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STUDY OF "CHOPPER/INVERTER AC DRIVES USING GATE TURN OFF THYRISTORS"

Summary:

This report examines the study of variable speed induction motor drives.It includes a comparison of the various systems avalaible and specifically examines the present state of the art of power electronics used in drives.The various mathematical methods used to derive the parameters values are also examined.

Two specific control methods,the Voltage/Frequency and vector controlled methods,are designed,constructed and then compared. The vector control method was not however constructed satisfactorly and in view of the limited time scales it was decided to leave the complete development of this aspect for further work.

CONTENTS TABLE

LIST OF SYMBOLS

V.S.I	Voltage source inverter
C.S.I	Current source inverter
P.W.M	Pulse width modulation
G.T.O	Gate-turn-off thyristor
V_s	Stator voltage, V
I_s	Stator line current, A
I_m	Magnetizing current, A
I_r	Rotor current, A
R_s	Stator resistance, Ω
R_r	Rotor resistance, Ω
ω_1	Frequency supply, Hz
ω_r	Shaft speed, rad/s
s	Slip frequency, Hz
L_s	Stator leakage inductance, H
Lr	Rotor leakage inductance, H
Φ	Main flux, Wb
E_m	E.M.F, V
Q_x	Reactive power, W
V_{as}	Phase stator voltage (A), V
V_{bs}	Phase stator voltage (B), V
I_{as}	Stator current in phase (A), A

I_{bc} Stator current in phase (B),A

I_{cs} Stator current in phase (C),A

V_{ar} Rotor voltage of phase (a),V

V_{br} Rotor voltage of phase (b),V

V_{cr} Rotor voltage of phase (c),V

p (d/dt) Laplace-operator

ψ Flux linkage,Wb/T

G_{qs} Switching function

V_{rec} Voltage at the output of the rectifier,V.

I_{rec} Current delivered by the rectifier,A.

L_d Filter inductance,H.

C_d Filter capacitor,F.

I_{qs} Active or torque component of the stator current,A.

I_{ds} Reactive or field component of the stator current,A

V_{qs} Voltage vector aligned along the stator flux,V

V_{ds} Voltage vector in quadrature with the stator flux,V.

CHAPTER ONE

1.0 INTRODUCTION

This report examines the study of variable speed Induction Motor drives.It includes a comparison of the various systems available and specifically examines the present state-of the art power electronics used in drives.The various mathematical methods used to derive the parameters values are also examined.

Two specific control methods,the Voltage/Frequency and vector controlled method,are designed ,constructed and then compared.

The vector control method was not however constructed satisfactorly and in view of the limited time scales it was decided to leave the complete development of this aspect for further work.

1.1 OVERVIEW

Variable-speed AC drives are increasing in popularity and numerous attempts have been made to enhance their performance capability and reliabilty while maintaining reasonable cost,in order that they will compete more favourably with DC drives.

Over the last decade,inverter-fed induction motor drives for industrial application have been developed to a usable level. The low cost,high intrinsic reliability and manufacturing simplicity make the induction motor very attractive to the designers of AC variable speed drives.

However the situation is unfortunately far more complicated when speed control is of interest.

This project starts by considering the relative merits of the AC motors,by reviewing in general the AC variable speed drives and particularly the induction motor drives.

Various control techniques adopted to date are analysed and compared to give a basis for the choice of a method suitable, with regard to its implementation,in speed control of an induction motor.

An experimental drive system using a V/Hz control is described and results from the static and dynamic tests are given.

1.2 Structure of the project

Chapter one gives an overview of AC variable speed drives.

Chapter two surveys the actual state of the art in variable AC drives with an emphasis on the use of the induction motor.

Chapter three looks at the different methods of Speed. control in AC drives.

Chapter four gives a review of different semiconductor devices and the choice of the type of inverter power circuit to be designed in this project.

Chapter five describes the system built and presents some results.

Chapter six looks at the achievements of the work carried out in the project.,and the recommandations for further works.

CHAPTER TWO

2.0 STATE OF THE ART IN VARIABLE-SPEED AC DRIVES

2.1 Introduction

Progress in traditional power semiconductors such as rectifiers and thyristors has by-and-large been a step by step evolutionary progress,with gradual but steady improvement being made in operating characteristics,rating and packaging concepts.The cumulative effect,when viewed in relationship to the modest earlier begining of these devices some 30 years ago has however,been quite dramatic.

Superimposed onto this traditional mainstream power semiconductors technology has been the emergence of several new types of power semiconductors.

The most significant and revolutionary of these is undoutedly the power M.O.S.F.E.T,Bipolar transistor,and the Gate-Turn-Off thyristor(G.T.O).

These new devices have had a significant effect on power electronic converters.Many restrictions have been removed on system complexity,cost,speed and power range,and opened the way for a new integrated approach to digital control of AC drives.

Before this revolution,numerous methods of obtaining stepped or continuous change have evolved,some of which resulted in complicated ac-machines(Schrage motor) and some in the use of auxiliary machines (Scherbius and Kramer systems).Few of these systems have been very efficient and all have required more maintenance than the induction motor alone.The basic principles of those classical methods of speed control that are relevant to systems using power semiconductor converters are discussed briefly in the following sections.

Nowadays,adjustable speed AC drives has become more common in variable-speed applications.

In this chapter ,a review of variable-speed AC drives will be presented in the first section,which will examine various drives systems used in the recent past.This will be followed by a second section dealing with converter-fed induction motor drives,and finally an attempt will be made to draw comparison between these drives.

2.2 Survey of variable AC drives.

In adjustable AC drives,the static power converters constitute an interface between the primary source of energy and the machine itself as shown in FIG 2.1.

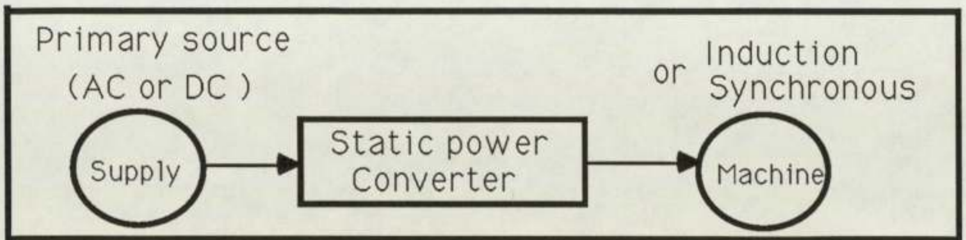


Fig 2.1 Energy Conversion Diagram

Generally a static-power converter adapts and controls the enegy flow to provide an AC source at its output with variable voltage and frequency.This ability of the converter allows the variation of the motor speed by acting on the two machine parameters (voltage,frequency).

The best established of these drive systems will be presented briefly in the following subsections.

2.2.1 AC voltage controller.

The use of the AC voltage controller,is a simple and economical technique for speed control,generally used with induction motors.A number of such circuit configurations are available and the literature fully documents these arrangements and their performances [2.1],[2.2]

The voltage controller shown in FIG 2.2 is one such circuit,it is an amplitude controller of the applied stator voltage by symmetrically phase controlling the trigger angles of line commutated antiparallel thyristors.This variation in terminal stator voltage can be used to give a measure of speed control.This method called phase control,provides continuously variable speed control.However,phase control has the dual disadvantage in that the output voltage is load dependant and the line power factor keeps deteriorating as the triggering angle is retarded more and more,[2.3],[2.4],[2.5].

Also,the harmonics generated by this method causes more heating in AC rotating machines.These harmonics may travel along the line thereby causing disturbance to other equipments operating on the same line,[2.6].

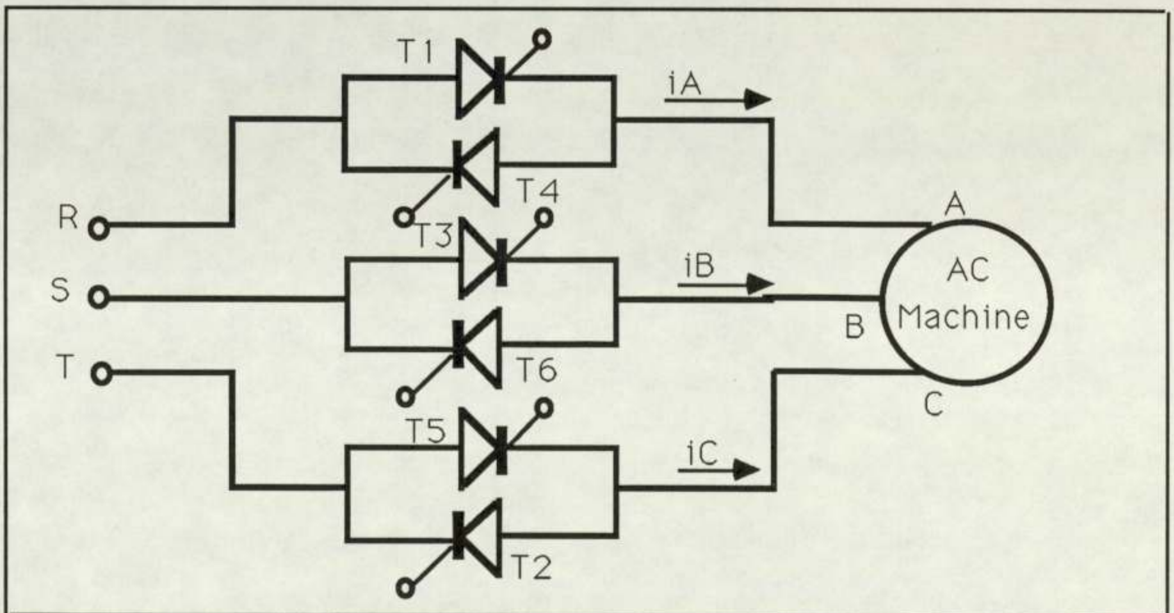


Figure 2.2 AC voltage controller

This drive has a severe drawback particularly at low speed (as much as 18%-20% of the full load power is dissipated as heat within the machine),[2.7].

The application of such drives is commonly restricted to low power,for pump and fans,and must be used with caution as mentioned by CHAUPRADE,R et al in [2.8].

2.2.2 Static KRAMER drive or Subsynchronous drive.

The static KRAMER or subsynchronous cascade system is one of the family of so called slip-power recovery drives in which the speed of a slip ring induction motor is controlled by the application of a control circuit to its rotor winding.

The static KRAMER,is based on a similar principle of speed control as in a wound rotor machine,where the rotor circuit resistance is mechanically varied.This method has been used for a time,but this old technique is innefficient as a result of large slip power dissipation.This problem is overcome by electronically varying the equivalent rotor resistance and returning back the rotor slip power to the AC line through a voltage and frequency changing system consisting of a rectifier and inverter with an intermediate dc-link,thus permitting speed regulation as shown in FIG 2.3 [2.9],[2.10].

The most serious drawback of this form of drive has been its high reactive power requirement,particularly at the upper end of the speed,and the consequent high cost of power factor correction [2.11].

This drive also generates low order harmonics which are reflected back to the AC line,and create overheating and torque pulsations at low speed.

The static KRAMER system is one of the few which can be used in very large power applications,and where a limited range of speed is required,they are suitable for centrifugal pumps such as boiler feed pumps,waterwork pumps,mine ventilation,as mentioned in [2.12].

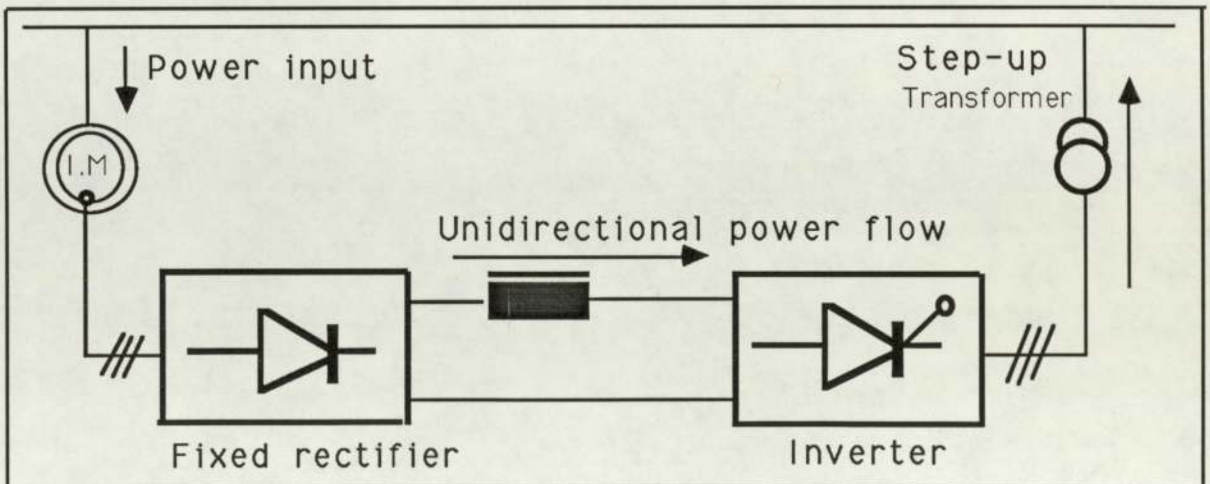


Figure 2.3 Static KRAMER drive or subsynchronous cascade

2.2.3 Static SCHERBIUS drive or sub/super synchronous drive.

The static SCHERBIUS drive can be used either with a wound rotor induction motor or with a synchronous motor.This drive was first proposed by POLGE.R.J,[2.13].The development of the silicon rectifier renewed the interest in a system consisting of a wound induction motor and frequency converter in the rotor,[2.14],[2.15].

If the slip power is fed back to the line,the drive operates in subsynchronous mode(similar to the static KRAMER),on the other hand if the slip power is supplied to the rotor,the drive operates in supersynchronous mode.

This type of drive is generally used in very high power ranges such as,pumps and blower applications as mentioned in[2.16],[2.17].A static SCHERBIUS drive configuration is shown in FIG 2.4.

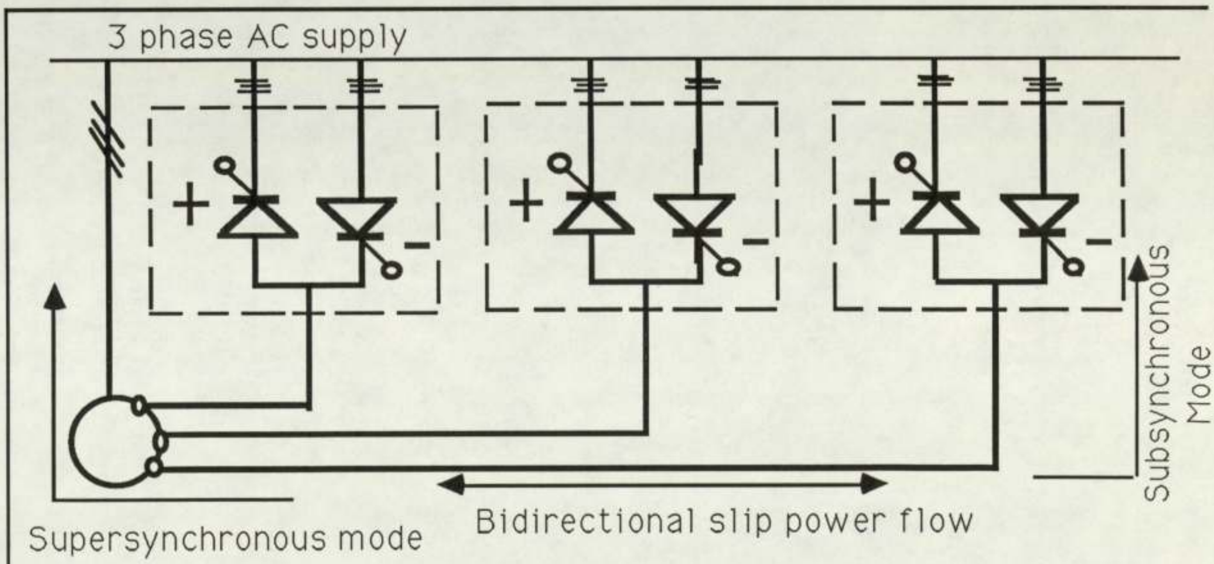


Figure 2.4 Static SCHERBIUS or Sub/Supersynchronous drive

2.2.4 Cycloconverter drive.

The principle of a controlled cycloconverter has been described by Mc.MURRAY,in [2.14].The function of the cycloconverter is to convert a high input frequency alternating current or voltage into a lower output frequency alternating current or voltage.The cycloconverter consists of six groups of three-phase half wave circuits.One half of these groups are called the "positive group"(+),and the other half are called the "negative group"(-).The function of the positive group is to carry current during the positive half cycle of the output frequency wave,and the negative group to carry the current during the negative half cycle of the output frequency wave.The ouput frequency is determined by the length of time the positive group and negative group are allowed to carry the load current.

The cycloconverter has the facility for control of both its output frequency and voltage,independantly of one another.Since in industrial applications,the primary power supply is almost invariably a fixed frequency AC supply,it is possible to connect this supply directly to the input terminal of the cycloconverter as shown in FIG-2.5,thus providing a variable speed AC drive system,with only one single stage of power conversion.

As mentioned with the previous drives,the cycloconverter can be used with a squirrel cage induction motor or with a synchronous motor.This drive can be easily designed for bidirectional power flow [2.19].

The input to such a drive contains complex harmonics as indicated in the previous drives,but they are attenuated during normal operation by the machine inductance.The power factor is also poor.The upper output frequency limit of the cycloconverter is usually less than the supply frequency (typically around 2/3 of the supply frequency).This restriction does,then,place a practical limitation on this type on drive system,as indicated by B.MOKRYTZKI in [2.20].

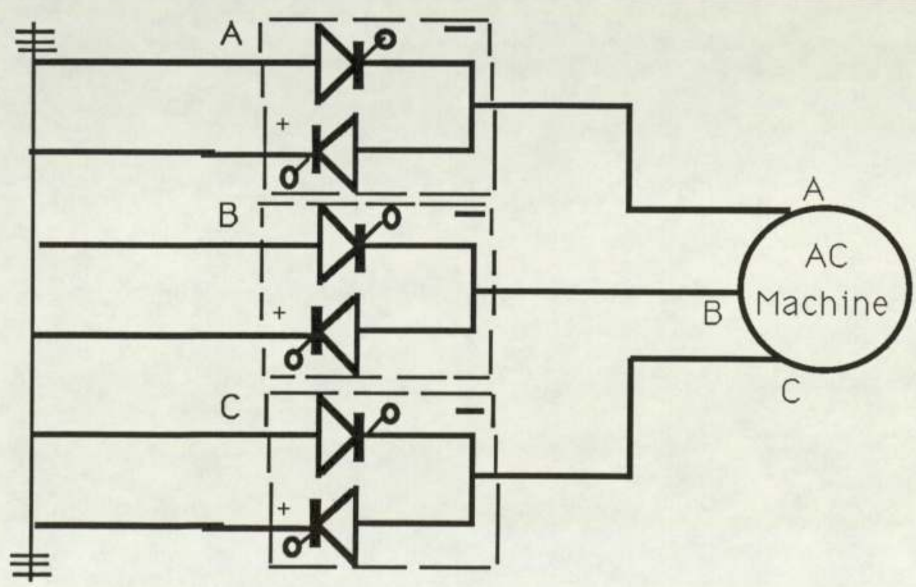


Figure 2.5 Cycloconverter drive

2.2.5 Voltage source inverter (V.S.I).

The voltage source inverter converts the DC-input supply to variable frequency,variable voltage.This means that the system

may be fed from a DC-busbar,such as an overhead DC-traction system,battery source or by using either a controlled or uncontrolled power converter followed by a chopper.

In the configuration shown in FIG 2.6,[2.21],the normal three phase AC supply is first converted to a variable DC-link voltage,then inverted to a variable voltage,variable frequency AC supply,in order to vary and control the motor speed.

There are two basic modes of inverter operation:

a) the six step square wave (SSSW)inverter.

b) the pulse wave modulation (P.W.M) inverter.

The SSSW-inverter are mainly applied in low and medimu power range.

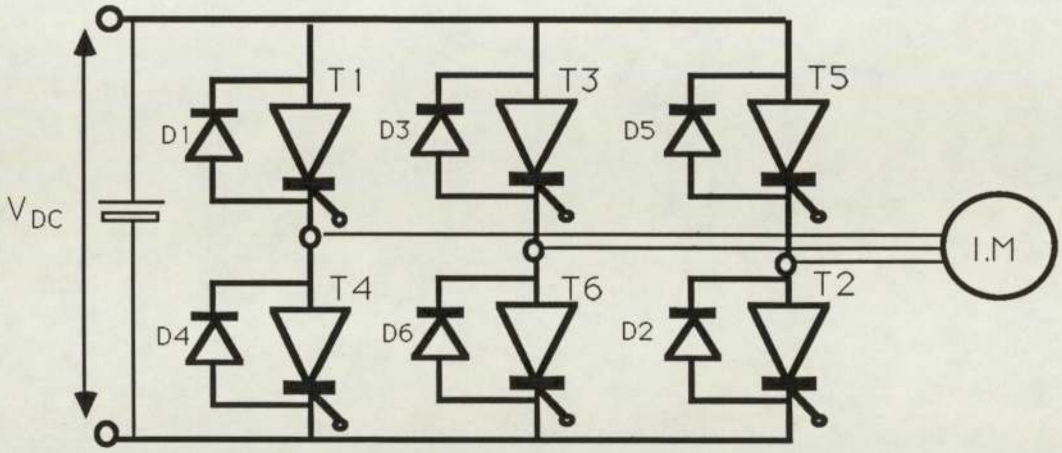
The basic voltage source inverter circuit is shown in fig 2.6.The DC-link voltage is imposed at the input of the inverter via a capacitor.The voltage output is of variable frequency and amplitude to control the speed and torque of the induction motor.

In this drive,the inverter output voltage is three square waves,phase shifted by 120 degrees.

These waves are rich in low order harmonics which have two effects,they create torque pulsations and overheating of the motor especially at low speed operation.

The feed back diodes shown in fig 2.6 are used to allow the flow of reactive energy from the motor.

At any time there will be three thyristors in conduction,follwing the switching conduction pattern shown in fig 2.6(a).Alternately the thyristors are turned on and will produce a square wave as shown in fig 2.6(b).



Basic power circuit of a -V.S.I

a)

Sequences	1	2	3	4	5	6	1'	2'	3'	4'	5'	6'
Positive group-ON	1,3	3	3,5	5	5,1	1	1,3	3	3,5	5	5,1	1
Negative group-ON	2	2,4	4	4,6	6	6,2	2	2,4	4	4,6	6	6,2

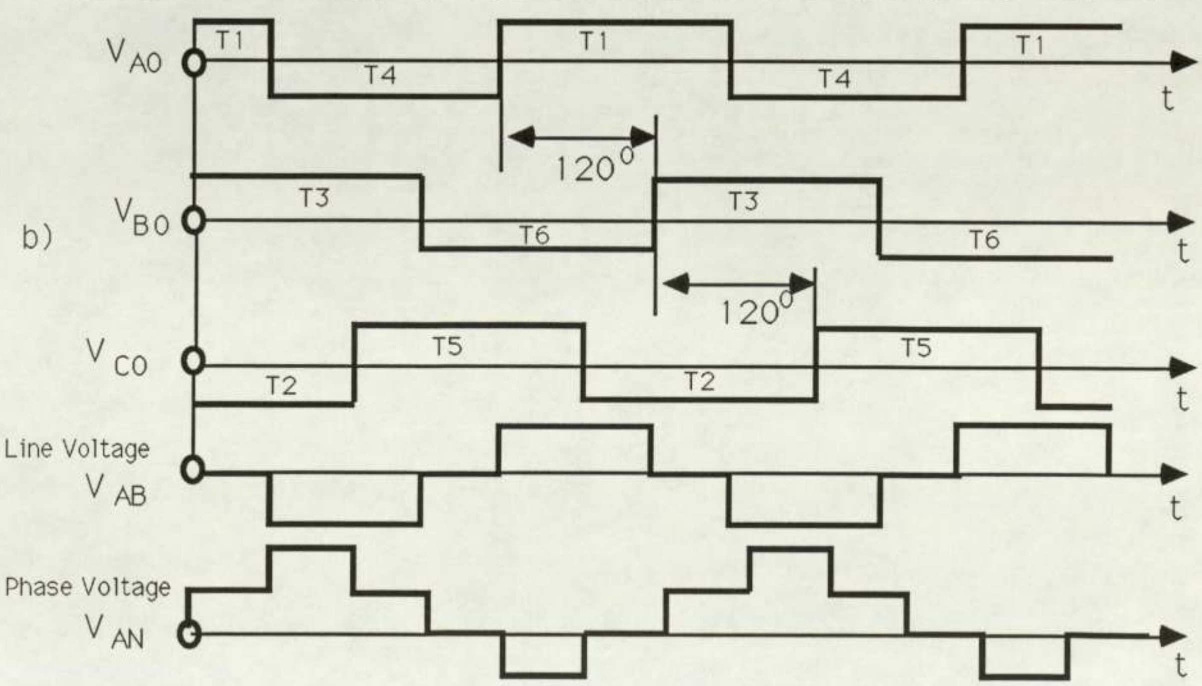


Figure 2.6 Voltage waveforms in a V.S.I

In conventional inverters where slow turn-off thyristors are used,there is sufficient delay time between the turn-off time of thyristor T1 and the turn-on time of thyristor T4,and with the machine reactance.,dc-short circuit of the line supply is prevented.

