

DOKUZ EYLÜL UNIVERSITY
GRADUATE SCHOOL OF NATURAL AND APPLIED SCIENCES

HARMONIC ELIMINATION IN INDUCTION MOTOR DRIVE USING SPACE VECTOR METHOD

38540

A Thesis Presented to
the Graduate School of Natural and Applied Sciences
Dokuz Eylül University

In Partial Fulfillment
of the Requirements for the Masters Degree
in Electrical and Electronics Engineering

by

Taner AKKAN

Advisor: **Dr. Yavuz ŞENOL**

January 1995
İZMİR

THESIS ABSTRACT

HARMONIC ELIMINATION IN INDUCTION MOTOR DRIVE USING SPACE VECTOR METHOD

**BY
TANER AKKAN**

Master Of Science, January 1995
(B. Sc. Dokuz Eylül University, 1991)

72 Typed Pages
Supervised by Dr. Yavuz ŞENOL

In this thesis, a new PWM (Pulse Width Modulation) method which is called Space Vector PWM Method, is applied on adjustable speed control of squirrel cage induction motor (IM). This method has many superiorities when compared with the other methods. First of all, the harmonic reduction is more effective, and PWM generation is more simple and requires only small look up table to control speed of the motor.

At the first step of the study, the Space Vector Method is simulated on a Personal Computer (PC) to understand and analyze the method. The software programs that developed on the PC was written in QUICK BASIC language because of its easy programming.

At the second step, microcontroller board has been designed to perform the method in real-time. In this board, INTEL 8098 microcontroller is used to generate output PWM voltage waveforms. The most important reason to select this microcontroller is the High Speed Output (HSO) unit that performs output operations with less CPU effort. Also the on-chip auxiliary units; serial channel, timers, analog digital converter (ADC), High Speed Input unit (HSI), and input/output (I/O) ports, simplify the design of the system. This board is actually an emulator board to develop the assembler program easily controlled by

serial channel. Also this board includes the inverter driver logic and liquid crystal display (LCD) to observe the system status.

The third step was to develop an 8098 assembler program. This program uses HSO unit to generate six PWM outputs to drive the inverter board. The look-up table data for each frequency were obtained from a Space Vector Simulation Program written on a PC. Moreover, the frequency and modulation index parameters, motor start, motor stop, motor direction are controlled from the Serial Channel Emulation Program written on the PC.

At the fourth step, an inverter board was designed to obtain sinusoidal three phase voltage waveform for drive 3-phase IM. The main switching element was selected as IGBT (Insulated Gate Bipolar Transistor) due to its ease of drive and high switching frequency. Ease of drive simplifies the board design. High switching frequency of the Space Vector Method is achieved by the high speed feature of the IGBT.

Briefly, Space Vector Control Method was applied. Computer simulation results were obtained. Experimental results were generated at different frequencies by the microcontroller board, inverter board, and 0.5 HP (0.37KW) AC Squirrel Cage Induction Motor. Finally, theoretical and experimental results were compared with each other, and the obtained results were presented.

ÖZET

Bu tezde, yeni bir PWM (Puls Genişlik Modülasyonu) metodu olan Space Vector PWM metodu sincap kafesli asenkron motorların ayarlanabilir hız kontrolüne uygulanmıştır. Bu metod diğerleri ile karşılaştırıldığında birçok üstünlüğe sahiptir. En başta, harmonik indirgeme daha efektif, PWM üretimi daha kolay ve motorun hız kontrolü için sadece küçük bir tetikleme açıları tablosu yeterlidir.

Çalışmanın ilk basamağında, Space Vector Metodu bir kişisel bilgisayar (PC) üzerinde metodu daha kolay anlayabilmek ve analiz etmek için simüle edilmiştir. PC üzerinde geliştirilen yazılımlar kolay programlanmasından dolayı Quick BASIC ortamında geliştirilmişlerdir.

İkinci olarak, Space Vector PWM metodunu gerçek zamanda uygulayabilmek için bir mikrokontrolör (microcontroller) kartı geliştirilmiştir. Bu kartta, çıkış PWM voltaj dalga şekillerini elde edebilmek için kolaylık sağlayan INTEL 8098 mikrokontrolörü kullanılmıştır. Bu işlemcili seçmekteki en önemli sebep, çok az CPU (Merkezi İşlemci Birimi) zamanı harcayarak çıkış işlemleri yapabilen HSO (Hızlı Çıkış Birimi) birimini iç yapısında barındırmasıdır. Ayrıca çip üzerinde seri kanal, zamanlayıcılar, analog dijital dönüştürücü (ADC), Hızlı Giriş Birimi (HSI) ve normal giriş/çıkış (I/O) portları bulunmaktadır. Çip üzerine tümleştirilmiş bu birimler sistem dizaynını kolaylaştırmaktadırlar. Bu mikrokontrolör kartı makina dili programları kolay geliştirilebilmek için bir PC'nin seri kanalı tarafından kontrol edilen bir emülatör kartı olarak kurulmuştur. Çalıştırılmak istenen program, PC üzerinde yazılıp derlenmekte ve programın makine dili kodları mikrokontrolör kartına aktarılmaktadır. Kartın, okunup yazılabilen RAM hafızasına yazılan bu kodlar, kart tarafından sanki kalıcı hafızasına (ROM) yazılmış gibi çalıştırılabilmekte ve program hataları kolayca düzeltilerek yazılım geliştirilebilmektedir. Ayrıca bu kart üzerinde inverter (evirici) kartını sürme mantığı ile sistemin durumunu gözlemek için bir LCD (Lıkd Kristal Display) de bulunmaktadır.

Üçüncü basamakta, 8098 makina dili programı geliştirilmiştir. Bu program, inverter kartını sürmek için gerekli altı adet PWM çıkışını HSO birimi yardımıyla üretmektedir. Her çalışma frekansı için gerekli tetikleme açıları tablosu PC'de yazılmış Space Vector simülasyon programından elde edilmektedir. Ayrıca, frekans ve modülasyon indeksi parametrelerinin transferi ve motor çalıştırma, durdurma, yön değiştirme gibi işlemler 8098

programında yapılabilmekte ve bu işlemler de PC' de çalışan Seri Kanal Emülasyon programı tarafından kontrol edilebilmektedir.

Dördüncü olarak, üç fazlı asenkron motoru sürmek için gerekli 3 fazlı gerilim dalga şekillerini üretecek anahtarlama elemanlarını, bunları sürme devresini, üç fazlı şebeke gerilimini doğrultarak DC gerilim elde eden 3 fazlı diyot doğrultucuyu içeren inverter kartı dizayn edilmiştir. Anahtarlama elemanı olarak kolay sürülmesi ve yüksek anahtarlama frekansına çıkabilmesi avantajları nedeniyle, IGBT (Yalıtılmış Kapılı Çift Kutuplu Transistör) seçilmiştir. Kolay sürülebilmesi inverter kartı dizaynını basitleştirmiş ve Space Vector Metodunun istenilen frekans için bir periyotta yer alan yüksek sayıda anahtarlama palslerini oluşturabilmiştir.

Kısaca, Space Vector PWM metodu uygulanmış ve bilgisayardan simülasyon sonuçları elde edilmiştir. Deney sonuçları ise PC tarafından kontrol edilen mikrokontrolör kartı, inverter kartı ve 0.37 KW ' lık bir asenkron motor ile çeşitli frekanslarda alınmıştır. Son olarak, simülasyon ve deney sonuçları birbiriyle karşılaştırılmıştır.

ACKNOWLEDGEMENTS

I would like to thank to my supervisor Dr. Yavuz Şenol for his precious support. I am grateful to Sinan Pravadalıoğlu for his help at every step of this study. I wish to express my appreciation to Prof. Dr. N. Süha Bayındır for his supervising.

I am especially indebted to my dear family for their encouragement and support.



TABLE OF CONTENTS

LIST OF FIGURES.....	ix
LIST OF PRINCIPAL SYMBOLS	xi
ACRONYMS	xii
1. INTRODUCTION	1
2. PWM SIGNAL GENERATION FOR USING SPACE VECTOR METHOD.....	4
2.1. INTRODUCTION.....	4
2.2. SINUSOIDAL PWM METHOD	4
2.3. SPACE VECTOR METHOD	6
2.3.1. PWM SWITCHING ANGLES CALCULATION.....	6
2.3.1.1. Performance Function	7
2.3.1.2. Quasi-Circular Locus Method.....	8
2.3.1.3. Performance Index	9
2.3.2. CHARACTERISTICS OF THE SPACE VECTOR PWM METHOD.....	10
2.4. HARMONIC ELIMINATION.....	12
2.4.1. Output Harmonics	12
2.4.2. Performance Parameters	15
3. THE SPACE VECTOR METHOD SOFTWARE.....	16
3.1. INTRODUCTION.....	16
3.2. 8098 16-BIT MICROCONTROLLER.....	16
3.2.1. High Speed Output (HSO) Unit	17
3.2.2. Software Timers Interrupt.....	18
3.2.3. Serial Channel.....	19
3.2.4. A/D Converter.....	19
3.2.5. Watchdog Timer.....	20
3.3. THE ALGORITHM OF THE ASSEMBLER CONTROL PROGRAM	20

4. SPACE VECTOR INDUCTION MOTOR DRIVE CIRCUIT	25
4.1. INTRODUCTION.....	25
4.2. THE SPACE VECTOR PWM SYSTEM.....	25
4.2.1. 8098 MICROCONTROLLER BOARD	26
4.2.1.1. Monitor EPROM.....	26
4.2.1.2. Simulator RAM	27
4.2.1.3. Serial Channel Translator Logic.....	27
4.2.1.4. 7-Segment 1x16 Character Display.....	28
4.2.1.5. The Driver Signals Inverter with Dead Time Generator Circuit	28
4.2.1.6. Direction Control Circuit	29
4.2.2. INVERTER	30
4.2.2.1. IGBT (Insulated Gate Bipolar Transistor).....	33
4.2.2.2. IGBT Driver	34
4.2.2.3. IGBT Protection Circuit	36
4.2.3. IGBT DRIVER POWER SUPPLY	37
4.2.4. THREE PHASE DIODE RECTIFIER.....	38
4.2.5. INDUCTION MOTOR.....	38
4.2.5.1. Operating Characteristics.....	40
4.2.5.2. Speed Control of Induction Motor.....	41
5. SIMULATION AND EXPERIMENTAL RESULTS.....	43
5.1. INTRODUCTION.....	43
5.2. Simulation Results.....	43
5.3. Experimental Results.....	44
5.4. Comparison of the Obtained Results	54
6. CONCLUSION.....	55
REFERENCES	57
APPENDICES	58
APPENDIX A:	58
APPENDIX B:	59

LIST OF FIGURES

2-1. PWM inverter output waveforms	5
2-2. The basic scheme of voltage source inverter	6
2-3. a) Circle locus of U and U^* b) A complete loci for $\Delta\theta = 30^\circ$	8
2-4. Switching angles and timing diagram for the pulsewidths	9
2-5. a) Loci of U and U^* and error vector ε b) Definition of $r\text{-}\phi$ coordinate	9
2-6. Complete 360 deg period voltage waveforms	12
2-7. Output voltage with two notches per half-wave	13
2-8. Output voltage for modified sinusoidal PWM	14
3-1. The 8098 block diagram	17
3-2. Assembler Program Flow-Chart (1)	22
3-3. Assembler Program Flow -Chart (2)	23
3-4. Assembler Program Flow -Chart (3)	24
4-1. The Overall System Block Scheme	25
4-2. The Microcontroller Board Block Diagram	27
4-3. The Driver Signals Inverter with Dead Time Generator Circuit	29
4-4. Direction Control Circuit	29
4-5. Three-phase bridge Inverter a) Circuit b) Waveforms	31
4-6. The Overall Inverter Circuit Scheme	32
4-7. (a) IGBT structure (b) IGBT equivalent circuit (c) symbol	34
4-8. (a) Simple IGBT Driver (b) Typical current and voltage waveforms of IGBT	35
4-9. IGBT Driver	36
4-10. Parallel and series snubber circuit	37
4-11. Basic +15V,-15V Power Supply	38
4-12. a) Torque production of Induction motor b) Rotor construction c) Cutaway view ...	39
4-13. a) Rotor Circuit b) Stator and Rotor Circuit c) Equivalent Circuit	40
4-14. The operational characteristics of IM	41
4-15. Adjusting AC IM speed	42

5-1. (a) 10 Hz., M=.2, 0-360° voltage and current waveforms	45
(b) 10 Hz., M=.2, 0-95° voltage and current waveforms	45
5-2. (a) 20 Hz., M=.4, 0-180° voltage and current waveforms	46
(b) 20 Hz., M=.4, 0-360° voltage and current waveforms	46
5-3. (a) 30 Hz., M=.6, 0-360° voltage and current waveforms	47
(b) 40 Hz., M=.8, 0-360° voltage and current waveforms	47
5-4. (a) 50 Hz., M=1, 0-360° voltage and current waveforms	48
(b) 100 Hz., M=1, 0-360° voltage and current waveforms	48
5-5. (a) 150 Hz., M=.1, 0-360° voltage and current waveforms	49
(b) 200 Hz., M=.1, 0-360° voltage and current waveforms	49
5-6. Root locus graphics for the selected frequency and M values.....	50
5-7. For 10 μ s pulsewidth the Dead Time Generator output signal (Voltages are 5V) ...	51
5-8. For 30 μ s pulsewidth the Dead Time Generator output signal (Voltages are 5V) ...	51
5-9. For 50Hz the detailed phase-a and phase-b voltage output	51
5-10. For 50Hz the phase-a voltage output for a complete period	52
5-11. For 50Hz the phase-a voltage output for a half period	52
5-12. For 50Hz the phase-a current	52
5-13. Another the phase-a current for 50Hz operation frequency	53
5-14. 8098 Microcontroller board and PC	53
5-15. Inverter board, power supply, and the induction motor	53

LIST OF PRINCIPAL SYMBOLS

ε	Error Vector
β	Switching angle (beta)
$\Delta\theta$	Sampling angle
α_1	Switching angle 1 (alfa1)
α_2	Switching angle 2 (alfa2)
ΔT	Sampling interval
di/dt	current slope
dV/dt	voltage slope
E_r	Induced Phase voltage
f_0	Operating frequency
I_{FL}	Full-load current
I_r'	Rotor Current
M	Modulation Index
P_a	Performance Function (for phase-a)
$R_g(\text{OFF})$	IGBT Driver input OFF resistance
$R_g(\text{ON})$	IGBT Driver input ON resistance
R_m	Resistance for excitatiton (or core) loss
R_r'	Resistance of the rotor winding of an IM.
R_s	Per phase resistance of the stator windings
s	Slip
S_a	Switching function (for phase-a)
U	Performance Vector
U^*	Reference Performance Vector
V/Hz	Stator voltage to operation frequency ratio
V_{st}	Voltage applied to stator of Induction Motor for selected frequency
ω	Angular speed
ω_s	Stator frequency
X_m	Magnetizing reactance
X_r'	Leakage reactance of the rotor windings of an IM.
X_s	Per phase leakage reactance of the stator windings.

ACRONYMS

ADC	Analog-Digital Converter
CAM	Content Addressable Memory
DAC	Digital-Analog Converter
DF	Distortion Factor
DSP	Digital Signal Processor
FET	Field Effect Transistor
GTO	Gate Turn-Off Thyristor
HF	Harmonic Factor
HSO	High Speed Outputs
I.M.	Induction Motor
IGBT	Insulated Gate Bipolar Transistor
LCD	Liquid Crystal Display
LOH	Lower Order Harmonic
MOSFET	Metal Oxide Semoconductor Field Effect Transistor
PC	Personal Computer
PF	Performance Function
PWM	Pulse Width Modulation
RPM	Revolutions Per Minute
SF	Switching Function
SHM	Sub-Harmonic Method
THD	Total Harmonic Distortion
VFC	Voltage to Frequency Converter

CHAPTER ONE

INTRODUCTION

The purpose of this thesis is to develop a Space Vector PWM (Pulse Width Modulation) method based on AC Induction motor speed control system. In many industrial applications, adjustable speed drives are required. In the past, these drives are employed by DC motors. Although DC motors are easy to control, they are more expensive to purchase and require extra maintenance. Also DC Motors which have same power ratings in comparison with AC motors, occupy large frame size (20 to 40% heavier than equivalent AC motors). Moreover, DC motors require special DC power supplies which brings further cost. AC motors control system is more complex but additional cost is offset by the economies of using an AC motor.

In the recent years with the development of power switching semiconductor devices and microcontroller technology, the speed control of AC Induction motors become more popular. Among all the control methods, the direct or indirect vector controlled induction motor drives have gained the widest acceptance in high performance applications[12][13]. While implementing a vector controlled induction motor drive, time-varying co-ordinate transformations must be done to achieve excellent performance. This method requires heavy computations. Therefore, a high speed microprocessor equipped with floating point unit, especially Digital Signal Processor (DSP), is usually adopted to meet the need of calculation speeds. The software of high speed microprocessor is complicated, but this problem can be solved by using effective control algorithm.

A sinusoidal PWM technique is also required to generate sinusoidal three phase line current using an inverter. Since the AC Induction Motor as an inverter load has a low-pass filter characteristic because of windage inductive effect, so the high-order harmonics are eliminated in the output-currents. A special method that determines switching angles using special harmonic elimination is preferred to obtain high performance from inverter.

A new method called Space Vector is easily used for driving power switching devices to obtain sinusoidal 3-phase current. Another microprocessor must be used because of the density of calculations on the high-speed microprocessing unit. This thesis involves PWM voltage signals generations using relatively low speed microcontroller instead of a high speed microprocessor. The PWM voltage generation parameters are directed by high-

speed microprocessor or supervisor computer. In this work, this unit is selected as PC (Personal Computer). Only two parameters are transferred. These parameters are angular speed and modulation index.

Current harmonics due to generated voltages were high and very effective at low speeds which causes torque pulsations. Elimination of current harmonics is vital for the importance of total system performance. Smooth start and acceleration at low speeds can only be achieved by the elimination of low order voltage harmonics. The Space Vector PWM method has superiority on previous methods. Switching and motor losses are relatively low and speed adjustment area is wide. Also with this method, generating of three phase voltages and control of the drive system becomes more simple and effective.

The inverter switching element is selected as IGBT (Insulated Gate Bipolar Transistor). This device has many superiorities on other power semiconductor devices. The operating speed can be up to 45 kHz like a power bipolar transistor. This frequency range allows the use of more pulses which helps to reduce low-order harmonics in the output current. Moreover, the gate drive current requirement is respectively low in comparison to the bipolar transistors. This is achieved by MOSFET input structure of IGBT. Space Vector Method output waveform has many pulses, so this device is selected. IGBT, also simplifies the driver circuit design due to its low input current requirement.

The control system consists of microcontroller and inverter boards. At first, Space Vector PWM method theory is applied to the PC to observe the simulation results. This software program was written in QUICK BASIC programming language. This program generates the switching data to obtain the three-phase sinusoidal output current, but this is impossible because of the harmonics due to switching action. Space Vector method is an optimized method for obtaining an output signal similar to a sinusoidal waveform with less harmonic contents. This software program shows switching data numerically and graphically and also analyzes the effects of harmonic contents. The main aim of this program is to generate switching data in hexadecimal format in the range of 2Hz to 50Hz. The generated data is arranged in assembler look-up table format to be used by microcontroller assembler program.

Another software program is written in 8098 microcontroller assembly language. This program is similar to BASIC program logic structure, but there are many differences between two programs. In the BASIC program we can obtain floating point results, but we can not use these numbers in assembler language of the selected microcontroller. The 8098 microcontroller uses integer binary arithmetic logic unit (ALU). As a result, we must round floating point numbers and convert these numbers into their integer equivalents. This conversion affects the system performance.

The Inverter base driving signals are obtained by using the HSO (High Speed Output) unit of the microcontroller. This unit reduces the microprocessor process load and we can run the assembler program more effectively. If a general input/output (I/O) port is used for this purpose, I/O operations reduces the microprocessor performance. The most important reason to select 8098 microcontroller for this application is the HSO unit and the other on-chip peripheral devices. The other units are: Serial Channel for communication with the host computer, A/D converter for sampling the line currents of the motor for closed-loop system, Software Timers for real time interrupts, Counters for measuring the current that uses voltage to frequency converters (VFC), Watchdog timer for protecting the total digital system from noise sources and general I/O ports to control some digital circuits.

In the second chapter, PWM generation methods are introduced. After giving the general view about PWM method, the Space Vector PWM Method is introduced. Then, coefficients and angles equations are given. A harmonic elimination concept is given.

In the third chapter, the 8098 microcontroller introduced. The units that used for generating Space Vector PWM waveforms and controlling the system from a PC are described. These units are HSO (High Speed Output) , Timers, Software Timers, Serial Channel, A/D converter, Watchdog Timer. Then, the algorithm of assembler program is given and explained.

In the fourth chapter, the Space Vector PWM system is explained. First, the inverter board is introduced. A three phase bridge inverter is explained and the voltage and current waveforms are given to understand operation of the inverter. Then, the designed inverter board is introduced and explained briefly. The switching element, IGBT, is introduced, then IGBT driver requirements and the driver circuit is given. Power Supply, three-phase diode rectifier, over current protection and current measurement circuits are explained briefly. At the end of the chapter, the induction motor structure and its characteristics are given. Speed control of induction motor is explained using motor characteristics.

In the fifth chapter, simulation results that obtained in QBASIC program and experimental results are given. The simulation results are the graphic outputs of the QBASIC Space Vector simulation program and this program is explained briefly. The experimental results are given as some photographs. At the end of the chapter, some comparisons are made between the simulation and experimental results.

In the sixth chapter some conclusions are given. This chapter includes the overview of the study, the difficulties coped with, and some suggestions.

CHAPTER TWO

PWM SIGNAL GENERATION FOR USING SPACE VECTOR METHOD

2.1. INTRODUCTION

In this chapter, firstly the sinusoidal PWM method is briefly described. In this method, the switching angles to obtain sinusoidal output waveform are found from a comparison between reference sinusoidal waveform and a high frequency carrier triangle waveform. Sinusoidal output frequency and amplitude can be selected by changing the frequency of sinusoidal waveform. But these data are fixed and can not be changed.

Secondly, the Space Vector Method is explained. In this method, switching data are not fixed for each frequency. The amplitude and frequency of sinusoidal output waveform can be easily controlled.

At the end of the chapter, the general harmonic elimination method is explained. Also, from the general equation the harmonic magnitude of each frequency is calculated. The results can be seen from the Space Vector simulation program.

