n\alpha_k)], \qquad n=1, 3, 5, \ldots\ldots \qquad (2)$$

$a_{n-1}=b_n=b_{n-1}=0$, (half quarter-wave symmetry),

Where:

m = number of the commutations per quarter period.

α_k= switching angles, $0<\alpha_k<\alpha_{k+1}<\pi/2$.

With an appropriate selection of the m degrees of freedom, it is possible to determine the amplitude of the fundamental and to eliminate m-1 harmonics. The triplen harmonics are not available in the line voltage as cos = nπ/6 = 0, when n=3 thus reducing the harmonic content.

Numerically showing the system of nonlinear equations becomes more difficult by increasing m. when the required initial values for α_k are not close enough to the final values, but a certain routine in the development of the solutions was found, permitting-for any m- a direct prediction of the initial values that assures convergence. Due to the periodicity of the trigonometric functions, several solutions of the equations can be obtained; the number of physically correct solutions increases with m.

Then for a desired number of commutations per 90^o, the harmonic spectrum of the various solutions can be compared to each other and the best one is selected to be included in the pattern table. Considering the technical limitations of the inverter and taking into account that the higher order current harmonics are attenuated by the leakage reactance of the ac motor, it is of interest to eliminate-over the whole operating speed range-all harmonics below an absolute frequency value, for example 400Hz. Also, the low frequency torque pulsations are eliminated providing smooth low-speed rotation of the drive; neglecting flux harmonics, the dominant torque pulsations are those due to the interaction of fundamental flux in the air gap with the harmonic rotor currents [7].

Hence a switching pattern was computed making use of the near proportionality of voltage and frequency in ac machines operating with constant flux. At low magnitudes and low frequencies of the fundamental, many more harmonics are eliminated than at high magnitudes and frequencies the constration imposed by the inverter must be considered. Consequently, the number of commutation per quarter-period, m, is changed in accordance to the modulation index.

With the aim of limiting the maximum switching frequency to 300Hz and eliminating all harmonics below 400Hz, a pattern is computed with up to 23 commutations per quarter, cycle at low amplitudes; with m=23 all harmonics up to the 67th are eliminated.

In order to keep the inverter switching frequency constant over the output frequency range, m diminishes as the frequency of the fundamental increases. The switching pattern is subdivided into ranges with a different number of commutations (fig.2). The voltage waveform for a specific α_1 can be directly obtained from this figure; the shaded area corresponds to $-U_B$ and the rest to $+U_B$. In the range of very high magnitude ($\alpha_1>1.21$) no solution exists for the system of nonlinear equation; here the demand of harmonic elimination is replaced by a criterion minimizing harmonic distortion. In case of $\alpha_1 = 1.26$, the time between two commutations becomes smaller than the required minimum dwell time (170 µs) and m has to be reduced to m=1. [4].Performance of using ($30 \leq \alpha \leq 60$) in order to avoid the overlap appearance and besides decreasing the noise. Various output quantities are calculated by considering the fact that firing angle (α) is measured with respect to $\omega t = 60^0$and the device conducts up to an angle of ($120 + \alpha$), then $\alpha = 120/2\pi f_s$. [3].

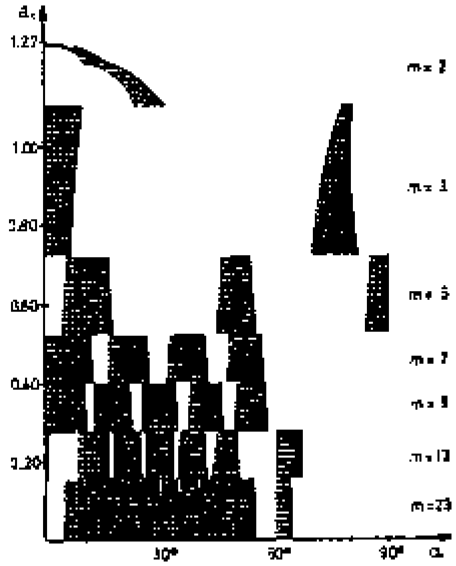

Fig.2 Switching pattern for PWM inverter control. Shaded area corresponds to - U_B, unshaded to + U_B, a_1 is amplitude of fundamental and m is number of commutation per quarter-period. [3].