Now with the rapid turn-off thyristor or with the G.T.O's which have smaller turn-off time,the delay times becomes shorter.This problem can be overcome with the use of a delay time circuit(protection circuit against short circuit on one leg).An exemple of such circuit will be presented later on in the chapter dealing with practical work.

The square wave mode is commonly used in such drives,where torque pulsations may be tolerated to some extent.

2.2.6 Current source inverter (C.S.I)

The current source inverter-fed induction motor (C.S.I) has been around for 15 years,for the reason it provides a precise control of electromagnetic torque with a minimum number of switching elements in its power circuit.

An immediate consequence of the current source inverter is the simplification of the control circuitry that is required to commutate the thyristors.The commutation circuit consists only of capacitors and diodes.Commutation transformers or reactors are eliminated except for a small value of inductance in series with each thyristor to limit the di/dt to a safe value.Elimination of commutation reactors not only allows increased operating frequencies but also drastically reduces audible mechanical noise levels in the operating equipment.

Because there is no requirement for special commutating transformers and premium grade thyristors,the current source inverter design is economically attractive at all power levels.But now with the self extinguishing ability of the G.T.O thyristor,the current source inverter has become less competetive.

The current source inverter drive can be used with induction motors or synchronous motors.In the configuration given in FIG2.7,[2.21],the variable DC voltage can be obtained from a controlled rectifier.At the output of the rectifier a smoothing "LC" filter is required to minimize "beating" motor current caused by the difference in inverter output frequency and rectifier line frequency.

An alternate means of adjusting the DC bus input to an inverter is a three-phase full wave bridge and a chopper.This circuit has the advantage of good power factor at the AC-input and the possibility of faster voltage response due to a smaller "LC" filter time constant on the output of a high frequency chopper controller.

This variable DC link voltage is then converted to a current source inverter by connecting a large inductor in series.The inverter conducts the current to the three phases of the machine to generate a six stepped current waveform.

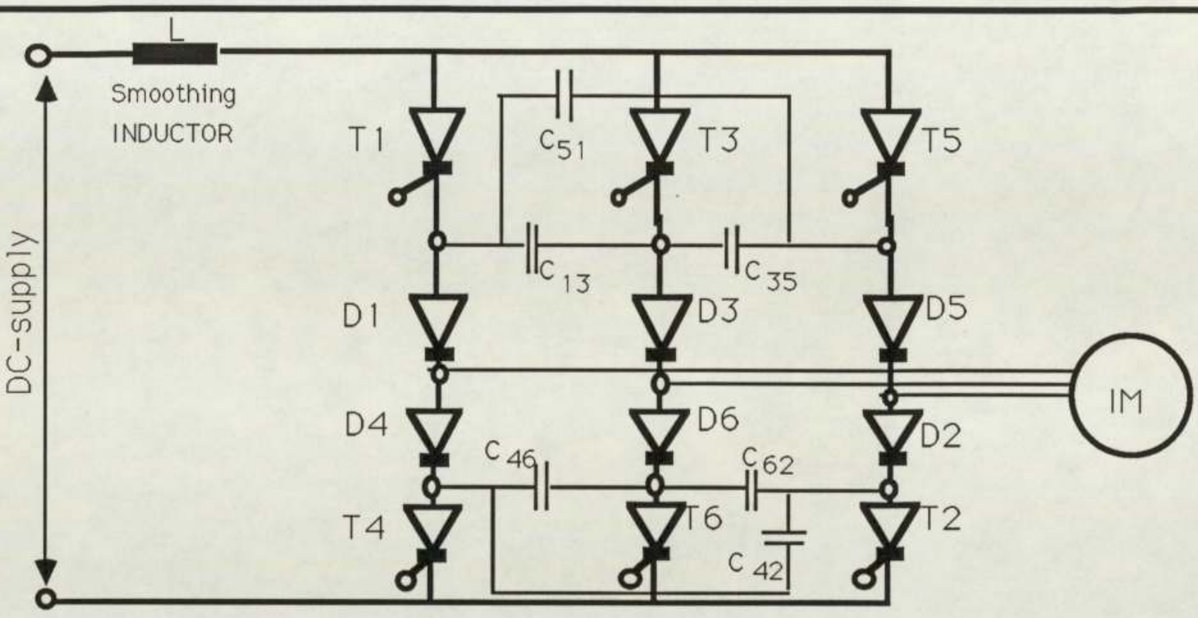
The machine terminal voltage is nearly sinusoidal with superimposed voltage spikes due to commutation.

A diode bridge rectifier followed by a chopper can also constitute the variable DC voltage.

In the configuration shown in fig 2.7,capacitors and diodes are used in order to help forced commutation in the inverter.In addition to this,diodes tends to isolate the capacitors from the load and to ensure their energy storage which is needed during commutation.

The six step current wave generated by the C.S.I may cause problems of harmonic heating and torque pulsations at low speed operation in the same way as for the V.S.I inverter.This problem can be overcome by modulating the DC link current,or by using two inverters which can be operated with phase shift to generatea twelve-step current wave.

These drives are widely used and have received much interest in recent years especially in traction application,particularly by the French Railways as mentioned by COSSIE.A in [2.22] and [2.23],and by Jacovide in [2.24].



a- Basic circuit of a C.S.I

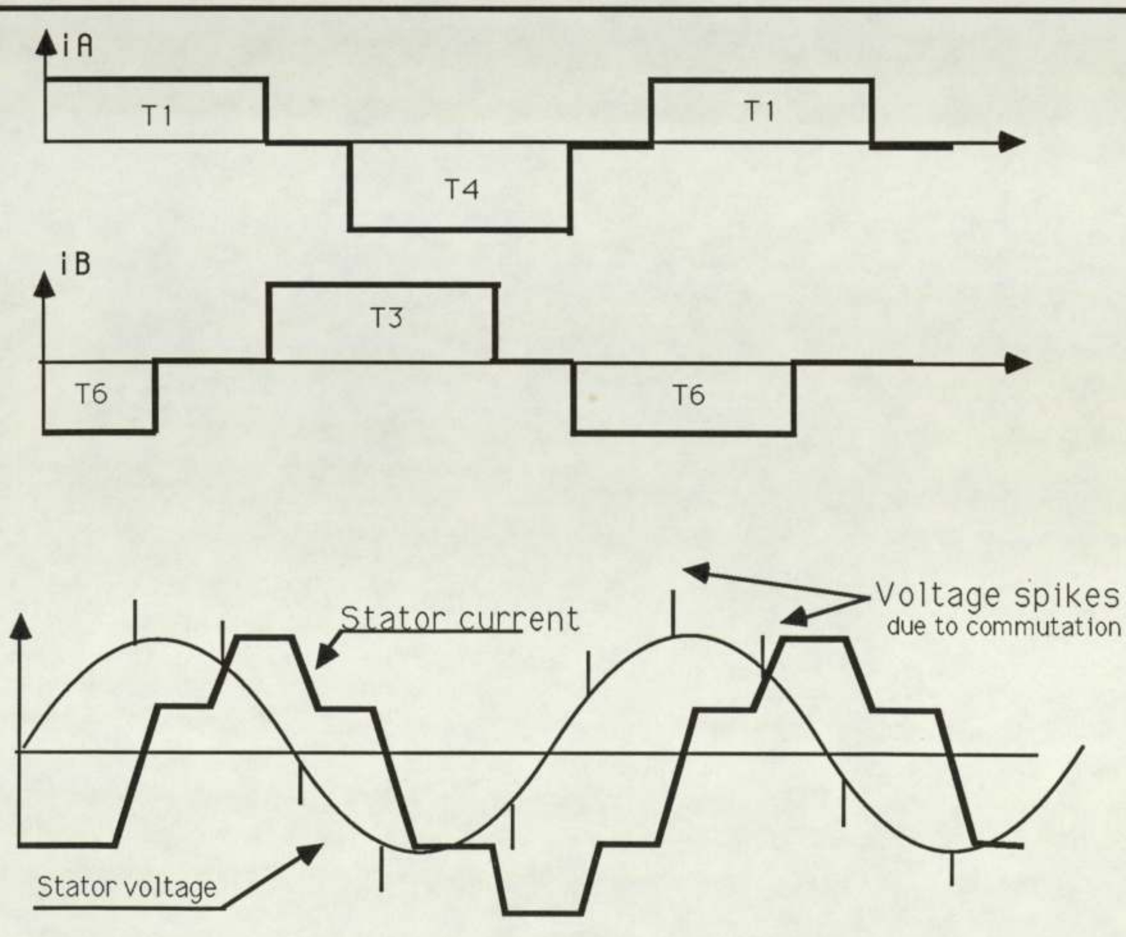


Figure 2.7 Current and voltage waveforms in a C.S.I

2.2.7 Pulse width modulated inverter (P.W.M).

The technique of Pulse Width Modulation (P.W.M),by which an inverter operating from a fixed-voltage DC-supply can generate an AC-output variable frequency and variable voltage is well known,[2.25].

Unlike the standart square wave inverter,presented in the previous section,the operational advantages of the pulse width modulated inverter are the elimination of undesirable low order harmonics,reduced torque pulsations [2.27],[2.28],[2.29].and they do not require a double stage of power conversion as was the case with the previous one as mentioned by B.K BOSE [2.21].

In an ideal P.W.M inverter,the output voltage waveform is a sinewave.

Practical P.W.M converters provide a series of rectangular pulses of constant magnitude equal to the DC supply voltage.In earlier systems,as exemplified in [2.26] P.W.M control signals were generated with the help of electronic hardware,but lately microcomputer have begun to play an important role in the computation of the switching points for P.W.M with selected harmonic elimination [2.27],[2.28].

The required switching sequences are determined on mainframe computers.Results are stored in lookup tables in

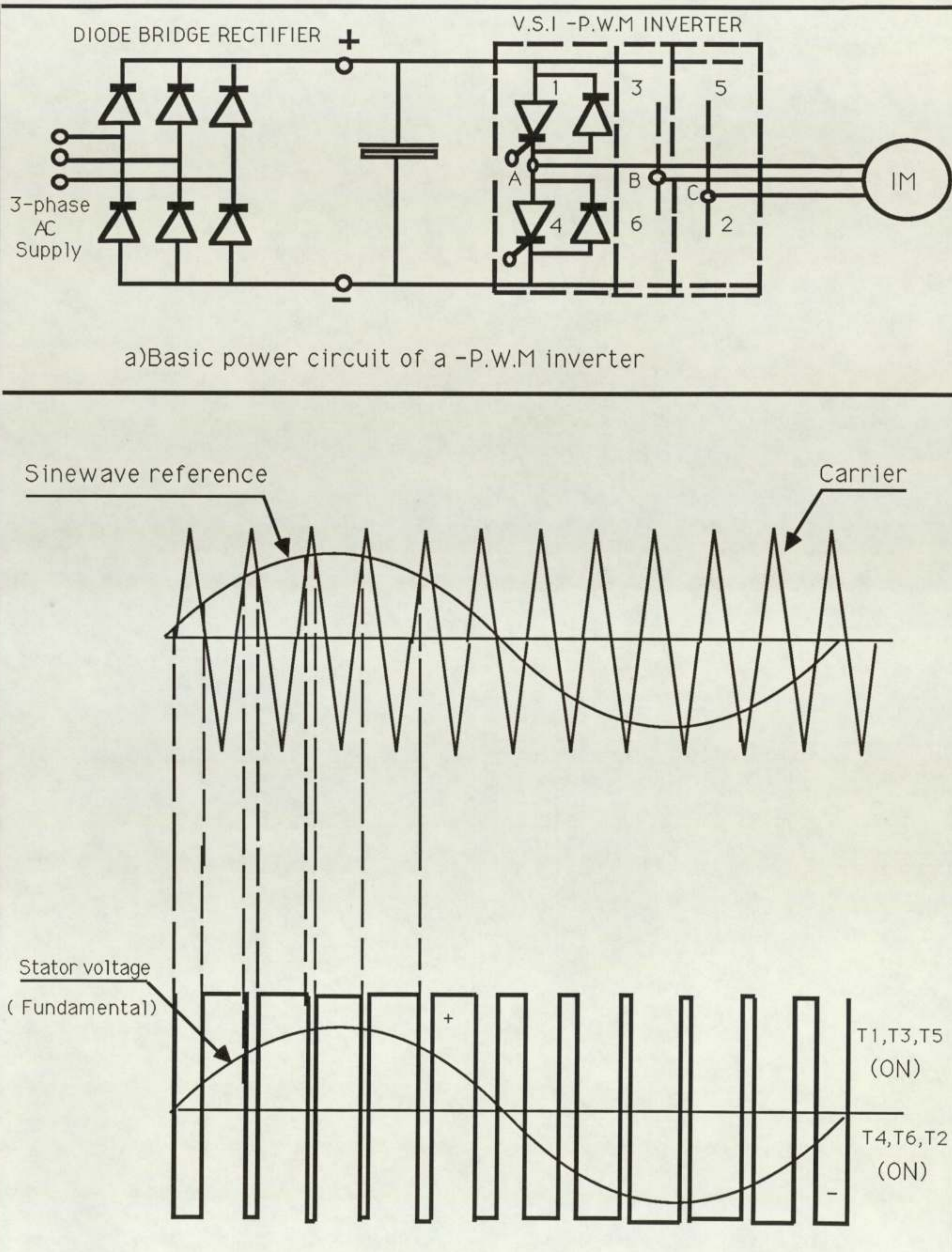
ROM and are used to control the switching of the inverter devices.Various P.W.M strategies are reported in the literature using software control with microprocessors.The notched square wave with fixed peak to peak amplitude and variable pulse width are obtained by comparing a modulated waveform with a high frequency carrier as shown in FIG 2.8.It is known that a higher switching rate produces a lower harmonic content,however the inverter switching losses increases with the increase of the switching frequency and becomes a restricting factor beyond a certain limit.In addition the maximum frequency is fixed by the finite turn-off time of the thyristors.

This restriction means that with the P.W.M inverter,High Speed Semiconductor switching devices are required.

In the configuration shown in FIG 2.8,[2.21],the P.W.M inverter is supplied by a fixed DC link voltage (there is no need for a controlled rectifier or a front end chopper),the switches are operated at higher frequencies so as to chop the output wave for the double purpose of voltage control and low harmonic elimination

Many modulation strategies exist,and all of them try to find a compromise between the requirement of waveform quality and inverter efficiency.

These types of drive are widely used from low to high power range where torque pulsations are not tolerated.



a)Basic power circuit of a -P.W.M inverter

Fig 2.8 Basic principle of P.W.M voltage generation

2.3. Voltage conversion classification

The power converters feeding three phase AC drives,presented in the previous section can be classified into two categories:direct and indirect conversion.The classification showing the different layouts of voltage conversion is given in FIG 2.9.

In order to select between these various configurations,some distinguishing criteria are needed.

The type of AC machine: induction or synchronous motor with reference to its power range,speed range,environmental conditions and its operating mode either continuous or intermittent.

The type of power supply: has to be defined according to its voltage range and specified as AC or DC supply ,single or three phase,fixed or variable frequency.

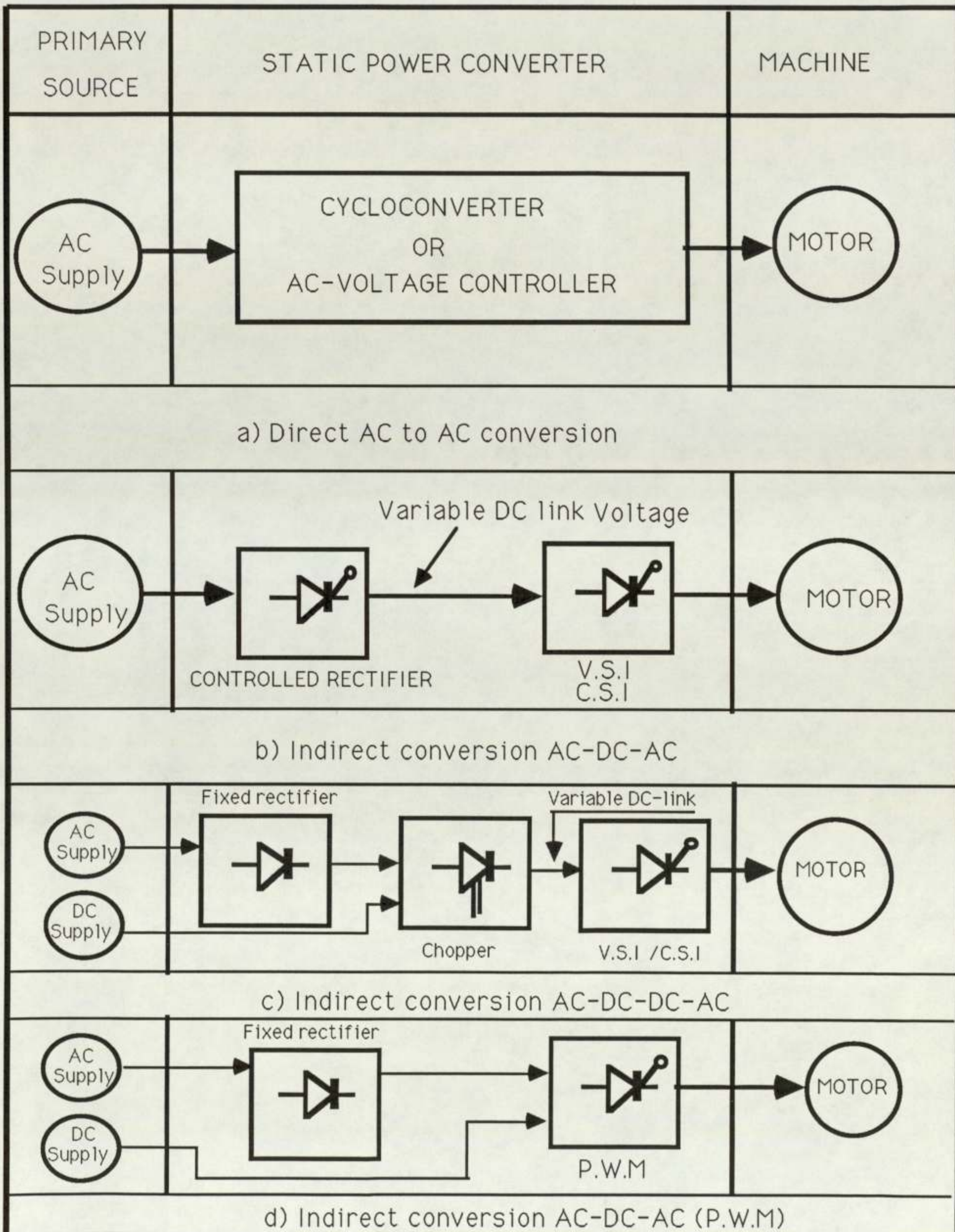


Figure 2.9 Different types of voltage conversion

2.4 Drive systems comparison

In this last section,a comparison of the different systems which have been presented,will be given and finally ,a conclusion will be drawn about the most suitable system in an induction motor drive application.

In the case of direct AC conversion,the voltage controller is the simplest of all the power converters [2.1],[2.2],however phase control used to control the output voltage has many disadvantages [2.3],[2.4].

With the cycloconverter,the power circuit is also simple due to the absence of commutation component,drive generation is simple and can be designed for four quadrant operation,due to the capabilty of bidirectional power flow[2.19].However it has an important drawbck created by the restriction required to avoid short-circuit between phases,and the output frequency is very limited [2.18].

The square wave Voltage Source Inverter is simpler than the previous case,but necessitates a variable DC link,in addition a filter "LC" is needed to attenuate the voltage and current ripples.Further to this,the square wave V.S.I has the weakness of generating a square wave rich in low order harmonics [2.21].These effect are not tolerated in some application especially in traction,firstly because of generating noises and disturbances on the signalling and telecommunication systems,and secondly because of creating torque pulsations and overheating at low speed operation.

The problems of harmonics can be reduced by using P.W.M inverters [2.27[,[2.28].In fact these systems are cheaper than the previous ones,because they do not need a contolled DC link voltage and use less components [2.21].

The output voltage and frequency from the P.W.M inverter can be varied according to a switching pattern strategy.These types of drives also have some limiting factors arising from the switching losses and the thyristor's finite turn-oo time.

In contrast to the cycloconverter and the V.S.I,the C.S.I has several good features such as simple design and control using slow turn-off thyristors thereby permitting cost saving.The power circuit of such a system is rugged and reliable,with no possibilty of short circuit due to large smoothing reactor [2.21].

In spite of the various merits mentioned above,the C.S.I inverter has some disadvantages especially at low speed operation as a result of pulsating torque caused by the square wave current which is also heavy in low harmonics. The effects are the same as for the V.S.I inverter.In addition a large DC link inductor is required to regulate and smooth the current at the inverter input which increase the weight and cost of the drive system.

2.4.1 Conclusions

To conclude,it is obvious through the review of these different systems,that all of them can be used with an induction motor,or synchronous motor,depending upon their specific applications.
It can be stated that cycloconverters are used generally in the low speed.
The V.S.I and the C.S.I systems,are used for high speed application and over the complete power spectrum.

Finally,with the advance in power electronic technology and the use of new techniques of modulation,both systems (V.S.I and C.S.I) in Pulse Width Modulation inverter configuration are becoming more competetive and attractive than the cycloconverter system,and there is a clear tendancy toward their use with induction and synchronous motors.

CHAPTER THREE

3.0 Methods of Speed/Torque control in AC drives.

3.1 Introduction

The nature of the AC induction motor presents a number of special problems in its application as a drive.Certainly as the frequency of the motor is adjusted,the voltage must also be adjusted to keep a nominal voltage to frequency ratio,thereby maintaining the flux at essentially the same level.The fulfilment of this requirement ensures the ability of generating the highest possible torque per ampere of the stator current and therefore results in the best possible utilisation of the available current capability of the drive,[3.1],[3.2].

In the case of the DC machine,sensing of armature and field current will yield a fairly accurate measure of the flux level in the machine and,hence the developed torque.However the flux level and,hence the torque in an induction motor are not as easily controlled,in fact the flux level in the air-gap depends,in a complicated way,on the currents in every circuit of the machine and on the relative position of the stator and rotor field.

Various methods of controlling induction motors have been used,such methods may be divided into two groups:the scalar and the vector control

In the following subsections,the control of induction motors is reviewed,this is followed by the method of flux sensing and finally a detailed description of the speed controller used in this project is given.

3.2 Characteristics of induction motor.

The induction motor drive system is basically a multivariable control system.The voltage and frequency are the control inputs and the output may be speed,position,torque,air-gap flux,stator current or a combination of them.

The dynamic performance is somewhat complex because of the coupling effect between the stator and the rotor phases,where the coupling coefficients vary with the rotor position,in addition the parameters of the machine may vary with temperature and saturation adding further complication to the system analysis.

Therefore the machine model can be described by differential equations with time varying coefficients.To simplify the system equations,the time varying parameters are eliminated by orthogonal transformation,which results in a set of (d-q) system equations and therefore the induction motor may be analysed in a similar way as the DC motor.The dynamic model of the machine can be expressed with respect to either a stationary or a rotating reference frame(a detailed presentation of these transformations will be given in chapter five,dealing with the simulation and modelling of the drive system).

3.3 Scalar control methods.

Scalar control relates to the magnitude control of a variable only,the commands and feedback signals are DC-quantities which are proportional to the respective variables.This is in contrast to vector control,where both magnitude and phase of a vector variables are controlled.

3.3.1. V/Hz-control.

A simple and popular open loop V/Hz speed control method for an induction motor is shown in fig(3.1).

From the equivalent circuit per phase, of an induction motor ,the relationship between the flux ,the stator frequency and the applied voltage can be derived as follows:

$$V_1 = R_1 . I_1 + j\omega_1 I_1 . L_1 + E_m$$

3.1

$$E_m = V_1 - (R_1 . I_1 + j\omega_1 . I_1 . L_1)$$

3.2

$$E_m = K . \phi.\omega$$

3.3

If the stator impedance is neglected ,an approximate flux value is obtained from equation (3.2) and (3.3).Hence

$$\phi = K (E_m / \omega) = K (V/\omega)$$

3.4

From equation 3.4 we can see that the flux level is proportional to the ratio of the supply voltage and the supply frequency.To keep the flux level constant ,the ratio (V/ω) should be maintained constant at all operating conditions.This appoximation is satisfactory when the drive system is operating at high frequencies and leads only to moderate machine underexcitation,however at the low end of the frequency range,the neglected stator resistive drop becomes important compared to the applied voltage thus ignoring it leads to severe underexcitation and excessive loss of the torque capability.

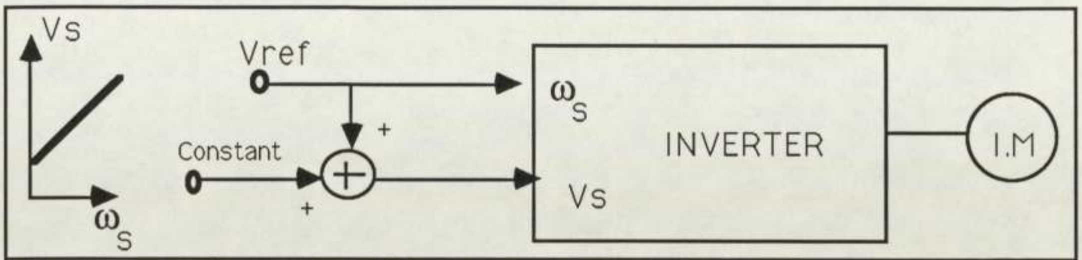


fig 3.1 Constant V/Hz control with linear shift

3.3.2 Voltage boost compensation

One of the methods used to correct the deficiency of the approximation presented above,is to apply a boost-voltage when the drive system is operating in the low frequency range to compensate for the voltage drop occuring in the stator resistance.