2.2. SINUSOIDAL PWM METHOD

In this method, the switches are turned-on and off several times during a half-cycle and the output voltage is controlled by varying the width of pulses[1]. The lower-order harmonics can be eliminated by selecting the number of pulses per half-cycle. Increasing the number of pulses would increase the harmonics order, so these harmonics could easily be filtered out by the nature of the inductive load. If there are p pulses per half-cycle of the same width, the maximum pulse width is limited to value of $360^\circ/p$. If these pulsewidths are chosen differently, the certain harmonics can be easily eliminated. There are many methods to determine the width of pulses, but the most known is sinusoidal pulse width modulation. If four pulses can be placed into a half cycle, then all the harmonics up to the seventh can be eliminated. In this case, seventh harmonic is the lowest order harmonic[1].

A sinusoidal reference waveform that determine the voltage and frequency output of the inverter is compared with a carrier triangle waveform. The intersection points of two waveforms determine the position of pulses and pulsewidths. The generation technique is

shown in Fig. 2-1. In the Figure 2-1a, the PWM waveform is obtained. In the Figure 2-1b, the magnitude of the reference sinusoidal waveform is reduced. So, the pulse widths become smaller. In the Figure 2-1c, the frequency of the carrier waveform and the reference sinusoidal waveform is raised, and the number of pulses per half cycle are increased. The output frequency is determined by changing the frequency of sinusoidal waveform. The pulse widths, so the magnitude of the output current, are varied by changing the amplitude of reference sinusoidal waveform or the modulation index, M .

$$M = (\text{reference signal amplitude}) / (\text{triangle waveform amplitude})$$

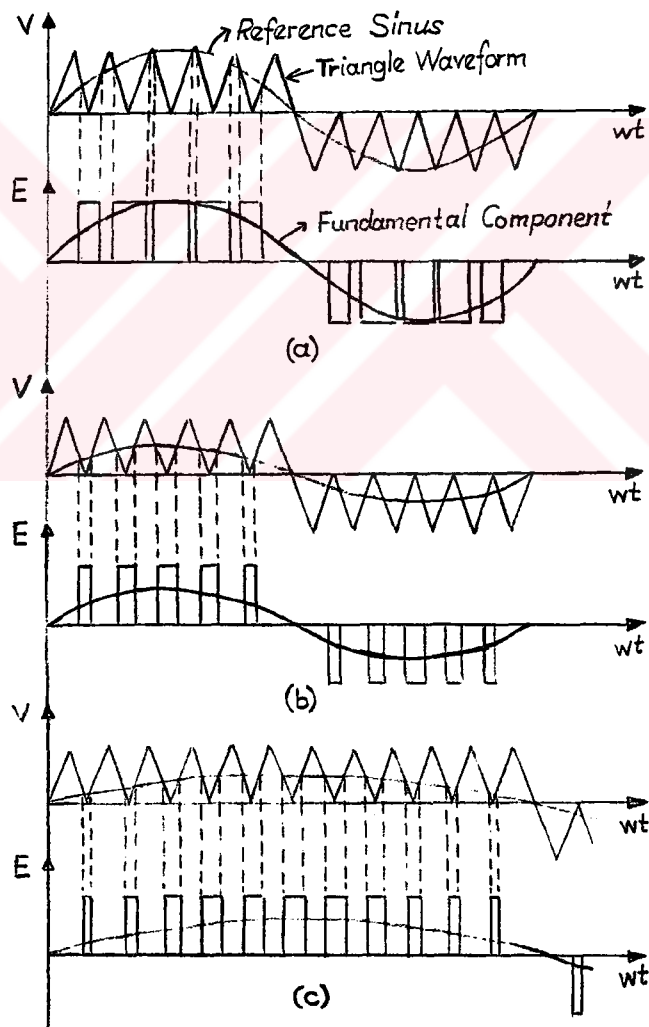


Fig. 2-1. PWM inverter output waveforms

2.3. SPACE VECTOR METHOD

Subharmonic modulation method (SHM) is an analog PWM method and a sinusoidal waveform is sampled by a triangle waveform at high sampling frequencies. The intersection points of both waveforms determine the PWM switching angles. SHM has been widely used for sinusoidal PWM voltage source inverters. A digital version of SHM is an alternative approach, because the SHM method has some drawbacks. The new digital method calculates PWM pulsewidths with a microcontroller in real time instead of using pre-calculated look-up tables. When we compared these two methods we can obtain some features of new method below[2].

- 1) Real-time PWM control of inverter system with a microcontroller is possible.
- 2) The maximum output voltage of new method is 15 percent greater and the number of switchings is 30 percent less than the output voltage of SHM at the same carrier frequency. Carrier frequency is known as sampling frequency in the Space Vector Method.
- 3) Synchronization of both carrier wave and reference signal wave is not necessary. As a result, the control circuit becomes simpler.
- 4) Low-order harmonics are well suppressed.

In the new method, a performance function (PF) that is a time-integral function of the output voltage, is introduced to eliminate low-order harmonics. Then, if the distortion factor of PF is minimized, optimal pulsewidths are determined by using instantaneous space vector representation of voltages.

2.3.1. PWM SWITCHING ANGLES CALCULATION

The PWM pulse patterns are controlled so that the performance function in the vector notation may draw a circular locus. A voltage source inverter scheme is drawn to explain the switching function in Figure 2-2.

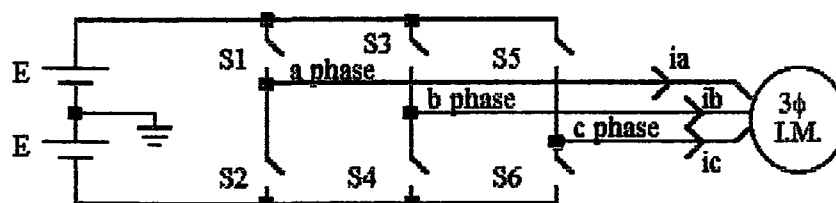


Fig. 2-2. The basic scheme of voltage source inverter

The switching function (SF) for phase-a is defined as

$$S_a = 1 \text{ when } S1 \text{ is ON and } S2 \text{ is OFF,} \quad S_a = -1 \text{ when } S1 \text{ is OFF and } S2 \text{ is ON}$$

The SF can be expanded in fourier series as

$$S_a = \sum A_k \cdot \cos k\theta, \quad k=1,3,5, \dots, \quad \theta = \omega.t = 2.\pi.f_0.t \quad (2-1)$$

where k is the harmonic order. Then, the output phase voltage is obtained from SF as

$$V_a = E.S_a, \quad V_b = E.S_b, \quad V_c = E.S_c \quad (2-2)$$

where E denotes the amplitude of the dc line voltage.

2.3.1.1. Performance Function

The high-order harmonics in the voltage are reduced in the line currents (ia, ib and ic) by the induction motors low-pass filter characteristics, so eliminating low-order harmonics is more important by using SF. A performance function (PF) is determined to explain this specification as

$$P_a = S_a \cdot dt = \sum [A_k / (k.\omega)] \cdot \sin k\theta \quad (2-3)$$

The Pb and Pc for other phases are defined similarly. If we compare both of PF and SF equations, PF frequency spectrum is multiplied by weighing factor 1/(k.ω) that is derived from integral function. When the harmonic order is bigger, the weighing factor is small, so the higher harmonics contribution will be low.

The SF and PF should be represented with Instantaneous Space Vectors. The voltage vector V_n is defined in terms of three phase quantities as

$$V_n = (2/3) \cdot [S_a + a.S_b + a^2.S_c], \quad a = \exp(j.2.\pi/3)$$

$$V_n = (4/3) \cdot \exp[j.(n-1).\pi/3], \quad n=1,2,3,4,5,6, \quad V_0=0, \quad V_7=0 \quad (2-4)$$

Using Equation 2-4, we have seven kinds of voltage vectors as shown in Fig.2-6b.

The performance vector U is defined as a time-integral function of V_n and its equation is given in Equation 2-5. U draws a locus that is determined by the voltage vectors and U moves with a constant speed. If V_0 and V_7 selected, U does not move.

$$U = (2/3) \cdot \int (V_n \cdot dt + U_0) = (2/3) \cdot (P_a + a.P_b + a^2.P_c) \quad (2-5)$$

If SF is ideal, it should be a pure sinusoidal and expressed by for phase-a.

$$S_a^* = (2/\sqrt{3}) \cdot M \cdot \cos \omega.t \quad (2-6)$$

From Equation 2-5, If the output voltages are pure sinusoidal, the performance

vector locus will be

$$U^* = -j \cdot \left[\frac{2}{\sqrt{3}} \right] \cdot (M/w) \cdot \exp(jw t) \quad (2-7)$$

where w is the output frequency, M is the modulation index ($0 < M < 1$) for controlling amplitude of the output voltage. The U^* which is called the reference performance vector, draws a pure circle locus and corresponds to the signal in SHM. The U^* rotates in counter-clockwise direction with a radius $\left(\frac{2}{\sqrt{3}} \right) \cdot (M/w)$ and an angular speed w .

2.3.1.2. Quasi-Circular Locus Method

The U locus can be controlled by selecting V_n , and the time width of V_n so that U may follow U^* locus as close as possible. This is called quasi-circular locus method. U corresponds to a stator flux vector.

From the symmetry, $0 < \theta < \pi/3$ period is sufficient to obtain a complete 2π period. The θ denotes the phase angle output voltage. The selection of V_n and adjustment of its timewidth are conducted at every sampling interval, ΔT . The ΔT corresponds to a carrier period in SHM, also this period is divided into $\pi / (3 \Delta \theta)$ intervals.

If we want to compare U and U^* , the enlarged one sampling interval can be seen in Figure 2-3a. For $\Delta \theta = 30$ deg., a complete loci of U and U^* illustrated in Figure 2-3b. The U^* locates at the position $P(x_0, y_0)$ at $\theta = \theta_0$ (or $t = t_0$), at the position $Q(x_4, y_4)$ at $t = t_0 + \Delta T$ or ($\theta = \theta_0 + \Delta \theta$). These positions and their co-ordinates are defined by

$$\begin{aligned} x_0 &= \left(\frac{2}{\sqrt{3}} \right) \cdot (M/w) \cdot \sin(w \cdot t_0) & y_0 &= \left(\frac{2}{\sqrt{3}} \right) \cdot (M/w) \cdot \cos(w \cdot t_0) \\ x_4 &= \left(\frac{2}{\sqrt{3}} \right) \cdot (M/w) \cdot \sin w(t_0 + \Delta T) & y_4 &= \left(\frac{2}{\sqrt{3}} \right) \cdot (M/w) \cdot \cos w(t_0 + \Delta T) \end{aligned} \quad (2-8)$$

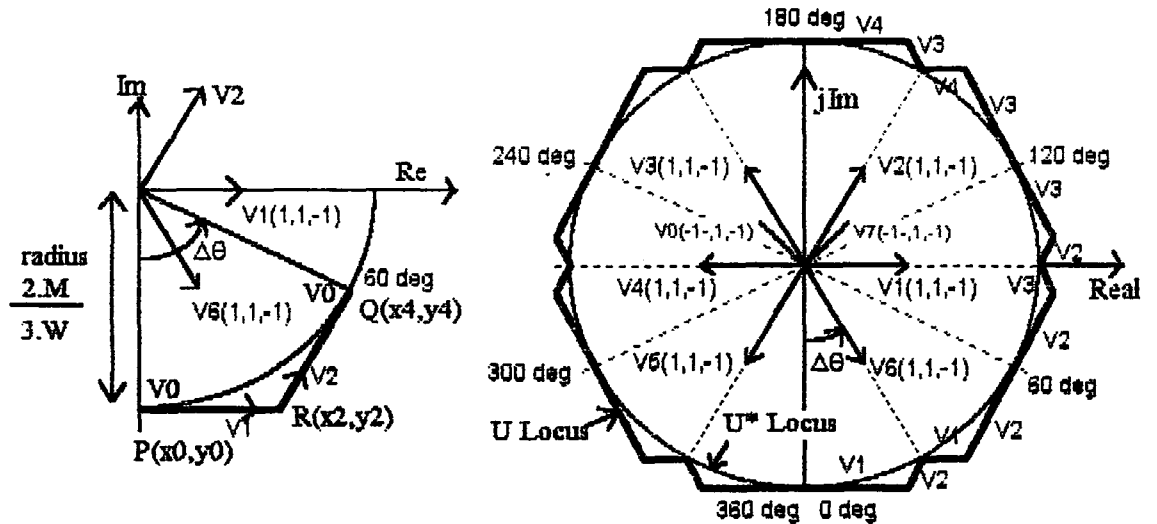


Fig. 2-3. a) Circle locus of U and U^*

b) A complete locus for $\Delta \theta = 30$ deg.

The timing diagram for each voltage phases are illustrated in Figure 2-4. The U locus is adjusted so that U coincides with U^* at the beginning point P ($t=t_0$) and at the end point Q ($t_0+\Delta T$) of every sampling interval. In a ΔT period, two zero vector V_0 or V_7 , and two nonzero vectors which sequentially selected to make U locus close to U^* locus. There is no switching, during ΔT interval in V_a voltage waveform. In the SHM method, switching occurs in every phase and in every interval.

$$\alpha_1 = t_3 - t_1 = K_1(\theta_r).M \quad \alpha_2 = t_2 - t_1 = K_2(\theta_r).M \quad \beta = t_1 - t_0$$

where

$$K_1(\theta_r) = \Delta T \cdot \cos(\theta_r - \pi/6), \quad K_2(\theta_r) = \Delta T \cdot \cos(\theta_r + \pi/6)$$

with $\theta_r = \theta_0 + \Delta\theta/2$ and $\sin \Delta\theta/2 \cong \omega \cdot \Delta T/2$ (2-9)

The θ_r symbol represents the phase angle of the voltage during a period ΔT .

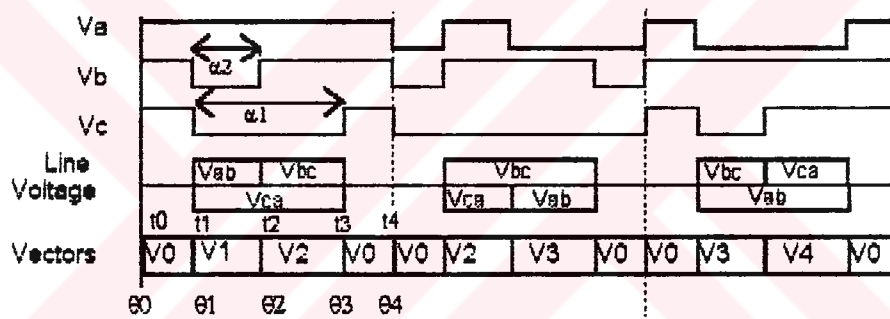


Fig. 2-4. Switching angles and timing diagrams for the pulsewidths

2.3.1.3. Performance Index

Error vector is introduced in Equation 2-10 and if ε is represented in the r - ϕ co-ordinate, which rotates with the reference vector U^* as shown in Fig 2-5. Its two perpendicular components are shown in Equation 2-10.

$$\varepsilon = U - U^* = \varepsilon_r + j\varepsilon_\phi \quad (2-10)$$

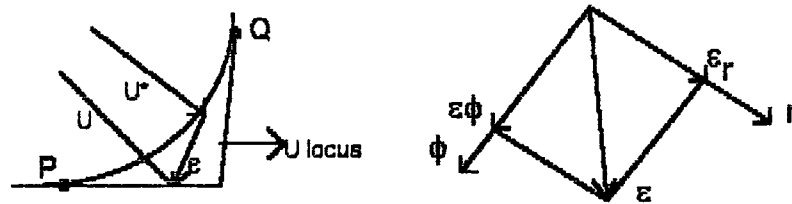


Fig. 2-5. a) Loci of U and U^* and error vector ε b) Definition of r - ϕ co-ordinate.

α_1 and α_2 can be determined easily once the modulation index M is given. But β can be found using the error vector properly. The performance index for the m th sampling period is given in Equation 2-11. If J_m is integrated over a complete period, the total performance index, J , is obtained and given in Equation 2-12.

$$J_m(\beta_m) = \int |\varepsilon|^2 . dt = \int (\varepsilon_r^2 + \varepsilon_\phi^2) . dt = J_{mr} + J_{m\phi} \quad (2-11)$$

$$J = \int |\varepsilon|^2 . dt = \sum J_m = \sum (J_{mr} + J_{m\phi}) = J_r + J_\phi \quad (2-12)$$

J corresponds to the distortion factor of the output current and represents the copper loss and also the magnetic noises of driven motors. Optimal PWM pulse patterns can be obtained by minimizing J . If J_m values are minimized, the optimal β time can be found. β_m can also be obtained analytically by using U^* circular locus and U locus. As $\Delta\theta$ is very small the U locus will be very closer to U^* circular locus. This means if the operation frequency is small or the speed is low, the U locus will be more similar to U^* locus. K_3 coefficient is given in Equation 2-13. β pulse location is obtained using K_3 coefficient and this is given in Equation 2-14.

$$K_3(\theta_r) = (\sqrt{3}/2) \Delta T . \{ [(1/6) . \sin^2 \theta_r + 1] . \cos \theta_r - (1/2) . \cos^3 \theta_r \} \quad (2-13)$$

$$\beta = \Delta T/2 - K_3(\theta_r) . M \quad (\beta = 0, \text{ if } \beta < 0) \quad (2-14)$$

2.3.2. CHARACTERISTICS OF THE SPACE VECTOR PWM METHOD

It can be easily realized that the K_1, K_2 and K_3 coefficients are the only function of the θ_r phase angle that changes with $\Delta\theta$ steps. The other unknown parameter is ΔT , and if f_s is selected, then we can calculate this value as

$$\Delta T = 1/f_s = 1/3600 \text{ Hz} = 277.78 \mu\text{sec} \approx 278 \mu\text{sec}.$$

So these coefficients can be calculated in terms of the current θ_r phase angle and listed in the ROM memory of the Space Vector control board. From this table microcontroller actually calculates the α_1, α_2 , and β times. Also we need the $\Delta\theta$ phase angle steps that can be calculated using f_s (sampling frequency) and f_0 (operation frequency). The $\Delta\theta$ angle resolution can be easily obtained from the equation below.

$$\Delta\theta = \omega . \Delta T = 2 . \pi . f_0 . (1/f_s) = 2 . \pi . f_0 / f_s \quad (2-15)$$

The operation frequency period is divided by sampling frequency period or in angular terms total θ (360° or 2π) is divided $\Delta\theta$ phase angle steps. This division is equal to the interval number as

$$\text{interval number} = g = \theta / \Delta\theta = 2\pi / (2\pi \cdot f_0) / f_s = f_s / f_0 \quad (2-16)$$

From the 60° locus symmetry, total 360° period can be obtained using this specific 60° period data. So the interval numbers for the desired frequency is divided six for 60° period. If we want to calculate the ROM table size, first we determine the angle resolution and how many bytes for K1, K2 and K3 coefficients are required.

If our operation frequency low limit is selected as 2 Hz, we can calculate the angle resolution and ROM size. From the equation 2-15, $\Delta\theta$ is calculated as

$$\Delta\theta = \text{angle resolution} = (2\pi \cdot f_0) / f_s = 0.2 \text{ deg.}$$

and the interval number is obtained as

$$\text{interval number} = g = f_s / f_0 = 3600 / 2 = 1800 \quad (\text{for } 360^\circ \text{ period.})$$

$$\text{required interval number} = g / 6 = 1800 / 6 = 300 \quad (\text{for } 60^\circ \text{ period.})$$

Also, this calculation can be proved using the previous $g = \pi / (3 \cdot \Delta\theta)$ formula. The interval number for 60° period means that if K coefficients can be written in 1 byte, the total memory requirement is equal to 900 bytes ($300 \cdot 3$).

This size of memory is lower than the memory size that obtained from SHM. In SHM, the pulse locations cannot be calculated using coefficients for desired operation frequency and it is necessary to create ROM table for total 360° period. If the maximum frequency is selected as 50 Hz, the total size is calculated as

$$\begin{aligned} \text{SHM ROM size} &= \text{interval number} \cdot 3 \text{ switch angles} \cdot (50 \text{ Hz} / 2 \text{ Hz}) \\ &= 1800 \cdot 3 \cdot 25 = 135000 \text{ bytes} \end{aligned}$$

This size exceeds our microcontroller maximum memory address space 65536 bytes. So the proposed PWM method is superior when compared to SHM.

The complete formulas to obtain K1, K2, K3 and $\alpha_1, \alpha_2, \beta$ angles are given below and also their graphics is sketched using MATHCAD mathematics program. The MATHCAD program and the graphics are given in Appendix A.

The all switching 360° deg period waveforms for each phase and inter-phase voltages are shown in Fig 2-6. Two phases are changed with each other to obtain a new turning direction and comparisons between two direction is shown in same figure.

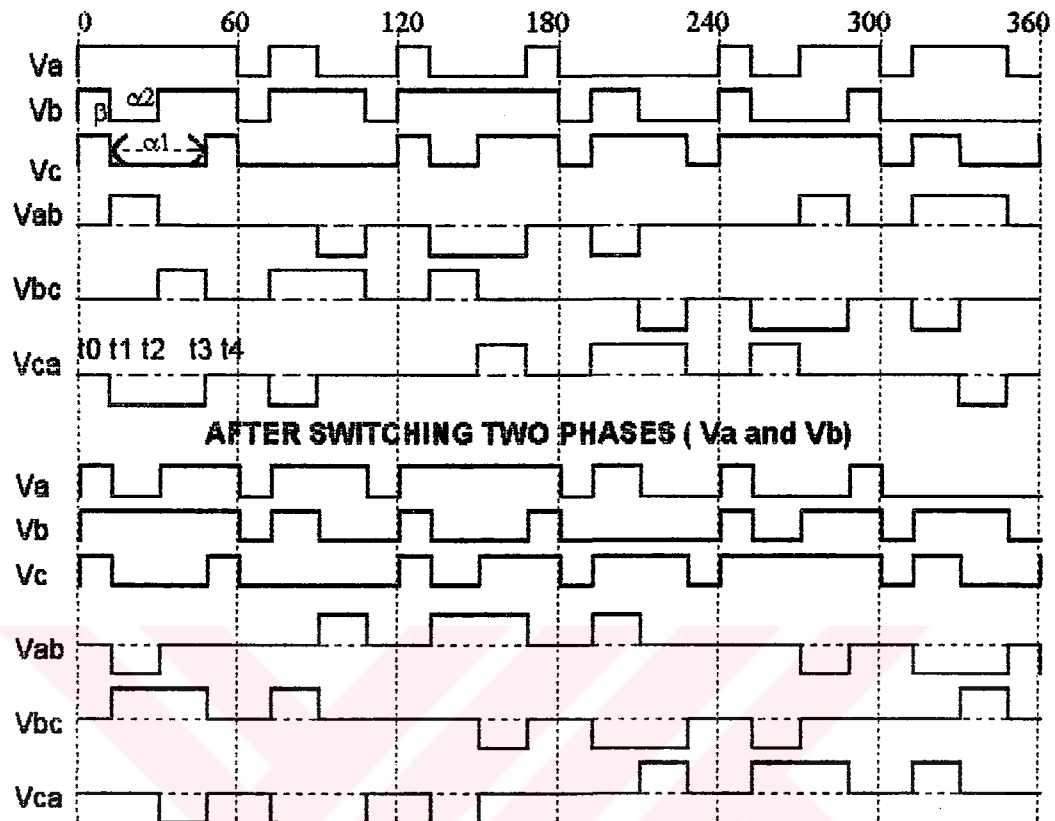


Fig. 2-6. Complete 360 deg period voltage waveforms

2.4. HARMONIC ELIMINATION

All inverter output voltages are not pure sinusoidal, they contain some certain harmonics that effects the output performance. When these inverter outputs applied to the three-phase induction motor as a load, non-sinusoidal phase currents flows into phase windings. The harmonic content of the output voltage of an inverter may be of interest from the point of view of its effect on the load, or for other reasons such as the possibility of interference with communications systems. So the elimination of harmonics is required.