Fig.3 shows the measured line-to-line voltage spectra for different amplitude levels referred to their fundamentals. From pattern using $m \leq 23$ (number commutation per quarter period), cycle at high amplitudes and harmonics are eliminated, and fundamental frequency is increased, when m is decreases, thus for the selected pattern using the inverter switching frequency (f_s) and the frequency of the lowest order nonzero harmonic (f_s) are illustrated in fig.4. In order to maintain low-speed torque the (V/f) ratio must be increased at a frequency below about 10 Hz, This can be done by adding a constant value to the (V/f) ratio. The constant is chosen to be equal to the voltage which produces the nominal flux zero speed. The inverter output waveform, which is nonsinusoidal when input to the motor results in current harmonics and torque pulsation problems especially at low speed. Various techniques for achieving voltage control, frequency control, and reduction of the harmonic effect and torque pulsation are investigated.

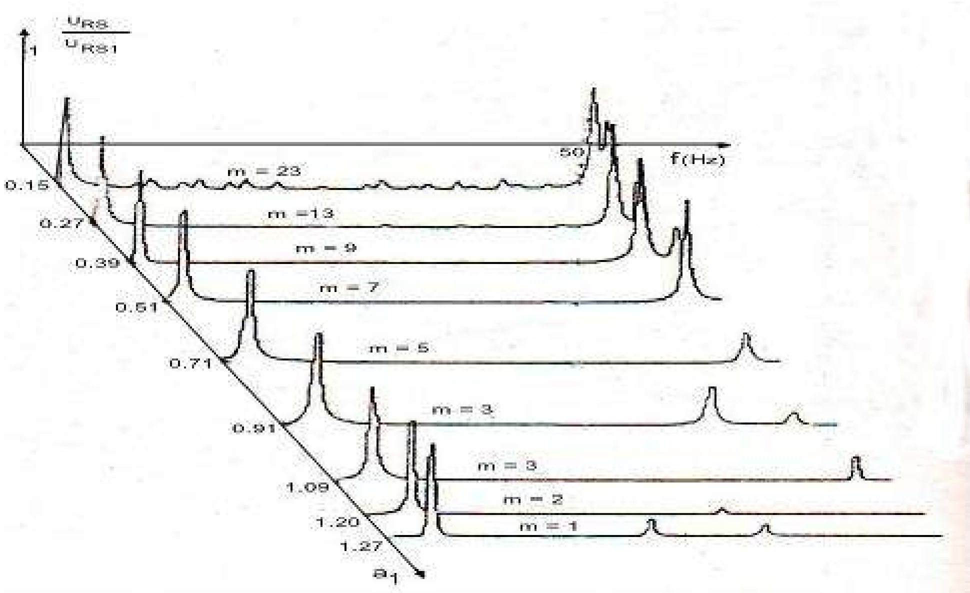

Fig.3 Measured voltage spectra for pattern of fig.2.

Fig.4 (a) (V/f) characteristics for constant flux operation.
(b) Inverter switching frequency f_s.
(c) Frequency of lowest order nonzero harmonic f_z.

The proposed PWM modulator:

A modulator with 6-kbyte Programmable Read Only Memory (PROM) and a 12-bit Digital-Analog Converter (DAC) is implemented (see fig.5); a theoretical analysis, computer simulation, and experimental results for a three-phase bridge inverter controlled by an 8085 microprocessor-based system are presented.

A period is subdivided into 1536 parts (divisible by four for quarter-wave symmetry and by three for 120^o phase shift), so that an angle resolution of 0.23^o, is obtained. The voltage range is subdivided into 128 amplitude levels for amplitude step a quarter-period of one phase has been stored in memory as binary data. As the memory array as divided into three groups, each one contains the information for 30^o.Angle resolution can be calculated by (α /128) = 0.23^o, when $\alpha = 30^o$

The switching information for the three-phases can be obtained simultaneously. The clock for the counter is provided by a linear Voltage-Controlled Oscillator (VCO); which receives its input voltage from the DAC; thus the resolution of frequency of the inverter output is given by the resolution of the DAC (for 12-bits, f = 0.024Hz).[5]