This approach generates high torque at low speed,but it has a major drawback due to heating problems of the motor,in fact at no load the magnetizing current due to the boost voltage is very important.If current and speed feed-back control are used ,the drive will be very sensitive at low frequency operation because of the high level of the no-load current due to the applied voltage boost.The V/Hz control with voltage boost at low speed is represented in fig 3.2.

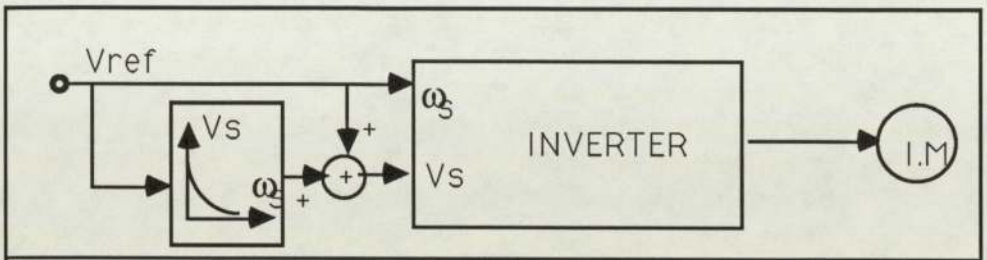


Fig 3.2 Constant V/Hz control with voltage boost

3.3.3 Current boost compensation

From the equivalent circuit and equations 3.2 and 3.3,when the induction motor is driven from a fixed voltage and a fixed frequency the flux level decreases with an increase in the load current because of the voltage drop across the stator.To neutralise the effect of this drop ,the stator voltage should increase proportionally to the load current ,one common technique used is the current-dependant boost.The control signal V_1 is obtanined by full wave rectification of the stator current.This approach causes an overexcitation but the torque capability is safeguarded.A basic "V/Hz" control method is briefly described by the following scheme:

The power circuit consists of a phase controlled rectifier with a single or three-phase controlled rectifier,LC-filter,and six-stepped inverter.In this scheme the frequency ω_e is the command variable and it approximates to the motor speed by neglecting the small slip frequency.The scheme is defined as V/HZ control because the rectifier voltage command V_s is generated directly from the frequency signal through a V/HZ gain constant "G".In steady-state operation,the machine airgap flux Ψ_m is approximately related to the ratio V_s/ω_e .Therefore maintaining the rated airgap flux will provide maximum torque.However at low speed,the stator voltage will tend to be zero and it will essentially be absorbed by the stator resistance.To compensate for this deficiency,an auxiliary voltage V_0 is injected to overcome the effect of the stator resistance so that rated airgap flux and full torque are available even at very low speeds.With open loop voltage control,the AC line voltage fluctuation and impedance drop will cause fluctuation of the airgap flux.This fluctuation can be prevented by providing closed loop voltage control of the rectifier,in addition if the speed drifts with the variation in load torque a speed loop is added.

An alternative V/Hz control scheme with slip regulation is shown in fig(3.3).Here the error of the speed control loop generates the slip command ω_{sl} through a proportional and integral(PI) controller and limiter.The frequency command also generates the voltage command through a V/Hz function generator which incorporates the low frequency stator drop compensation.Since the slip is proportional to developed torque,the scheme can be considered as torque control within a speed control loop.

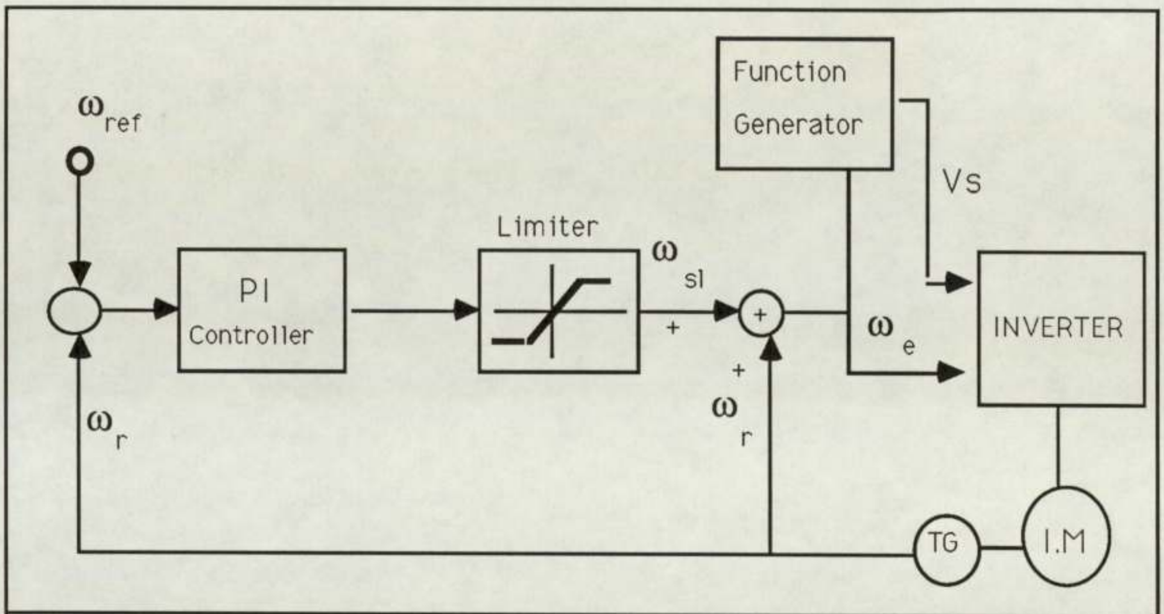


Fig 3.3 Constant V/Hz control with slip regulation

3.3.4 -V/Hz control with flux feed-back.

As described earlier,the V/Hz control scheme has the disadvantage that the airgap flux may drift,and as a result the torque sensitivity with slip or stator current will vary.If the correct V/Hz ratio is not maintained,the flux may be weak or may saturate the magnetic circuits.

The stator circuit parameters may also vary due to temperature and saturation,causing drift in the airgap flux.If the airgap flux decreases,the slip ω_{sl} has to increase for the same torque demand.As a result,the machine's maximum torque capability will decrease and the transient response will deteriorate.To compensate for this effect an additional flux loop control is added..The flux control loop compares the command flux and the feedback flux and generates the voltage command for the inverter.

The flux level may be sensed directly or by synthesis,this feature will be discussed in the following subsections.

3.4 METHODS OF FLUX SENSING.

3.4.1. Direct flux sensing.

In order to effect accurate flux and,hence torque regulation without sensitivity to motor parameters,a direct measurement of the flux is effected by the use of Hall probes.The measured air-gap flux signal is fed-back to the processing unit and used to decouple the torque producing component of the stator current from the flux producing component.

This method is considered more costly and perhaps impractical,because of the special mechanical work required on the motor.In addition this method suffers from the unreliability of the flux measurement,since Hall sensors are very temperature sensitive.Finally this technique would make the drive incompatible with standard motors.

3.4.2 **Search coil sensing.**

Another method of flux measurement is effected by inserting small coils made of very thin wire which are placed around the slots of the stator of the machine.The voltage induced in these coils is sensed and then integrated to produce a measure of the flux linking the coils.In order to effect accurate torque and flux measurement proper placement of these coils in the motor is required to obtain the correct signal of the air-gap voltage E_m.If the centre line of the coils is misaligned (stator and sensing coils) this will result in an incorrect phase relationship between the flux and the stator current.In addition flux coils require extra-work and such work will increase the cost of the drive system.Since the coils are formed of very thin wire (approximately 40 gauges), they are subject to breakage due to vibration.Repair of the flux sensors involves the removal of the tooth top insulation strip from the stator slots containing the flux coils.This method like the above Hall sensor approach is incompatible with the use of the standard induction motors,in addition the method suffers from the inaccuracy of integration at low speed [3.3].

3.4.3 Indirect method of flux sensing.

The flux feed-back information is not obtained by direct measurement as discussed previously,because of their associated practical difficulties and costs,but is derived by processing the stator voltage and the line current.

The main parameters required to derive the flux signal are:

a) the magnetizing current I_M

b) the line stator current I_1

c) the motor inductive parameters L_1,L_2,L_M

d) the stator frequency ω_1

e) the reactive power Qx

L_1,L_2,L_M can be obtained from experimental tests on the induction motor.

I_1 is obtained from the current transducer.

ω_1 is known (speed reference).

Qx is measurable using a V.A.R-meter.

Hence the magnetizing current can be derived from the equivalent circuit,as follows:

$I_M = E_M/\omega_1 L_M$ 3.5

$I_2 = E_M/[(\omega_1 L_2)^2+(R_2\omega_1/\omega_2)^2]^{1/2}$ 3.6

$\cos\theta = \omega_1 L_2/[(\omega_1 L_2)^2 + (R_2\omega_1/\omega_2)^2]^{1/2}$ 3.7

-

from the phasor diagram represented in fig (3.4-a)

$$I_{2x} = I_2\cos\theta = I_2^2\,\omega_1\,L_2/E_M \qquad 3.8$$

$$I_1^2 = I_M^2 + 2I_2^2\,L_2/L_M + I_2^2 \qquad 3.9$$

In equation 3.9 the current "I_2" cannot be measured directly therefore this quantity will be substituted by its equivalent expression as follows:

$$I_2^2 = K_1(I_1^2 - I_M^2) \qquad 3.10$$

where $K_1 = L_M/(L_M + 2L_2)$

As mentioned earlier Qx represents the reactive power,to obtain the analytical expression equivalent to Qx,first consider the phasor diagram shown in fig(3.4-a).

When the drive is supplied by a three-phase voltage (Va,Vb,Vc),the respective currents (Ia,Ib,Ic) are delayed by an angle θ.Hence:

$$Va = V_M \sin\omega_1 t$$

$$Vb = V_M \sin(\omega_1 t - 2\pi/3)$$

$$Vc = V_M \sin(\omega_1 t + 2\pi/3) \qquad 3.11$$

-

$$Ia = I_M \sin(\omega_1 t - \theta)$$

$$Ib = I_M \sin(\omega_1 t - 2\pi/3 - \theta)$$

$$Ic = I_M \sin(\omega_1 t + 2\pi/3 - \theta) \qquad 3.12$$

-

The phasor diagram shown in fig(3.4) is used to reduce the three-phase system to a two-phase system.

Equating m.m.f's on the 2-phase axes,the axis of phase "a" having an angle "θ" with the axis of phase "α",to obtain:

$$I_\alpha N/2 = N/3\{I_a\cos\theta + I_b\cos(\theta + 120) + I_c\cos(\theta + 240)\}$$

i.e

$$I_\alpha = 2/3\ \{I_a\cos\theta + I_b\cos(\theta + 120) + I_c\cos(\theta + 240) \qquad 3.13$$

Similarly

$$I_\beta = 2/3\ \{\ I_a\sin\theta + I_b\sin(\theta + 120) + I_c\sin(\theta + 240) \qquad 3.14$$

The equations (3.13 ,3.14) represent the transformation of current vector expressed in term of a,b,c components into the same quantity expressed in α,β components.

Transformations between rotating and fixed axes :

Now $I\alpha$ and $I\beta$ are instantaneous values of balanced 2- phase sinusoidal current I.

$$I_d = I_\alpha \cos \omega_1 t + I_\beta \cos(\omega_1 t + 90) \qquad 3.15$$

$$I_q = I_\alpha \sin \omega_1 t + I_\beta \sin(\omega_1 t + 90) \qquad 3.16$$

With I_α at its maximum value when $\omega_1 t = \gamma$,and for a mechanical rotational frequency equal to the frequency of the currents ($\omega_r = \omega_1$),from equations (3.15 . 3.16) we get:

$$I_d = I_M\{ \cos(\omega_1 t - \gamma)\cos\omega_1 t + \cos(\omega_1 t - \gamma +90)\cos(\omega_1 t +90)\}$$

we get finally:

$$I_d = I_M\cos\gamma$$

$$Iq = I_M\sin\gamma \qquad 3.18$$

In a similar way the expression for the voltages in the stationary frame are :

-

$$Vd = V_M\cos\gamma$$

$$Vq = V_M\sin\gamma \qquad 3.19$$

By taking the crossproduct of equations 3.18 and 3.19,we obtain:

-

$$V_d I_q = [V_M \sin\omega_1 t] x [[-I_M \cos(\omega_1 t - \theta)] \qquad 3.20$$

$$V_d I_q = -1/2[V_M I_M \sin\theta] - 1/2[V_M I_M \sin(2\omega_1 t - \theta)]$$

similarly,

$$V_q I_d = 1/2[V_M I_M \sin\theta] - 1/2[V_M I_M \sin(2\omega_1 t - \theta)] \qquad 3.21$$

-

As can be seen from these products the component ($1/2 V_M I_M \sin\theta$) represents the reactive power.It is necessary to cancel the AC-components by substracting equation (3.20) from equation (3.21).

The final result is:

-

$$2Qx = V_q I_d - V_d I_q = V_M I_M \sin\theta \qquad 3.22$$

-

The expression represented in equation (3.22) can be processed by electronic means,to obtain an information on the reactive power.The bloc digram shown in fig 3.5-a represents the scheme for the derivation of the reactive power.

To derive the flux signal which is the useful information,a scheme similar to the one shown in fig 3.5-a,however the difference being that the multiplier voltage inputs are corrected for a stator drop accross the inductance.The equations,obtained from manipulation of the previous expressions are used to implement the flux feed-back method which is independant of voltage drop when the drive system is loaded.

Instead of voltage V_1,V''_1 is used:

$$V''_1 = V_1 - j\omega_1 L' I_1$$

$$V''_d = V_d - j\omega_1 [L_1 + L_M L_2/(L_M + 2L_2)] I_d$$

$$V''_q = V_q - j\omega_1 [L_1 + L_M L_2/(L_M + 2L_2)] I_q \qquad 3.23$$

On this basis ,the flux regulation scheme takes the form of the block diagram of fig 3.5-b.

Some of the control principles of voltage-fed inverters as discussed above are also valid for current-fed inverters.

3.6 Volt/Hz control with current feed-back.

This drive can be used in electric-vehicle-type applications where torque instead of speed control is required.Here the DC-link current and inverter frequency are the two control parameters where the current can be varied by modulating the firing angle of the front-end rectifier.In contrast to the voltage fed inverter,a current fed inverter cannot be controlled in an open loop manner.A closed-loop control system of a current fed inverter where the current and slip are controlled independantly is shown in fig(3.5).The principal disadvantage of this system is that the machine airgap flux has no control,and as a result is rarely used in practice.

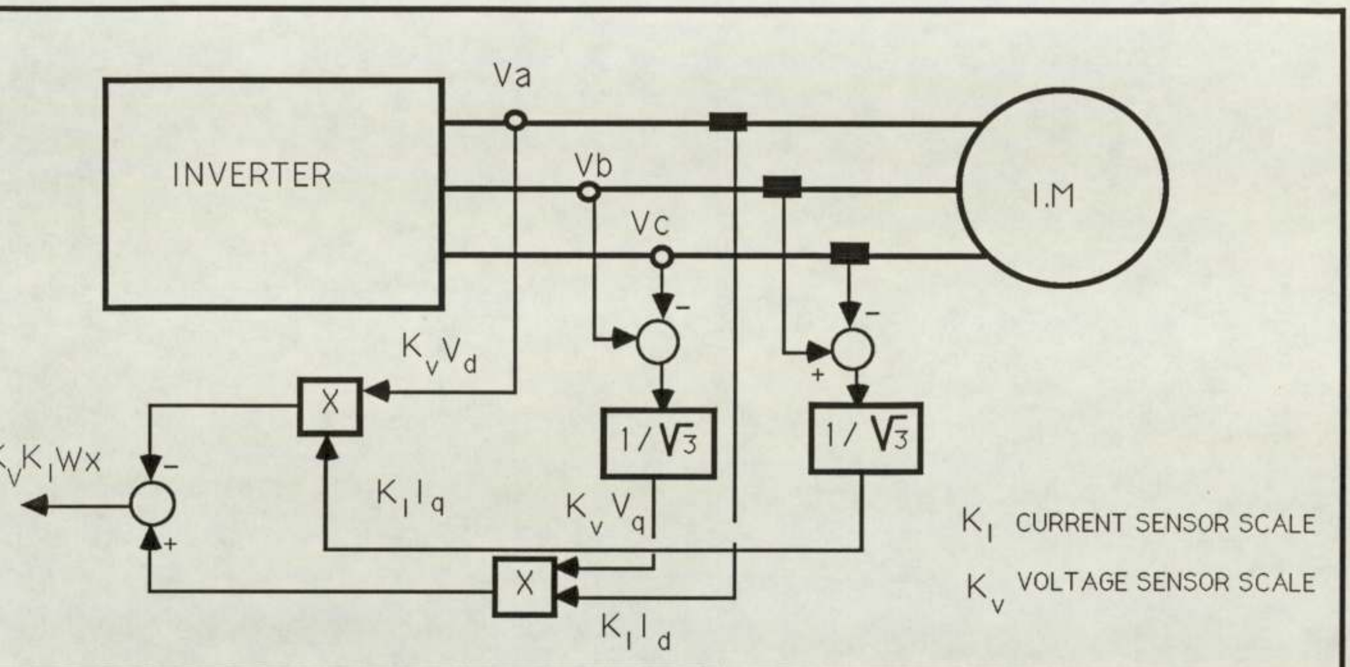


Fig 3.5-a SCHEME FOR DERIVATION OF THE REACTIVE POWER SIGNAL

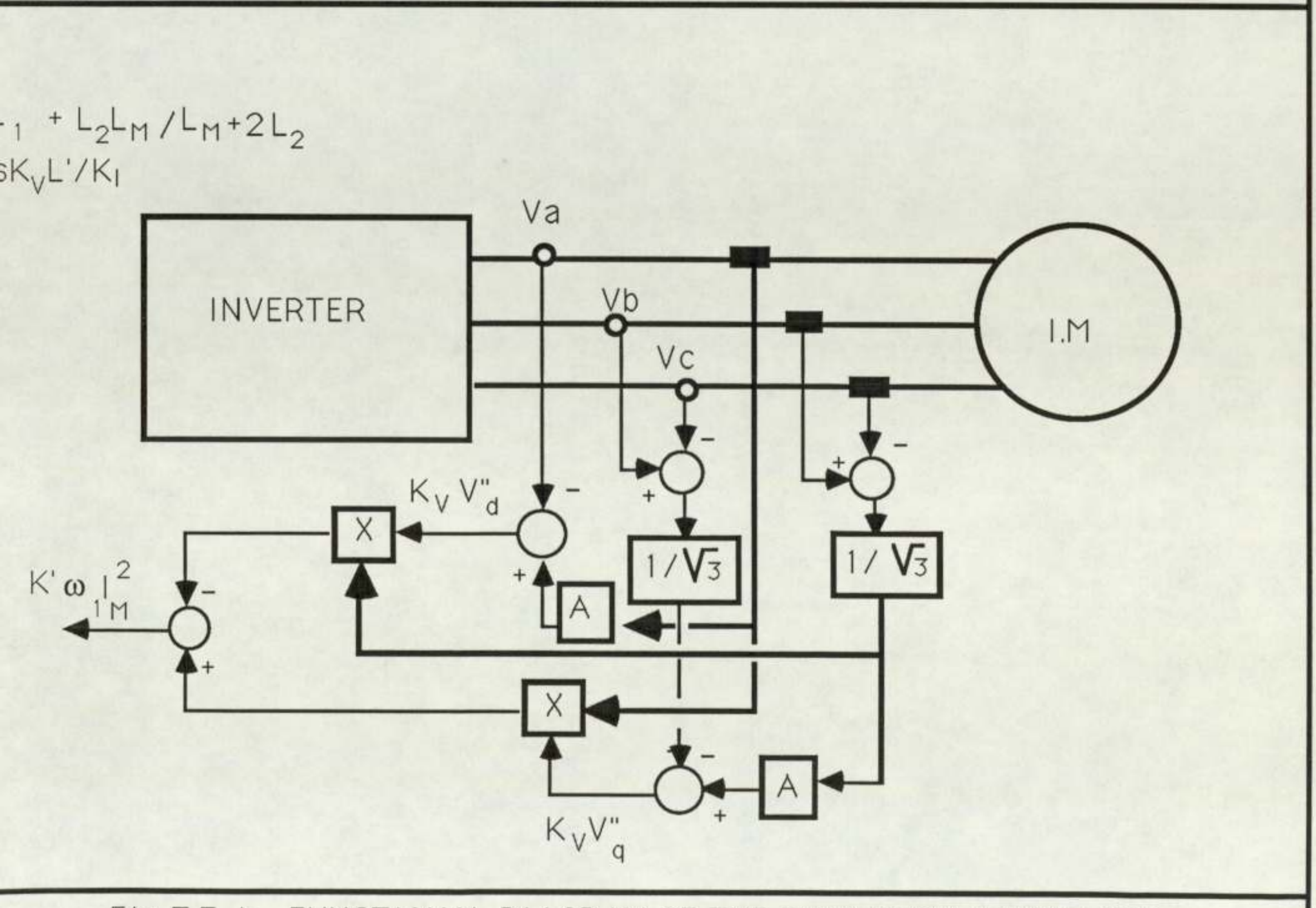


Fig 3.5-b FUNCTIONAL DIAGRAM OF THE FLUX REGULATING LOOP

The circuit diagram represented in fig(3.5-a)is designed to obtain the reactive power information according to the equations derived above (eq.3.22).

The input signals (stator currents and the the stator voltages) are :

a) sensed through current and voltage sensors

b) the voltage and current signals obtained from the sensors are processed through amplifiers(RS 741) according to equations given in (eq 3.14 and 3.15).

c) the following processing consists in multplying the signals obtained from the amplifer as expressed by equations (3.20 and 3.21).

d)finally the output signals from the multipliers are processed in an analog subtractor according equation 3.22

The circuit diagram given in fig(3.5-b) is similar to the one given in fig (3.5-a) with further processing as given by equation 3.23.

3.5 **Slip power recovery control.**

In this approach the slip measurement is used to represent the motor speed reduction with load.For evaluating slip accurately,a digital speed transducer coupled to the shaft is required.In its simplest implementation the output signal from the transducer (DC tachogenerator) is substracted from the analog frequency reference signal to provide the slip measurement. When the tachogenerator is not needed for speed regulation purposes,its use for only flux control makes such a solution expensive and often impractical.In addition this technique still fails to provide proper compensation when operating at low speeds.

A speed control system using the static Kramer method is shown in fig(3.6),this type of drive has characteristics similar to those of a separately excited DC machine,and therefore the control configuration is analogous to a phase controlled rectified DC drive system.With a constant airgap flux,the torque is proportional to DC link current I_d ,which is controlled by a feedback loop.The airgap flux remains approximately constant during the whole operation, as dictated by the stator voltage and frequency.

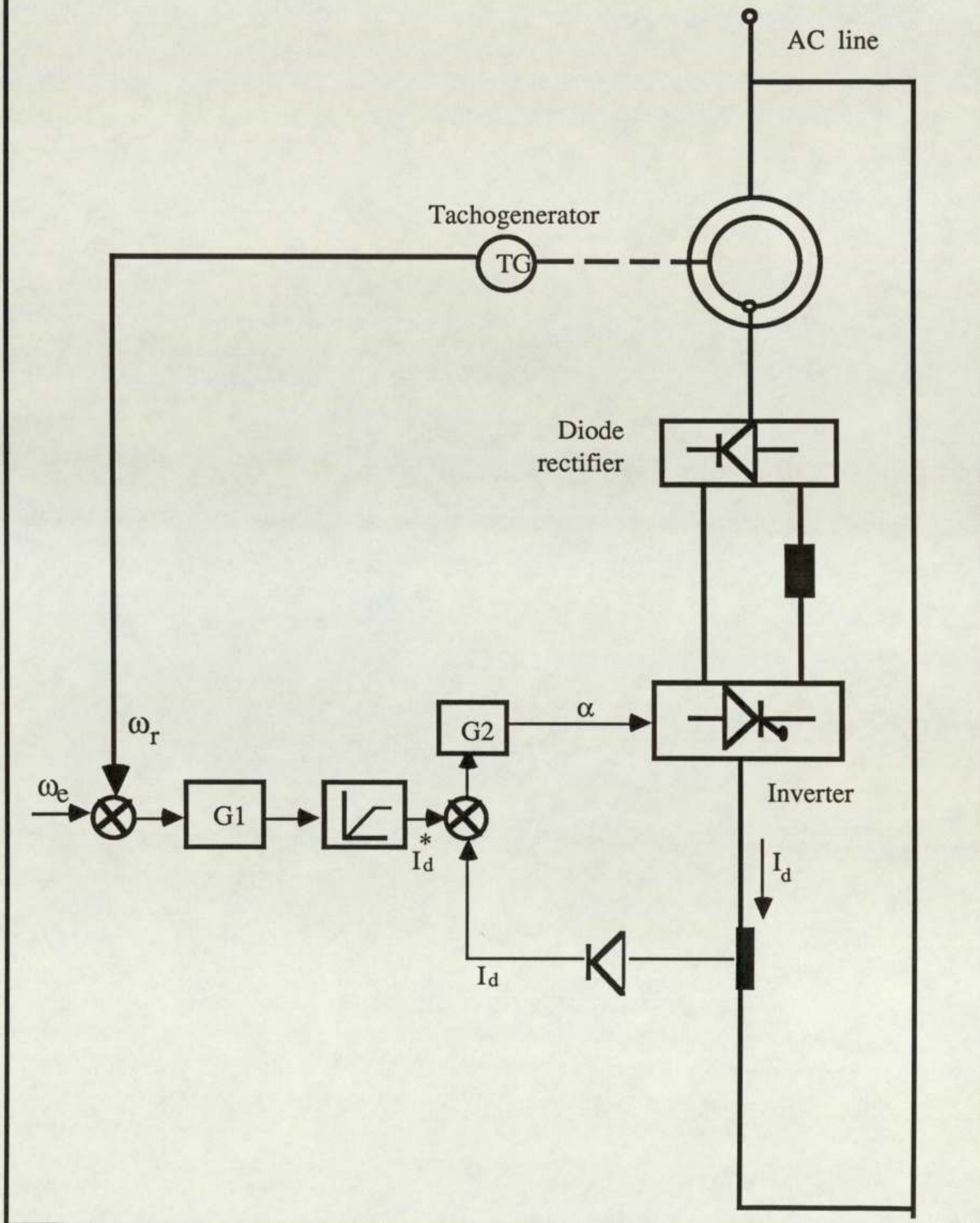


Fig 3.6. Speed control of static Kramer drive system

The control is simple however an associated disadvantage is that the drive system can be controlled in one quadrant only.