2.4.1. Output Harmonics

The harmonic currents and main component of inverter output voltage can be analyzed by using Fourier analysis. The harmonic currents basically determine the the copper losses of the motor which is the major part of total losses. The rms harmonic current does not only depend on the performance of the pulsewidth modulator, but also on the internal impedance of the motor. Also harmonic currents create the torque harmonics, but lower torque ripple can be obtained at higher current harmonics[11].

There are only odd harmonics in a symmetrical supply voltage. The current harmonics can be calculated by observing induction motor equivalent circuit. Iron and copper losses due to the voltage and current harmonics are observed in addition to supply losses for the desired speed and moment. Iron losses can be eliminated when compared losses for the desired speed and moment. But, copper losses cause that the heat in the motor windings and increase of resistance.

Harmonics effect the losses increase of IM, and the existence of moment variation at the low operation frequencies. The percentage of harmonics in the IM voltage and currents are determined by pulse number in a half-period.

$$V_{ab} = \sum_{n=1,3,5}^{\infty} (4V_s) / (n.\pi) . \sin (n.\beta/2) . \cos n.(wt - \beta/2) \quad (2-17)$$

The Equation 2-17 indicates that the n th harmonic can be eliminated by a proper choice of displacement angle, β , if $\sin(n.\beta/2)=0$ or $\beta = 360/n$ and their harmonic will be eliminated if $\beta=360/3=120$. The unwanted harmonics at the output of single-phase inverters can be eliminated by introducing a pair of symmetrically placed voltage notches as shown in Figure 2-7.

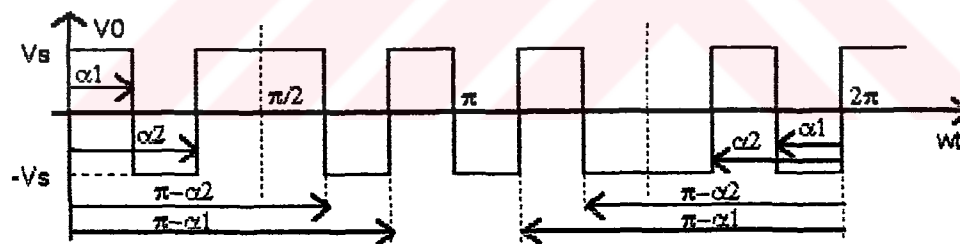


Fig. 2-7. Output voltage with two notches per half-wave

The Fourier series of output voltage can be expressed as

$$V_o = \sum_{n=1,3,5,\dots}^{\infty} A_n \sin (nwt) \quad (2-18)$$

where

$$\begin{aligned} A_n &= (4V_s / \pi) . \left[\int_0^{\alpha_1} \sin (nwt) . d(wt) - \int_{\alpha_1}^{\alpha_2} \sin (nwt) . d(wt) + \int_{\alpha_2}^{\pi/2} \sin (nwt) . d(wt) - \dots \right] \\ &= (4V_s / n.\pi) . \left[1 + 2 \sum_{k=1}^M (-1)^k . \cos(n.\alpha_k) \right] \quad (k=1,2,3,\dots,M) \end{aligned} \quad (2-19)$$

$\alpha_1, \alpha_2, \alpha_3, \alpha_4, \dots, \alpha_M$ are the *firing angles* per quarter cycle. A_1 is the fundamental frequency coefficient while a_3, a_5, a_7, a_9 are the harmonic coefficients. Also k is the number of switches per quarter-cycle.

The modified PWM techniques can be applied to generate pulses which would eliminate certain harmonics effectively in the output voltage, as shown in Fig. 2-8.

If we extend Equation 2-19 for m pulses per quarter-wave.

$$A_n = (4V_s / n\pi) \cdot [(1-2\cos(n\alpha_1) + 2\cos(n\alpha_2) - 2\cos(n\alpha_3) + 2\cos(n\alpha_4) \dots] \quad (2-20)$$

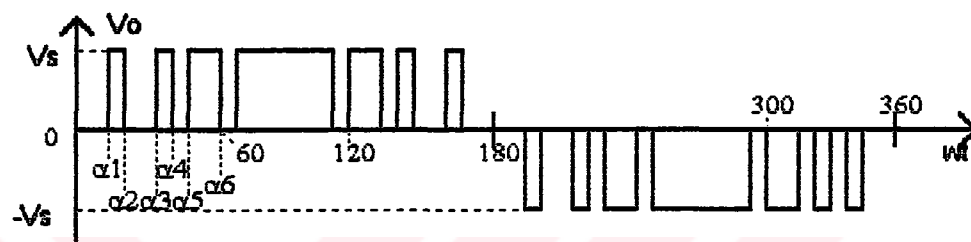


Fig. 2-8. Output voltage for modified sinusoidal PWM

The harmonic elimination methods, which are only suitable for fixed output voltage, eliminates the low-order harmonics. However, switching and magnetic losses (iron) increased in the transformer due to higher harmonic frequencies. It is not always necessary to eliminate the third harmonic and the harmonics that multiples of three (triplen), which are not normally present in three-phase systems. Therefore, in three-phase inverters, it is preferable to eliminate the 5th, 7th and 11th harmonics of output voltages, so that the lowest-order harmonic is the 13th. For elimination the 5th, 7th, 11th harmonics, we can find angles assuming that $A_5=A_7=A_{11}=0$ using equation 2-20. Solutions of these equations yields that $\alpha_1=10.55$, $\alpha_2=16.09$, $\alpha_3=30.91$ deg.

$$1-2\cos(5\alpha_1) + 2\cos(5\alpha_2) - 2\cos(5\alpha_3) = 0$$

$$1-2\cos(7\alpha_1) + 2\cos(7\alpha_2) - 2\cos(7\alpha_3) = 0$$

$$1-2\cos(11\alpha_1) + 2\cos(11\alpha_2) - 2\cos(11\alpha_3) = 0$$

The induction motor filters the low-frequency harmonics, so we think that it is sufficient to eliminate low-order harmonics that close to main component. But, also increasing harmonic frequencies accelerates skin-effect in the conductors that carry currents and causes the current to be stacked on the conductor surface. This changes the load resistance and parameters that used in the system. So if we can do, the most effective way is to eliminate harmonics as many as possible.

2.4.2. Performance Parameters

The output of practical inverters contain harmonics and the quality of an inverter is normally evaluated in terms of the following performance parameters[7]. These parameters are displayed in the Space Vector Simulation program for the desired frequency.

HARMONIC FACTOR (HF): The harmonic factor for nth harmonic, which is a measure of individual harmonic contribution, is defined as

$$HF_n = (V_n / V_1) \quad (2-21)$$

where V_1 is the rms value of the fundamental component and V_n is the rms value of the nth harmonic component.

TOTAL HARMONIC DISTORTION (THD): The total harmonic distortion is a measure of closeness to fundamental component of the waveform, is defined as

$$THD = (1/V_1) \cdot \left(\sum_{n=2,3,\dots}^{\infty} V_n^2 \right)^{1/2} \quad (2-22)$$

DISTORTION FACTOR (DF): The THD gives the total harmonic content, but it does not indicate the level of each harmonic components. If a filter is used at the output of inverters, the higher order harmonics would be attenuated more effectively.

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, because they are divided by n^2 . As a result, 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

$$DF = (1 / V_1) \cdot \left[\sum_{k=2}^n (V_k/k)^2 \right]^{1/2} \quad (2-23)$$

where V_k denotes the amplitude of the kth harmonics of line voltages. The distortion factor of an individual (or nth) harmonic component is defined as

$$DF_n = V_n / (V_1 \cdot n^2) \quad (2-24)$$

LOH: The lowest order harmonic is the harmonic whose frequency is the closest to the fundamental.

CHAPTER THREE

THE SPACE VECTOR METHOD SOFTWARE

3.1. INTRODUCTION

A Space Vector PWM generation method has been implemented by a simulation program written in BASIC language. The coefficients which determine the switching angles are generated as an output file using this program and used in the assembler program as a look-up table for the Space Vector control board. Also the same logic for generating desired frequency PWM waveforms that used in BASIC program, is applied with some changes to the assembler program.

The microprocessing unit was selected as 8098 microcontroller to obtain PWM patterns to drive six power switches using special High Speed Output (HSO) unit of this microcontroller. This unit consists of six high speed outputs which takes less effort from CPU when compared with normal I/O device logic. The features of this unit are described more detailed in this chapter.

3.2. 8098 16-BIT MICROCONTROLLER

From 1970, microprocessor technology has been developed very fast. Manufacturers have produced 16,32,64 bit microprocessors to build more powerful personal computers that can be used for general purpose. Moreover, the stand alone systems are built using these microprocessors for industrial control purpose. Another branch is represented by microprocessors that have been combined with RAM, ROM, various I/O capabilities, programmable timers, interrupt handler unit on a single chip to perform real-time control applications. These microprocessors are commonly referred to as microcontrollers. Unlike microprocessors, microcontrollers are generally optimized for specific control applications in a small, compact and cheap system.

The 8098 [3][4] is a member of Intel's second-generation MCS-96 family that is designed to be used in applications which require high speed calculations and fast I/O operations. The 8098 is a microcontroller that combines many I/O features with powerful

CPU. The 8098 is a 16-bit internal and 8-bit external data bus version of 8096 microcontroller. In the Figure 3-1, functional block scheme of the microcontroller is shown.

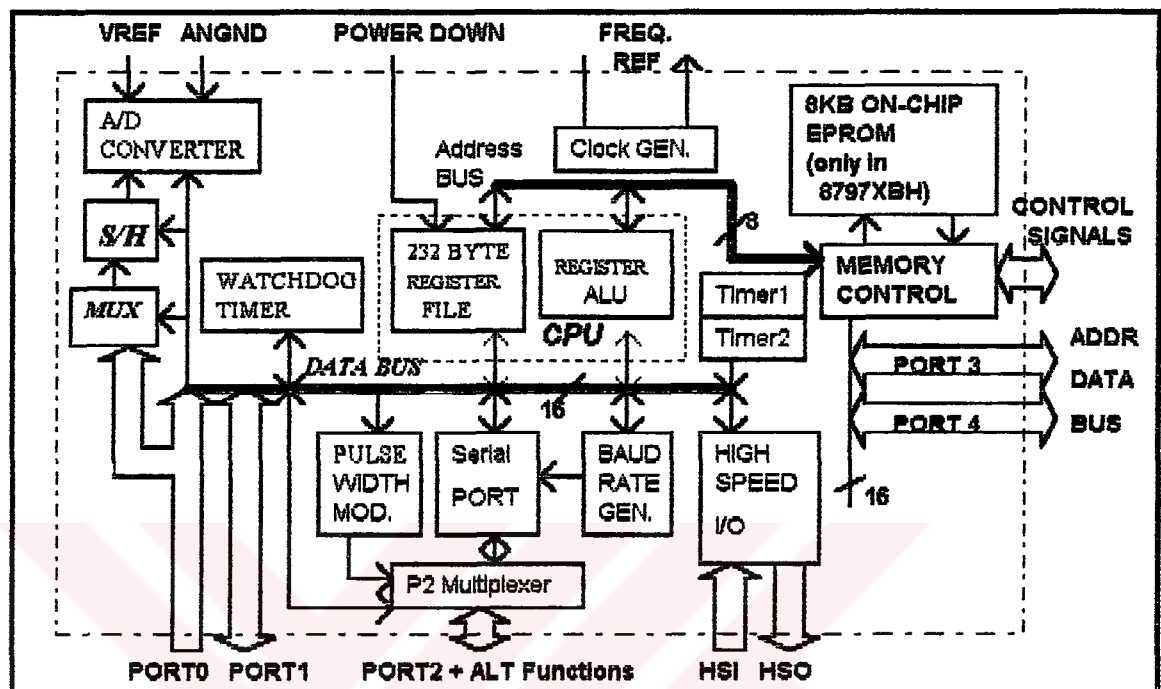


Fig.3-1. The 8098 block diagram

The 8098 CPU uses 16-bit ALU that operates on a 256-byte register file instead of an accumulator. This is called as register to register architecture. Many of the instructions can use bytes or words from anywhere in the 64K memory as operands. The 8096 has 100 instructions and this instructions can operate on bits, bytes, words and double words. Standard logical and arithmetic operations can be performed for byte and word operations. The 8098 has many I/O features to operate with little CPU intervention. The I/O features used in this study are described in the following sections[3][4].

3.2.1. High Speed Output (HSO) Unit

Eight channel HSO unit is ideal for inverter motor control, pulse width modulation applications used in industrial systems. At times determined by the software, the HSO unit outputs can be 1 or 0, to have some devices turn-on, turn-off, start or stop some process. Programmed times can be referenced to either Timer 1 or Timer 2 and HSO unit can schedule all the events that programmed before at specific times. Generally, HSO unit works with Timer 1 to implement the output compare function, also works with Timer 2 to implement the pulse accumulator function. The main component of the HSO is a 23x8 CAM (Content Addressable Memory) file. This file stores up to eight events that are

waiting to occur at specific times. The time value of these entries in the CAM is compared with actual time in free running timer 1 every $2\mu s$. If a match occurs the specified event will be triggered and that location of the CAM will be made for another pending event.

In this thesis, HSO unit is used to generate software interrupts in every ΔT times and to create 3 switching pulse output. (HSO.0, HSO.1 and HSO.2). The other inverter driver signals are the inverted copies of these signals and a dead time generator ensures there will be no short circuit on the line voltages. A dead time generator generates a time delay at the rising and falling edges of three driver signals (described in detail in Chapter 4). CAM has 8 entries at the same time. In the program one of them is reserved for generating software interrupts, the other 7 seven locations are used to program α_1, α_2 and β times properly. So there are no CAM locations left for other purposes.

3.2.2. Software Timers Interrupt

The software timers are used to generate an interrupt after a certain amount of time. These timers are actually 4 software flags in I/O Status Register (IOS1) and controlled by HSO_COMMAND register[3][4]. Timer 1 or 2 are used to trigger interrupt routines which must occur at regular intervals. By using this specification a multitasking process can easily be set up.

In this thesis, Software Timer 0 is used to generate a periodic interrupt in every ΔT times. This must be used to know the start of ΔT times and therefore generate $\alpha_1, \alpha_2, \beta$ switching times. Simply, Software Timer 0 uses free running timer 1 to compare its pre-programmed time written in one of the CAM location. When its time comes, this event data is extracted from CAM and an interrupt generated. In the entrance of this interrupt routine, new Timer 1 value must be set up for the following ΔT time. After process time has been done, a next interrupt comes and we set new ΔT time and so on. But the interrupt subroutine running time must not exceed ΔT time, otherwise a new triggered interrupt comes when the former interrupt routine is on process. After certain amount of time for this confusion CAM is filled up and doesn't accept a new command, hence the program control is lost. The total time that is required to calculate switching angles is obtained and it is seen that ΔT time is sufficient to perform all operations for the selected frequency.

3.2.3. Serial Channel

Controlling a device from a remote location is an important task and generally requires additional hardware. But on-chip serial port of 8098 allows the microcontroller to

transfer data from/to other devices efficiently. In this work, a PC controls the operation frequency and modulation index by its own serial channel and a communication channel is formed between PC and 8098. In a process, generally several motors or other system that can be controlled by an Inverter system are not close to control system. Using the serial channel, all process can be easily controlled from a control room computer.

The serial port has synchronous and asynchronous modes. In the asynchronous mode, maximum 187.5 KBaud, and in the synchronous mode 1.5 MBaud are available. In this thesis, the Inverter system is controlled by PC using the asynchronous serial channel for developing the system. If the vector control algorithm is applied to DSP processor, to use fast synchronous mode has more advantage.

In the Asynchronous Mode, transmit and receive operation is performed on the SBUF (Serial Buffer) register[3][4]. The data that received from other device is detected when RI (Receive Interrupt bit) is set. Also transmit process started by writing the data to SBUF. Eight data bits, one start, one stop bit and the 9600 BAUD rate is selected. Serial port control and baud rate selection can be done using SPCON/SPSTAT and BAUD registers[3][4].

The serial channel is used for two purposes. First, to load program object codes that form new developed program from PC to microcontroller RAM memory, and then execute them as a monitor ROM program. Second, to receive control codes from PC such as frequency and modulation index.

3.2.4. A/D Converter

A/D converter is used to read motor current as a feedback analog signal. The 8098 has a 10-bit A/D converter that can use any of four input channels. The input comes through a sample and hold (S/H) circuit. The conversions are done using the successive approximation method, and require 22 μ sec, permitting a maximum sampling rate of 45 KHz. The A/D converter is controlled by AD_RESULT and AD_COMMAND registers[3][4].

Also the Timer 2 can be used for reading motor current using a Voltage Controlled Oscillator (VCO). The Timer 2 is an 16-bit event counter and counted on both the rising edges and the falling edges of the input signal and the minimum time between edges is 2 μ sec. This corresponds to a square wave input having a maximum frequency of 250 KHz. Once, we convert the current to frequency the Timer 2 CLK input counts these pulses and calculate the current level that corresponds to counted frequency value. So, this is a simple and cheap A/D converter. It is easy to isolate motor current ground from digital ground than A/D converter. The reason is that we cannot use opto-isolator to isolate

analog signals because of nonlinearities of the light emitting diode, but it can be used for square wave transfer optically.

3.2.5. Watchdog Timer

In noisy environments that especially motors or inductive loads driven, there are many voltage spikes on the line voltage. Even, you prevent this noise using efficient grounding and PCB techniques, still there may be unwanted results. If this happens, microcontroller's PC (program counter) or stack space corrupted, so program execution fails and the microcontroller executes random codes.

When the system is resetted, the Watchdog Timer disabled and writing 1EH and E1H sequentially to the same address 000AH enables the timer. Once enabled, it cannot be disabled again. The time-out interval (with 12 MHz crystal) is 16.384 ms. We must reset the Watchdog Timer before time-out to write same bytes to 000AH. If microcontroller lost its control, it can not reset watchdog timer, so it is overflowed and reset the system. Microcontroller is resetted and entered a known state, therefore the problem is solved.

3.3. THE ALGORITHM OF THE ASSEMBLER CONTROL PROGRAM

There is a ROM table that consists of K1, K2 and K3 coefficients. This table is created in QUICK BASIC program and they are verified by observing the graphic results. Then, this table is converted to their binary equivalents and is stored in the microcontroller assembler program. All of the switching angles are obtained from these coefficients.

The assembler program is built on the real-time interrupt that performed in every period of sampling frequency. Also this interrupt based program watches the serial channel for host commands.

The program algorithm is shown in Figure 3-2 to 3-4. If the program is explained part by part there are many sections which making the whole program. These are:

a) INITIALIZATION SECTION: Asynchronous serial channel is established with the 8 data bit, 1 stop bit, no parity bit, 9600 BAUD selections. All constants and variables take their initial values.

b) MAIN PROGRAM LOOP: In this section, received serial channel commands (Direction, Start, Stop, Hertz, Modulation Index) are evaluated. After processing one ΔT time, (for 3600 Hz sampling frequency, approximately 278 μ sec.) serial channel interrupt is enabled and the program returns to this loop. If 2-bytes serial channel command is received, the

serial command register is equal to number 2, and this loop comes to end. Then the commands that mentioned above, are processed. This loop can be broken by either the Software timer interrupt or the serial channel receive interrupt.

c) SOFTWARE INTERRUPT: In this subroutine, firstly the next software interrupt time (ΔT time later) is set up. Then the serial channel interrupt is disabled to prevent the missing HSO commands. The writing HSO commands for three channel has the first priority in the program. The K_1, K_2 and K_3 coefficients for the desired frequency are fetched from the look-up table. The look-up table was prepared by the QBASIC simulation program before. The coefficients look-up table is based on calculated 2Hz data for 60° of its total period. The higher frequencies can be easily obtained changing the steps. For example, for 2Hz, the look-up table pointer is incremented by three after reading three coefficients. For 4Hz, the pointer is incremented by six $[3 * (300/600/4) - 1]$. For 50Hz, it is incremented by 24 $[3 * (300/600/50) - 1]$. Then the α_1, α_2 , and β switching angles calculated by multiplying with the modulation index parameter M . Now we are ready to write next ΔT time switching data on HSO channels. At the same time the data that written before has been processed by the HSO unit. After writing all the data for the next software interrupt, the serial channel interrupt is enabled again that gives some time to receive command bytes. When the interrupt return command has reached, the program control is left to the main loop section.

d) SERIAL INTERRUPT: In this interrupt subroutine, the bytes that received from the PC serial channel are stored after the count register has reached number 2. At the end of the subroutine, program control passes to the main loop section. When this command is operated the count register is loaded by number 0 and now we are ready to receive another command.

e) FETCHSW SUBROUTINE: This subrotline is very important. The pointer that controls the switching data look-up table position (swpoint) is loaded for the next adress. The switching data are changed for three phases every 60° of the total period. These data are shown in the Figure 2-3b and Figure 2-7 of the second chapter. If the one 60° is finished, the swpoint is resetted to the initial value. Then the switching data are put into the beta, alfa1 and alfa2 registers.