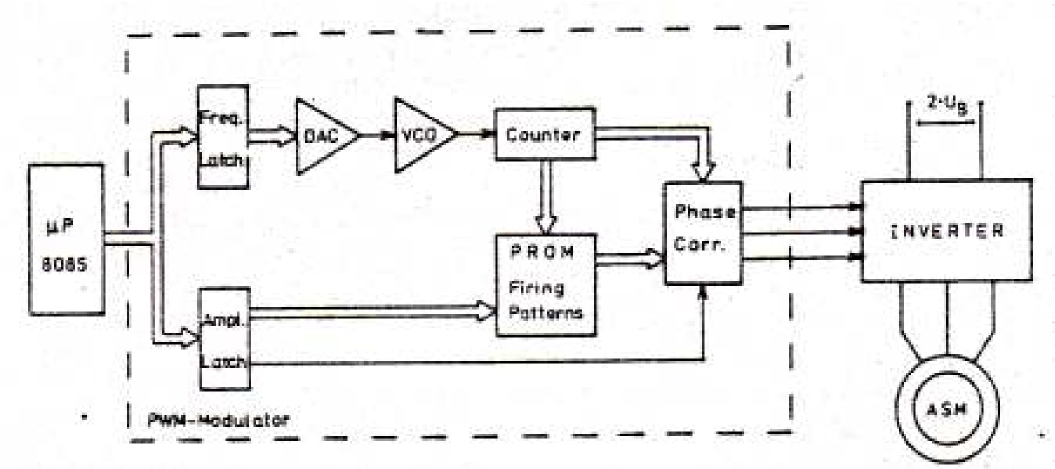

Fig.5 Digital Pulsewidth Modulator.

Transistors between two patterns can occur at any moment and without phase discontinuity. Fig.6, shows the measured voltage and current for an amplitude change from α_1= 0.25 (m =5) to α_1= 0.51 (m =7). An additional circuitry avoids commutations within the minimum dwell time. The digital nature of the modulator results in great flexibility. Other optimization criteria can be easily implemented by changing the switching information stored in PROM's.

- The accuracy of the fundamental voltage amplitude and frequency depends only on memory size, VCO, and the DAC.
- Frequency and amplitude of the fundamental can be controlled independently.

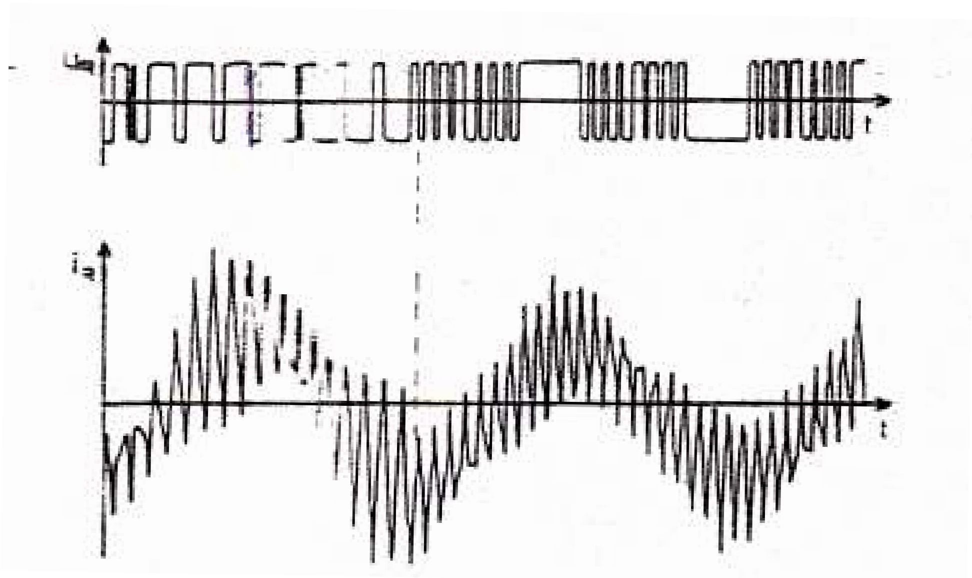

Fig.6 Measured $U_R(t)$ and $i_R(t)$ for f_o= 20Hz and amplitude change from α_1= 0.52 (m =5) to α_1= 0.51 (m =7).

An 8085 microprocessor matches the amplitude to the frequency in order to obtain a constant flux operation characteristic which permits the development of high starting torques at low frequencies as shown in fig.4 (a). The curve is a staircase waveform and ΔU corresponds to one percent of U_B. Parameter changes of the symreized curve can be realized by software (e.g., slope).To get zero output frequency (f_o= 0Hz, α_1= 0.0), all phases are switched to $+U_B$ and $-U_B$ with f_s= 3Hz in order to avoid the discharge of the commutation capacitors.