3.6 **Vector control method.**

In the scalar control methods of voltage-fed and current-fed inverter drives discussed so far,the voltage or current and the frequency are the basic control variables of the induction motor.In vector control,an AC machine is controlled like a separately excited DC machine,see fig(3.7).In a DC machine,the torque is given by :

$Te = K.I_a.I_f$,where the control variables "I_a" (armature or torque component of current) and "I_f" (field or flux component of current) can be considered as orthogonal or decoupled "vectors".Since the current I_f or the corresponding field flux is decoupled from the armature current I_a,the torque sensitivity remains maximum under both transient and steady-state conditions.This mode of control can be extended to an induction motor if the machine operation is considered with respect to a synchronously rotating frame where the sinusoidal variables appear as DC quantities.

3.6.1-**Induction machine representation for vector**

control

The dynamic behaviour of a three-phase induction motor with a squirrel-cage rotor or a short-circuited wound rotor can be conveniently described by vector equation in a d-q rotating reference frame.

With the appropriate constraints,the d-q frame can be made to rotate synchronously with the stator or rotor voltage,current or flux vectors.

The balanced system of three stator currents can be represented by a synchronously rotating stator current.

This is also equivalent to a system of two-phase stationary currents $I_{s\alpha}$ and $I_{s\beta}$ .The air-gap vector Ψ_m is also a synchronously rotating vector lagging behind the stator current vector by the torque angle σ.

The component of I_s along the flux axis is the field component I_{sd} and the other component in quadrature is the torque producing current I_{sq}.

With these resolved stator current components,the following relation depicting the induction motor operation can be derived.

$$V_{ds} = (R_s + L_s p) I_{ds} - \omega_s L_1 I_{qs} \qquad 3.18$$

$$V_{qs} = L_s \omega_s I_{ds} + (R_s + L_1 p) I_{qs} \qquad 3.19$$

$$E^2=V_{ds}^2+V_{qs}^2 \qquad 3.20$$

$$I_{ds}=I_s\cos\theta \qquad 3.21$$

$$I_{qs}=I_s\sin\theta \qquad 3.22$$

$$I_s^2=I_{ds}^2+I_{qs}^2. \qquad 3.23$$

In an induction motor ,$R_r^2 >> \omega_{sl}L_2L_m$

Hence

$$\psi_m = L_mI_{ds} \qquad 3.24$$

$$\omega_s = (R_r/L_m).I_{qs}/I_{ds} \qquad 3.25$$

$$T_e = (2/3)(L_mI_{ds})^2.R_r\omega_{sl} \qquad 3.26$$

$$T_e = (2/3)L_mI_{ds}.I_{qs} \qquad 3.27$$

Equations (3.24)and (3.27) would be the equations of a dc-motor if I_{ds} is replaced by the field current and I_{qs} by the armature current.In the case of a dc-motor,the two components exist physically and distinctly.They can be separately measured at the dc-terminal and controlled,whereas in the induction motor I_{ds} and I_{qs} are fictitious.The quantities to be controlled are the dc-link voltage and the operating frequency.These quantities must be manipulated so that the desired level of I_{ds} and I_{qs} are maintained inside the motor.

The two control inputs I_{ds} and I_{qs},the direct-axis and quadrature-axis component are respectively analogous to the field currrent "I_f "and armature current"I_a".Therefore the torque can be expressed as:

$$Te = k\psi_m I_{qs} = K.I_{qs}.I_{ds}$$

This basic concept of how I_{qs} and I_{ds} can be established as control vectors in the vector control method is explained in Fig(3.8),with the help of phasor diagrams in a synchronously rotating d^e-q^e reference frame.The phasor diagram is drawn with the air gap voltage V_g aligned with the q^e axis The stator current I_s lags the voltage V_g by (90-θ),i.e $i_{qs} = I_s \sin\theta$ is in phase with V_g and $i_{ds} = I_s \cos\theta$ is in quadrature with V_g

The current i_{qs} is the active or torque component of the stator current.The current i_{ds} is the reactive or field component of the stator current and is responsible for establishing the air gap flux ψ_m

From the phasor diagram ,the developed torque across the air gap is given $Te = k|\psi_m| i_{qs} = k i_{qs} i_{ds}$ where iqs and ids are shown in fig(3.8).The torque equation is therefore identical to that of a DC-machine.

The fundamentals of vector control implementation with the machine model can be explained as follows

The phase currents I_a,I_b,and I_c are converted to direct and quadrature components Isd,Isq,these components are then converted to a synchronously rotating frame by the unit vectors $\cos\omega_e t$ and $\sin\omega_e t$.

The basic principle of this control is to maintain the instantenious magnitude of the secondary flux constant and that asssures the independent control of flux and torque current.

A detailed analysis of the different methods used in the vector control is given in the follwing subsections.

3.6.2- **Orientation of stator current.**

The magnitude and phase of the stator current determine the flux produced and the torque generated in the machine.

These two parameters can be controlled by properly orienting the stator current and it can be done with respect to the stator mutual,or rotor vectors.From the equivalent circuit of an induction motor the stator mutual and rotor flux are identified.

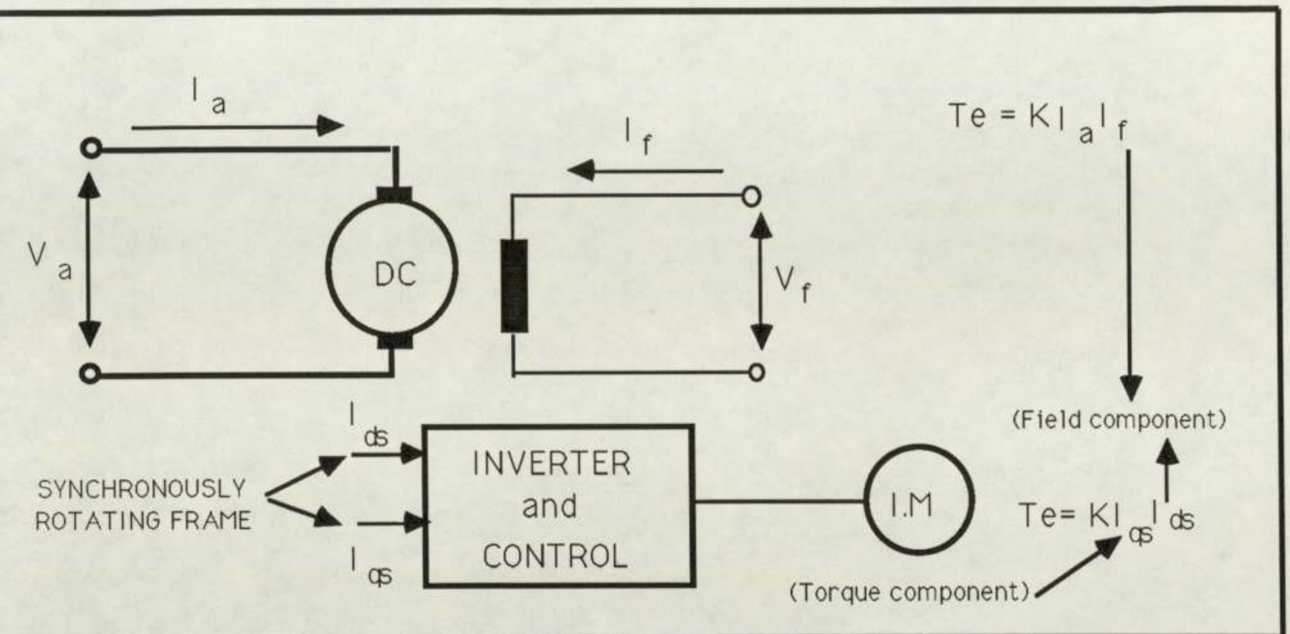


Fig 3.7 Induction motor and DC machine analogy in vector control

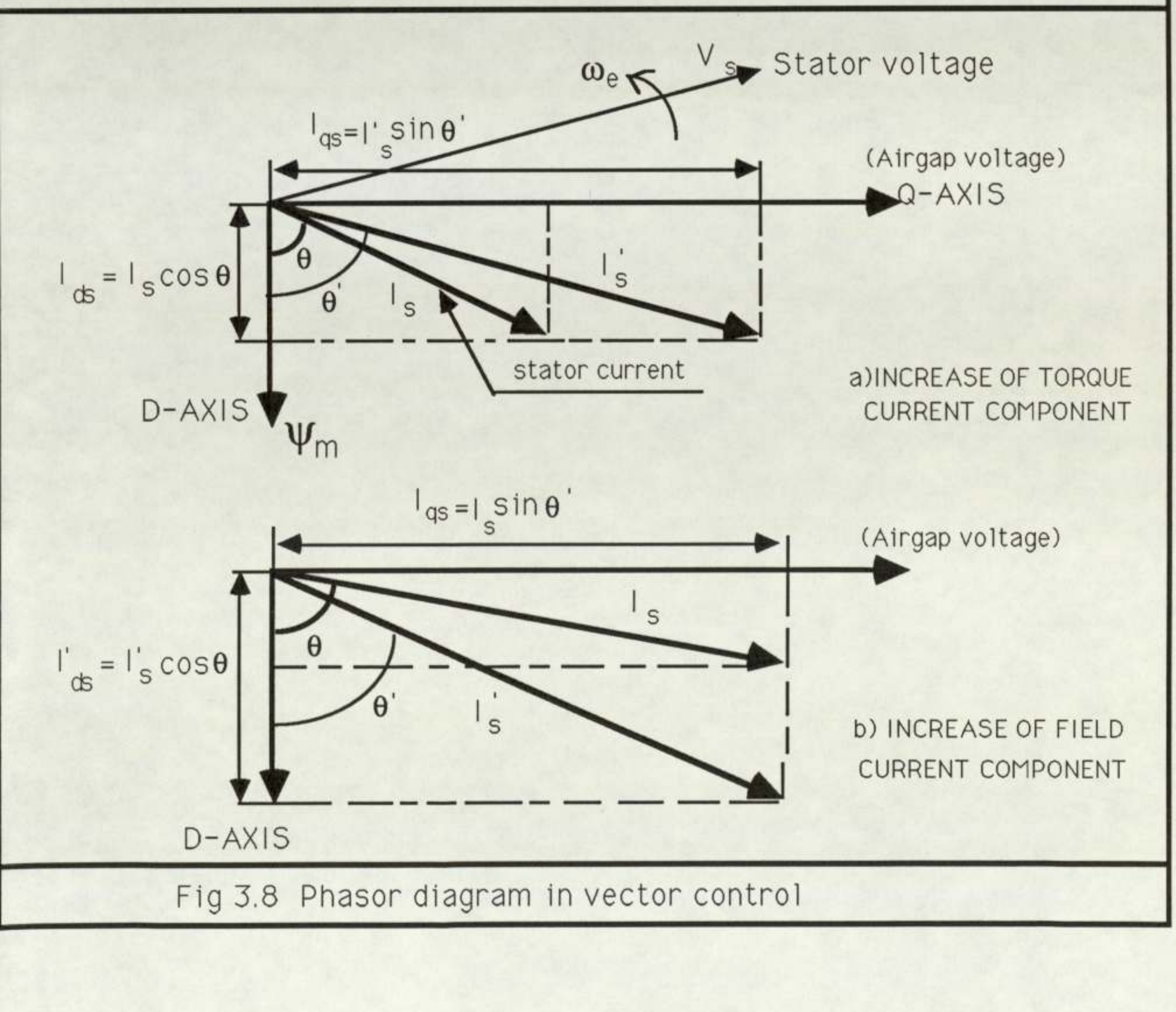


Fig 3.8 Phasor diagram in vector control

The equations for a balanced symmetrical three-phase induction motor can be written:

$$U_s = R_s I_s + d\Psi_s/dt \qquad 3.6.a$$

$$U_r = R_r I_r + d\Psi_r/dt \qquad 3.6.b$$

3.6.2.1 **Orientation with respect to stator flux**

From the equivalent circuit of the inductiom motor given in the previous section:

The stator flux is given by:

$$\Psi_s = L_s I_s + L_m I_r = L_m I_{ms} \qquad 3.6.c$$

where $I_{ms} = (1+\rho_s) I_s + I_r$ 3.6.d

and $\rho_s = [(Ls/Lm)-1]$

Projecting the stator current in phase and in quadrature to the stator flux:

$$I_s = I_{\phi s1} + j I_{\phi s2} \qquad 3.6.e$$

Substituting equations (3.6.d) and (3.6.e) in equations (3.6.a) to (3.6.c),and separating out the real and imaginary parts,

$$R_s I_{\phi s1} + L_m dI_{ms}/dt = U_{\phi s1} \qquad 3.6.f$$

$$R_s I_{\phi s2} + L_m \omega_s I_{ms} = U_{\phi s2} \qquad 3.6.g$$

$$I_{\phi s1} + \rho T_r dI_{\phi s1}/dt = (T_r/L_r)d\psi_s/dt + \psi_s/L_s + \omega_{sr}\rho T_r I\phi_{s2} \qquad 3.6.h$$

$$\omega_{s1} = (1+\rho T_r d/dt)L_{\phi s2}/[(T_r/L_s)\psi_s - \rho T_r L_{\phi s1}] \qquad 3.6.i$$

$$Te = 3/2[L_m(I_{ms}I_{\phi s2})] = (3/2)[\psi_s]X[I_s] \qquad 3.6.k$$

Equation(3.6.i)defines the build-up of the flux and equation (3.6.k)defines the electromagnetic torque produced.It is seen from equation (3.6.h),that the flux build-up not only depends on $I_{\phi s1}$,but also on $I_{\phi s2}$ and thus a coupling effect exists between the currents in the two axis.

3.6.2.2 **Orientation with respect to mutual flux.**

The mutual flux is defined by:

$$\Psi_m = L_m(I_s+I_r) = L_m I_m \qquad 3.6.m$$

where $I_m = I_s + I_r$

It can be shown in a similar way that the relationships between the stator voltage and currents,flux build-up and the torque are given by:

$$I_{\phi m1}+T_{s1}d(I_{\phi m1})/dt=$$

$$= U_{\phi m1}/R_s + \omega_s T_{s1} I_{\phi m2} - 1/Rs[d(\Psi_m)/dt] \qquad 3.6.n$$

$$I_{\phi m2}+T_{s1}d(I_{\phi m2})/dt=$$

$$=U_{\phi m2}/R_s - \omega_s T_{s1} I_{\phi m1} - (\omega_s/R_s)\Psi_m \qquad 3.6.p$$

$$I_{\phi m1}+T_{r1}d(I_{\phi m1})/dt=$$

$$=(T_r/L_m)d(\Psi_m)/dt + \Psi_m/L_m + \omega_{s1} T_{r1} I_{\phi m2} \qquad 3.6.q$$

$\omega_{sl}=(1+T_{r1}d/dt)I_{\phi m2}/(T_r/L_m)\Psi_m-T_{r1}I_{\phi m1}$ 3.6.r

$Te=(3/2)I_m I_{\phi m2}=(3/2)[\Psi_m][I_s]$ 3.6.s

From equation (3.6.n),it can be seen that the flux dynamics depend on both the current components $I_{\phi m1}$ and $I_{\phi m2}$.

3.6.2.3 **Orientation with respect to rotor flux.**

The rotor flux is defined by:

$\Psi_r=L_rI_r+L_mI_s=L_mI_{mr}$ 3.6.t

where $I_{mr}=(1+\rho_r)I_r+I_s$

and $\rho_r=[(L_r/L_m)-1]$

When the current is oriented toward the rotor flux,it can be shown that:

$I_{\phi r1}+\rho T_s d(I_{\phi r1})/dt=$

$=U_{\phi r1}/R_s+\omega_s\rho T_s I_{\phi r2}-(1-\rho)T_s d(I_{mr})/dt$ 3.6.u

$I_{\phi r2}+\rho T_s d(I_{\phi r2})/dt=$

$=U_{\phi r2}/R_s-\omega_s\rho T_s I_{\phi r1}-(1-\rho)\omega_s T_s I_{mr}$ 3.6.v

$T_r d(I_{mr})/dt=I_{\phi r1}-I_{mr}$ 3.6.w

$\omega_{sl}=I_{\phi r2}/T_r I_{mr}$ 3.6.z

From equation (3.6.w),it can be seen that the flux dynamics are very simple and the stator component responsible for the flux and the torque are completely decoupled.

In spite of the different structure configurations of the three compared methods described above,there are common points that should be noted.They are regarded as the most ideal from the view point of the dynamic characteristics (a quick response adjustable speed AC motor drive system is highly expected in industrial applications).But from the view point of practicability sensing of the actual flux vector poses a problem.

CHAPTER FOUR

4.0 Choice of semiconductor device

4.1 Introduction

The various inverter configurations reviewed in [chapter two,subsections(2.3.2/2.3.4)],are of the switching type of inverter.In this form of inverter the semiconductor components are used as switches.

This is the most efficient way of using semiconductor component,since their losses approach those of a perfect switch ie. zero.

In practice,semiconductors do dissipate some power when conducting,due to a finite voltage drop across them.In addition the finite time taken for a semiconductor to switch,means that there are times when the semiconductor device is carrying current whilst there is a significant voltage across it.

The dissipation during switching can be minimised by ensuring that the device switches as fast as possible,but the resulting rate of change of voltage and current can damage the device.Protection circuits,known as snubbers,can be added to

prevent this but some power is dissipated by the snubber and the power loss is proportional to the switching frequencies.Hence the efficiency of a switching mode inverter is,in practice,less than 100%.
A switching inverter is ideally suited to produce quasi-square waveforms at the fundamental frequency required by the load motor.When used in this manner,the square wave amplitude can be controlled by adjusting the dc-input to the inverter as mentioned in chapter 2.3.2.

A major point in favour of voltage source inverters is the ease with which the output voltage can be sensed and compared with the reference signal.The main disadvantage of a voltage source inverter is that it is not inherently short circuit proof and some current overload protection is necessary,the simplest form being fast acting semiconductor protection fuses.

In contrast a useful feature of a current source inverter is the fact that it is short circuit proof.However,this is offset to some extent by the need to incorporate current sensors to enable the control circuit to hold the output current at the desired value.A current sensor is more difficult to implement than a voltage sensor,partly because the current sensor has to be located in series with the load.

4.2 Choice of the inverter type

Having studied the various inverter option described before,it was decided that the simplest type was a switching inverter producing a voltage controlled,quasi-square waveforms at the frequency required by the load motor.The chain of decisions involved in this choice can be summarised as follows:

A switching inverter is most suited for the production of square waves since this is the basic action of a switch.Furthermore,if the switching frequency is kept to the minimum possible,the losses in the inverter are minimised and the driver circuits for the gating of the semiconductor switches can be simpler because of the low frequency operation.

The techniques (such as P.W.M) needed to synthesize sinewaves,require additional complexity in the signal portion of the inverter and higher output stage frequencies.Finally,it was decided that a voltage source inverter was preferable to a current source inverter in order to avoid the need for current sensors.

The consequences of the choice of a quasi-square wave voltage source inverter was the need for a means to control the magnitude of the voltage waveforms.Since the inverter cannot control the magnitude of the output voltage waveforms,it is necessary to supply the inverter from a variable voltage source which must be able to handle the total inverter power requirements.

4.3 Choice of the semiconductor device

Four semiconductor switches were considered for use in the inverter.They were:

a)Thyristor

b)Bipolar power transistor

c)M.O.S.F.E.T power transistor

d)G.T.O thyristor

All the devices are suitable for use in inverter applications,however the required power level and switching frequency will determine which is the most appropriate device to use.

The devices obviously must be rated for the voltage and current levels that they have to switch.They may also require what is known a snubber protection,to prevent secondary breakdown effect during switching.

Secondary breakdown is a condition where imperfections in the device structure cause uneven current distribution during switching,resulting in hot spots due to increased current density in those regions.The device can fail due to secondary breakdown.Snubber networks limit the rate of change of voltage and current in the device and prevent secondary breakdown.An excessive rate of change of voltage is especially troublesome,since it can cause the device to avalanche into conduction at a time when they should be off,and the resulting uneven current distribution leads to secondary breadown.

Snubber designs are well documented and fully reported in several papers [5.1],[5.2],[5.3],[5.4].Subsection 5.9 contains the design details for the snubber networks.

The manner in which thyristors and bipolar transistors work can be found in most electronic textbooks,for example [5.5],[5.6].However M.O.S.F.E.T's,and G.T.O's are relatively new devices and their ratings are constantly being improved.

The acronym M.O.S.F.E.T stands for Metal-Oxide-Semiconductor-Field-Effect-Transistor,while G.T.O stands for Gate-Turn-Off thyristor.Technical articles [5.7 to 5.21

inclusive],and manufacturers application notes are more helpful at present.

The main characteristics of each device have been summarised in the following subsections,so that the type of device chosen for the inverter can be justified.

4.3.1 Thyristor chartacteristics

Salient features of a thyristor include:

a) The turn-on time of a thyristor is relatively slow,of the order of 2μs to 10μs,similarly the turn-off time is very slow,being about 10μs to 20μs for a high speed thyristor.

b) A certain minimum time is required when commutating a thyristor,during which the blocking action re-establishes itself.

c) Conditions (a) and (b) place a maximum practical switching frequency limit of between 1kHZ to 10 kHZ on thyristor circuits.

d) The gate current required to switch large anode currents is relatively small,typically 100mA for a 30 A thyristor.

e) The device may require an inductive snubber in series with it to limit the rate of rise of anode current (dIa/dt) at switch-on.'Hot spots' within the thyristor structure can occur if the current is allowed to rise too fast before the thyristor is fully conducting.

f) The device can be triggered on by large rates of change of voltage across the anode-cathode terminals (dVac/dt).Therefore snubber circuits are needed to limit the maximum dVac/dt.

g) They need commutation circuitry to turn them off in inverter applications.

4.3.2 Bipolar transistor characteristics

The important features of power bipolar transistors can be summarised as follows:

a) The structure of PNP transistors makes them less suitable for power applications than NPN transistors.Therefore NPN transistors often have to be used in parts of circuits where ideally PNP transistors would be used.This can result in the need for complicated driver circuits for the NPN transistors.

b) The collector current is a function of the base current.The ratio of these two currents (the current gain β) is in the order of 5 to 100 for power transistors.current gain of up to 250 can be achieved using Darlinton power transistors but at the expense of slower switching speeds.

c) When the base current is sufficiently large to take it into saturation,the transistor is effectively a closed switch with a collector to emitter voltage drop of between 0.1V to 2.5 V depending on the transistor design.

d) An excess stored charge builds up inside the transistor if the base current is larger than the value required to just hold the transistor in saturation.The stored charge has to be removed before the collector current can fall to zero,and the design of the base drive must allow for this.

e) Current conduction in a bipolar transistor is essentially by means of the minority carriers.The recombination times associated with the minority carriers limit the switching times of the bipolar transistor.The turn-on and turn-off times of a power bipolar transistor are typically 1μs when the base is correctly driven,although much faster devices with switching times of the order of 300ns are avalaible.

f) The transistor can be turned-off at will by removing the base current.The switching times of 150 to 200ns are achievable(frequency range of about 5MHz)as mentioned by

Brian Kelly in [5.7].

g) The need for a continuous base drive current to maintain the transistor in an on-state,results in a significant base drive power requirement.

h) The rate of change of collector current (dIc/dt) may need to be limited at switch-on (as for the thyristor),in order to prevent hot spots and consequent damages.

i) To prevent avalanche breakdown of the collector-base junction,and also to limit power dissipation during swich-off,it can be necessary to use snubber circuits to limit the rate of change of voltage (dVce/dt) across the collector-emitter terminals.

j) Bipolar transistors can handle voltages of up to 1000volts and currents of up to 100amperes,but in general their power handling capability is less than thyristors.