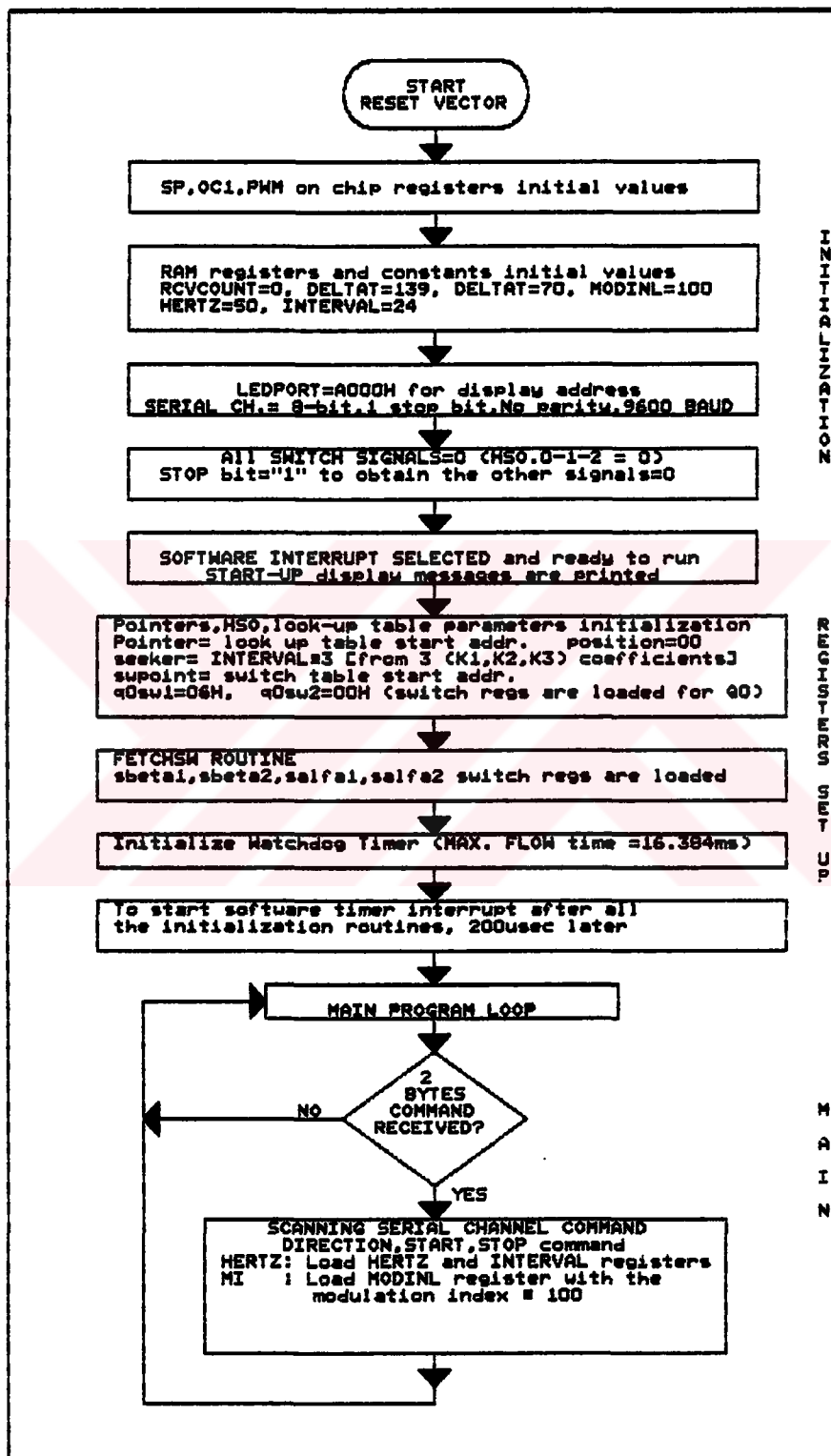


Fig.3-2. Assembler Program Flow-Chart (1)

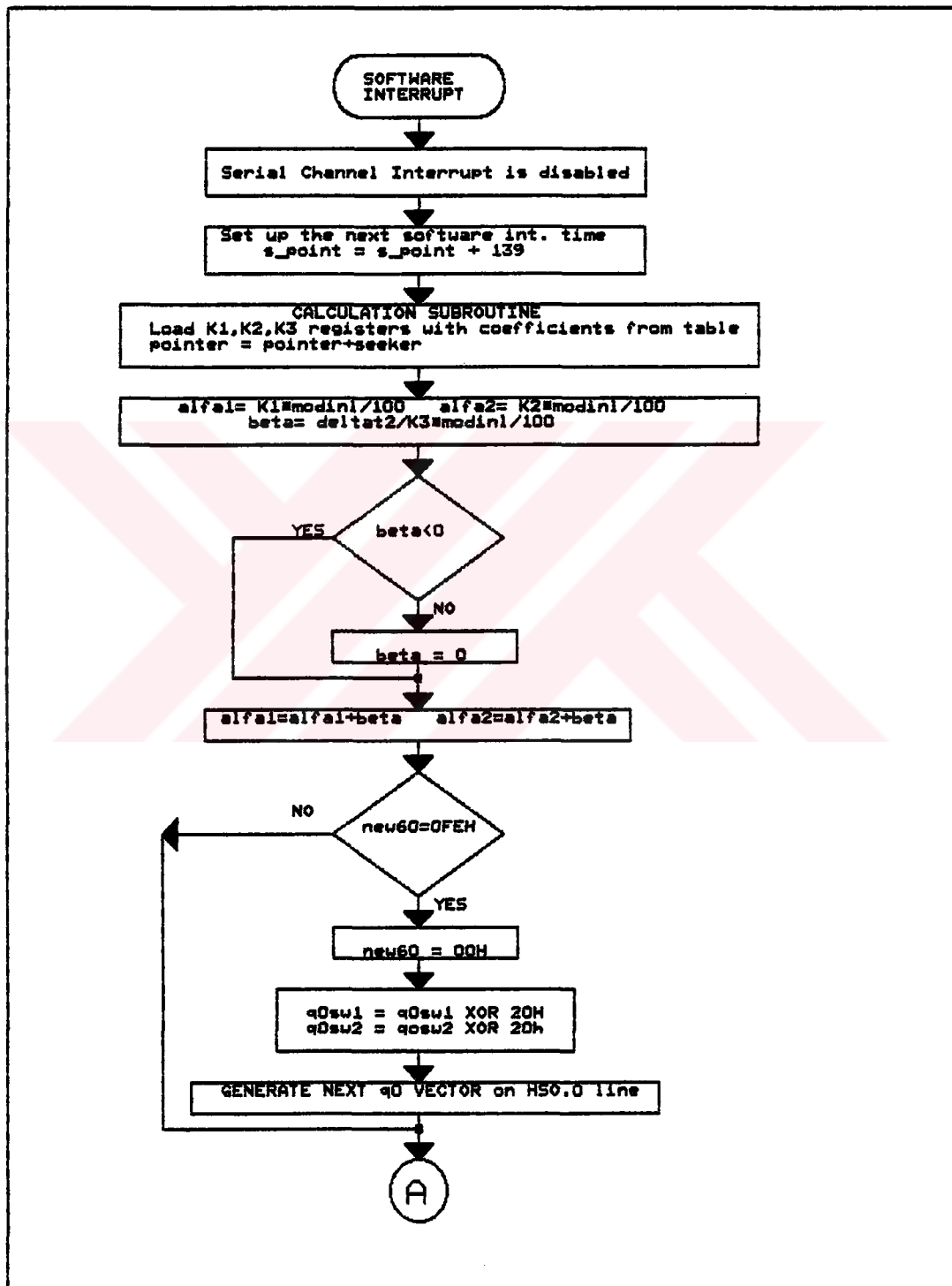


Fig 3-3. Assembler Program Flow -Chart (2)

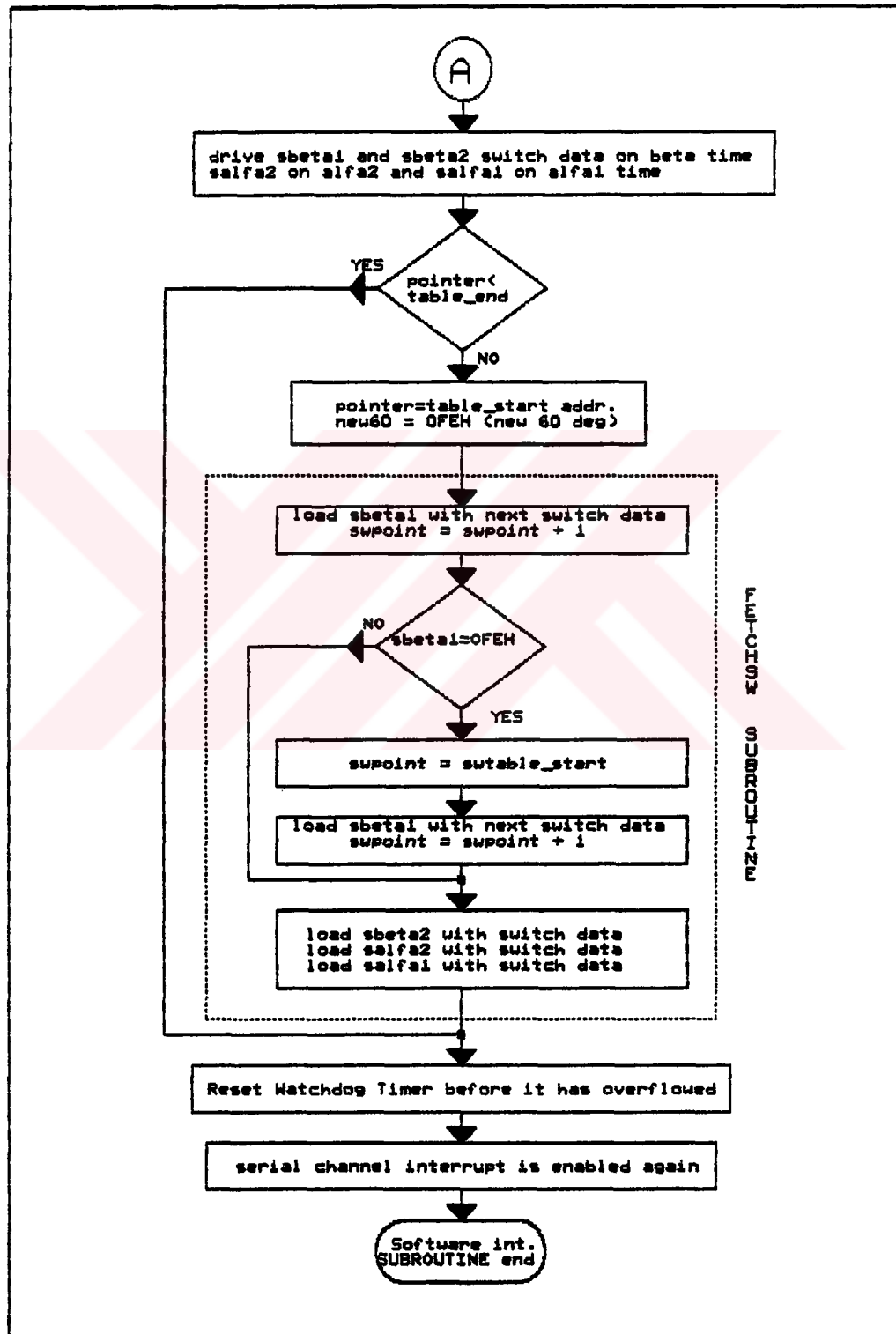


Fig 3-4. Assembler Program Flow-Chart (3)

CHAPTER FOUR

SPACE VECTOR INDUCTION MOTOR DRIVE CIRCUIT

4.1. INTRODUCTION

In this chapter, the overall induction motor drive system is described. Five main parts of the total drive circuit are given. First part is the microcontroller board. Second part is the inverter circuit that includes IGBT driver circuit. Third part is the induction motor. Fourth part is the system power supply. Fifth part is the 3-phase AC-DC rectifier circuit.

A voltage source inverter is introduced and its specifications are explained with some illustrations. The IGBT characteristics and drive requirements are described briefly. IGBT driver circuit and IGBT protection circuit are given. The induction motor structure, the equivalent circuit, operating characteristics and speed control of the induction motor are explained.

4.2. THE SPACE VECTOR PWM SYSTEM

This system consists of five blocks. These blocks will be explained in this chapter. The overall system block scheme is shown in Figure 4.1.

The first block is the microcontroller board and contains a 8098 microcontroller, 32KB EPROM, 8 KB Simulator RAM, 8 KB Data RAM, RAM and EPROM switch logic, serial channel voltage translator logic, the driver signals inverter with dead time generator circuit and a LCD display. The second block is the inverter board and contains IGBT gate driver circuit with opto-isolator, opamp, transistors, and IGBTs with fast recovery diode, RC snubber circuit, and heat sinks. The overcurrent protection and current measurement circuits are explained briefly. The third block is the 3-phase diode rectifier section. The fourth block is the IGBTs driver and microcontroller board power supplies energized from one of the phase voltage. The fifth block is the Three-Phase Squirrel Cage Induction Motor. Also in the thesis, a Personal Computer controls the microcontroller board with its serial channel. This PC can change the AC motor speed and direction from a remote location. The start and stop condition also can be controlled.

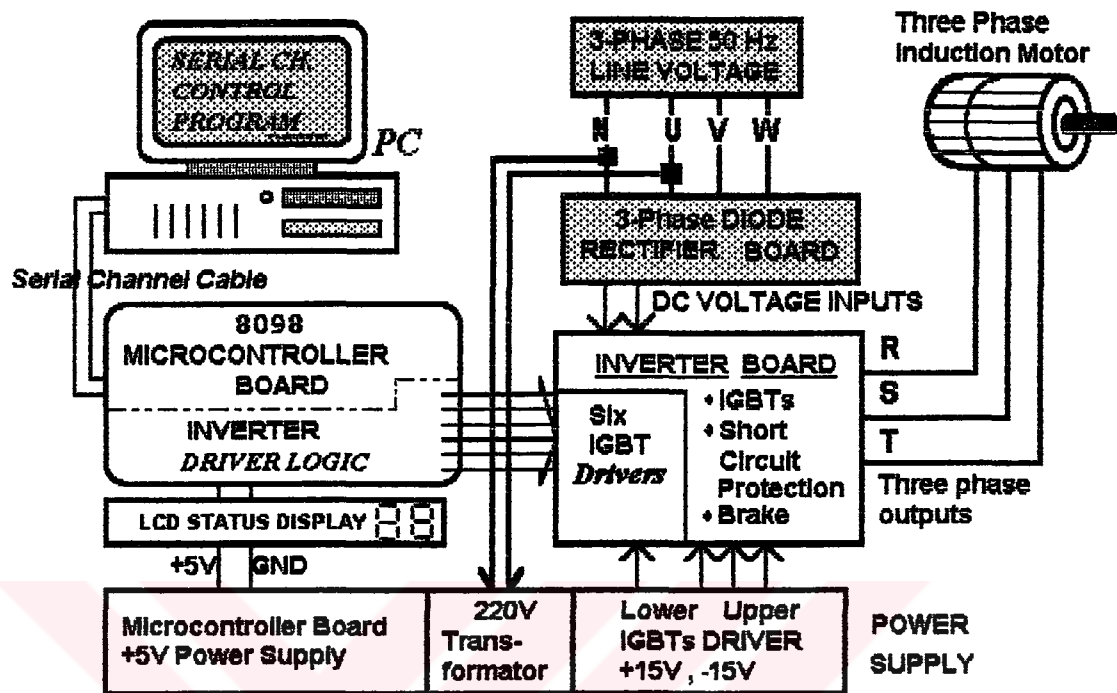


Fig. 4-1. The Overall System Block Scheme

4.2.1. 8098 MICROCONTROLLER BOARD

In this section, the all external devices in the microcontroller board, explained briefly and their circuit scheme are given. The control board block diagram is shown in Fig. 4-2. This board is designed as a software emulator which is loaded from PC using serial link. In the emulator board, a PC serial link controlled logic switches the monitor program memory and the program which has been under development process. After downloading from PC to emulator board has been done, the microcontroller sees no monitor program memory, but our simulated ROM memory.

The microcontroller board includes 32K monitor program EPROM, 8K simulator RAM, 8K data storage RAM, serial channel voltage converter logic, 16 digit numeric LCD display module with serial data input to display system conditions and results, and the deadtime generator to obtain six inverter driving outputs. All microcontroller board components will be explained briefly.

4.2.1.1. Monitor EPROM

This is a 32KB EPROM and in this EPROM memory, there is a program which controls the serial channel till receive a character that initiates load process of the Space

voltage level translation for transmitted data (TXD), received data (RXD), also switch logic inputs and reset control line are supplied by TLC232 and 1489 receiver integrated circuits.

4.2.1.4. 7-Segment 1x16 Character Display

In the microcontroller board, we need a display unit that shows the conditions of the control system. This unit is a complete self contained liquid crystal display module which displays one line of 16 seven-segment digits including decimal points[8].

The device consists of a liquid crystal display in twisted nematic technology and all necessary driving circuitry. It is packed in a compact transparent housing which protects the whole LCD. Data to be displayed are transferred via a serial data bus. A built-in refresh is provided (internal RAM). The controlling external microcontroller has to write information into the drivers only if the displayed pattern has changed. The CPU overhead for driving the module is dramatically reduced in this way. The data input format can be chosen to display numeric information with some possible alphabetic words that can be defined in a 7-segment display. The controlling microcontroller is connected to the display module via a 4-line serial databus (I²C-bus compatible).

4.2.1.5. The Driver Signals Inverter with Dead Time Generator Circuit

This circuit is used to prevent short circuit on power switches at the same arm, when switching occurs at the same time. In other term, one switch control input has high to low transition and the other switch control input has low to high transition at the same time. That means, one switch is closed and the other is opened. But these two power switches are not ideal switches, they require small amount of time to turn-off. There is no problem with the one which will conduct, but to turn-off the other switch takes some time, so there is a small amount of time that short circuit occurs. This is called as commutation. The driver signals inverter circuit has two functions. One of them is to obtain lower arm switches drive signals, and the other is provide some delay that allows to turn-off one of the power switch. The circuit scheme and the function of this circuit is shown in Figure 4-3. HSO.0, HSO.1 and HSO.2 generate the upper arm IGBTs driver signals. Lower arms signals are the inverted and dead-time processed copies of the upper arms signals. The dead time width is selected as 1 or 2 μ s. The HSO.3 is logic 0 when the microcontroller is resetted, so at the initialization of the system all the six driver signals are guaranteed equal to zero. This is called as STOP condition. For obtaining the driver signals, simply we make HSO.3="1", and this condition is called START condition. For braking, we make all

the upper arms driver signals are "1", and lower arms are "0" at the START condition. Therefore the rotating magnetic field is stationary and the induction motor cannot turn.

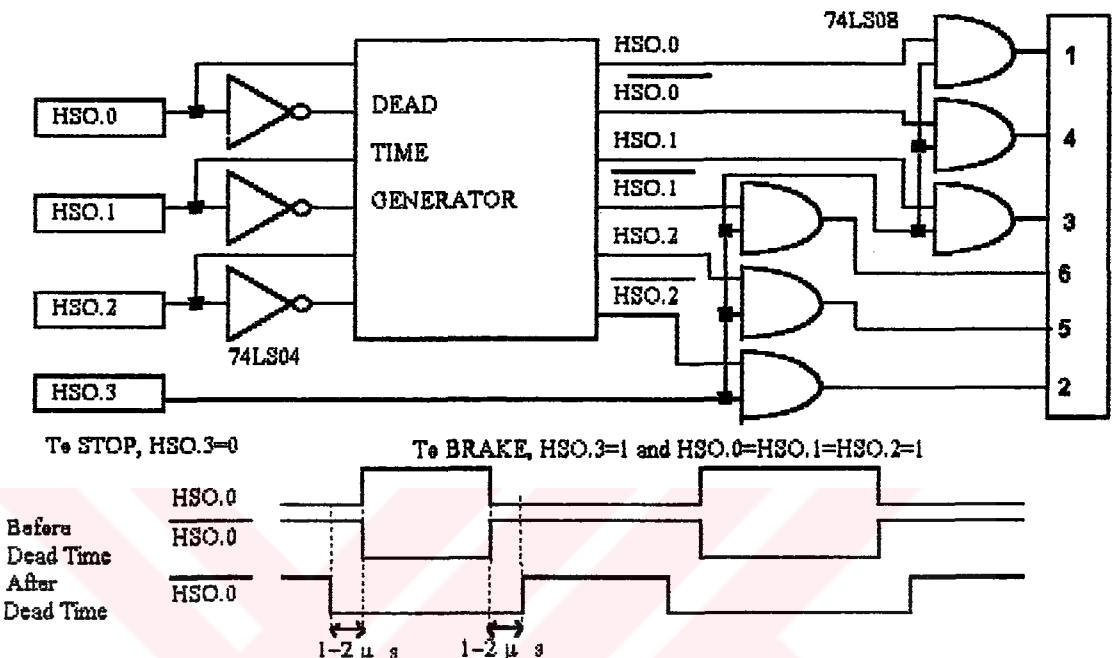


Fig. 4-3. The Driver Signals Inverter with Dead Time Generator Circuit

4.2.1.6. Direction Control Circuit

The three-phase, IM is well suited to reversing rotation. This reversal is accomplished by switching any two of the phases. A Quad 2-input selector IC is used to realize this control. It is basically a 4-pole 2-position switch for logic signals. This IC has a select (SEL) input. When SEL is LOW, the A inputs are passed through to their respective Q outputs. For SEL HIGH, the B inputs appear at the outputs. ENABLE (E') HIGH disables the device by forcing all outputs to LOW. This circuit is shown in Figure 4-4.

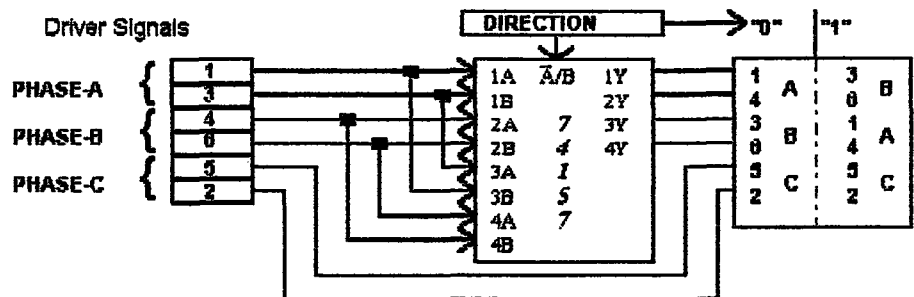


Fig. 4-4. Direction Control Circuit

This switching of two phases also can be achieved in the software. The only change is the HSO.0 and HSO.1 respectively. This is very simple and requires no circuit.