Results:

The proposed modulator is tested with a commercial PWM thyristor inverter and a 22KW induction machine. The inverter is connected to a constant dc link voltage of 300V. Fig.7 and Fig.8 show the measured waveforms of $U_R(t)$,$U_{RS}(t)$,$i_R(t)$,and $i_R(f)$, for different values of α, f, m and i reff. Where:

$U_R(t)$ = voltage between phase R and the center tap of the dc link voltage
$U_{RS}(t)$= line-to-line voltage.
$i_R(t)$ = line current.
$i_R(f)$ = linear spectrum of i_R from 0 to 300Hz.

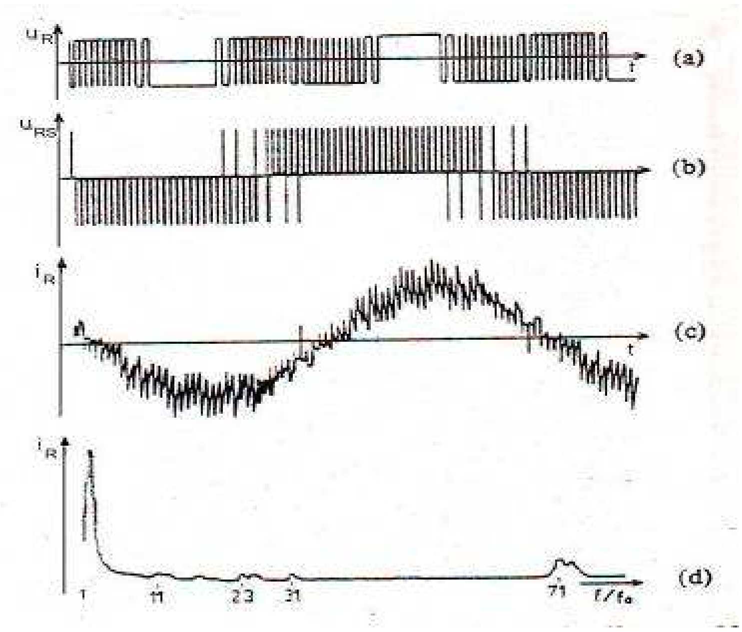

Fig.7 Measured waveforms for α_1=0.14, f=6.1Hz, m=23, and i_{reff}=14.5A. (a) $U_R(t)$. (b) $U_{RS}(t)$. (c) $i_R(t)$. (d) $i_R(f)$.

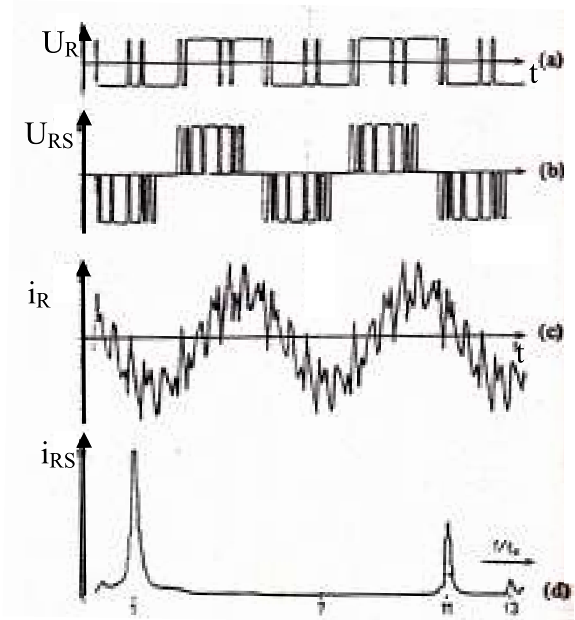

Fig.8 Measured waveforms for α_1=0.98, f=37Hz, m=3, and i_{reff}=14A. (a) $U_R(t)$. (b) $U_{RS}(t)$. (c) $i_R(t)$. (d) $i_R(f)$.

$U_R(t)$ = voltage between phase R and the center tap of the dc link voltage.
$U_{RS}(t)$= line-to-line voltage.
$i_R(t)$ = line current.
$i_R(f)$ = linear spectrum of i_R from 0 to 300Hz.

The spectral analysis shows that the desired harmonic elimination is achieved. In fig.7, m = 23, and the seventy-first is the first significant appearing harmonic; some low-order components (eleventh and twenty-third) are not totally eliminated because of the angle discretization, which can become important for large values of m. fig.8.