4.3.3 M.O.S.F.E.T power transistor

The power M.O.S.F.E.T characteristics can be summarised as follows:

a) The control terminal by which the switch is controlled is called the gate.

b) When zero voltage is applied to the gate with respect to the source,the M.O.S.F.E.T presents a very small resistance between the drain and the source.

c) If the gate to source voltage is raised above a fixed threshold voltage V_{GS},the M.O.S.F.E.T turns-on and the drain and source are connected by a low value resistance R_{DS},typically 0.01Ω to 2.5 Ω,depending on the voltage rating of the M.O.S.F.E.T.

d) The gate current Ig drawn when the device is on is typically nanoamperes due to the high input impedance.During turn-on or turn-off,the gate current is large as it is essentially the charging current for the gate capacitance.

e) The ratio of the drain current to the gate current is extremely large (theoritically infinite),and the device can be considered as a voltage controlled switch.

f) The current conduction is by means of majority carriers.The switching times are therefore very fast (10 to 100ns),and are essentially only limited by the time taken to charge and discharge the gate capacitance.

g) The higher current ratings are avalaible at the lower voltage ratings ,e.g 60 V devices are rated to 30 A,whereas 1000 V devices are only rated up to about 4.7 A.The higher voltage rating devices are very expensive, and have a high turn-on resistance (e.g 2Ω for a 800 V device).

h) In certain applications some snubber protection may be necessary to limit the rate of change of voltage across the drain and source (dV_{DS}/dt).However manufacturers stated that M.O.S.F.E.T's can easily withstand 50 kV/μs.Therefore in most practical applications no snubbing circuit is needed because rates of change of voltage are rarely of these magnitudes.

k) The faster switching times and the ability to operate without snubbers,results in a reduction in switching losses compared to bipolar transistors.

4.3.4 Gate-Turn-Off-Thyristor-(G.T.O)

Briefly,the G.T.O thyristor characteristics can be summarised as follows:

The G.T.O thyristor has the advantages of the bipolar transistor

transistor(turn-off capability).Besides, it is expected that the G.T.O will be applied increasingly to inverters and because of this, there is a possibility that in the future the G.T.O will become less expensive,[5.19],and increase system efficiency by 2%-4% as reported by J.Karvinen in [5.21]

CHAPTER FIVE

5.0 Design and tests of the Chopper/Inverter power converter

5.1 Description of the drive system.

A block diagram of the G.T.O-V.S.I induction motor drive system which was designed and built is shown in FIG-(5.1).The induction motor is fed from a voltage DC link system,which consists of a chopper power converter connected to a three phase G.T.O inverter through a filter and a DC link capacitor.

The chopper power converter is fed from a fixed DC voltage supply,to produce the variable DC link voltage needed by the drive system.The inverter triggering pulses needed to control the output frequency of the drive system are produced by the controller which will be described later in subsection 5.6.

As can be seen from fig-5.1 the block diagram shows only a schematic view of the control signal for one phase (the principle is similar for the other remaining phases.

Both the inverter and the chopper power converter are separated from the low signal processing controller units by an optoisolator.

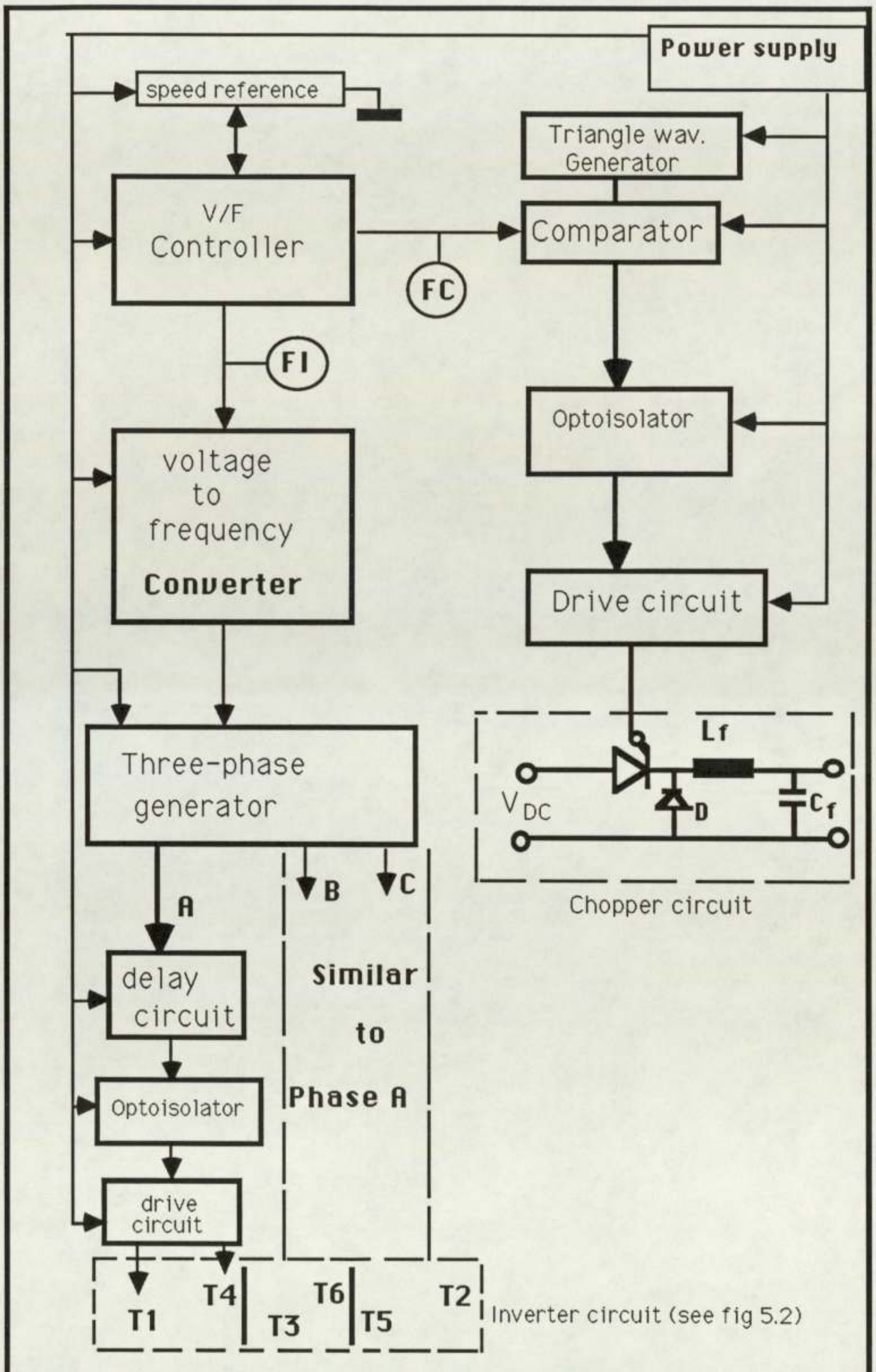


Figure 5.1 Bloc diagram of Chopper/Inverter drive

5.2 G.T.O-V.S.I design.

5.2.1 Inverter power circuit.

As stated earlier in subsection 5.1,the inverter produces a three-phase square waves output.The inverter is made up of six switch units,with pairs of switches connected as a half bridge to each phase winding.The inverter output circuit is arranged as shown in FIG(5.2),where the 'D'blocks represent the gate drive circuits,and the 'S' blocks represent the snubber-circuits.

The various parts of the inverter are described in the following subsections.

The device type:BTW58-1300R is employed and has an average current rating of 6.5 A,and further details concerning this device can be found in RS data sheet(4478-JUL-1983)

5.2.2 Gate drive circuit.

Each G.T.O thyristor in the inverter requires an individual control signal and so there are six drive circuits.It is well established that the G.T.O thyristor combines the high current,high blocking voltage capability of the conventional thyristor with fast and easy driving features of a transistor.

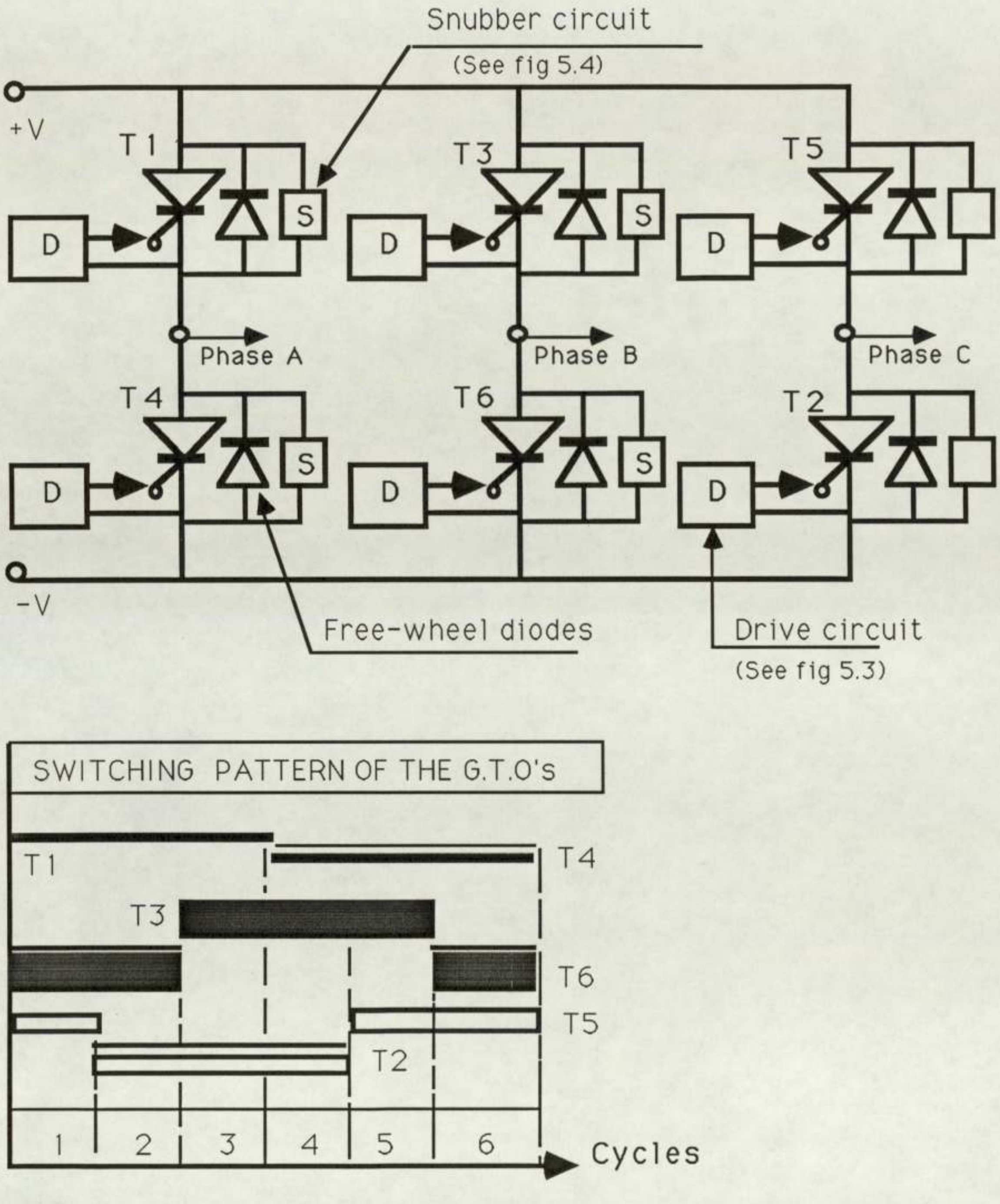


Figure 5.2 Basic circuit of the inverter power circuit

This implies that the device can operate with very short switching times for both turn-on and turn-off,less than 0.5 μs for the G.T.O type "BTW58-1300R"used in this application.

As given in the data sheet,the switch latches-on when a positive supply (of 1.5 V,200mA,for less than 10 μs is applied to its gate.

Turn-off is achieved by extracting from the gate a short pulse of between 20% and 100% of the anode current (for a few nanoseconds) and this can be done by applying a negative voltage of between (-5V and -15V).

To achieve this performance practically,it is essential that the device is driven correctly.The practical drive circuit built for this application is shown in fig(5.3) and a basic circuit used to check the drive circuit is depicted in fig(5.3a),finally gate current waveforms are shown in fig 5.3(b,c).

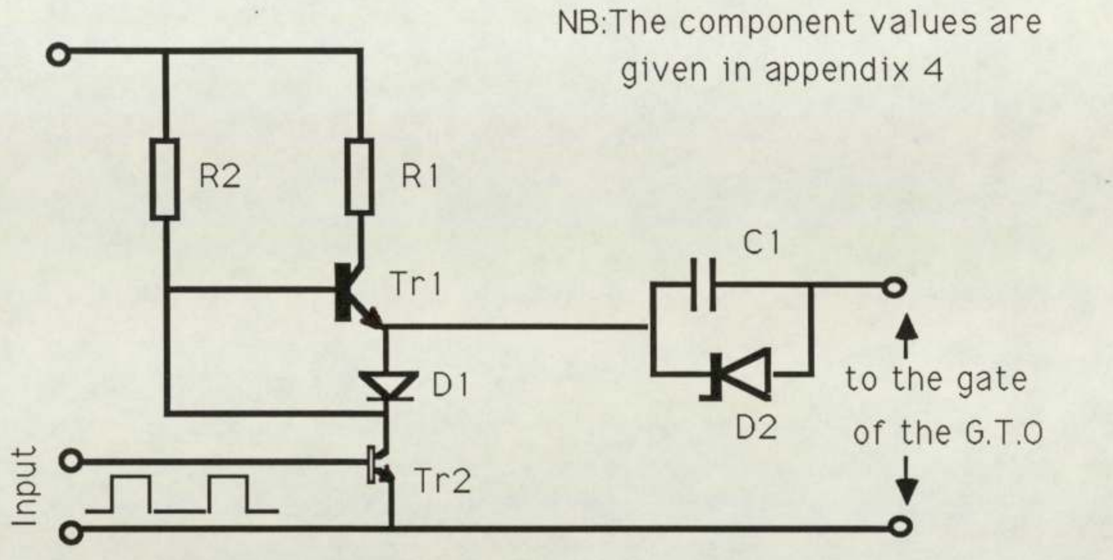


Figure 5.3 Gate drive circuit

The gate drive circuit operates as follows: with TR2 turned-off,positive gate drive current flows from the DC supply rail to the G.T.O via the emitter follower TR1,when the transistor TR2 is turned-on,TR1 turns-off and a negative voltage of about 10 volts,set by the diode zener,is applied to the gate of the G.T.O thyristor.As long as TR2 is on,the gate voltage will remain negative,because the reverse gate resistance of the G.T.O in the non conducting state is high and the capacitor C1 will therefore only discharge slowly.

Each drive circuit has its own isolated power supply to ensure that there is no interaction between the different switch units.

For this reason three transformers (one for each arm) with two independant outputs from each of them,were used to step down and to isolate the 240V-AC supply to 20V,which is then rectified ,filtered and regulated with the voltage regulator before feeding the gate drive circuit.

To prevent ground loops and other interferences between the control board and the main power circuit,optoisolators type "RS 307-979" were used to couple the control circuitry and the power board.These device are suited to TTL and analogue interfacing,and are often recommanded in control circuits.
A pulse transformer solution was rejected because the amplitude of the output voltage of a pulse transformer can suffer significant decreases for a given control signal mark/space ratio and repetition rate condition.The arrangement for a single switch unit is similar to the other remaining switches.

5.2.3 G.T.O- protection circuit

When a semiconductor device mentioned in section 5.3 is used in a switching mode inverter drive,it is generally necessary to provide some snubber protection to prevent damage occuring to the power semiconductor.
A thyristor or transistor dissipates very little power when is

an ON or OFF state,since in the former state the voltage drop across the device is close to zero.However ,during the time that a transistor or thyristor is switching from one state to the another it carries some current and has a potential difference across it.Hence significant power can be dissipated,especially if the device switches many times per second.

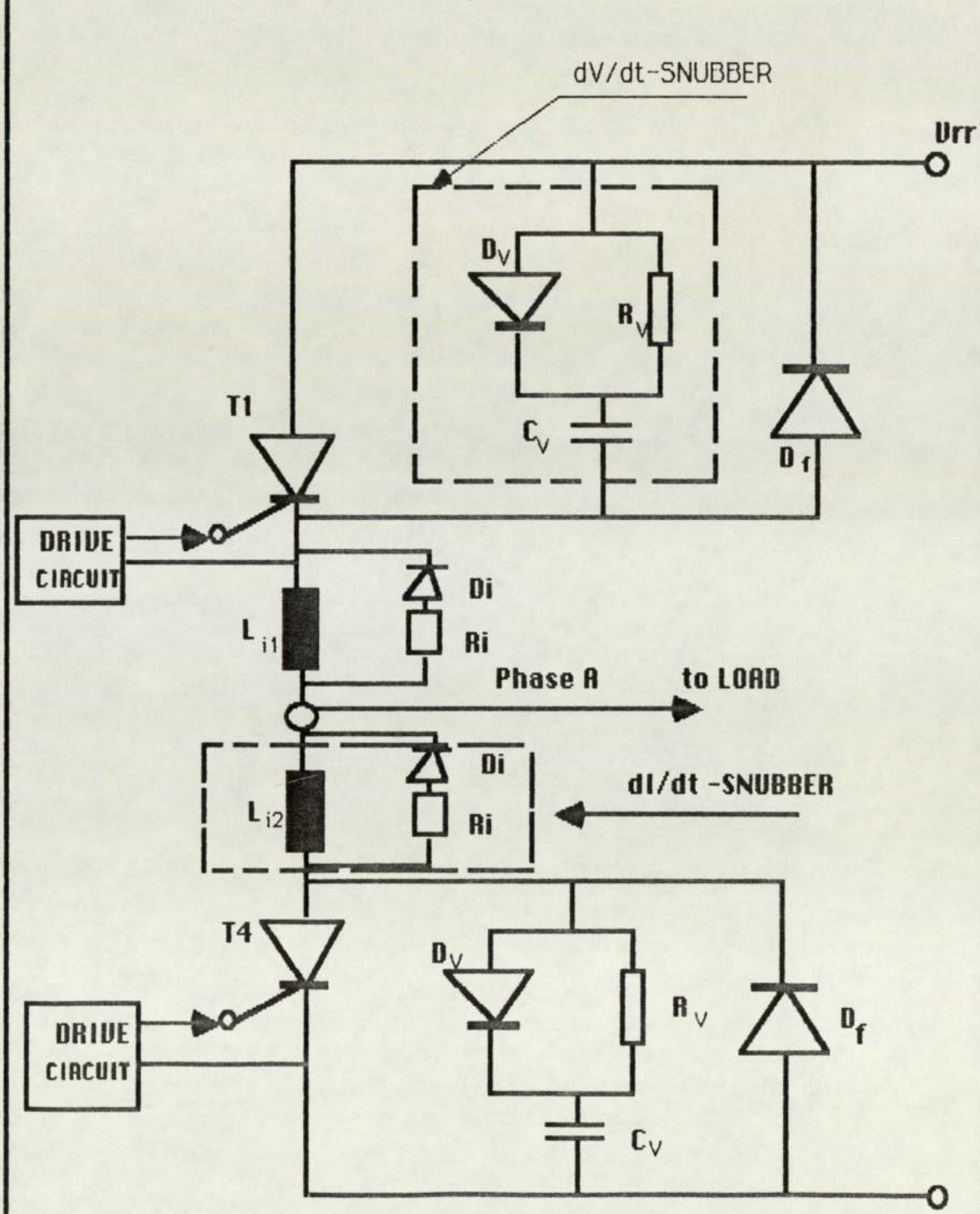
A1 - "dV/dt" snubber circuit : is connected in parallel with the power electronic device to limit the rise of voltage across it as it turns-off.

This reduces the power dissipation within the device.A dV/dt snubber also reduces the chances of avalanche breakdown within the device brought about by excessive rates of change of voltage.

A2 dI/dt snubber circuit: is connected in series with the power switch to limit the rate of rise of current in the thyristor as it turns-on.This reduces the power dissipation because the voltage across the device can fall to zero before a significant current begins to flow.In addition,a dI/dt snubber prevents "hot spot" in the semiconductor as the device starts to conduct.Hot spots can lead to device breakdown.

Much has been written on the application and design of snubber circuits [5.1],[5.4].The snubber networks connected to the power electronic switch are shown in fig(5.4)

The Values of the SNUBBER components are in section 6.4



FIG(5.4) INVERTER OUTPUT STAGE (SHOWING THE SNUBBERS)

The value of the snubber components are calculated in chapter 8 .It should be noted that the snubber component values are approximate.This is mainly because the G.T.O switching times used in this project are the typical times that can be achieved with good gate drive circuit.

5.3 Variable dc-supply (Chopper power converter).

This supply is required to provide steady regulated output voltage in order that the inverter voltage output matches the output frequency.

The phase controlled rectifier was rejected because it requires far more components to build it and mainly because a fixed DC supply was available on the spot.Instead a chopper power circuit to produce the DC variable voltage supply required by the drive system was built.A basic form of a chopper power circuit will be given later.with its control circuitry,necessary to keep the width of the pulse proportional to the reference dc signal generated by the V/Hz controller.

5.3.1 Mode of operation of the DC chopper regulator

The DC chopper regulator can be regarded as a switch having a high operating frequency.By periodic opening and closing of the switch,the load receives voltage pulses the mean value U_{out} of which can be controlled from zero to the input voltage U_{in} by means of the "ON" time T_{on} of the switch as expressed below.

$$U_{out} = (T_{on}/T)U_{in}$$

$$T = T_{on} + T_{off} = 1/F_c$$

where: T_{on} is time during which the G.T.O thyristor is conducting, T_{off} is the time during which the free-wheeling diode "D" is conducting.

F_c is the chopping frequency.

If the switch "S"(or the G.T.O thyristor is turned ON)is closed,the load current rises according to an exponential function governed by the time constant of the load circuit.

When the switch is opened,the load current can continue to flow via the free-wheeling diode "D",due to energy stored by the inductance of the load.The current then decreases according to an exponential function until the switch "S" is reclosed.

The result is a pulsating DC current,the ripple content of which at a constant switching frequency depends on the time constant of the load circuit and the "ON" time T_{on}.

By varying the firing times of the G.T.O thyristor,it is possible to vary the width of the voltage pulse applied to the load as, and thus control the mean load current.

When the DC chopper regulator is off,the problem arises of drawing pulsating current from the source.Some means must be provided to permit the system current to continue to flow since sudden interruptions of the system current will produce high current peaks due to the source inductance,which may lead to failure of the G.T.O thyristor in the DC chopper regulator.The most suitable storage element for the system current when switched off by the DC chopper regulator is a capacitor C_{in} which is connected across the input supply terminals.The source inductance together with the capacitor C_{in} forms an oscillating circuit the resonance frequency of which is;

$$Fo = 1/[2\pi(L_s.C_{in})^{1/2}]$$

Where, L_s is the inductance of the source,and C_{in} the filter capacitor connected across the supply terminals.

Care should be taken that this frequency does not lie within the range of the superimposed harmonic frequency emanating from the operating frequency of the DC chopper,otherwise an undesirable resonance condition may occur.

The closer the resonant frequency of the system to the switching frequency of the DC chopper regulator,the higher will be the amplitude of the oscillatory currents and voltages.At resonance,i.e when the two frequencies are the same the oscillation amplitudes are limited only by the ohmic resistance of the circuit.The resonance frequency in the system can be avoided by selecting a switching frequency well above the resonant frequency of the input circuit.

A basic operation of a chopper power converter is illustrated in figure 5.5 with the output voltage and the load current in steady state represented respectively in figure 5.5(b) and 5.5(c)

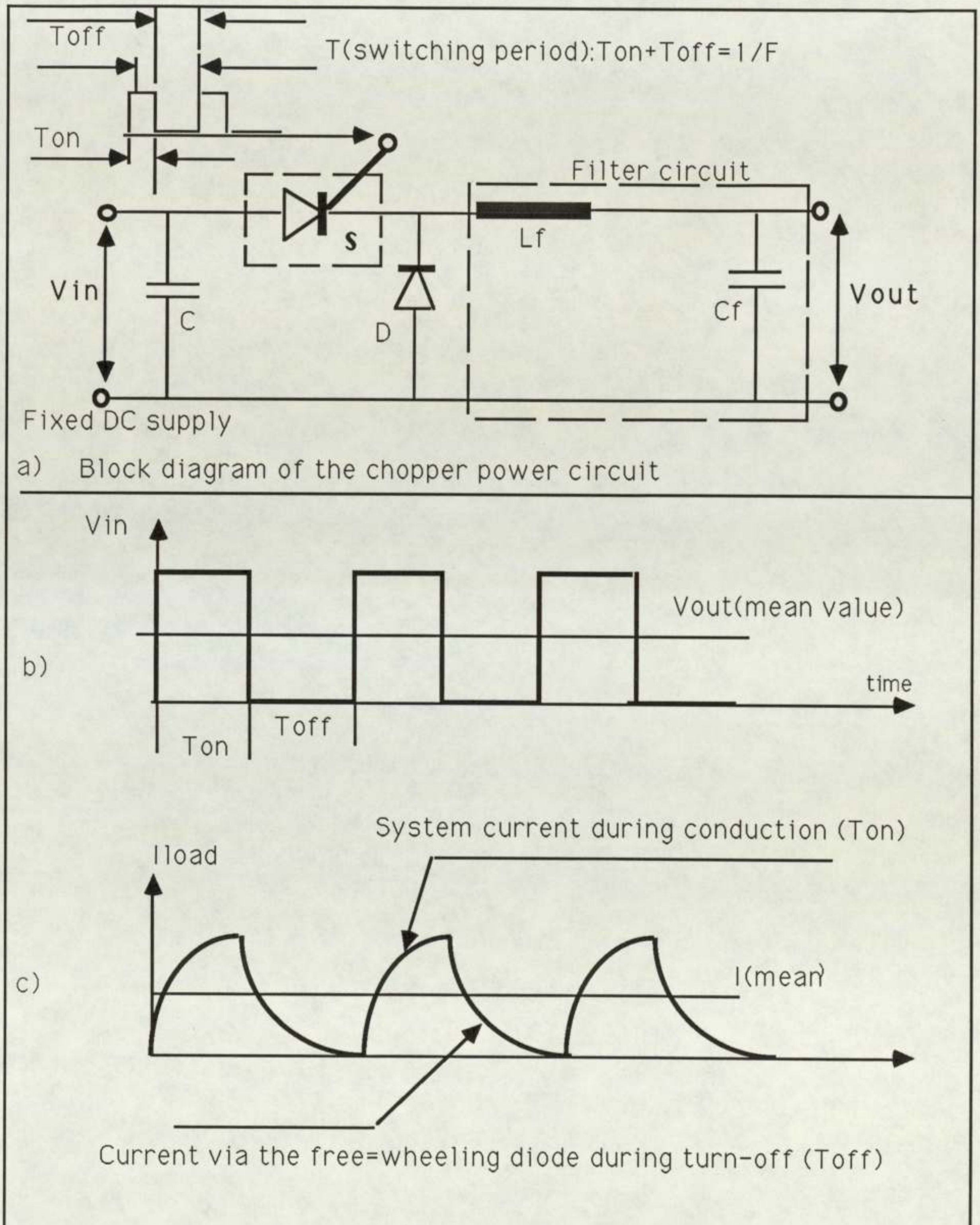


Figure 5.5 Basic operation of the chopper power circuit

To minimise the size of the filter inductance the chopping frequency should be very high.