When AC Induction motor runs, we can not turn the direction until the motor stopped. If we want to turn the direction, first we must stop the motor for using STOP procedure described before. Then, we start the motor after switching of two phases.

4.2.2. INVERTER

The Inverter is a device that changes a DC input voltage to a symmetrical AC output voltage of desired magnitude and frequency [5][6][7][8]. Generally, the DC input voltage is fixed, so a variable output voltage can be obtained by varying the gain of inverter, which is accomplished by pulse width modulation (PWM) control of the inverter.

The ideal inverter output voltage waveforms should be sinusoidal, but it contains certain harmonics. For low and medium power applications square wave and quasi-square wave voltages may be acceptable, but for the high power applications the distortion of sinusoidal waveforms must be kept low. Using high-speed power semiconductor devices, the harmonic contents of output voltage can be reduced by switching techniques.

Inverters are widely used in industrial applications such as variable speed AC motor drives, induction heating, uninterruptible power supplies (UPS). The input may be a battery, fuel cell, solar cell, or DC line voltage that obtained from AC voltage.

In voltage source inverter type input voltage remains constant, in current source inverter input current is maintained constant, and in variable dc linked inverter input voltage is controllable. In this thesis, 3-phase voltage source inverter is used. Three-phase bridge inverters[7] are normally used in high power applications. A three-phase output can be obtained using the inverter that shown in Figure 4-5.

In this circuit, the switching elements conduct for 180°. When transistor Q1 is ON, terminal-a is connected to the positive DC voltage. When Q4 is ON, terminal-a is now negative DC voltage. In this figure we obtain six division in a period. The gate signals are also shifted from each other by 60 deg to obtain three-phase balanced voltage.

The instantaneous line-to-line voltage, v_{ab} , can be expressed in a Fourier series, recognizing v_{ab} shifted by $\pi/6$ and the even harmonics are zero. We can notice from equation 4-1, the triplen harmonics would be zero in the line-to-line voltages.

$$v_{ab} = \sum_{n=1,3,5,\dots}^{\infty} V_n \cdot \sin [n \cdot (\omega t + \pi/6)] \quad , \quad v_{bc} = \sum_{n=1,3,5,\dots}^{\infty} V_n \cdot \sin [n \cdot (\omega t - \pi/2)] \quad , \quad v_{ca} = \sum_{n=1,3,5,\dots}^{\infty} V_n \cdot \sin [n \cdot (\omega t - 7\pi/6)]$$

$$\text{where } V_n = (4V_g)/(n\pi) \cdot \cos (n\pi/6) \quad (4-1)$$

With resistive loads, the diodes across the transistors have no functions. If the load is inductive, the current in each arm of the inverter would be delayed to its voltage. When Q2 is OFF, negative line current i_c flows through D5. Hence the load terminal-a is connected to the DC source through D5 until the load current reverses its polarity at $t=t_1$. During the period for $0 \leq t \leq t_1$, Q5 will not conduct. Similarly Q2 will start to conduct at $t=t_2$. Line-to-neutral voltage and line current waveforms that is similar to sinusoidal waveform due to the low-pass filter nature of RL load are shown in Figure 4-5.

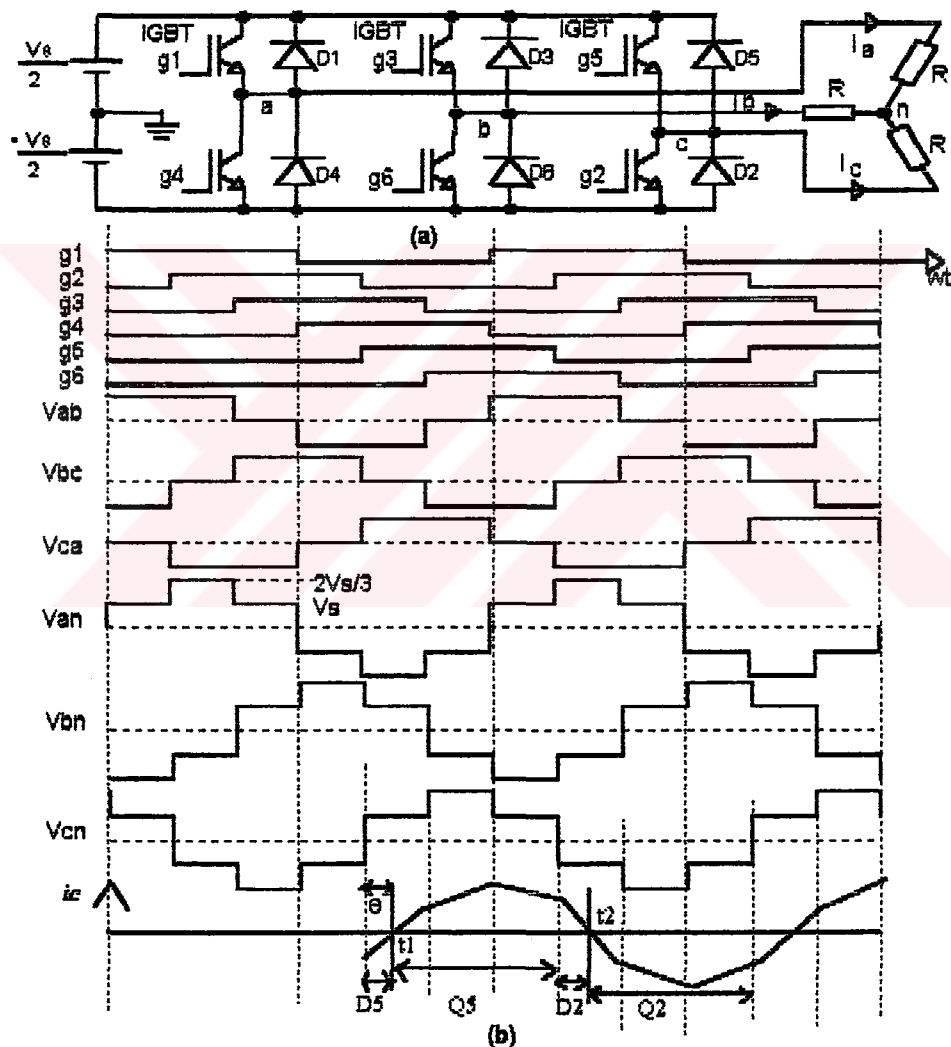


Fig. 4-5. Three-phase bridge inverter a) Circuit b) Waveforms

The inverter circuit that used in this thesis, consists of two main blocks. The first part is the 3-phase AC to DC diode rectifier, and the second part is DC to AC inverter section. The main switching element is IGBT, and also IGBT is use in current limiting and brake circuits. The overall inverter circuit scheme is shown in Figure 4-6.

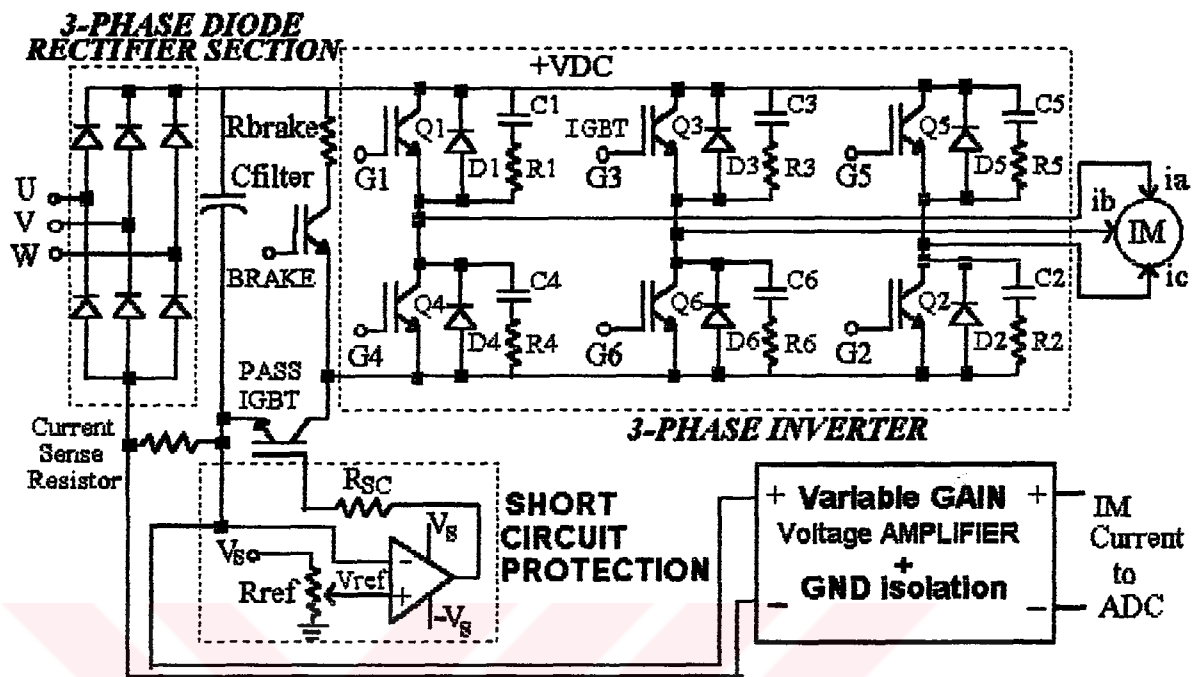


Fig. 4-6. The Overall Inverter Circuit Scheme

The load is Induction motor and its equivalent circuit shows that it is Inductive load. The Inductance wants to continue its own current in a counter direction after the applied voltage is switched off. In the Inverter circuit, the IGBT devices are switched-on and off many times, and after switching off the current spikes may be harmful for the IGBT devices. A parallel commutation diode recovers these spikes and conducts when the IGBT switched off and prevents the switching elements from high currents. Also the dV/dt and dI/dt limiting circuit will be described later.

In the inverter circuit, a current sense resistor which the resistor value is small, converts the current that drawn by induction motor. This resistor must be selected as high wattage resistor because of the high current that flows through. The converted voltage simply compared with reference voltage to prevent short circuit currents. If the reference voltage is adjusted to 5 amperes, that means this value is the short circuit current value, and the current cannot be over this value. This is achieved by the opamp comparator circuit. If the current is lower than the short circuit value, opamp inverting input voltage is lower than the non-inverting input voltage, and the output of the opamp is the positive power supply ($V=15V$). When this conditions exists, the pass IGBT is turned-on and the current can flow, so the Induction motor runs. But, the non-Inverting input voltage exceeds the reference voltage, the opamp output voltage is the negative power supply. ($-V=-15V$). At this time IGBT is turned-off and the overall circuit is protected from the high current.

We can measure this current value by using variable gain voltage amplifier that amplifies the current sense resistor voltage. If the output of amplifier is adjusted to 0,+5V range, we can use ADC converter to read the actual current. We must isolate inverter and microcontroller board ground line by an isolation circuit. That can be an isolation transformer or opto-coupler circuit. The current measurement is described in the 8098 ADC converter section more detailed.

The brake circuit is necessary when we want to stop the induction motor suddenly. At this time, we turn-on the upper arms IGBTs and turn-off the lower arms IGBTs, so the magnetic field is not rotating now. For this reason, the rotor of the induction motor stops, but the induced counter emf in the windages, is harmful for the switching devices. To eliminate this extra voltage, simply we turn-on the brake IGBT. Then, all the currents is passed through the collector resistor, and extra power is consumed. After that, we can turn-off the brake IGBT. Now, we can run the induction motor again.

4.2.2.1. IGBT (Insulated Gate Bipolar Transistor)

The ideal semiconductor for power switching applications[7] should exhibit a low forward voltage drop in order to keep the forward conduction losses small and also a high turn-on and turn-off speed in order to keep the switching losses small. In addition, the gate drive power requirements should be low and the device should operate at a high current handling density in order to minimize both the size of the device and required cost.

For applications, which do not require gate turn-off, the thyristor is the best choice because of its high forward conduction current density, even at the high operating voltages. However, many thyristors require reverse anode potential to achieve turn-off. The power bipolar transistor has been extensively used for applications where the load current is controlled by the input signal. Bipolar transistors can be turned-off less than 1 μ sec, but requires high base drive current both during the on-state and during turn-off. For applications which requires gate turn-off capability, gate turn-off thyristor (GTO) and the Power MOSFET have been developed. The GTO operates at higher current densities than the bipolar transistor, but cannot operate at high switching speeds and requires very high gate turn-off currents. In contrast, the Power MOSFET is capable of very high switching speeds and requires low gate drive currents. However, these devices operate at much lower current densities than the bipolar transistor and have no more costly and complex driving requirements.

In this thesis, Insulated Gate Bipolar Transistor (IGBT) is selected as a switching element of the inverter section. This IGBT is the IXYS 10N100A. The IGBT [6][9][14], has

the drive requirements of a power MOSFET and the voltage and current handling capabilities of the bipolar transistors. Also, IGBT overcomes the high on-state loss power MOSFET's while maintaining the simple gate drive requirements of MOSFET's. The IGBT combines both the bipolar transistor and MOSFET structure as shown in Fig. 4-7a.

Since the IGBT has a low power gate drive requirement, a high current density capability, and a high switching speed while maintaining ON voltage equal to bipolar Darlington transistors, it is preferred over other devices in many high power applications. As a result, it combines the speed, ease-of drive, and wide safe operating area of a MOSFET, with the low saturation voltage of a bipolar device and has short circuit characteristics superior to both.

IGBT has several families and generations. Devices are now available with ratings of 1200V, 10-400 A, current fall-times of about $1\mu\text{s}$ and on state voltage drops at full current of the order of 2.5-5V depending upon current rating. At lower power levels, devices having ratings of 500V, 5A, a saturation voltage of about 2V and a current fall-time of 250ns are also available[9][14].

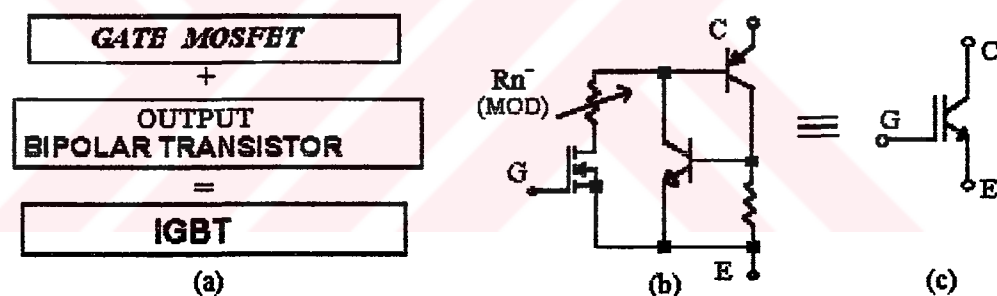


Fig.4-7. (a) IGBT structure (b) IGBT equivalent circuit (c) symbol

When a voltage is applied to the gate terminal, the gate-emitter voltage is increased to exceed a threshold level and a conducting channel is formed. Current is then introduced to the base of the pnp transistor via the FET, and this current initiates IGBT conduction. To turn-off the IGBT, the channel is removed by reducing gate-emitter voltage to zero. This breaks FET channel current that removes base drive to the pnp transistor. The cross-section, equivalent circuit and IGBT symbol are shown in Fig. 4-7b and 4-7c.

4.2.2.2. IGBT Driver

Transistors require certain turn-on and turn-off times. During turn-on, the collector current rises and the slope is called di/dt . During turn-off, the collector-emitter voltage must rise in relation to fall of the collector current, and the slope is called dV/dt .

The simple IGBT driver [6][14] is shown in Fig. 4-8a. The typical voltage and current waveforms of an IGBT is shown in Fig. 4-8b. When S_{on} is closed, the gate capacitance begins to charge up to t_0 , the gate threshold is reached and the collector current begins to rise [9]. During this time the voltage seen across the device (V_{CE}) drops due to the voltage drop across the power circuit inductance. At time t_1 , the current in the IGBT reaches the level of the load current and the direction of diode current flow reverse until at time t_2 , the diode begins to recover its reverse blocking capability. At time t_2 , the peak IGBT current is reached a level that should not exceed twice its nominal rating. Also, the *snappiness* of the diode recovery (effectively di/dt between t_2 and t_3) can cause a number of application problems; such as overvoltage on the diode, high dV/dt , oscillation, dV/dt turn-on of other devices. Therefore, we need fast recovery diodes when the IGBT is turned on fast. The di/dt can be controlled with increasing $R_g(ON)$, but the turn-on switching loss increases.

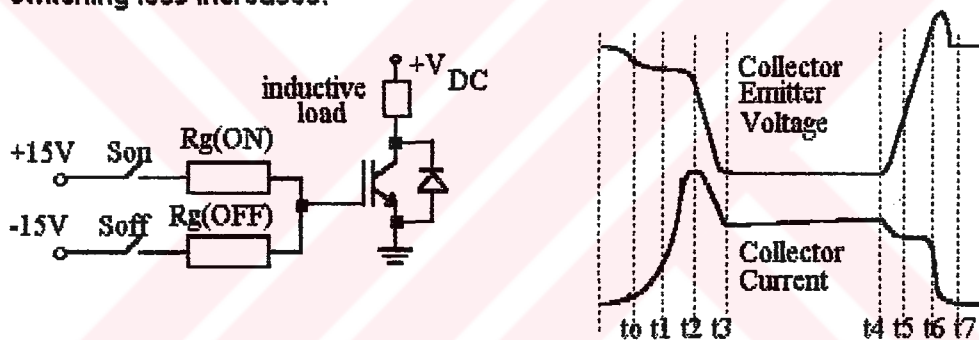


Fig. 4-8. (a) Simple IGBT Driver

(b) Typical current and voltage waveforms of IGBT

When S_{off} closed and S_{on} opened, the gate capacitance is discharged and the voltage across the device begins to rise to the supply voltage level(15). At this point, the free-wheeling diode begins to conduct, so the IGBT current can now fall. The fast di/dt with which the IGBT switches off in combination with the circuit stray inductance causes an over voltage across the device. At t_6 , the IGBT tail current appears. The current fall time and tail time is relatively independent of the value of $R_g(OFF)$. It is necessary to limit the dV/dt across the device by increasing $R_g(ON)$ on the device switching on.

Ready-made drivers are available today, but these drivers can be expensive and may not allow a user to optimize the performance of the IGBT for a specific application. Like a MOSFET the IGBT is a voltage driven enhancement mode device, with a capacitive input. This means that driving IGBT requires pulses of current to charge and discharge the gate capacitance during switching, but almost no power in steady state.

For using PSPICE analysis program, an IGBT switching test circuit is analyzed and its switching characteristics are obtained. An IGBT is driven by selected pulse

waveform and an inductive load is switched by this device. The proposed circuit and the analysis results can be seen in Appendix B.

There are two ways to isolate microcontroller board from inverter board with respect to ground[7]. The first one, Pulse transformers, have primary and secondary winding. Simply this is a transformer that transmits the input signal to the output terminals by magnetic field. So, there are no electrical connections between the input and the output signal. The transformer should have a very small leakage inductance and the rise time of the output pulse should be very small. At a relatively long pulse and low switching frequency, the transformer would saturate and the output would include distortions. The second method is using optocouplers.

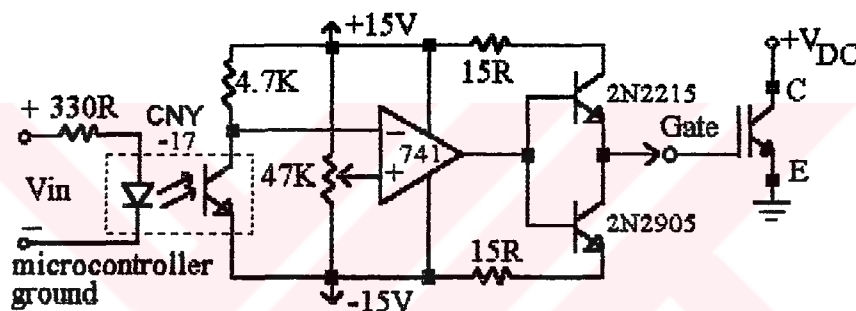


Fig. 4-9. IGBT Driver

In this thesis, the CNY-17 optocoupler is used for isolation. The transfer of the gate signal to that IGBT gate is achieved by the light emitting diode and a photo transistor. The rise and fall times of photo transistor are very small (turn-on time is typically 2-5 μ sec, and turn-off time is 300nsec). So we select a high speed optocoupler to transfer the high-frequency signals. Also the isolation voltage is typically 1-2kV. This ratings satisfy the driver circuit design. An OPAMP is used to invert optoisolator output and amplify the signal to driver transistors. This OPAMP is selected as 741. This opamp slew rate is .5 V/ μ sec and this gives us 10-15 kHz speed. This speed is very low, but the important parameter is the output current that drives the transistors. This current is approximately 20mA. In this thesis the driver circuit that is used is shown in Figure 4-9.

4.2.2.3. IGBT Protection Circuit

Due to the reverse recovery process of power devices and switching actions in the presence of circuit inductances, voltage transients occur in the circuits. Or short-circuit conditions may exit, resulting in an excessive current flow through the devices. Protection circuits are normally required to keep the operating di/dt and dv/dt within the allowable

limits of the IGBT. The RC network across the IGBT is known as the *parallel circuit* and limits the dV/dt . Also the series inductor limits the dI/dt and called a *series snubber*. The snubber circuits is required to maintain the voltage and current relationship of IGBT during turn-on and turn-off. It protects the IGBT from supply and load side voltage transients. It also reduces the magnitude of peak recovery voltage. The circuit is shown in Fig. 4-10.

With a RC-snubber circuit [7][8], the voltage across the IGBT will rise exponentially. With this high slope, we apply the turn-on signal to the gate when the conduction channel is not completely formed. This sudden voltage change destroys the IGBT due to the excessive currents. If the rate of rise of collector current is very fast compared to the velocity of turn-on process, heating will occur due to high current density and device can be destroyed as a result of excessive energy in a small time. The serial snubber stores this energy due to the reverse recovery time of the IGBT.

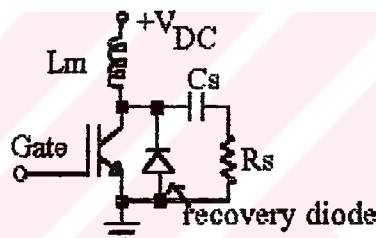


Fig. 4-10. Parallel and series snubber circuit

The recovery diode function is described before. The most important specification is the recovery speed. This diode is selected as IXYS 12-10A and the diode maximum voltage is 1200V and maximum current is 10A. The recovery time is 550nsec.