Fig.9, displays the measured frequency spectra of the line-to-line voltage and the current by a start of the machine from 0 to 50Hz at no load. The different ranges are easily recognizable and the transitions between different patterns and frequencies do not produce additional subharmonics.

The current spectra show the desired constant fundamental. By Dowling the installed memory capacity from 6 to 12-kbyte-PROM, an angular resolution twice as good could be achieved.

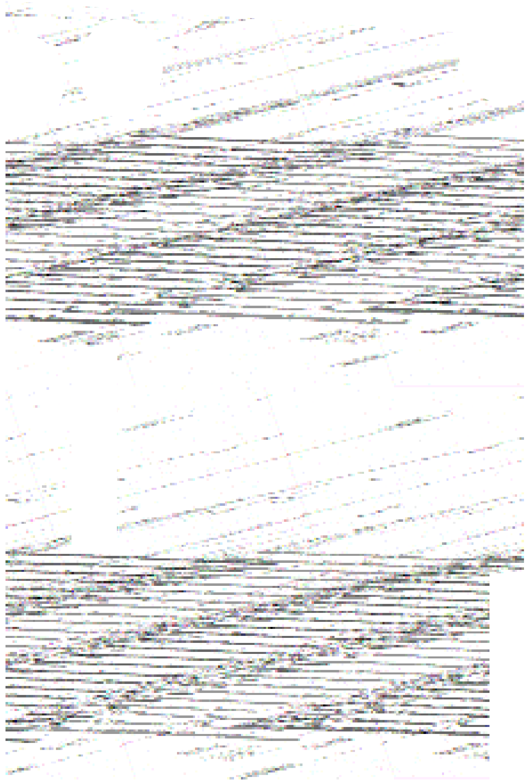

Fig.9 Measured voltage and current spectra by start of machine from 0 to 50Hz at no load with switching pattern of fig.2.

A different switching pattern results if other optimizing criteria are taken into account, which minimize losses or torque pulsations. Interesting observations have been made by driving the motor with the different solutions for equation (2); some of these are very near with the solutions for certain optimization strategies. In order to compare different PWM methods, measurements are made with the amplitude, load, frequency, and number of commutations in all cases.

Fig.10, shows the spectra of three PWM waveforms for α_1= 0.8, m= 4, and f_o= 40Hz, for harmonic elimination (solution 1), synchronized subharmonic PWM (triangle modulation), and harmonic elimination (solution 2), which represents almost the distortion minimization given by [6]. Within the limits imposed by the digitizing error, no difference between both results could be observed.