A simulation program was written to evaluate the effect of the chopping frequency " F " and the values of the filter inductance on the current output of the chopper.The results obtained by simulation are presented in section 5.7.3 and compared with the practical tests carried out on the chopper.

As can be seen,the simulation and the practical results are quite close.In this case it was found that a frequency of 2kHZ,and a filter inductance "L_f" of less than 100mH were quite satisfactory with regard to the output current ripple and fluctuation of the voltage output of the chopper.

5.3.2 **Controller design feature.**

The drive performance as mentioned in chapter three depends on the strategy used to control the speed/torque of the induction motor fed by a variable frequency,variable voltage static converter.In this project a simple "V/HZ" controller is to be considered.This technique is valid and leads only to moderate machine underexcitation,when relatively high frequencies of operation are of interest.At the low end of the frequency range,however,the neglected stator resistive drop becomes important compared to the stator supply voltage. Ignoring this fact leads to severe underexcitation and intolerable loss of torque capability.

To correct this deficiency at low speed,a boost of voltage is adopted.This technique which is simple to implement enables the motor to achieve high torque at low speed.

5.3.3 Description of the control system

The voltage reference signal U_{ref} is fed in to the voltage-frequency converter via a potentiometer,this reference voltage sets the reference speed at which the motor will run.

The pulses generated by the voltage to frequency converter are converted to a suitable level for the gating of the G.T.O thyristors,as will be explained later.Each frequency correlates to a given stator voltage,as a result the terminal voltage of the motor is regulated in a linear relationship with respect to the stator frequency.The relationship between the the voltage and frequency is determined by the ratio (volts/frequency) obtained from the V/Hz controller..

The controller is used :

a) to maintain a linear ratio between the stator voltage and the frequency in order to keep a constant flux,as explained in chapter three.

b) to provide the inverter triggering pulses and

c) to control the output voltage of the chopper (by varying the mark/space ratio) to match the frequency needed by the drive.

Briefly the controller operates as follows:

a) Firstly a speed reference signal is initiated from a potentiometer which is processed in the "V/Hz" controller.

From the output of the "V/HZ"controller two signals are generated,one is used to control the chopper (FC),while the second signal (FI) is used to control the switching frequency of the inverter,as depicted in fig 5.6

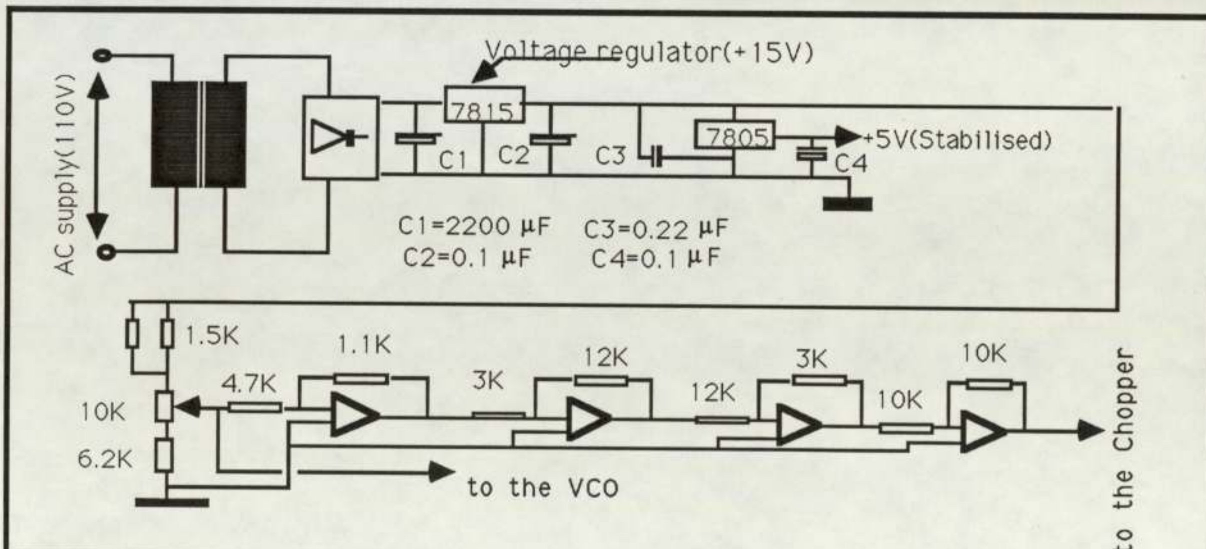


Figure 5.6 Circuit diagram of V/Hz ratio controller

b) The former signal[a variable DC signal (FC)]used to control the voltage output of the chopper is compared with a triangle waveform signal in a comparator.The comparator's output consists of pulses whose duty cycle varies from 10% to 100%

as the comparator's input voltage (FC) varies from the low value to the triangle voltage's peak value,as shown in fig 5.7.

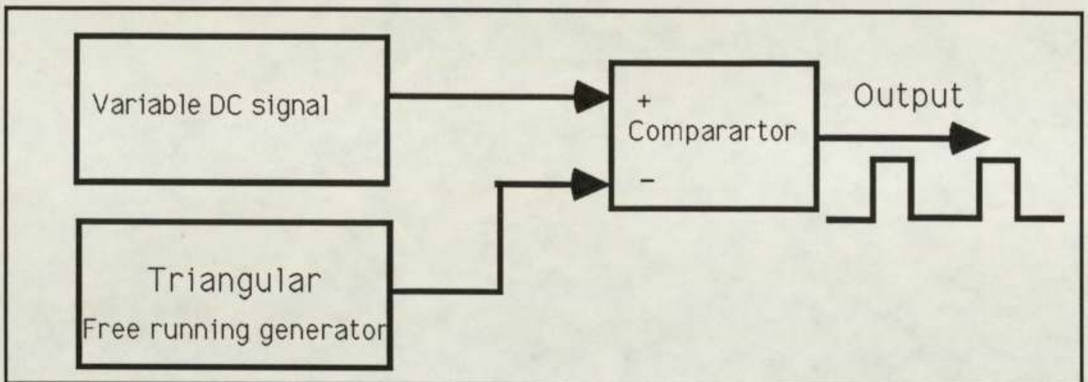


Figure 5.7 Circuit principle of the P.W controller

A detailed control circuit is given fig 5.7(a)

c) The pulses generated by the control circuit given in fig 5.7(a) are transmitted then through an optoisolator after which they are amplified in the driver circuit and finally used to trigger the GTO-thyristor of the chopper converter circuit ON and OFF.

The width of the pulses varies proportionally with the dc control signal applied at the input of the comparator.At the output of the comparator the pulse's width is varied from minimum up to the maximum width as shown in the scope waveforms obtained experimentally,fig 5.7(b)[G.T.O not connected].

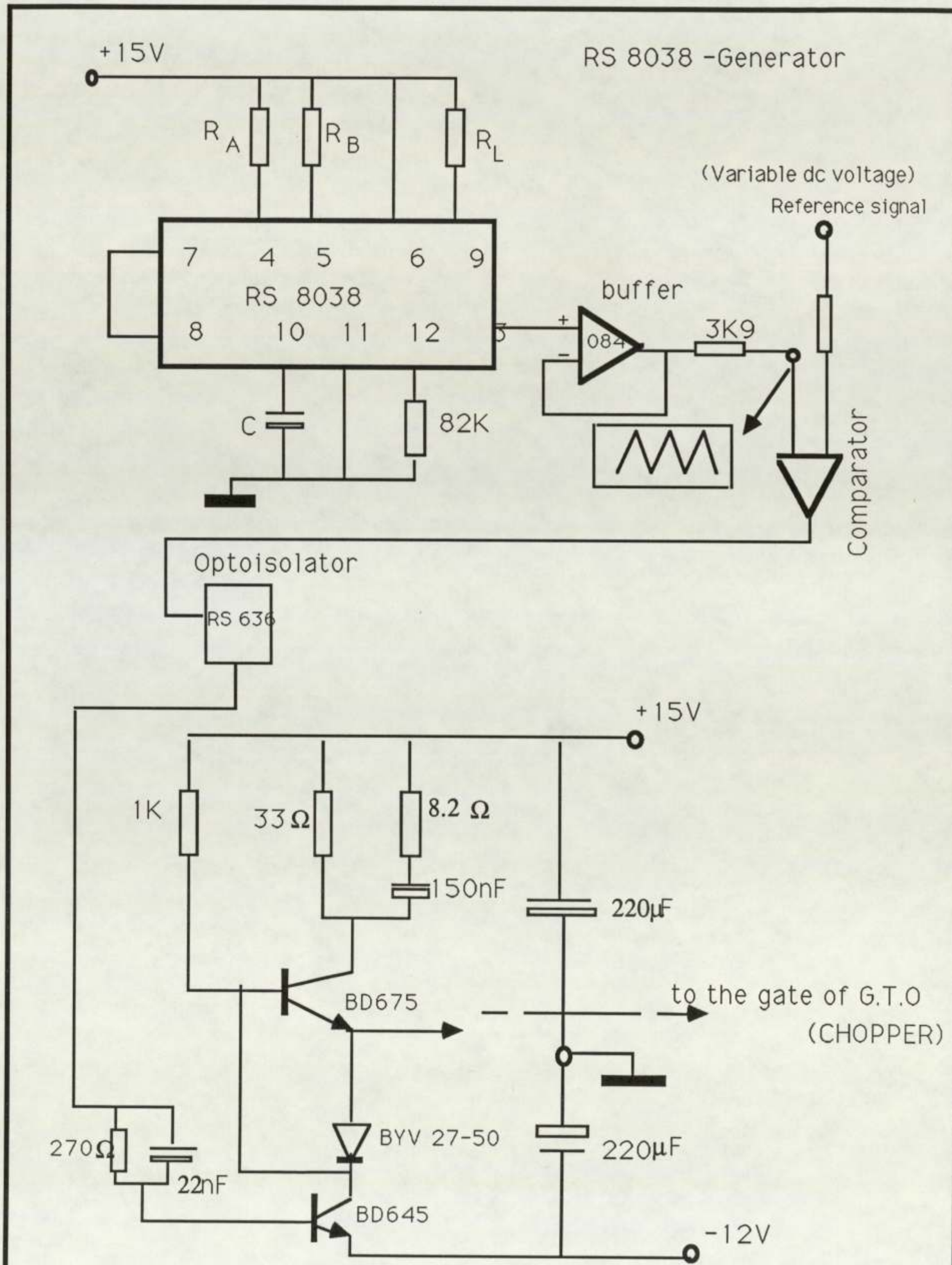
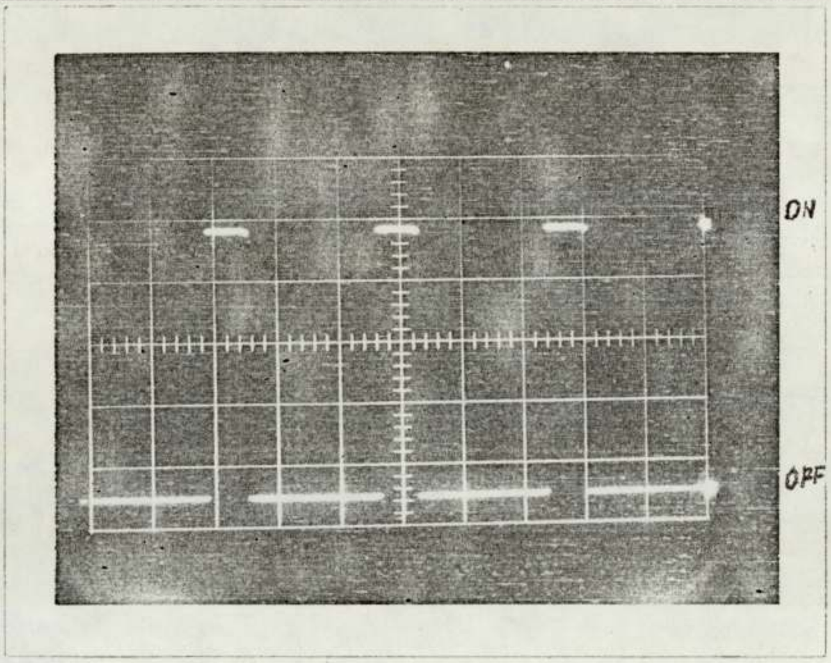
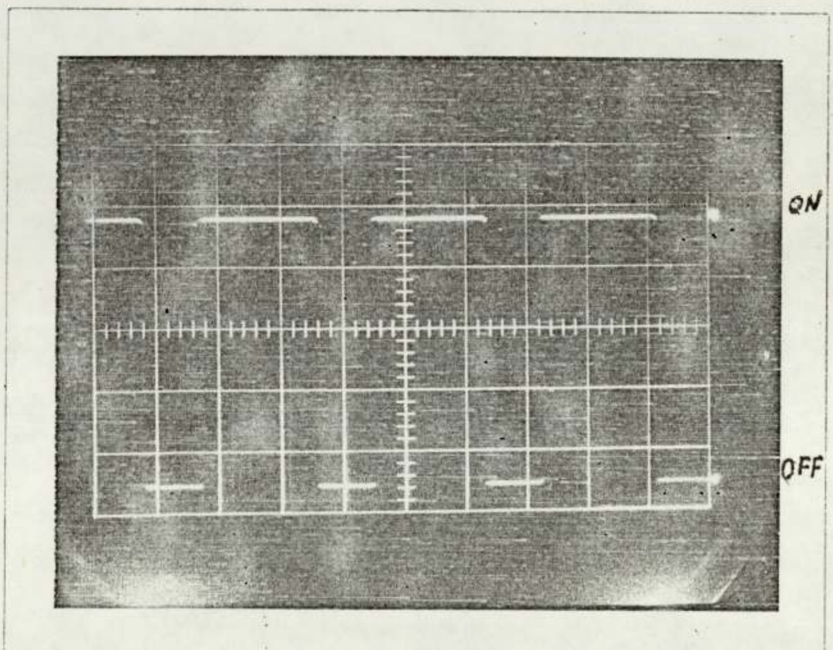


Figure 5.7a. Circuit diagram for the control of the Chopper



T=20%Ton



T=80%Ton

Figure 5.7(b) Voltage output from the drive circuit

d) The latter signal [a variable DC signal (FI)] from the "V/Hz" controller is converted into frequency pulses,by a "Voltage to frequency converter".

e) The frequency pulses generated by the V/F converter are transformed into a three-phase frequency signal,A,B,C.as shown in fig 5.8(a),fig 58(b)..

f) the phase signals A,B and C generated by the three-phase generator are duplicated into two frequency signals, respectively: (A1,A4),(B3,B6) and (C5,C2) with opposite sign,in a delay circuit composed of:

an inverter

one dual monostable,

one NAND gate and

one AND gate.

The rising and falling edge of the signals (A1,A4) respectively are delayed by a few microseconds to avoid the instantaneous conduction of the G.T.O's connected in series.

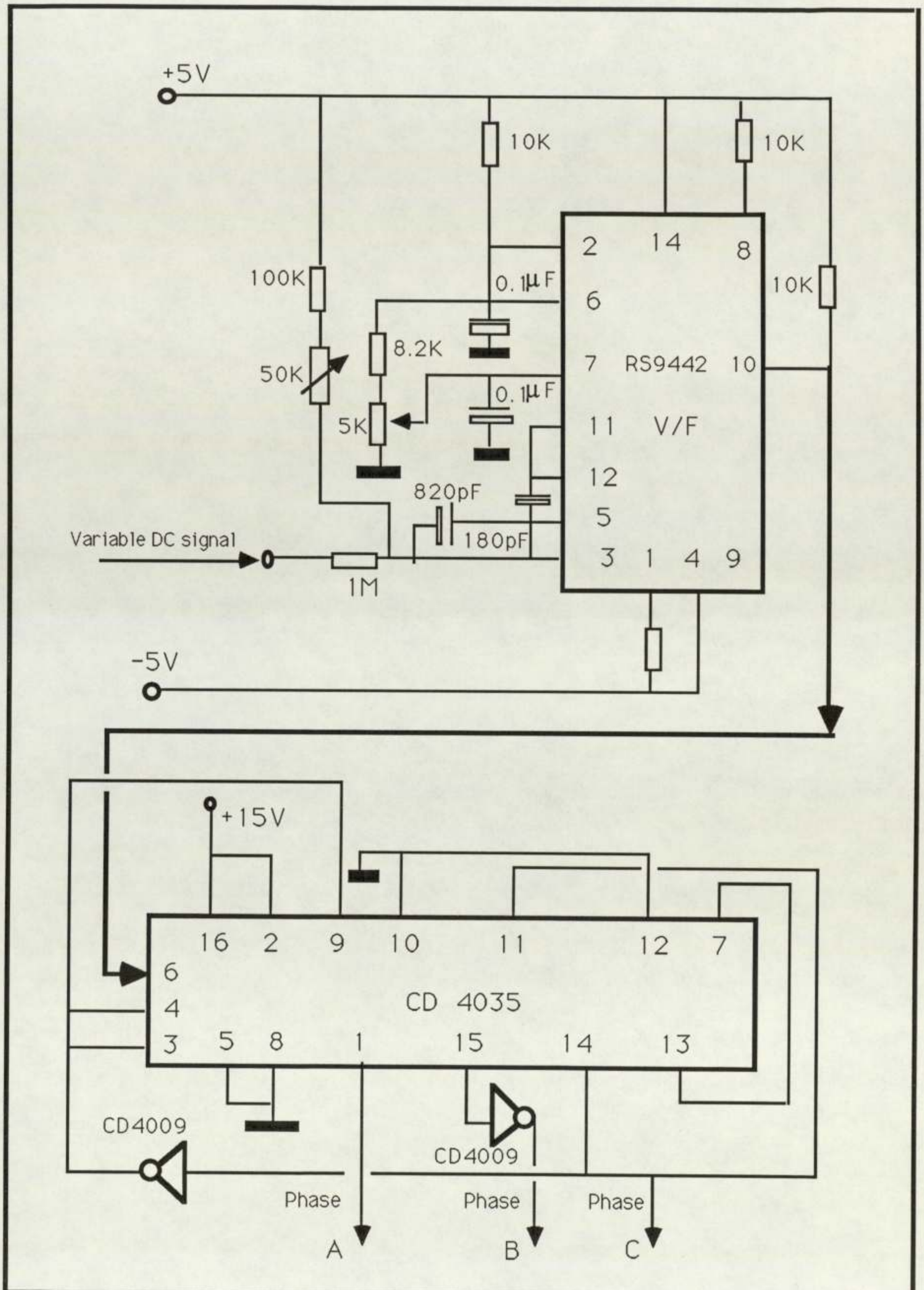
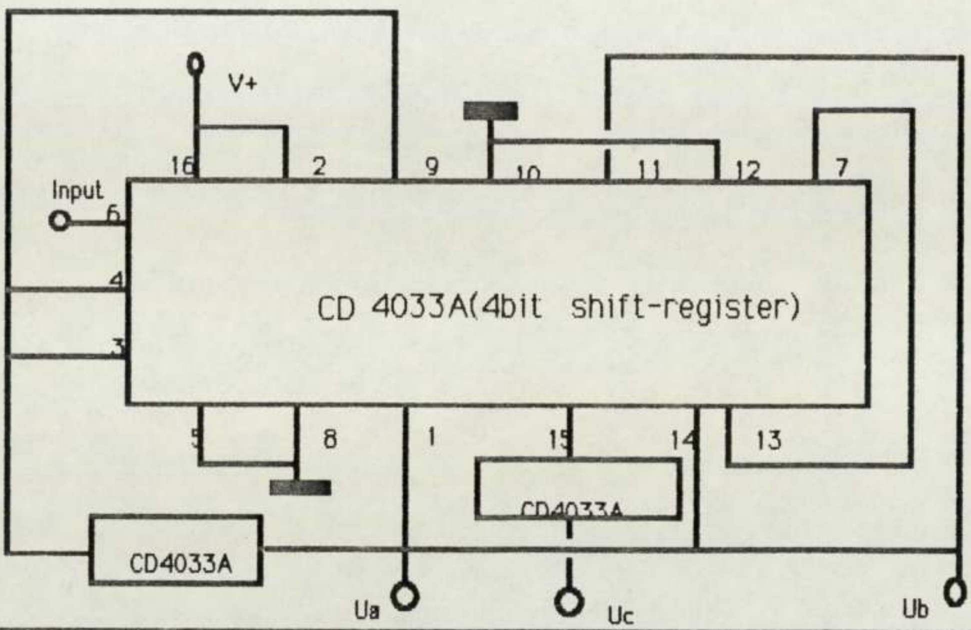
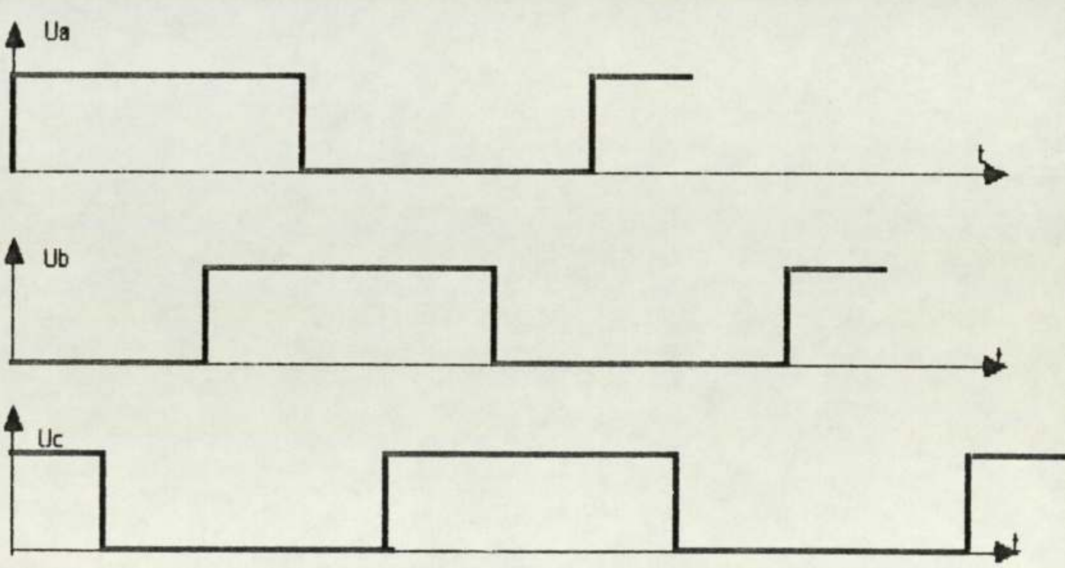


Figure 5.8 (a) Circuit diagram of the dc to three phase converter



a) CIRCUIT-PRINCIPLE-OF-THE(3-PHASE)-GENERATOR

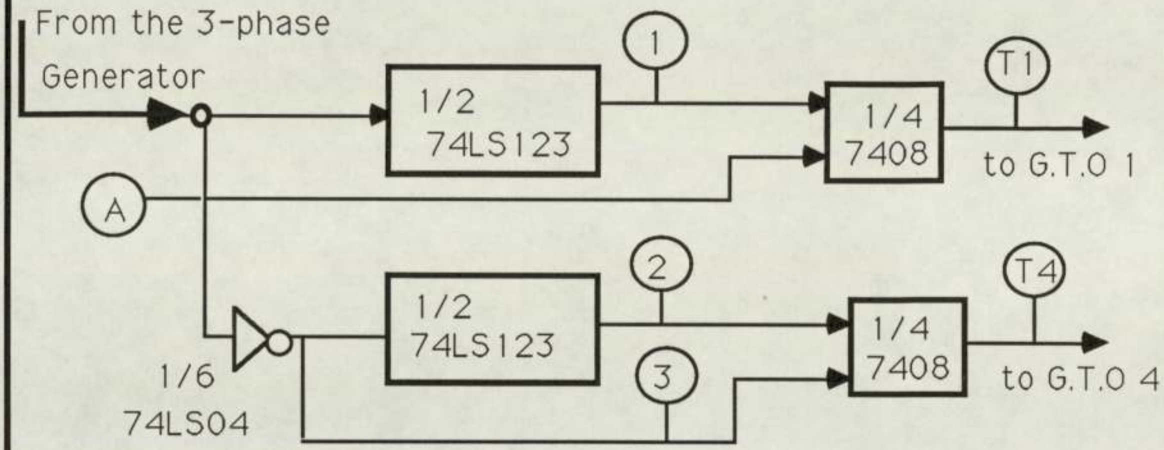


b) OUTPUT-VOLTAGE-WAVEFORMS

FIG 5.8.b THREE-PHASE -PULSE GENERATOR

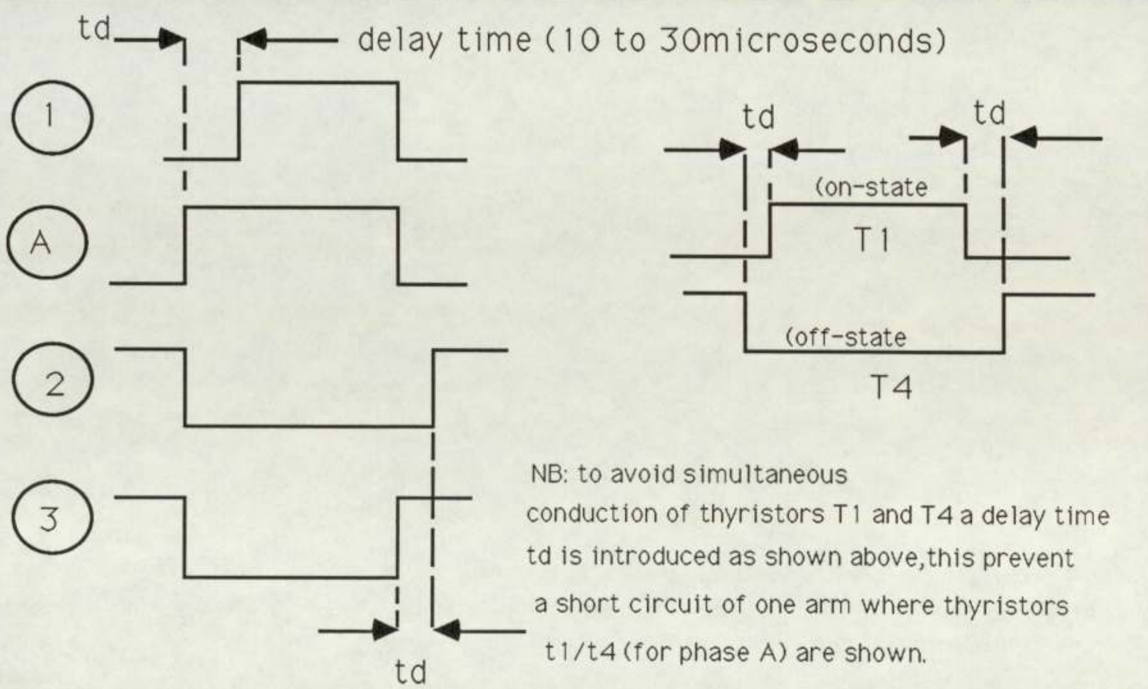
The delay circuit diagram is shown in fig 5.9(a) and the signal outputs from the delay circuit (or protection circuit) are shown in fig- 5.9(b).