Also the heat produced by losses in a semiconductor device must be dissipated sufficiently and effectively to operate the device within its upper temperature limit. We must ensure that all ratings should not be exceeded using protection against overvoltage, overcurrent, and overheating. The excessive heat can be dissipated by using heat-sinks.

4.2.3. IGBT DRIVER POWER SUPPLY

A special power supply unit is required to drive six IGBTs gate inputs. This power supply must be capable to source at least 1 ampere current each. For fast turn-on, the gate capacitance of IGBT must be charged very rapidly, and this requires high current. The other important specification about the power supply is isolation between upper arms IGBTs ground lines. To achieve this separate transformers are used. But there is no problem about lower arms IGBTs, because their grounds also at ground voltage. So, we can use only one transformer that can source three times IGBT gate current. Also, for microcontroller board we need extra +5V power supply. This +5V power supply must give

at least 1 ampere current. As a result isolated 5 transformers are required. The one of these transformers supply the lower arms IGBTs gate +15V,-15V driver circuits. This transformer must give the power three times the value of required power. These IGBTs ground lines are connected together. The emitter terminals of the lower arm IGBTs is also the ground of the DC voltage that obtained from the three-phase AC/DC converter. But, upper arms IGBTs emitter terminals are connected to the input lines of the three-phase induction motor. So, this power supplies must use isolated transformers.

The +5V, +15V, -15V regulated voltages are obtained from the regulator chips. For the positive voltages LM317, and for the negative voltages LM337 are used. The basic power supply scheme [10] is shown in Figure 4-11. The output voltage adjustment can be done by R2 resistor. The output voltage formula is given in the Figure 4-11.

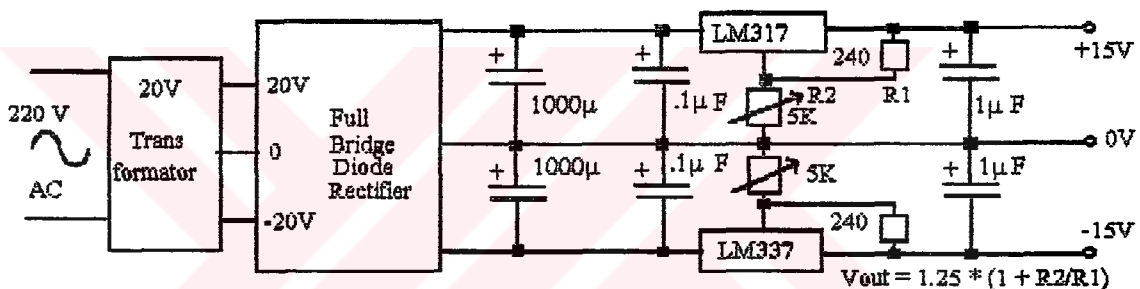


Fig. 4-11. Basic +15V,-15V Power Supply

4.2.4. THREE PHASE DIODE RECTIFIER

The three-phase diode rectifier converts the three-phase AC line voltage to the DC voltage and this DC voltage is not adjustable. The rectifier diodes are selected as IXYS DSI17-12A. The maximum current of this diode is 12A. The rectifier diodes have slow-recovery characteristics and are available in high voltage and current ratings such as 5000V and 2000A.

The filter capacitor ratings at the DC output of the rectifier are 3300 µF and 400V DC. This means, we can obtain a DC voltage under 400V DC. The block scheme of this rectifier is given in the overall inverter scheme. (Figure 4-6).

4.2.5. INDUCTION MOTOR

The interaction of conductors in a magnetic field, produces driving torque of both DC and AC motors. In the DC motor, the armature moves and the field is stationary. Currents in both field and armature windings create the magnetic fields that produce the torque. The rotor current comes from the supply through brushes and commutator. In the induction motor (IM), electromagnetic induction produces the rotor currents[12][7][8].

The rotor consists of axial conductors shorted at the end by circular connectors to form a squirrel cage. The stator is identical to the stator of a synchronous motor. As the stator field moves past the rotor conductors, emfs are induced by transformer action. Because of the reactance of the rotor windings, the current and therefore the rotor field lag the emf by angle θ , and the torque angle is greater than 90° . The induced emf is a maximum in the rotor conductors under the poles. Torque production of induction motor, rotor construction and cutaway view of induction motor is given in Figure 4-12.

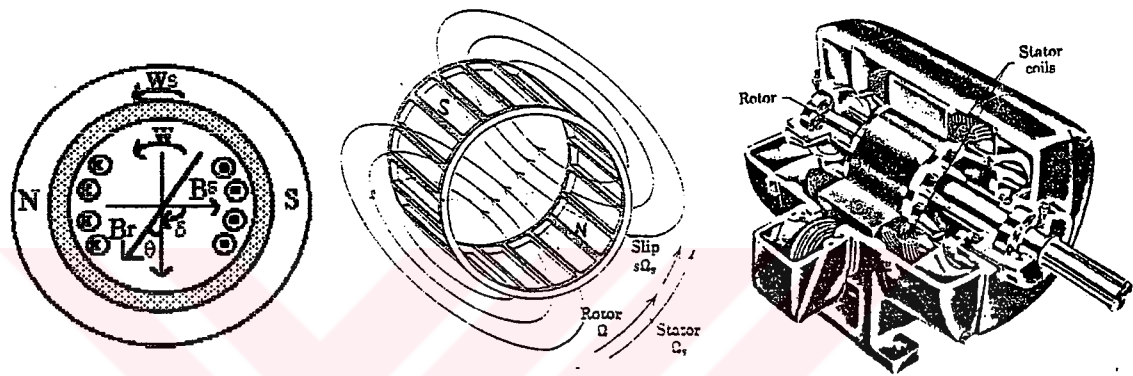


Fig. 4-12. a) Torque production of induction motor b) rotor construction c) cutaway view

The rotor must always rotate at a lower speed w from the synchronous speed. If the rotor could turn at the synchronous speed there would be no change in flux linkage, no induced current, and no torque. The relative speed of the rotor to the synchronous speed of the stator flux is known as the slip. By definition, the slip is

$$s = \frac{\text{syncro speed (stator flux speed)} - \text{rotor speed}}{\text{syncro speed}} = \frac{ws - w}{ws} = \frac{Ns - N}{N} \quad (4-2)$$

In this equation, slip will range between 0 and 1. Slip is usually expressed in percent. The slip at full load for a squirrel-cage induction motor is typically 2 to 5%.

The rotation speed of the stator field is called the synchronous speed.

$$w_s = (2\pi f) / p = (120f) / p \quad \text{synchronous speed} \quad (4-3)$$

where p is the number of poles and w is the supply frequency in rad/s.

The equivalent circuit [7][8] for one phase of the rotor is shown in Figure 4-13a, where R_r' is the resistance, the X_r' is the leakage reactance of the rotor windings, and E_r is the induced phase voltage when the speed is zero (or $s=1$). The rotor current is given by

$$I_r' = (s.E_r) / (R_r' + jsX_r') = E_r / (R_r' / s + jX_r') \quad (4-4)$$

The per phase circuit model of IM is shown in Figure 4-13b, where R_s and X_s are the per phase resistance and leakage reactance of the stator winding. The complete circuit model with all parameters referred to the stator is shown in Fig. 4-13c, where R_m represents the resistance for excitation (or core) loss and X_m is the magnetizing reactance.

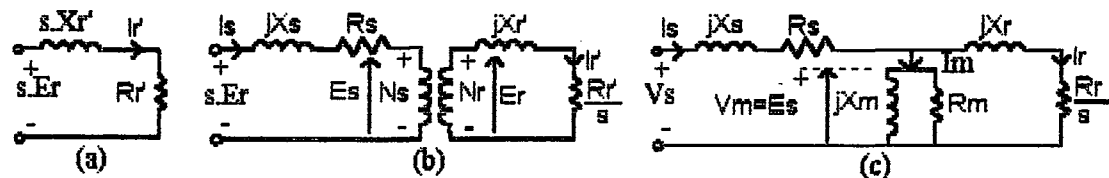


Fig. 4-13. a) Rotor Circuit b) Stator and Rotor Circuit c) Equivalent Circuit

4.2.5.1. Operating Characteristics

When the motor starts, the stator immediately goes to synchronous speed, the rotor is stationary and the slip is 1. Gradually, the rotor picks up speed, rotor currents decrease, and the rotor reaches a point where the generated torque equals the torque demanded by the load. The rotor speed stops to increase but remains as long as this condition exists.

In a no-load condition, the line current is limited to a small value called the exciting current that maintains the rotating field. Because the circuit is highly inductive, the power factor of the motor with no load is poor. The power factor for an unloaded IM may be up to 30% lagging. When unloaded, the rotor runs close to synchronous speed, and rotor current is very small. The interaction between the rotor flux on the rotating field is so small.

When the rotor shaft is loaded, the rotor slows down slightly. The rotating field, however, continues to rotate at synchronous speed. When IM slip increases, rotor current increases. The motor torque increases when the speed decreases and the power output increases. The increased rotor flux, caused by the increased rotor current, opposes the stator flux and reduces it slightly. The primary cemf, therefore, decreases slightly and stator current increases. This behaviour is similar to a transformer. Because of the low impedance of the stator windings, a small reduction in speed and cemf may produce large increases in motor current, torque, and power output.

If the motor is stalled, the resulting increase in rotor current lowers the stator cemf and causes excessive stator current that can damage the stator windings. At this time, the voltage applied to the stator should not exceed 50% of the rated voltage. For 230V operation voltage the maximum voltage we should apply to stator is 115V AC.

The typical torque and current curves for a 3-phase IM are shown in Figure 4-14. Rotor reactance increases with slip and increases its effects on rotor current and power factor as the motor load is increased. The pull-out point on the torque curve occurs at a slip of about 25%. The maximum torque at this point is about 3.5 times the full load torque and the rotor resistance is equals the rotor reactance. Rotor power factor is 0.707. Any additional load added to the motor beyond this torque value causes the motor to stall. In the stall condition, the stator current is about 5 times the full-load current value.

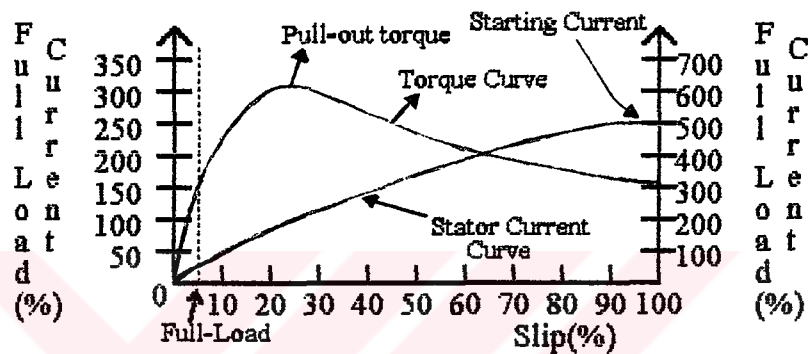


Fig. 4-14. The operational characteristics of IM

4.2.5.2. Speed Control of Induction Motor

The operating region of induction motors is normally with a small slip at just below its synchronous speed. At this speed, the developed torque is proportional to slip and the speed decreases with torque. The speed and torque of IM can be varied the supply frequency and stator voltage control[7][8]. The IM needs to have a constant flux in the rotor and stator to generate full torque. Therefore, the volts/hertz ratio between the stator voltage and the frequency of the stator voltage must be constant. To find V/Hz ratio, simply divide the rated voltage on the motor nameplate by the rated frequency.

$$\text{volts/hertz} = 230\text{V AC} / 50 \text{ Hz} = 4.6 \text{ V/Hz} \quad (4-5)$$

As an example we can find the stator voltage needed at 30 Hz as

$$V_{st} = \text{required stator voltage} = f \cdot (\text{V/Hz}) = 30 \text{ Hz} * 4.6 \text{ V/Hz} = 138 \text{ V} \quad (4-6)$$

If the voltage is fixed at its rated value, and frequency is reduced below its rated value, the flux will increase. At low frequencies, the reactance will decrease and the motor current may become too high. If the frequency is increased above its rated value, the flux and torque would decrease. If the ratio of voltage to frequency is kept constant the flux and the maximum torque remain approximately constant. At low frequencies, the air-gap flux is

reduced due to the drop in the stator impedance and the supply voltage has to be increased slightly to compensate stator impedance voltage drop and to maintain the torque level. If the motor operates in the constant V/Hz ratio, the torque remains constant, hence the power available from the motor will be directly proportional to the motor speed. The AC IM can be run at any speed, from a few revolutions per minute (RPM) to the top speed of the motor. We can see the effect of a change in stator frequency on the induction motor by looking at the speed-torque curve in Figure 4-15. Curve A is a typical curve for a four-pole machine with an applied stator frequency of 50 Hz. The motor will operate where the curve crosses the line labelled 100% load torque. If the stator frequency is reduced, the motor will shift its operating point to another curves. If both the voltage and frequency varied, the torque and speed can be controlled. The shapes of the torque curves are similar and the maximum torque value is independent of frequency.

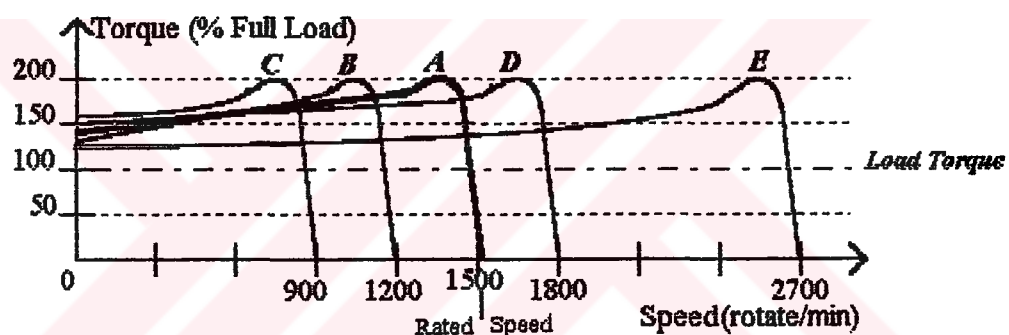


Fig. 4-15. Adjusting AC IM Speed

CHAPTER FIVE

SIMULATION AND EXPERIMENTAL RESULTS

5.1. INTRODUCTION

Firstly, the Space Vector Simulation program is introduced and the PWM waveforms for some selected frequencies and modulation indexes are given. These waveforms are printed using the simulation program.

Secondly, the experimental results are obtained using the oscilloscope. The photographs of the voltage and current waveforms can be seen in this part. Then, the simulation program results and the experimental results are compared with each other and evaluated briefly.

5.2. SIMULATION RESULTS

The Space Vector Simulation program is written in QUICK BASIC language to prove the selected pulse waveforms satisfying the sinusoidal output current with less harmonics. The theory is applied and the results are obtained graphically. In addition, analyzing harmonic contents of the simulated PWM waveforms shows us the effectiveness of the method.

The simulation program draws output voltage waveform for selected frequency and modulation index. The waveform for desired period range can be observe. The root locus of each frequency gives the closeness of U circle to the reference sinusoidal circle. This closeness of two circles is a measurement of the harmonic contribution. This circles are sketched in the simulation program graphically and we can compare two circles easily. The modulation index effect can be seen. Modulation index adjusts the output current magnitude by changing the PWM pulse widths.

Another option is the calculation of odd harmonic contents of the waveform and the simulation program gives percentage value of the harmonics over the fundemantal component. For each frequency these harmonic magnitudes are added together, and so we can see the sinusoidal waveform distortion graphically.

The phase and phase-to-phase voltage waveforms for selected frequency and modulation index values are shown in Figure 5-1 to 5-4. In these figures, if the waveforms

are seen clearly, the angle range is selected as 360 deg, otherwise the range is narrowed to see the waveforms in more detail. At the lower frequencies, number of pulses increases and the current sinusoidal waveform is much closer to sinusoidal waveform. But, at the higher frequencies the phase and phase-to-phase voltages contain less pulses, and the result current waveform is distorted with the harmonics.

As it seen from the volts/hertz control, when the operation frequency is low, the modulation index parameter is adjusted to get the same volts/hertz ratio. For example, if the frequency is 50 Hz, the M is selected 100% of the maximum voltage magnitude ($M=1$). But, if the frequency is 30 Hz, the M is selected as 60% ($M=.6$) of the maximum voltage.

In the Figure 5-5 and 5-6, the root locus of the selected frequency and modulation index can be seen. At the higher frequencies, the root locus is not close to reference circle, but at the lower frequencies, the root locus is more close to reference circle. As a result, at the lower frequencies the current waveform is more similar to sinusoidal waveform. If the M index is low, the output magnitude of the voltage and current is lower than the one that has higher M index. In the Figure 5-7, there is a table that compares the total harmonic distortion and distortion factor parameters. These results verify the precedence results.

5.3. EXPERIMENTAL RESULTS

Experimental results are given as photographs that obtained from the oscilloscope screen. In the Figure 5-7 and 5-8, the drive signals inverter with dead time waveforms can be seen. The dead time widths are approximately 3.5 μsec . In the Figure 5-9, the phase voltages detailed waveforms are given. These waveforms are not clear as simulation waveforms, but they give the idea about the generation of PWM waveforms. Figure 5-10 and 5-11 show a complete and half period of the PWM waveforms for one phase. In the Figure 5-12 and 5-13, the phase current waveform can be seen. These waveforms are similar to sinusoidal current waveform, but some harmonics effects can be easily determined. The photographs are not clear due to the oscilloscope trigger and hold adjustment. The camera and light adjustments was also effected the photographs quality.

The 8098 microcontroller board and the PC is shown in Figure 5-14. The inverter, power supply and the induction motor system can be seen in the Figure 5-15. The inverter input is DC voltage which is supplied from the three-phase diode rectifier circuit. This circuit is not shown. The rectifier circuit is fed from the 3-phase variac. The inverter three phase outputs are connected to the IM. The IGBT driver voltages are obtained from the power supply. The IGBT driver signals are received from the cable connected to the microcontroller board.

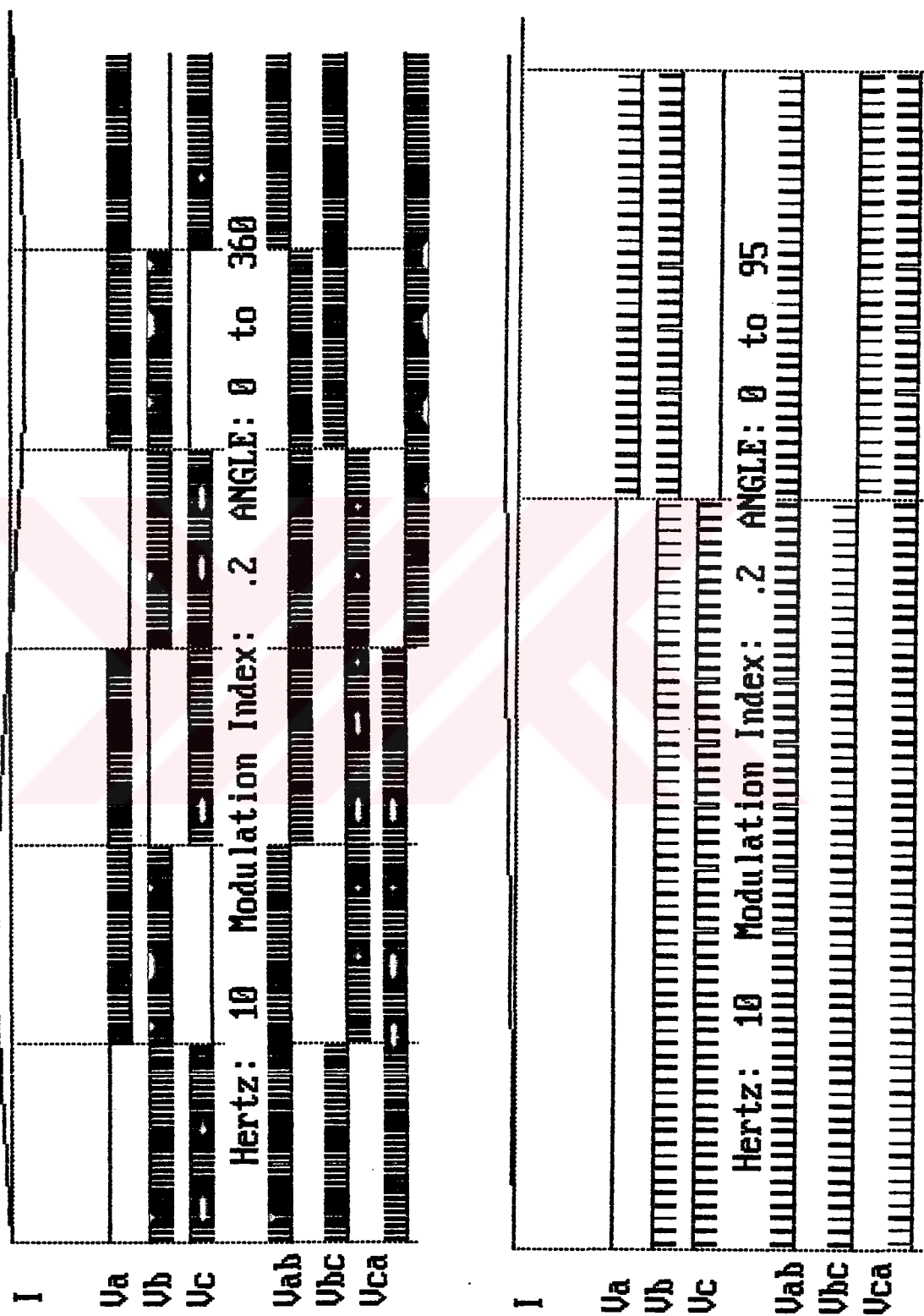


Fig. 5-1. (a) 10 Hz., $M=0.2$, 0-360° voltage and current waveforms
 (b) 10 Hz., $M=0.2$, 0-95° voltage and current waveforms