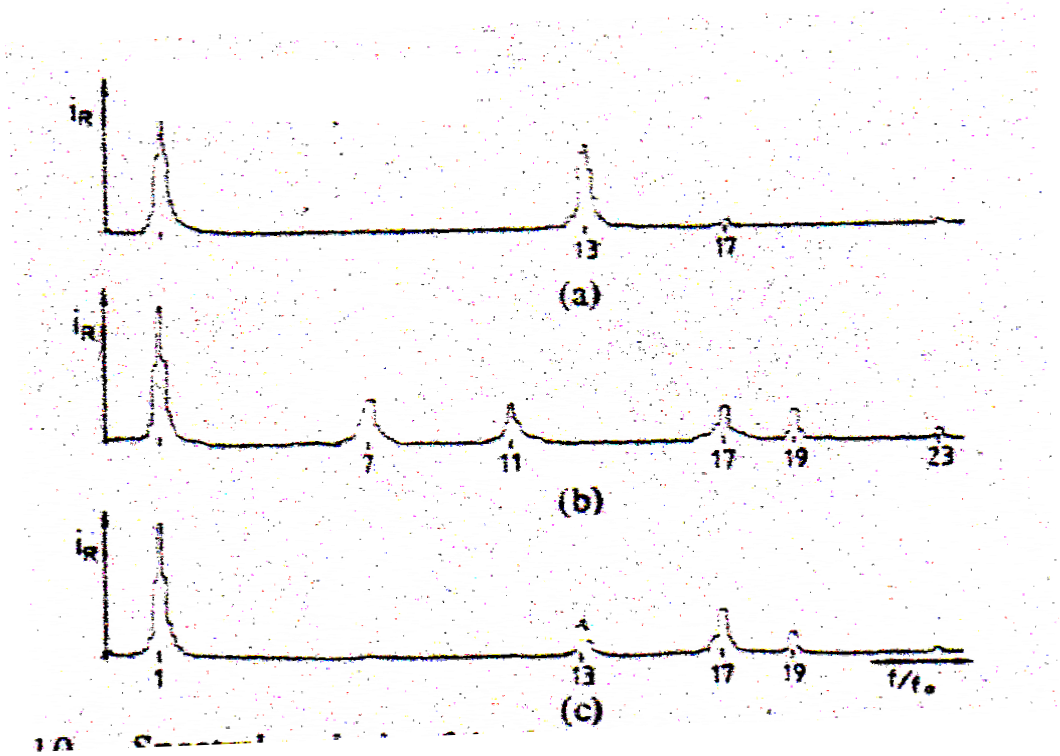

Fig.10 Spectral analysis of i_R for m = 4, α_1= 0.80, and f_o= 40Hz.

(a) Harmonic elimination (solution 1).

(b) Subharmonic PWM (triangle).

(c) Distortion minimization and harmonic elimination (solution 2).

This effect is not unexpected, because-in view of the inductive load- the low order harmonics have greater weighting factor than the higher order in harmonic distortion.

For the same m, the harmonic elimination method presents an approximately 50 percent wider clean spectrum than the subharmonic one, for which carrier and reference wave must be synchronized and have a frequency ratio being a multiple of three to obtain reasonable results; even at middle values of (f_c/f_o) this criterion should be considered. Fig.11, shows the current i_R for Harmonic elimination and Subharmonic PWM for the same conditions and a carrier to reference frequency ratio of 17 (m = 8).This scheme is similar to MPWM scheme except that in this case the width of the gating pulses in a cycle is not same. The width of the pulse depends upon the angular position of sinusoidal wave which is used as reference (modulating) signal. The modulating frequency (f_o) and carrier signals (f_c) are mixed in a comparator. The point of intersection between two signals determine the switching frequency (f_s) and commutating instant of the pulses per quarter period (m).

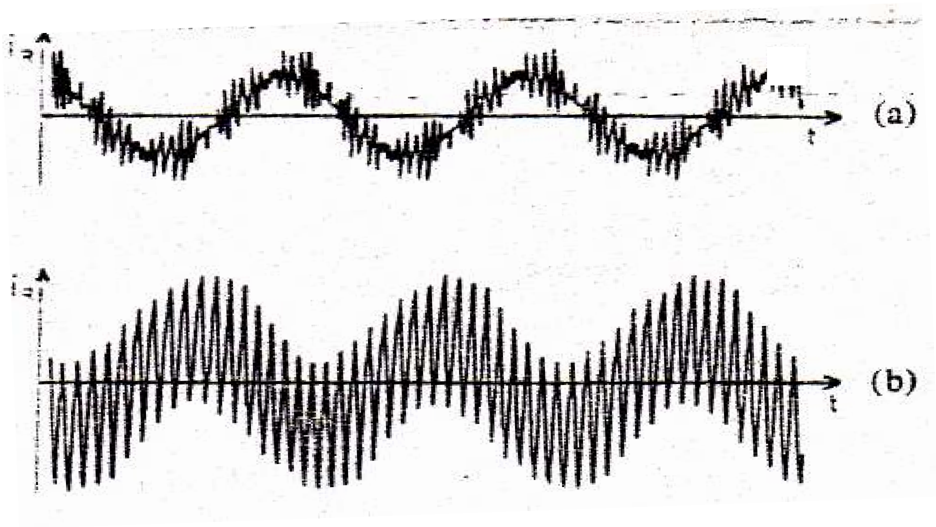

Fig.11 Measured current i_R for α_1=0.40, f_o= 15Hz, and m = 8, and (f_c/f_o=17). (a) Harmonic elimination. (b) Synchronized subharmonic (f_c/f_o is not multiple of three).

Conclusion:

An efficient and low-cost modulator has been implemented. Where advanced modulation techniques with optimized switching patterns can be employed. It allows operation over the whole range of the modulation index providing smooth transition between different patterns. Different strategies may be implemented by exchanging the PROM, requiring no hardware modifications.

The independence of the frequency and amplitude inputs also permits the realization of different V/f characteristics. Improvement in the output accuracy can be achieved by increasing the memory and the resolution of the DAC. By adding an 8 bit Analog – Digital Converter (ADC) for amplitude input and removing the DAC, the modulator can be connected to an analog control circuit as well.

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