Finally in a similar way as in (point c) the G.T.O's of the inverter are fired ON and OFF.



74LS123-monostable-7408:AND element-7404:NAND element

a- Circuit diagram of the delay network



b- signal waveform at different stage of delay

FIG.5.9 Protection network

5.5 Experimental tests.

This section describes the experimental tests carried out on a NECO induction motor drive,in which a three phase quasi-square wave voltage source inverter,a chopper power converter,and their respective control circuitry were designed and built to drive a three phase induction machine with the following parameters:

The motor name plate rating are:

NECO 3-PHASE MOTOR.

240 volts ,50HZ ,1425 rev/min ,250 W

stator : line voltage: 240 volt (rms)

----- full load current : 1.25 A (rms).

Prior to static and dynamic performance tests of the drive system,various individual test were carried out separately to check and to set-up each part of the system.

5.5.1 Testing of the switching unit

A single +15 V power supply ,and a driver circuit were built to check the circuit operation.

The power supply was found to give an output of 14.87 V on no load.The output voltage fell to 14.6 V when loaded with a variable resistance and a current of 800mA was drawn.The power supply was left running with a load current of 500 mA for 30 minutes with no problems,the voltage regulator type"RS-7515" was only slightly warm by the end of the test.

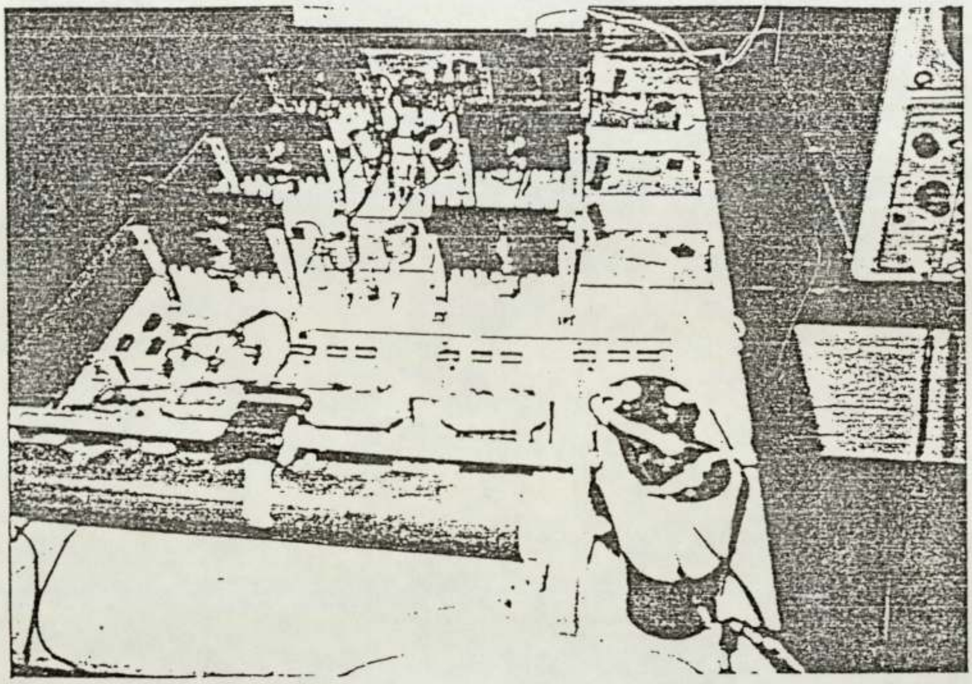
The driver circuit was then connected to the power supply.The power supply output voltage was 14.8 V,and the current drawn was 18.8 mA.

A G.T.O-thyristor was then connected to the drive gate circuit and was controlled with a range of frequencies from DC to 2 kHZ.It was found that the power supply voltage stayed constant at 14.8 V.

To check the G.T.O switching,a 500 ohms variable resistor was connected to the terminals device.As the resistor load was slightly inductive,a freewheel diode was connected across the load resistor to provide a freewheel path.The circuit was supplied from 0-150 V variable DC power supply.

5.5.2 **Inverter-Chopper assembly and initial testing.**

After all the switch units,power supplies,and control circuits were tested they were then assembled.A front view of the finished inverter and chopper are shown in the photographs with their respective layout in fig 5.10(a,b).



FRONT VIEW OF THE G.T.O's INVERTER

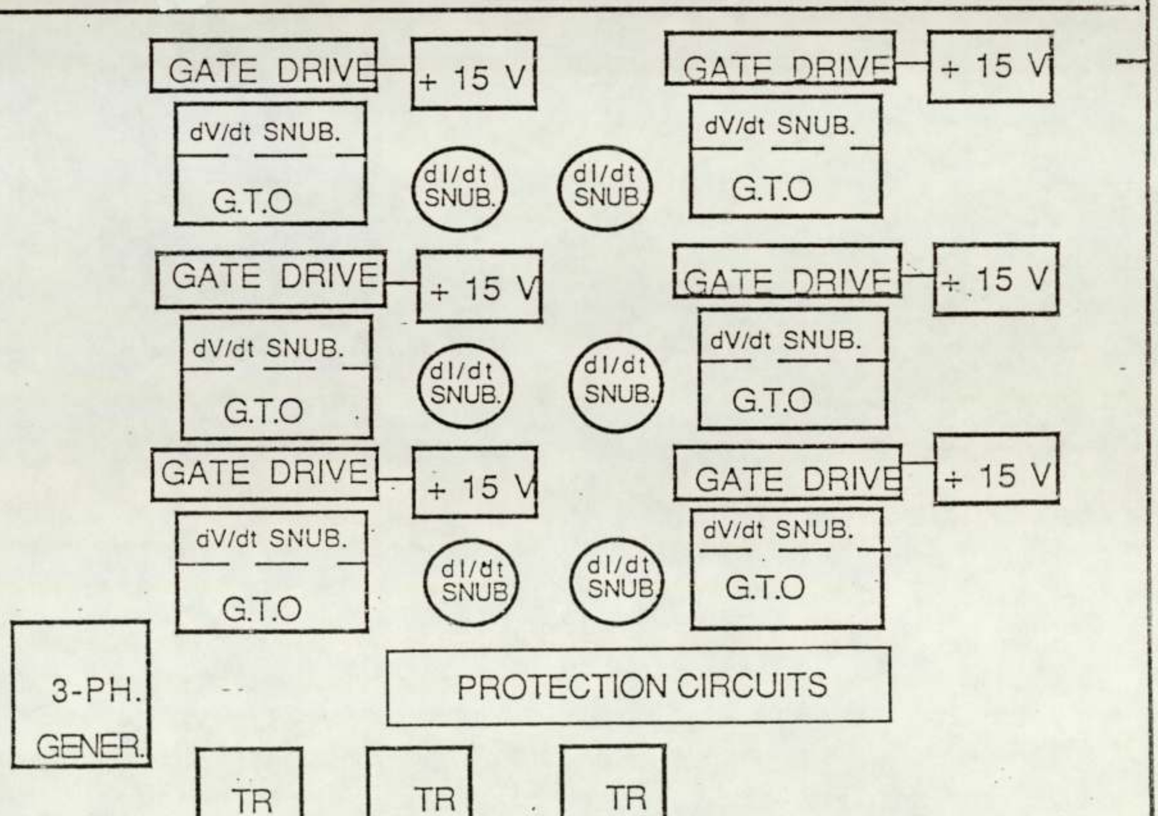
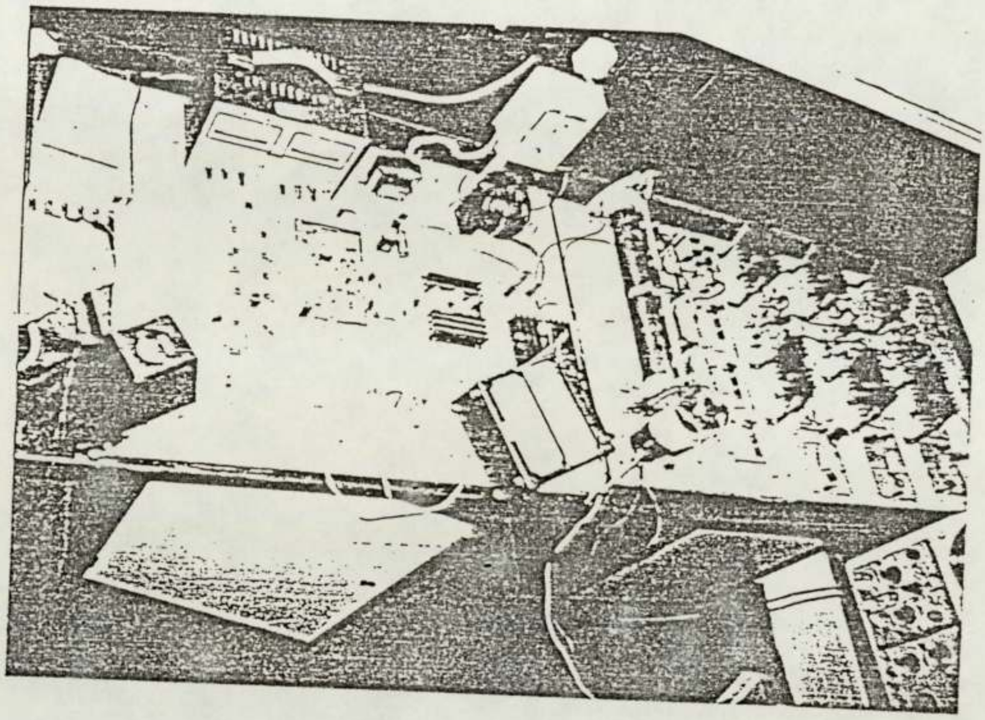


Fig 5.10(a) LAYOUT OF THE FINISHED INVERTER CIRCUIT



FRONT VIEW OF THE CHOPPER-INVERTER DRIVE

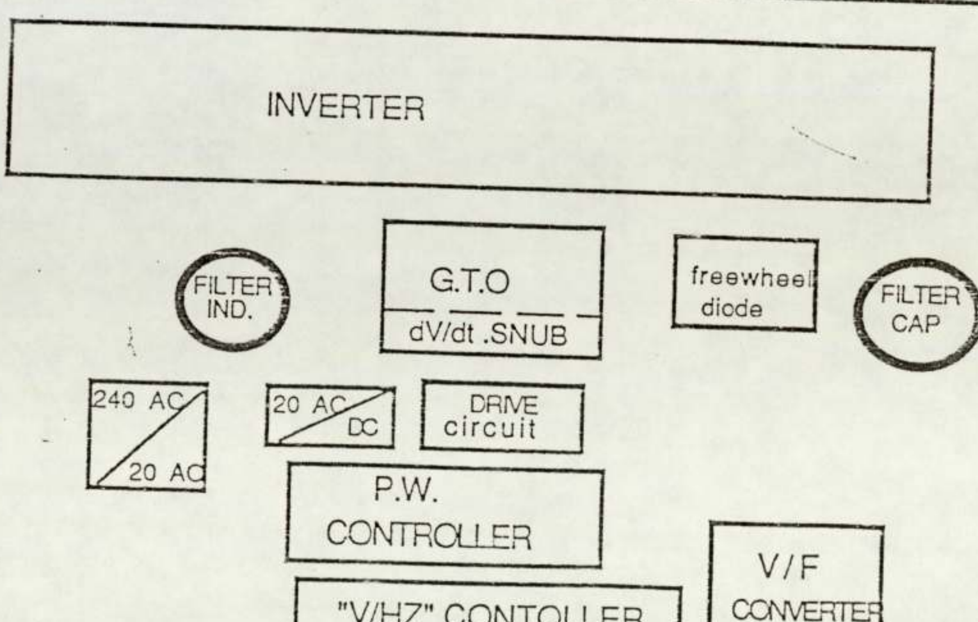


Fig 5.10(b) LAYOUT OF THE CHOPPER/ INVERTER DRIVE CIRCUIT

The chopper power converter was tested separately.For this test a variable resistance (0-200Ω) was used as a load.The test results are discussed in section 5.7.3

The inverter was then connected to the chopper,and tested with a variety of supply voltages and frequencies from 10Hz to 65Hz.,for more than twenty minutes,and no interaction between phases could be detected.The test results are summarised in section 5.7.3

5.7.3 TEST RESULTS

This subsection summarises the results obtained from the experimental tests carried out on the chopper ,the inverter and finally the complete drive system.

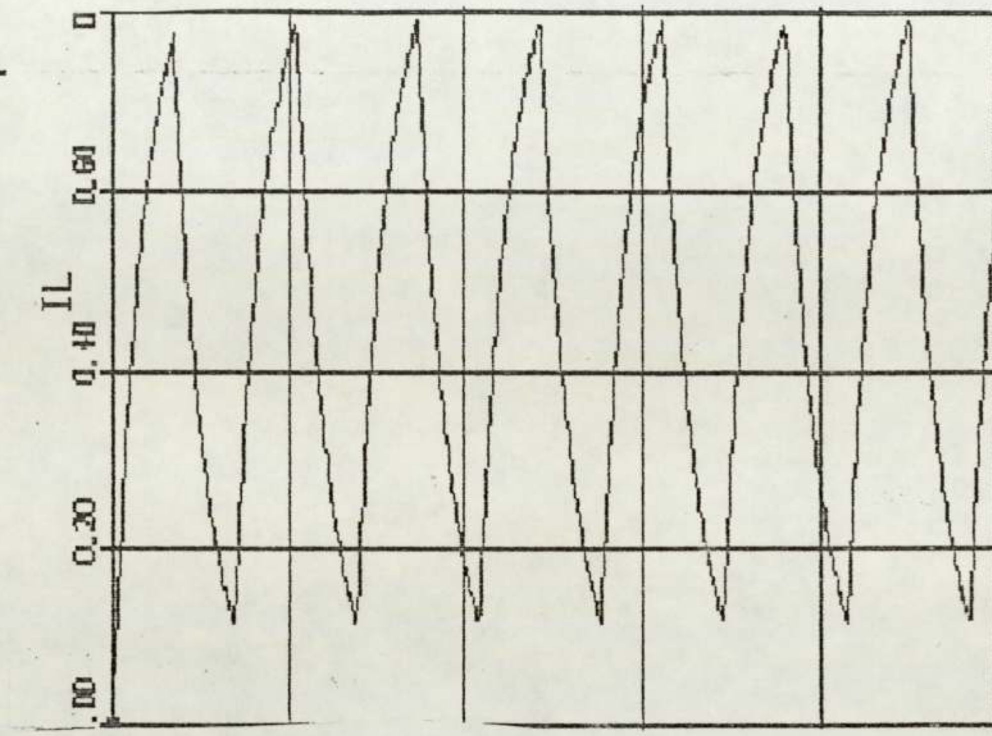
A)-First the chopper was tested for various load and pulse width control signals

The switching frequency has been set to 2kHz for a given filter inductance (100mH)to reduce the ripple current on the basis of the simulation program results.

Simulation tests were carried out for various values of the filter inductance and the switching frequencies to show the effect of these parameters on the ripple load current as depicted in fig 5.12(a-d),note they are in close agreement with the experimental results.It is worth noting that the load current is discontinuous in the case where the switching frequency and the filter inductance values are low as can be seen in fig 5.12(a).

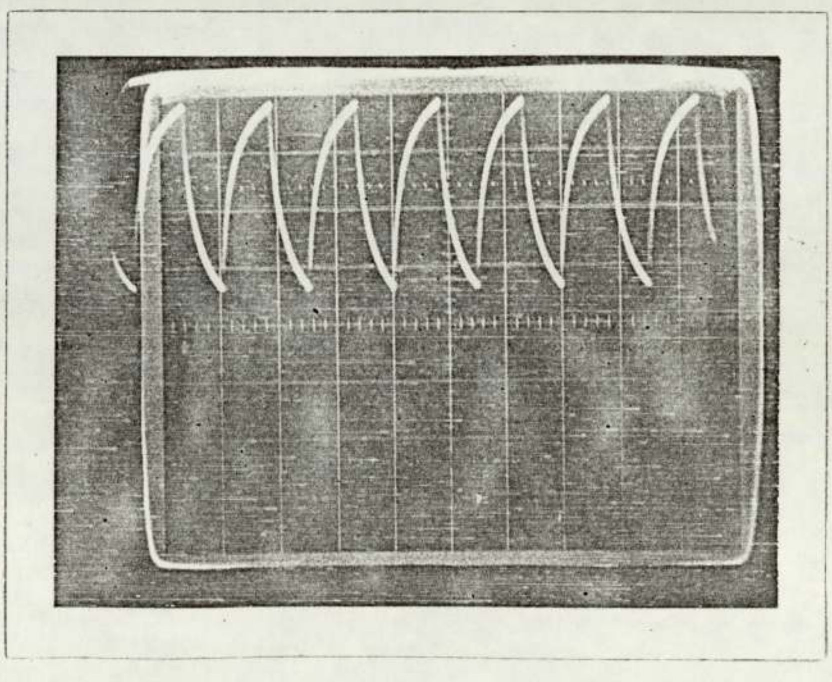
When the frequency is increased the load current is no longer discontinuous but presents a high ripple level,[,see fig-5.12(b,c)].This high ripple level may be reduced by connecting a large inductance in series, however this may lead to a bulky choke and increased losses in the inductor.
A second method of reducing the ripple level is to increase the switching frequency to the maximum switching capability of the G.T.O.This second approach have been adopted,thus when the frequency is further increased the ripple is dramatically reduced 5.12(d).Thus this simulation exercise has shown the advantages of modelling the real system before any real system is built.

B)-The inverter unit has been tested with a three-phase resistive load for various supply voltages and frequencies above the nominal drive frequency,i.e 65 Hz.
.The results obtained from the practical work shows that the line voltages (V_{AB},V_{BC},V_{CA})are in agreement with those predicted.



Simulation result

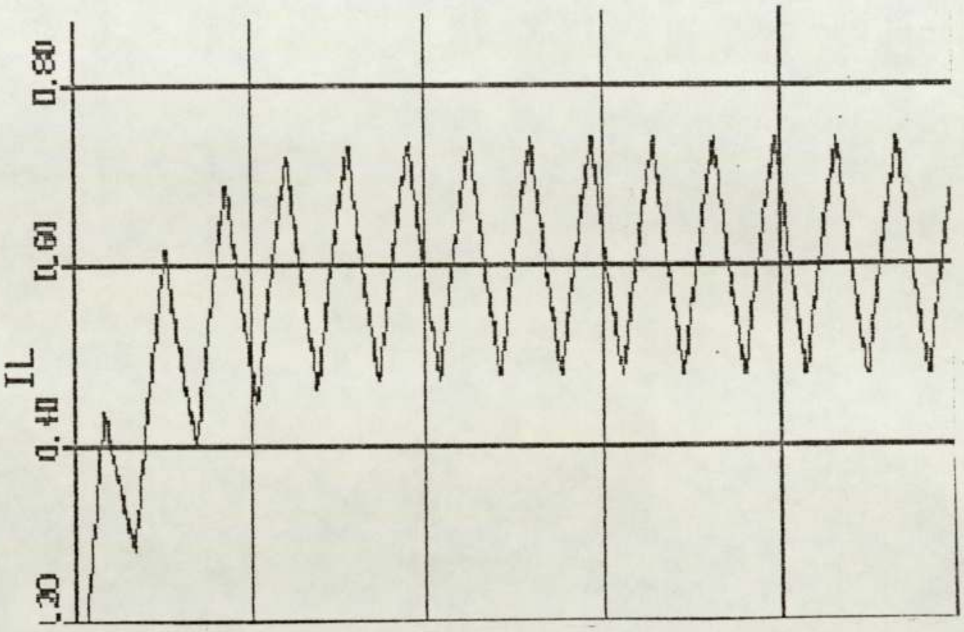
Amp/div=2mA Time/div=1ms Ton =25% T(period)=0.6ms



Practical result

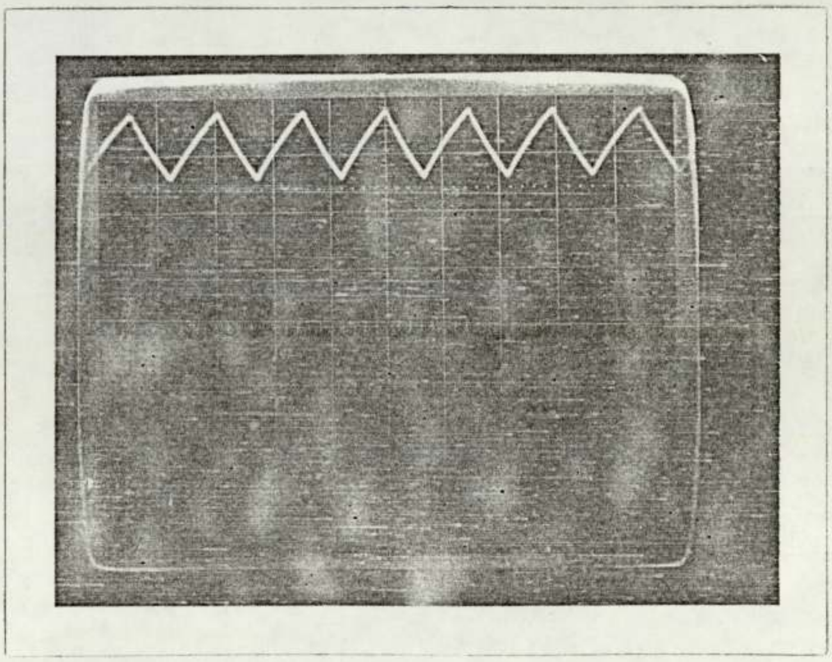
Amp/div=2mA Time/div=1ms Ton =25% T(period)=0.6ms

Fig 5.12(a) CHOPPER LOAD CURRENT



Simulation result

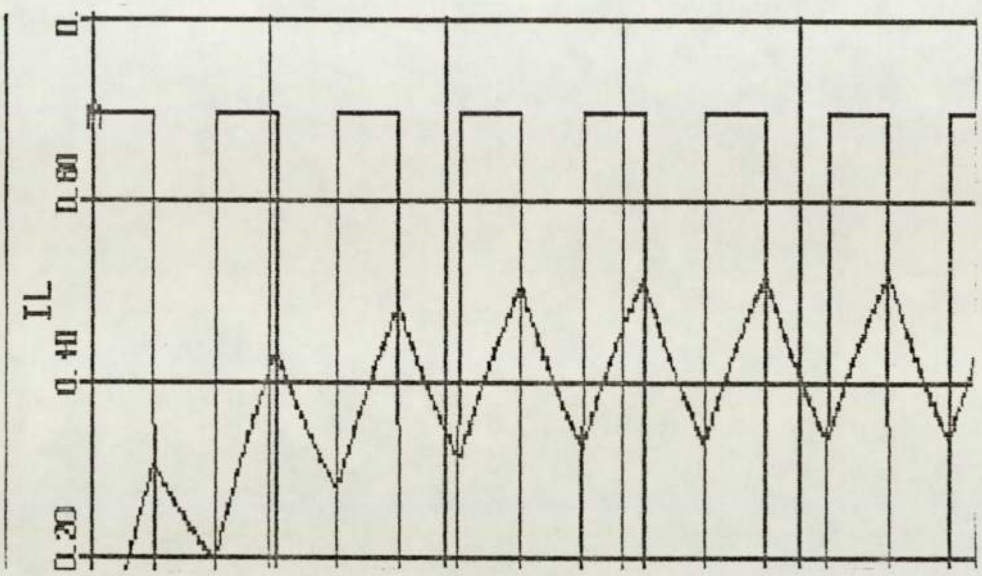
Amp/div=2mA Time/div=1ms Ton =50% T(period)=0.6ms