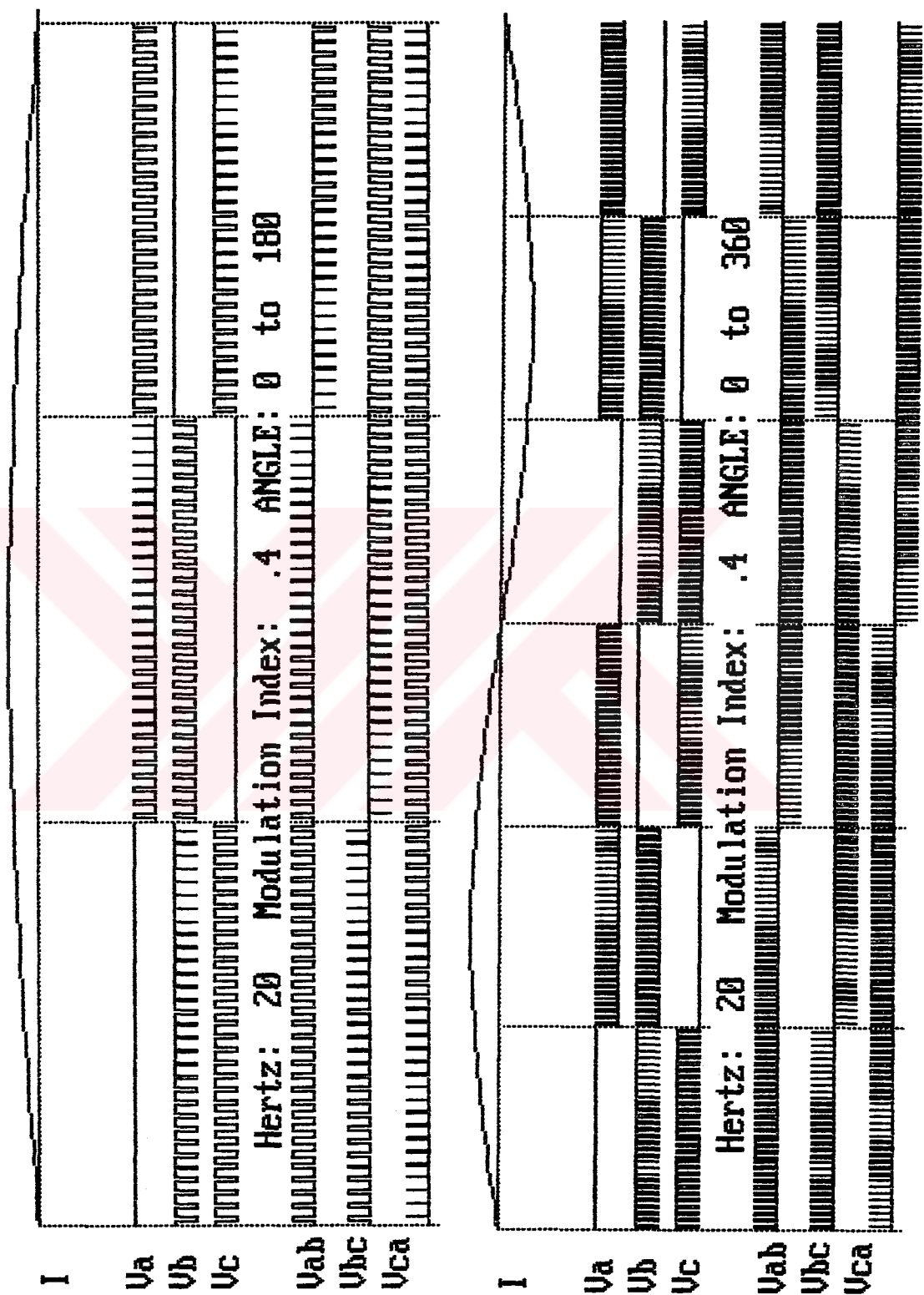


Fig. 5-2. (a) 20 Hz., $M=0.4$, 0-180° voltage and current waveforms
 (b) 20 Hz., $M=0.4$, 0-360° voltage and current waveforms

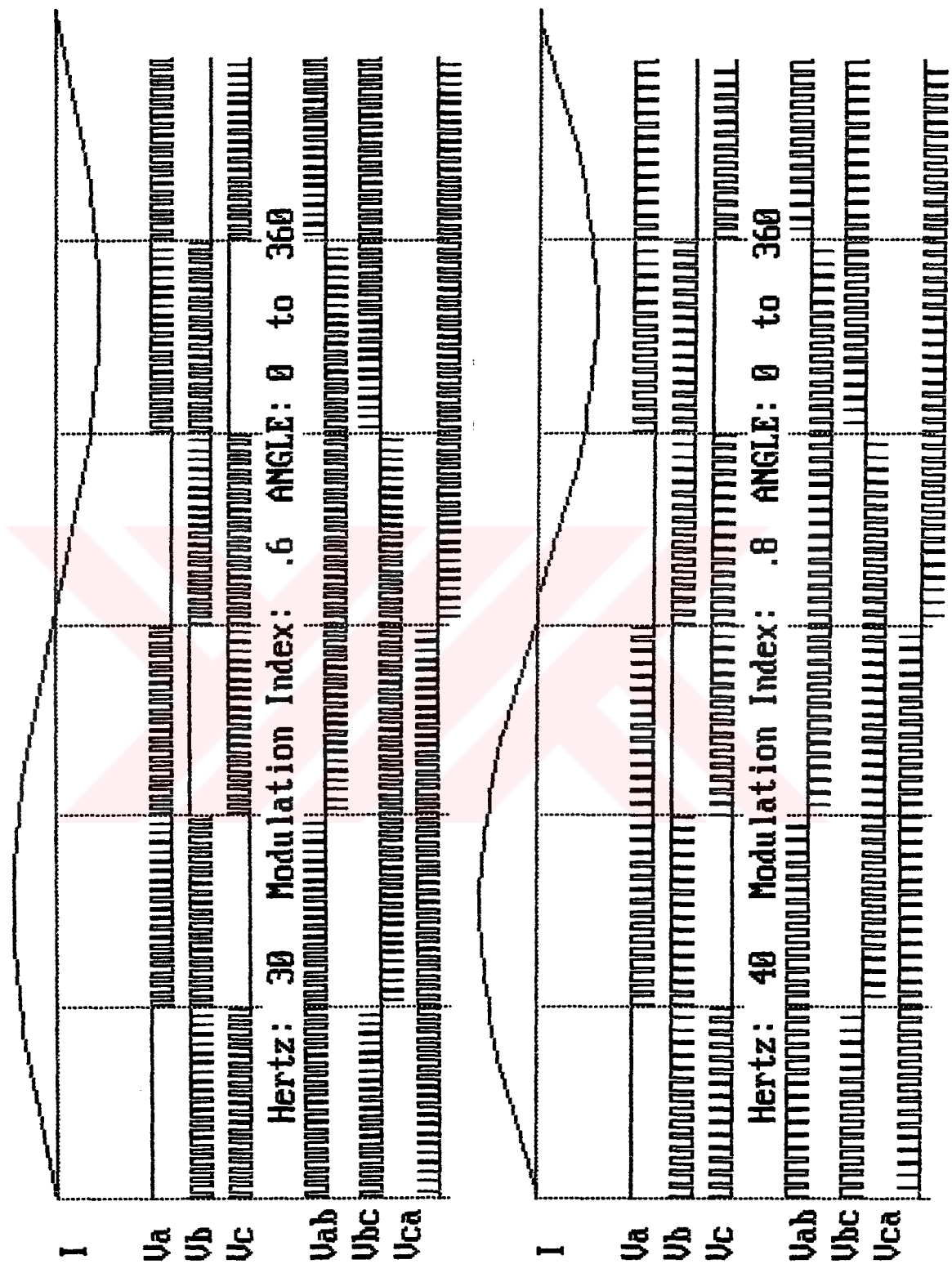


Fig. 5-3. (a) 30 Hz., $M=.6$, 0-360° voltage and current waveforms
 (b) 40 Hz., $M=.8$, 0-360° voltage and current waveforms

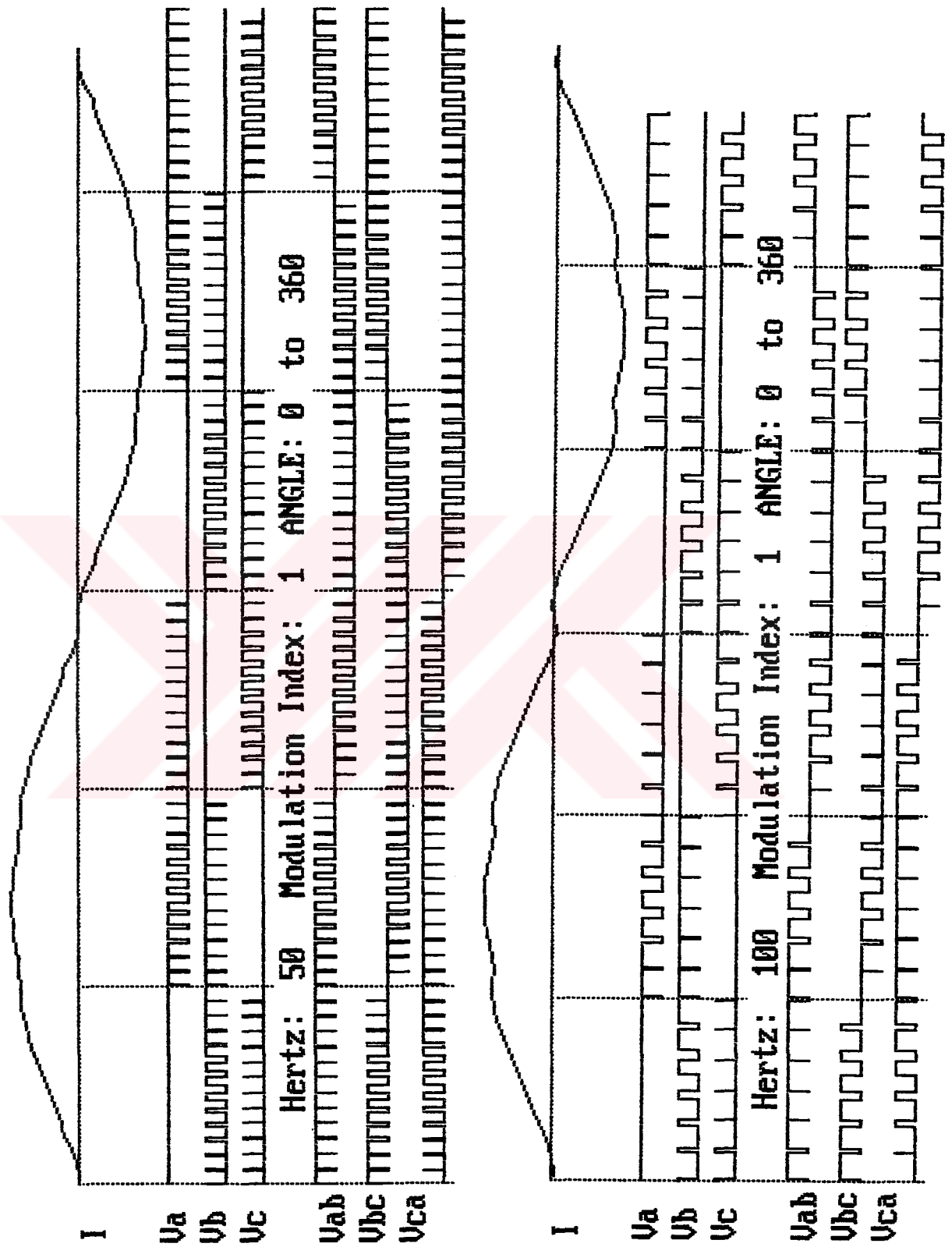


Fig. 5-4. (a) 50 Hz., $M=1$, $0-360^\circ$ voltage and current waveforms
 (b) 100 Hz., $M=1$, $0-360^\circ$ voltage and current waveforms

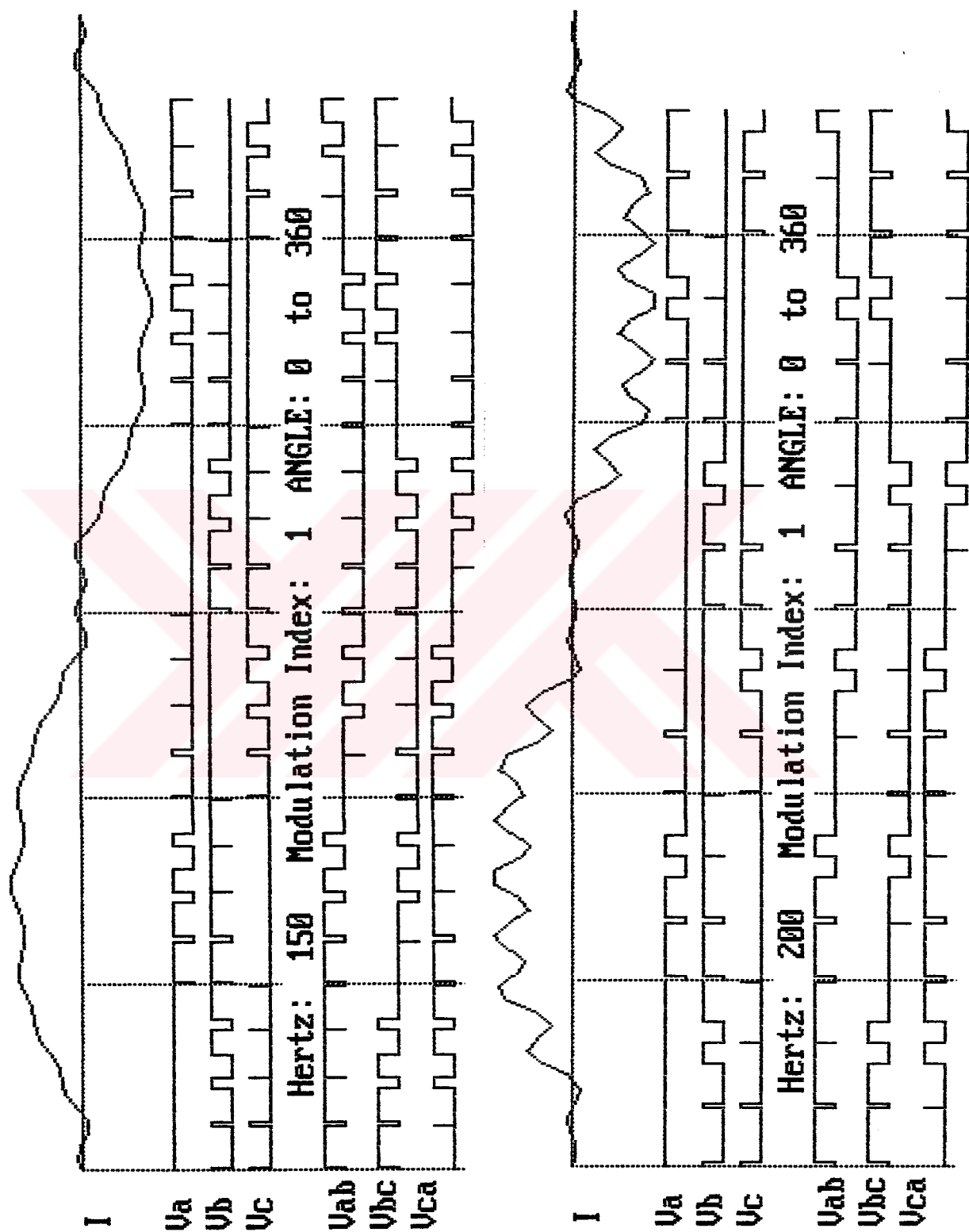


Fig. 5-5. (a) 150 Hz., $M=1$, 0-360° voltage and current waveforms
 (b) 200 Hz., $M=1$, 0-360° voltage and current waveforms

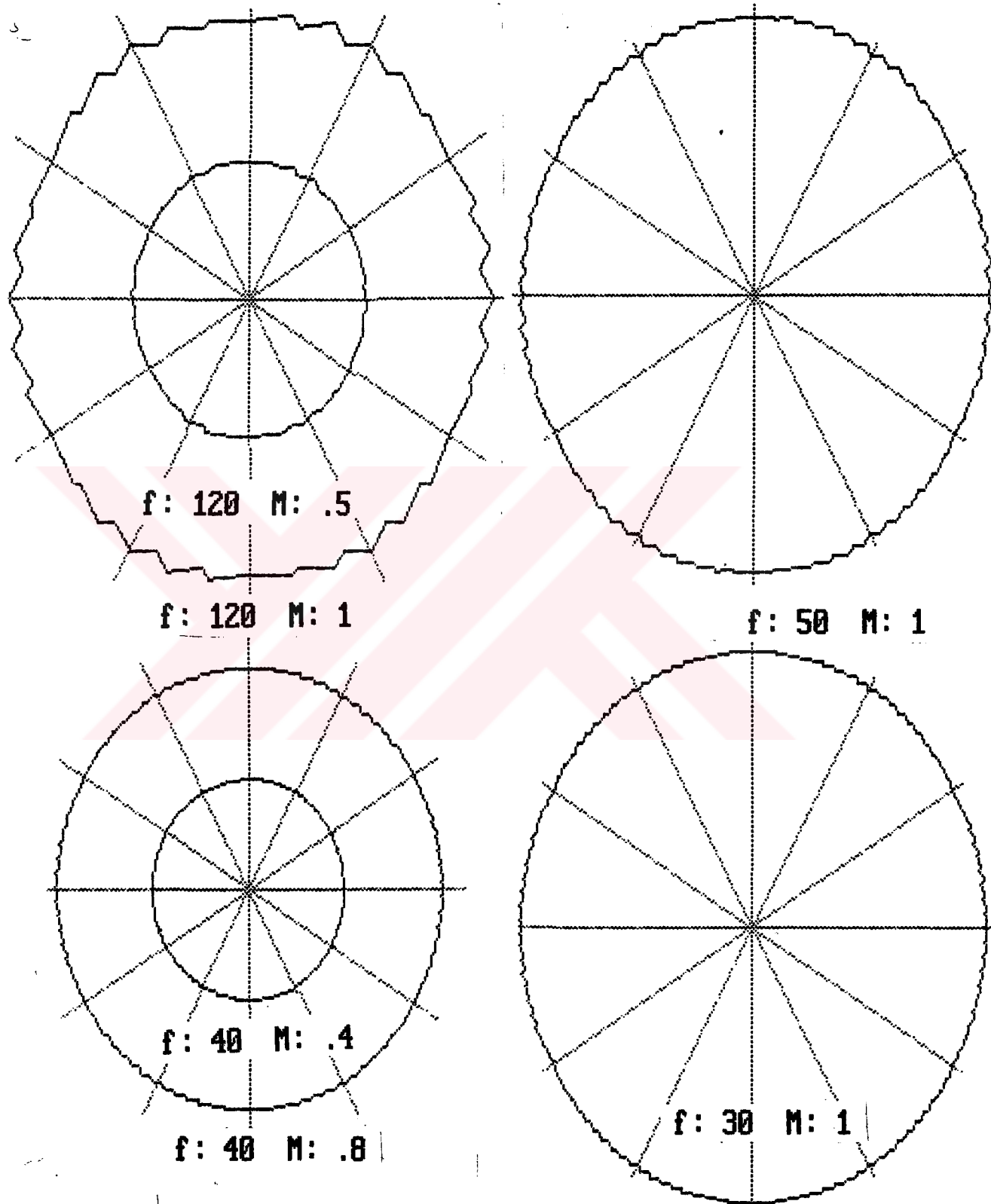


Fig. 5-8. Root locus graphics for the selected frequency and M values.

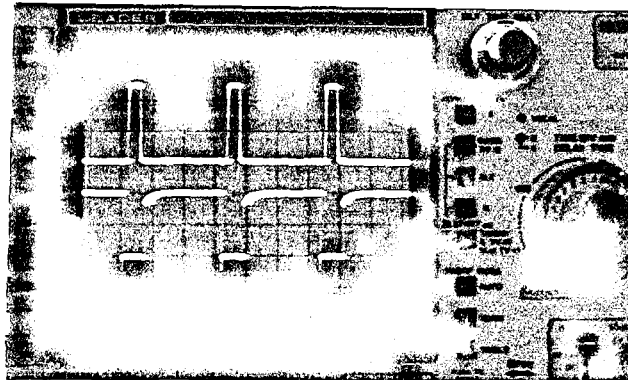


Fig. 5-7. For 10us pulsewidth the Dead Time Generator output signal (Voltages are 5V)



Fig. 5-8. For 30us pulsewidth the Dead Time Generator output signal (Voltages are 5V)

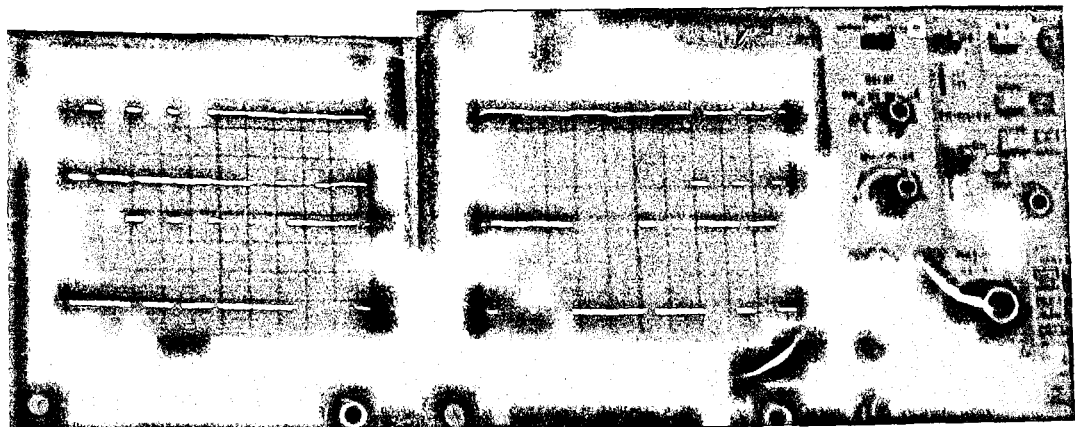


Fig. 5-9. For 50Hz the detailed phase-a and phase-b voltage output.

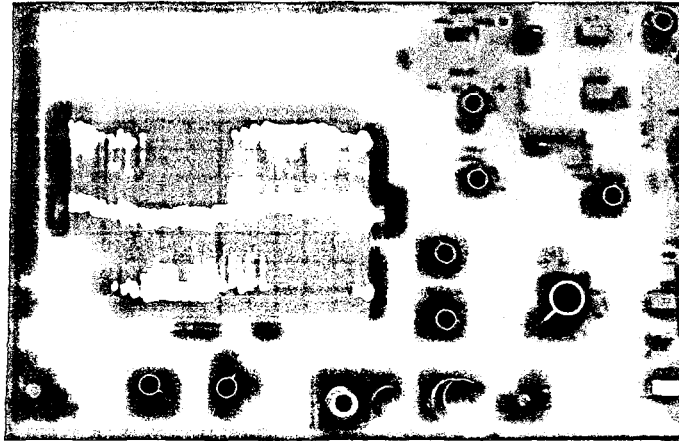


Fig. 5-10. For 50Hz the phase-a voltage output for a complete period.



Fig. 5-11. For 50Hz the phase-a voltage output for a half period.

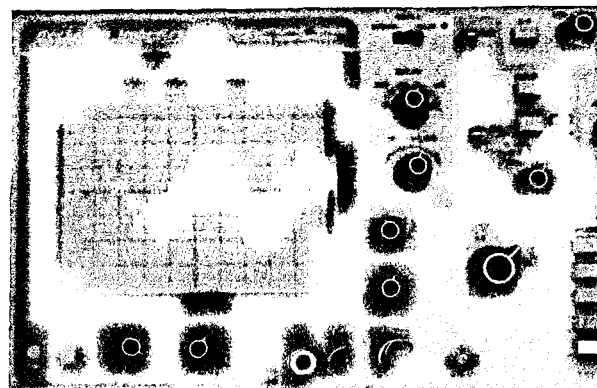


Fig. 5-12. For 50Hz the phase-a current

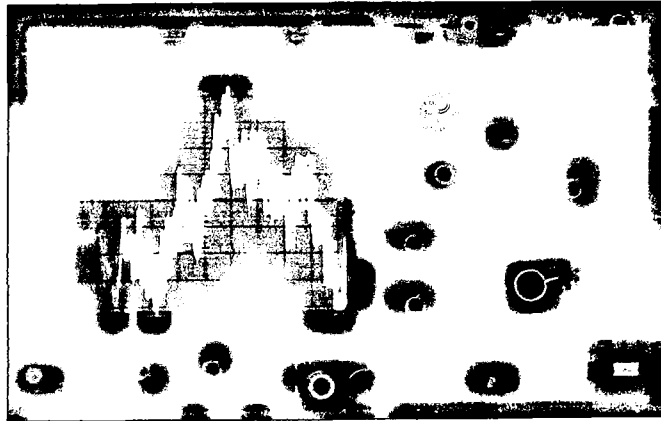


Fig. 5-13. Another the phase-a current for 50Hz operation frequency

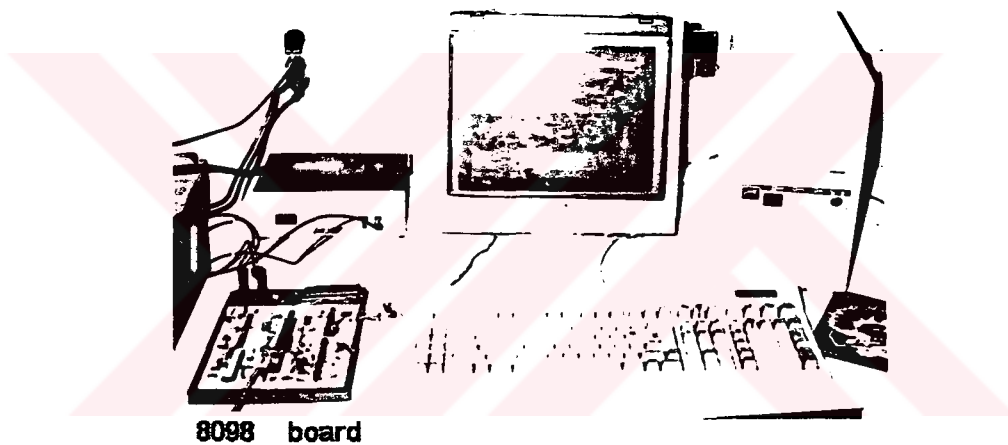


Fig. 5-14. 8098 Microcontroller Board and PC

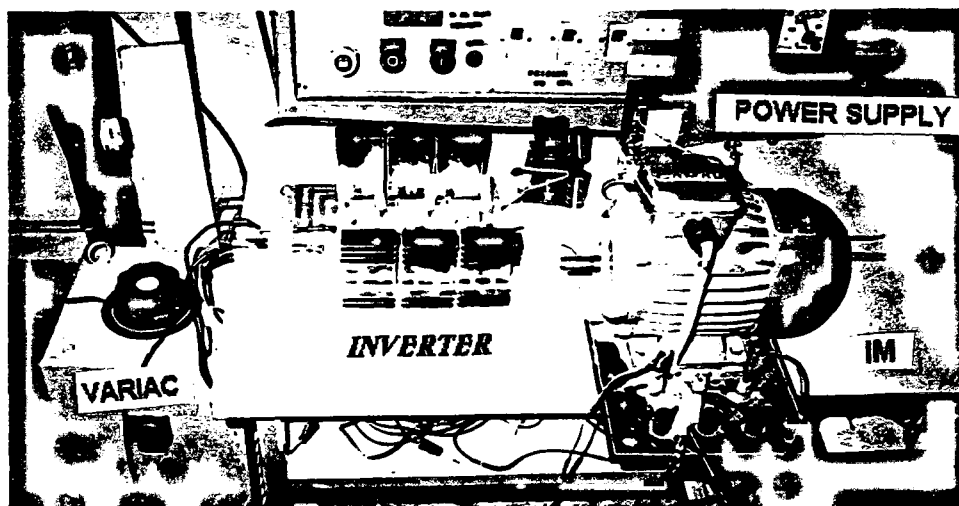


Fig.5-15. Inverter board, power supply, and the induction motor

5.4. COMPARISON OF THE OBTAINED RESULTS

When we compare the simulation and experimental results, we conclude that they are similar. But, the harmonics are effective at the experimental results. For instance, the closest harmonic to the fundamental is the 5th and the effects of this harmonic can be seen due to the generating PWM voltage waveforms.

At the low operation frequencies, the number of pulse in a period are increased, because in the Space Vector PWM method the number of pulses can be found as for operation frequency is 10 Hz, and sampling frequency is 3600 Hz.

$$g = (f_s / f_0) = (3600 / 10) = 360 \text{ pulses}$$

For total 360 deg period, the angle step is

$$\Delta\theta = 360^\circ / 360 = 1^\circ \text{ step}$$

For 20 Hz, g is 180 and $\Delta\theta$ is 2° . For 50 Hz, g is 72, and $\Delta\theta$ is 5° .

At the low operation frequencies, the shape of waveform contains more pulses and the IGBTs are switched more than the higher frequencies. In fact, this is necessary to obtain more sinusoidal waveform that gives required torque. Also, to provide the Volts/Hertz ratio at the low frequencies the percentage of modulation index must be decreased. All these things can be seen from the analysis results. (Fig. 5-1 to 5-5).

The simulation and experimental results are not similar, the reason for that can be 8098 software overhead. The Space Vector algorithm is complex and requires all the CPU time even the PWM outputs are obtained by HSO unit, and the program structure is based on real-time interrupts and serial interrupts. The second reason may be the opto-couplers that isolates the microcontroller and inverter board. The opto-couplers are slow when compared to pulse transformers, therefore the pulse shapes and times may be changed.

If we evaluate all the findings that obtained from the simulation and experimental results, we can conclude that this PWM method is an effective method to drive the induction motor. The most important advantage is the switching angles look-up table is so small and software controllable. The modulation index and operation frequency are adjustable. Induction motor direction reversal can be achieved by the software instead of using a digital logic circuit. In harmonic elimination methods, elimination is done by increasing the number of pulses in a period and this increasing must be adjustable for each operation frequency. This process can be easily done with the Space Vector PWM method.

CHAPTER SIX

CONCLUSION

In this study, an Inverter circuit has been built using a new type of power switching device, IGBT. Then, a new method that is called Space Vector has been used for obtaining a PWM sinusoidal waveform. Practical results were obtained from the hardware and software of the total system. The hardware system consists of a PC serial channel controlled 8098 microcontroller board, an inverter and induction motor. The 8098 microcontroller board calculates the PWM switching pulses for the desired frequency. Software part is the Space Vector PWM simulation, PC serial channel emulation and microcontroller real-time programs.

In the Space Vector PWM Method, the sampling frequency has been selected as 3600 Hz. This sampling frequency is required to obtain three-phase line currents that contain less low-order harmonics. Because of the high number of switching pulses, it can be observed that the switching losses would increase. If the sampling frequency is decreased, then the harmonic contents of the PWM waveform are increased. As a result, an optimum sampling frequency must be chosen.

The practical results are similar to the simulation results, but there are some differences between them. The inverter gate signals were obtained from a dead time generator. The dead time is required to protect the switching elements on the same arm from the short circuit. The output current waveform is distorted by the dead time process. The typical dead time is calculated as 1-2 μs . This time is sufficient to turn-off the IGBT. But the dead time is generated as 3-4 μs for providing safe operation in application. If the short circuit occurs at high voltages such as 200-300 V, the short circuit current becomes high and therefore might destroy the IGBTs. For example for the 2 Hz operation frequency, we have about 3600 number of pulses. The minimum time that applied to the IGBT element is 2 μs . So, it can be said that, the duration of dead time and normal pulses becomes equal to each other. For that reason, the output waveform is distorted so high when the number of pulses are selected high in a period. IGBT is more suitable for this type of high sampling frequency applications. The other switching elements like thyristors create more problems. If the minimum PWM pulse width is less than 6-7 μs , the output of

the dead time generator is zero and we can not obtain the inverted gate signals. For these reasons, the experimental results obtained different from the simulation results.

The microcontroller and the inverter board isolation was achieved by opto-couplers. The advantage of using this isolation method was to make the simple design. But, the opto-couplers signal transfer time is slow in comparison to the pulse transformers and also all opto-couplers have different transfer characteristics. These parameters effect the PWM voltage waveforms and therefore output current is distorted.

Designing an inverter is not easy. The six switches have some certain limits about their current and voltage ratings. These ratings can not be exceeded and the driver circuit must satisfy all the switching requirements. Some errors in the switch driver circuit and the RC snubber circuit may cause troubles. The dV/dt and dI/dt values must be limited to an appropriate value by using the RC snubber circuit. If the recovery diode speed is not sufficient, the induced high voltages at the induction motor windages are harmful for the switches. The gate drive voltages of IGBT switches must not be exceeded. If the voltages are higher than +15V or lower than -15V, the gate circuit of IGBT may be damaged.

An induction motor speed (RPM) control is achieved by the designed total system. This is not a robust control without using a feedback to obtain a closed loop system, so it must be supported by Vector Control Algorithm programmed on the host controller. For future work, Vector Control is easily applied to the Space Vector PWM system, because the system design satisfies this implementation. This controller must have a floating point processor and a powerful CPU. The total motor speed control system is created using a host system that observes the line currents as a feedback signal and calculates the frequency and modulation index parameters. Then these parameters are transferred to our PWM generating system via a serial channel.

A simulation software program was written in QUICK BASIC language to prove the selected angles for the turn-on and turn-off the inverter switches are similar to the theory. The line and line-to-line voltages can be observed for a desired frequency and modulation index. Root locus graphics also give some ideas about harmonic contents of the waveforms. In addition, simulation program analyzes harmonic contents of the output PWM waveform.

As a result, the Space Vector PWM method has many advantages. These are: easy control, small-sized switching data look-up table, adjustable frequency and adjustable output voltage magnitude, possibility of vector control. Moreover, if the sampling frequency is optimized the output current contains less harmonics.

REFERENCES

- [1] Şenlik, İrfan, "Eviriciler", Trabzon Güç Elektronikl Semineri, pp. 161-172, 1989.
- [2] Fukuda, S. , Yoshitaka, I., Hasegawa H. , "PWM Technique for Inverter with Sinusoidal Output Current", IEEE Transactions on Power Electronics, Vol. 5, No.1, pp. 54-60, February 1990.
- [3] Peatman, John B., "Design with Microcontrollers", McGraw-Hill Book Company, pp. 1-20, pp. 539-603, 1988.
- [4] Intel, "Embedded Control Applications", Intel Literature Update Service, 1988, AP-248.
- [5] Humphries, James T. and Sheets, Leslie P., "Industrial Electronics", pp. 174-211, pp. 302-312, 1989, Delmar Publishers Inc.
- [6] BIRD, B.M, King, K.G., Pedder, D.A.G., "An Introduction to Power Electronics",1993, John Wiley & Sons.
- [7] Rashid, M.H. , "Power Electronics", pp. 228-370, 1988
- [8] Bose, B.K., "Power Electronics and AC Drives", pp. 1-52, pp. 121-165, 1986, Prentice-Hall
- [9] Toshiba, "IGBT Driver- A Simple Low-Cost Solution" , March 1993.
- [10] Horowitz, Paul, "The Art of Electronics", Cambridge University Press, pp. 334-335, 1980.
- [11] Holtz, Joachim, "Pulsewidth Modulation - A Survey", IEEE Transactions on Industrial Electronics, Vol. 39, No.5, pp. 410-419, December 1992.
- [12] Smith, Ralph J., "Circuits, Devices and Systems", Wiley, pp. 655-680, 1976.
- [13] Ting-Yu Chang and Ching-Tsai Pan, "A Practical Vector Control Algorithm for μ -Based Induction Motor Drives Using a New Space Vector Current Controller", IEEE Transactions on Industrial Electronics, Vol. 41, No.1, pp. 97-103, February 1994
- [14] Toshiba, "IGBT User Guide", June 1992.

APPENDICES

APPENDIX A:

MATHECAD Program (SPACE.MCD) for proving coefficients K1,K2,K3 and $\alpha_1, \alpha_2, \beta$. Also this program is saved in the floppy diskette.

```

f0 := 50      M := 1      fs := 360
aci :=  $\frac{f0}{10}$       max :=  $\frac{60}{aci}$       deltata :=  $\frac{1}{fs}$       delteta := aci  $\frac{\pi}{180}$ 

tetar := aci  $\frac{\pi}{180}$       j := 1 .. max      cons_j := (j - 1)  $\cdot$  tetar

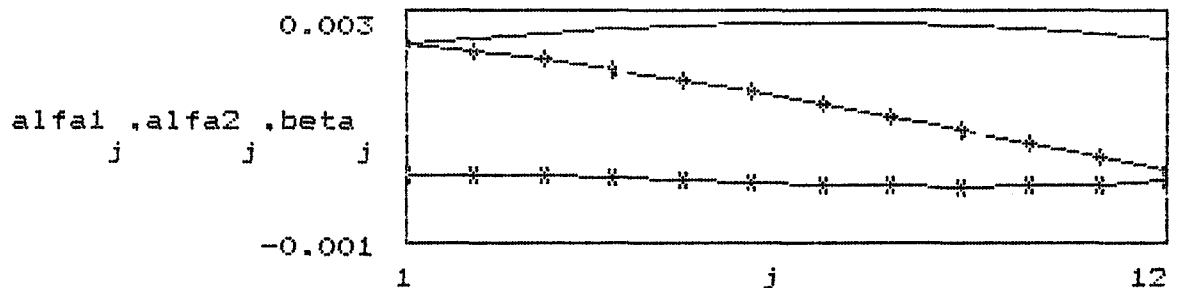
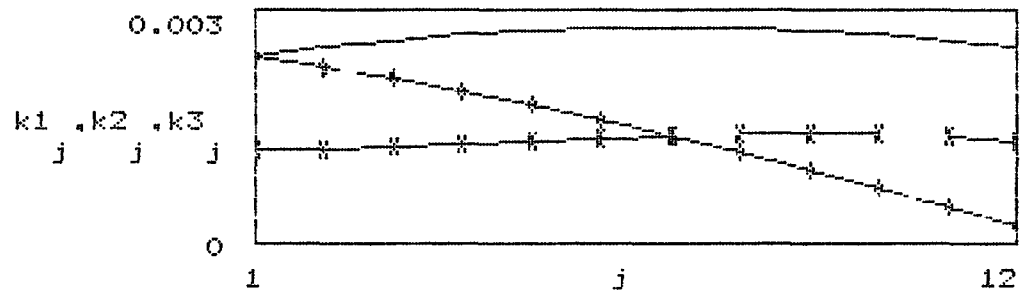
k1_j := deltata  $\cdot$  cos  $\left[ \text{cons}_j - \frac{\pi}{6} \right]$       k2_j := deltata  $\cdot$  cos  $\left[ \text{cons}_j + \frac{\pi}{6} \right]$ 

ext_j :=  $\frac{1}{2} \cdot \cos \left[ \text{cons}_j \right]^3$ 

k3_j :=  $\frac{\sqrt{3}}{2} \cdot \text{deltata} \cdot \left[ \left[ \frac{1}{6} \cdot \sin \left[ \text{cons}_j \right]^2 + 1 \right] \cdot \cos \left[ \text{cons}_j \right] - \text{ext}_j \right]$ 

alfa1_j := M  $\cdot$  k1_j      alfa2_j := M  $\cdot$  k2_j      beta_j :=  $\frac{\text{deltata}}{2} - k3_j \cdot M$ 

```



APPENDIX B:

IGBTDIR.CIR Pspice Listing.

***** IGBT switching circuit response for RL-load *****

**** IGBT collector-emitter circuit

```
V_0      1  0      200V    ; DC voltage
L_load   1  2      100uH    ; Inductance
R_load   2  3      5        ; resistor
X_diode  0  3      Diode    ; Recovery diode
C_snub   3  4      22nF     ; snubber cap.
R_snub   4  0      4.7      ; Last
X_igbt   3  6  0      IGBT    ; Switching element-IGBT
```

**** IGBT gate circuit

```
V_pulse  5  0      Pulse ( -15 15 200u 10n 10n 400u 1000u )
R_gate   5  6      10       ; Gate resistor
```

.Temp 25

.Inc IGBT25g.mod ; IGBT model file

.Inc diode1.mod ; Diode model file

.IC V(6) = 0V

* print step final time no print step ceiling use initial cond.

.tran .5u 2e-3 200u 500n UIC

.probe

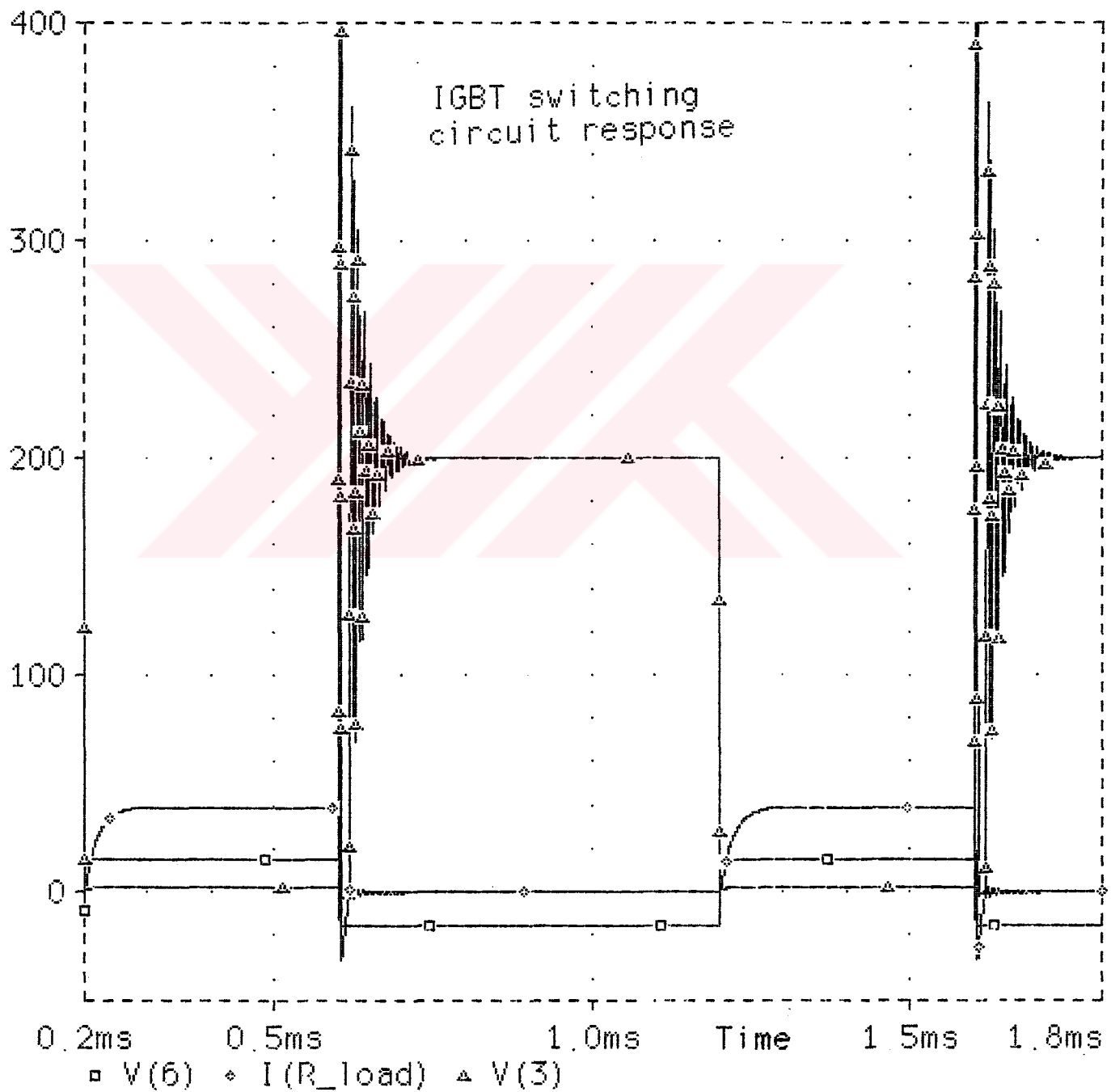
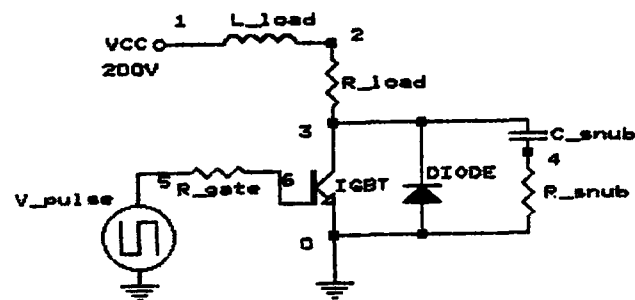
* Iteration ; Rel. Tolerance ; Temperature

.options itl5=0 reltol= 0.002 Tnom=125

.options itl4=50

.end

The drive circuit and the analysis results are shown in the following figure.



The drive circuit and the gate drive signal, collector voltage and current graphics