Practical result

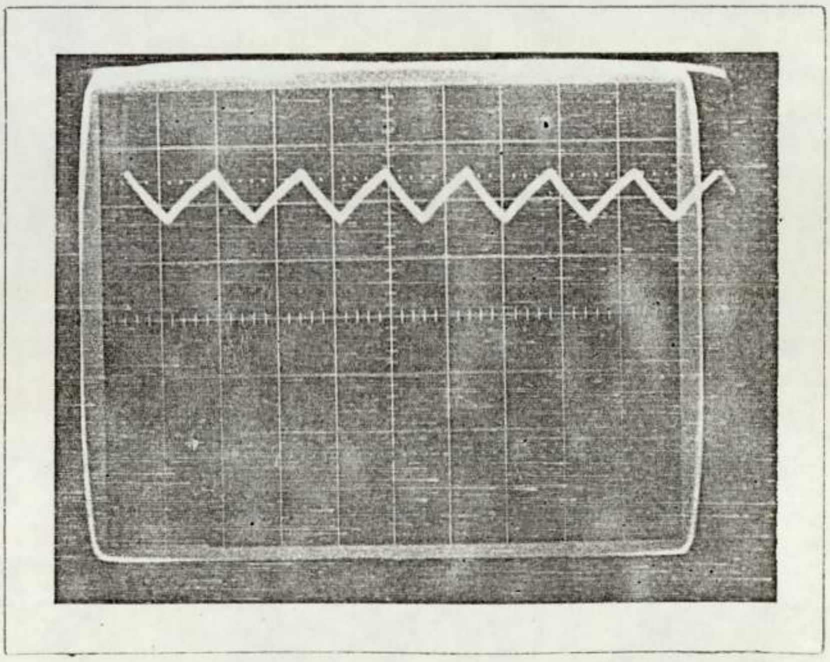
Amp/div=2mA Time/div=1ms Ton =50% T(period)=0.6ms

Fig 5.12(b) CHOPPER LOAD CURRENT



Simulation result

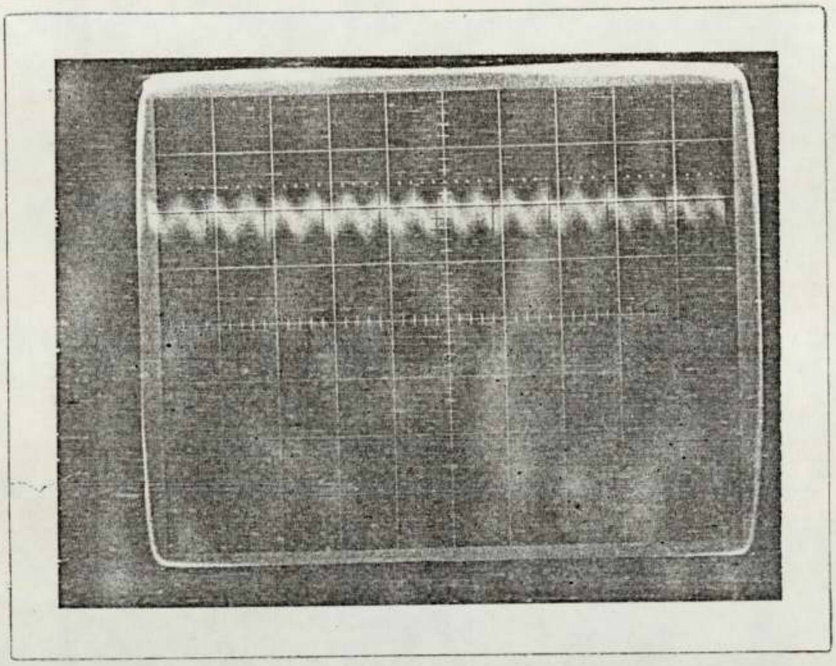
Amp/div=5mA Time/div=1ms Ton =75% T(period)=0.6ms



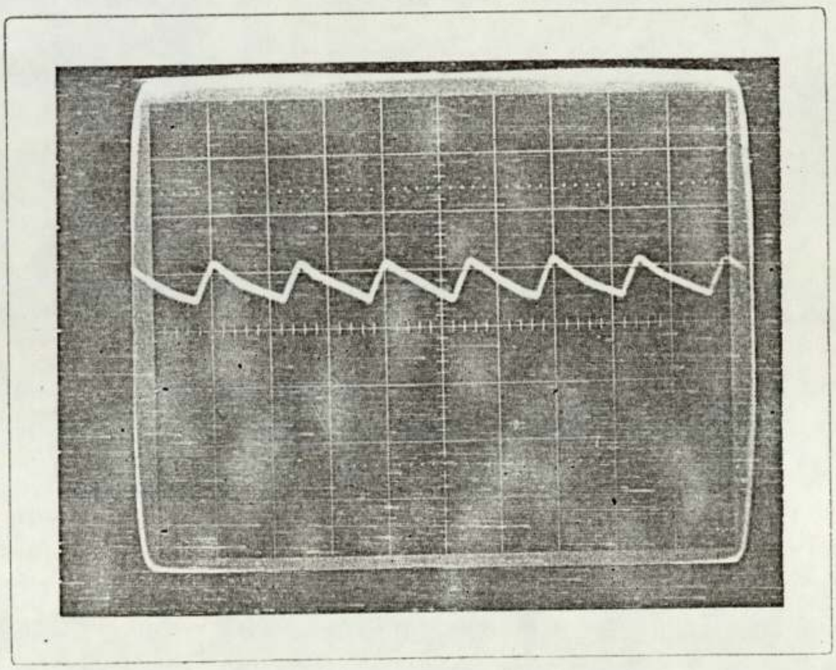
Practical result

Amp/div=5mA Time/div=1ms Ton =75% T(period)=0.6ms

Fig 5.12(c) CHOPPER LOAD CURRENT



Amp/div=5mA Time/div=1ms Ton =50% T(period)=0.6ms



Practical result

Amp/div=5mA Time/div=1ms Ton =25% T(period)=0.6ms

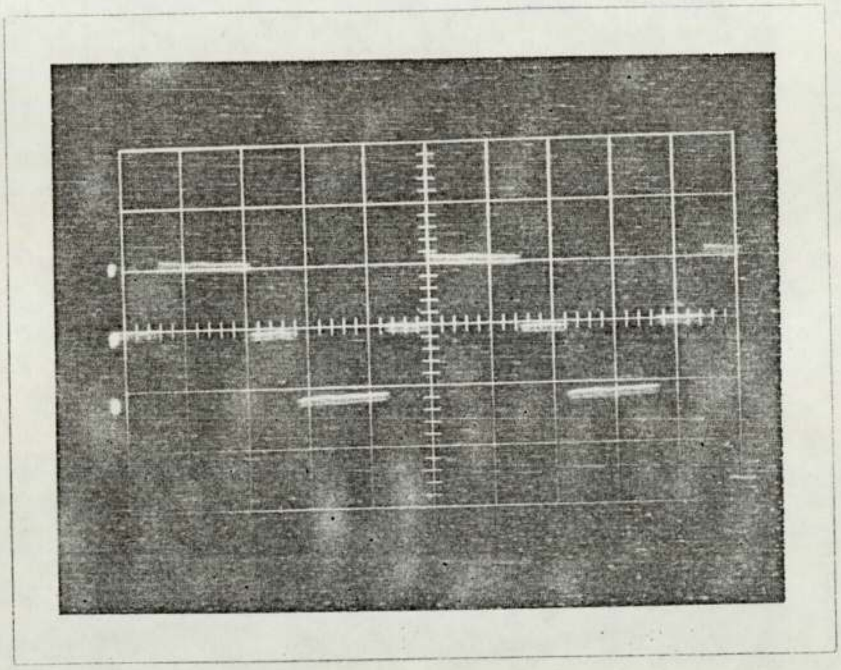
Fig 5.12(d) CHOPPER LOAD CURRENT

Finally tests on the drive system were conducted with an induction motor.The speed of the motor being controlled from 200 rev/min up to the maximum speed (1700 rev/min)steadily.

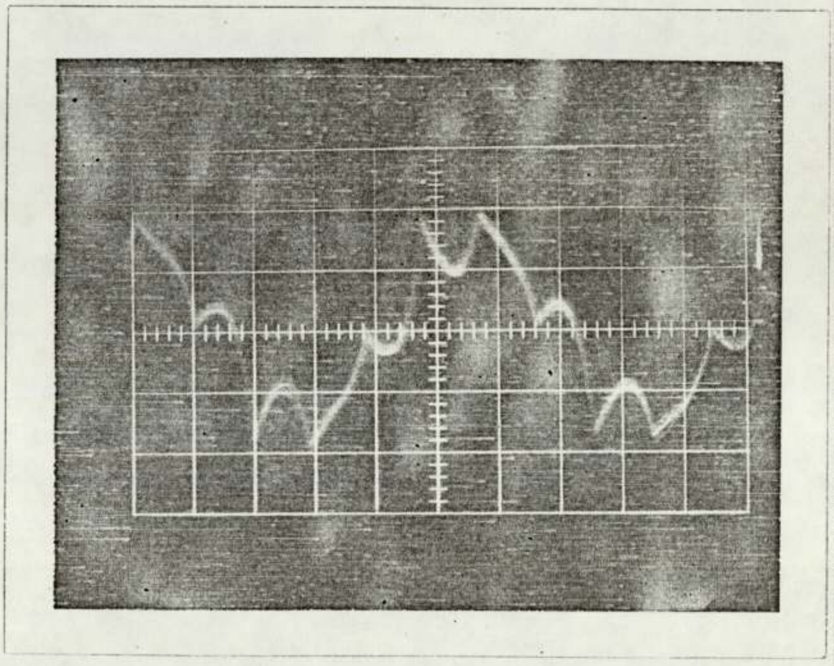
The line current and line voltage waveforms are shown in fig 5.13.,for the case when the induction motor not loaded is connected to the output of the inverter.
As can be seen the waveforms are "clean"(from any voltage spikes).The phase current waveform has the six step form that is a characteristic of a voltage source inverter.

The example waveforms shown illustrate that the inverter was switching quickly and cleanly.

The absence of overshoot voltage (spikes) shows that the diodes used (in the freewheel path and the snubber) were fast enough to cope with the rate of voltage change present in the system.



b)Line to line voltage
U=140 volts
f=30Hz



a)CURRENT WAVEFORM
Frequency :24 Hz
U=59 volts

Fig 5.13 VOLTAGE and CURRENT WAVEFORM of the INVERTER OUTPUT

5.8-Speed Control system with separate control of torque and magnetizing current components.

In the block diagram of the final drive circuit shown in fig 5.14,The starred (*) quantities indicate reference voltages,ω^*_r is the speed refence setting.

Speed feed-back ω_m is obtained from the tachogenerator and is compared with the refence setting.

a)Speed control:The slip frequency or the slip-speed.is generated as the ratio of the reference signals i^*_{1q}/i^*_{1d}

The synchronous speed reference ω_s is derived as the sum of actual motor speed and the slip speed.

The output signal of the summator is amplified and then converted into pulses(V/Hz converter),these pulses are used to trigger the GTO's thyristors of the inverter..The flux level is set manually,the vector sum of i^*_{1d} and i^*_{1q} is calculated and is used as the reference signal to control the output voltage of the chopper.

The tests results of the drive system are summarised in the next section.

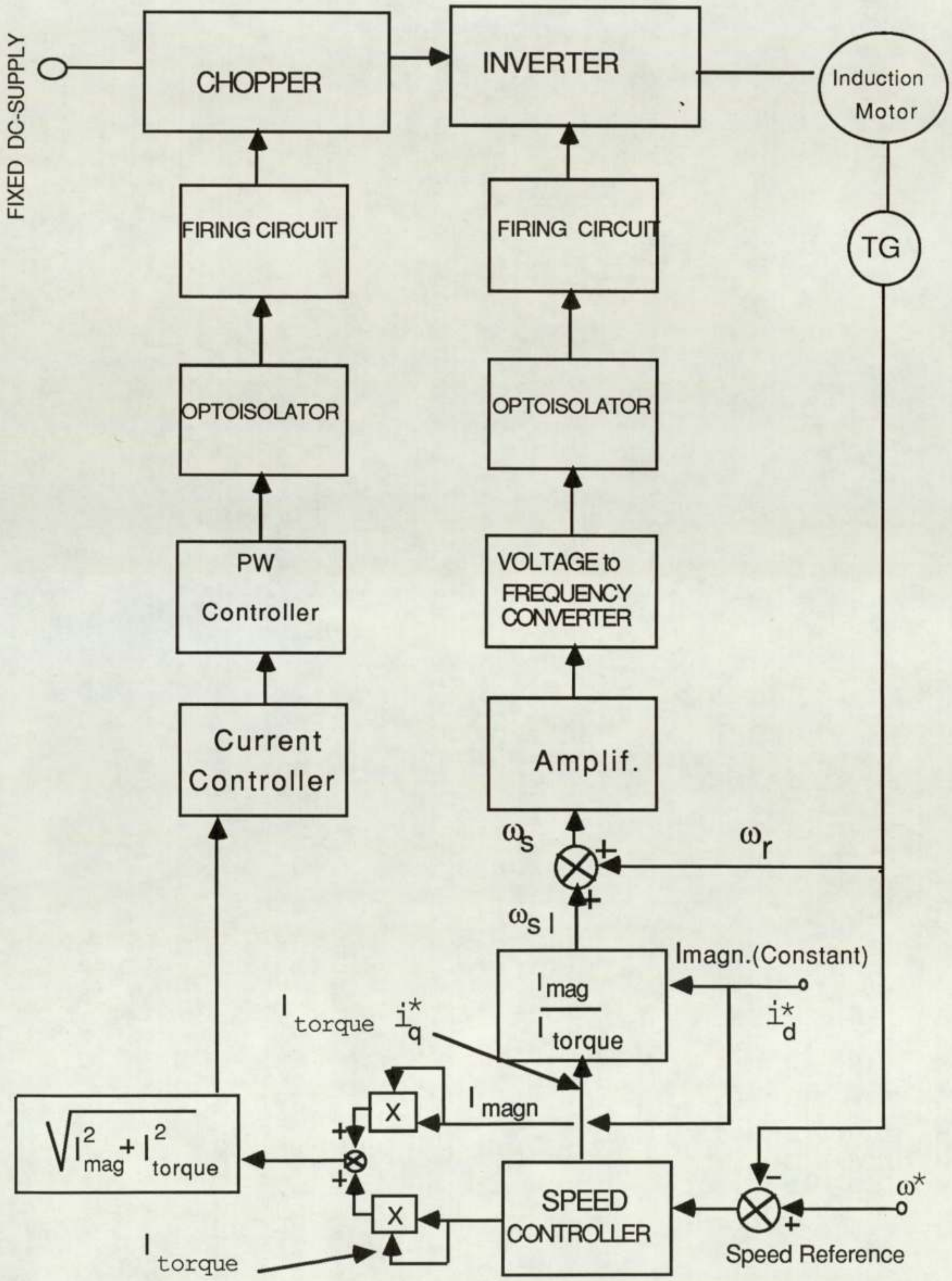


FIG 5.14 Circuit diagram of the separate control of the flux&torque

6. Conclusions.

The work carried out has demonstrated that G.T.O thyristors are;

a) relatively easy to incorporate in an inverter and chopper design.

b) their low power gate drive requirements reduce the amount of power electronic components required compared to a forced commutated inverter.

c)ACSL (Advanced computer simulation language) has been used to analyse the effect of the filter (L&C)components on the drive system

d) the results of the practical work [load currents for different chopping frequencies shown in fig 5.12(a-d)] are in agreement with the simulation results..

e) the use of this package (ACSL)has greately enhanced the design procedures during the project studies(choice of the filter inductance and capacitor values,the setting of the chopping frequency).

d) **Resistive load tests**:

The system (Chopper-Inverter) has been tested under various resistive loads and measurements made at the output of the inverter of the phase currents and the line voltages are shown in fig (6.1&6.2].

As can be seen from these results the line voltages(V_{AB}, V_{BC}, V_{CA})are in agreement with those predicted.

The efficiency of the "Chopper-Inverter" converter at full load was found to be very high "96%".This shows that only small losses occur in the GTO thyristor when it is fully on.as expected (the voltage drop of 2V across the device is given by the manufacturer-RS data sheet).

e) **No load tests**

The drive system composed of the chopper, the inverter power circuit and the induction motor were simply controlled using a constant ratio "volt per hertz",[V/Hz].The maximum speed reached by the induction motor (with no load) with this preliminary controller was 1700 rev/min.at 65 Hz

As can be seen from this no load test ,the speed up to the maximum of the rotor shaft is linearly related to the frequency increase,the motor have been driven beyond the nominal

Fig 6.1 Line current versus DC supply

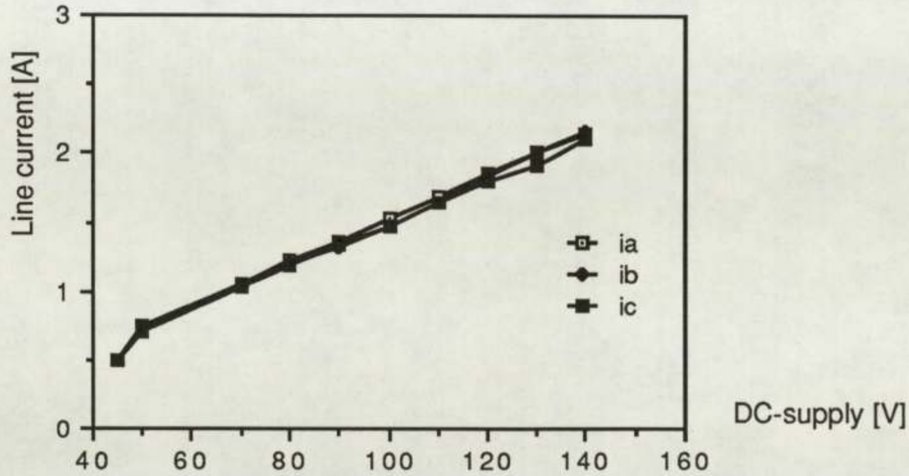


Fig 6.2 Inverter output voltage

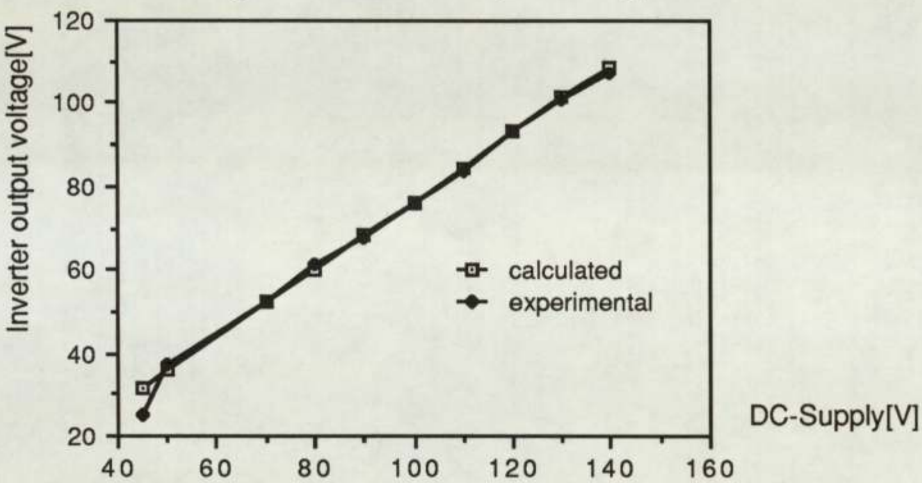
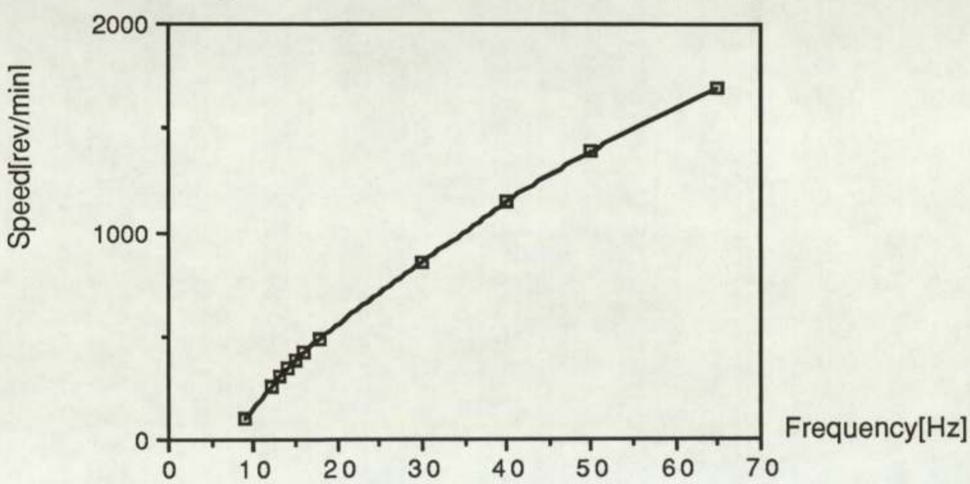


Fig 6.3 Rotor speed versus Frequency



frequency(50Hz-1400rev/min) up to 1700rev/min with a frequency up to 65Hz.

The result obtained is depicted in figures(6.3)

f) **Load tests with square-wave power supply**.

The drive system has been tested under various base frequencies(50Hz,40Hz,30Hz and 65Hz) at different loads. The reference frequency(base frequency)have been set to 50Hz and measurements of the torque developed by the motor ,the speed of the rotor shaft and the load current have been plotted.

Similar experiments have been carried out at different frequency settings(40Hz,65Hz) and the results have been plotted.

The Speed/Torque characteristics at 50Hz,40Hz and 65Hz shown in fig (6.4)and fig(6.5) are as expected.

The load characteristics (line current versus speed) are shown in fig(6.6 & 6.7) these characteristics show that with increasing load torque the speed drops and consequently an increase in current is observed.

This increase in current may be explained by the fact that the torque developed by the motor is proportional to the flux(which is maintained constant under V/Hz-control)and the current according to the relationship[$T=k\phi I$).

Fig 6.4 Torque-Speed characteristics

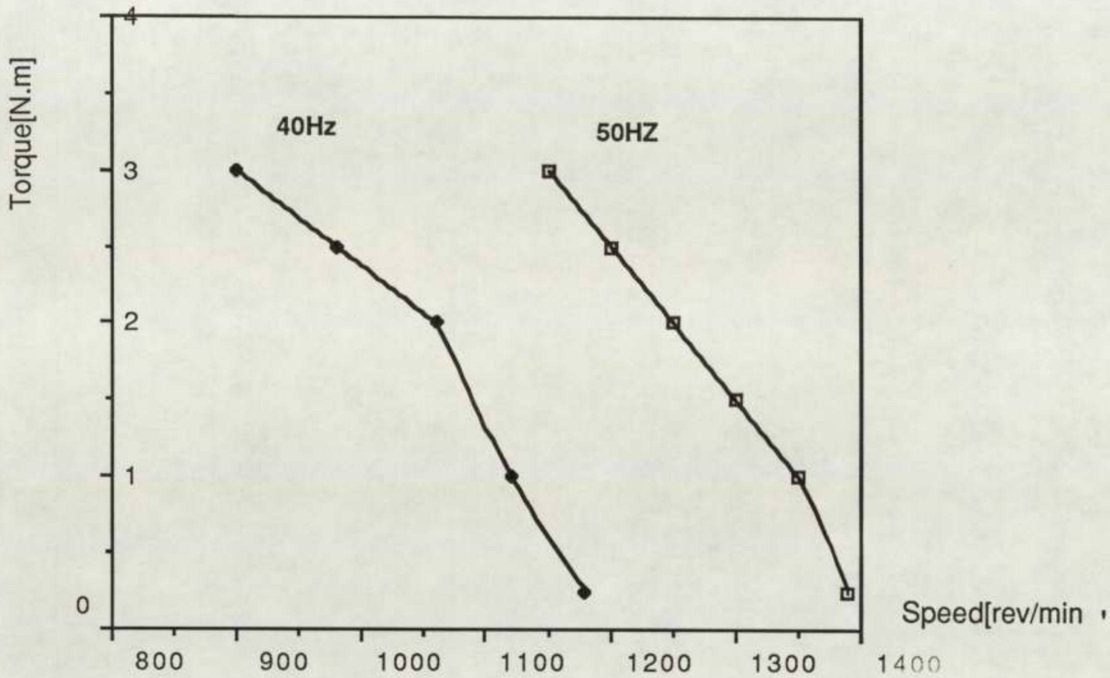


Fig 6.5 Torque-Speed characteristics

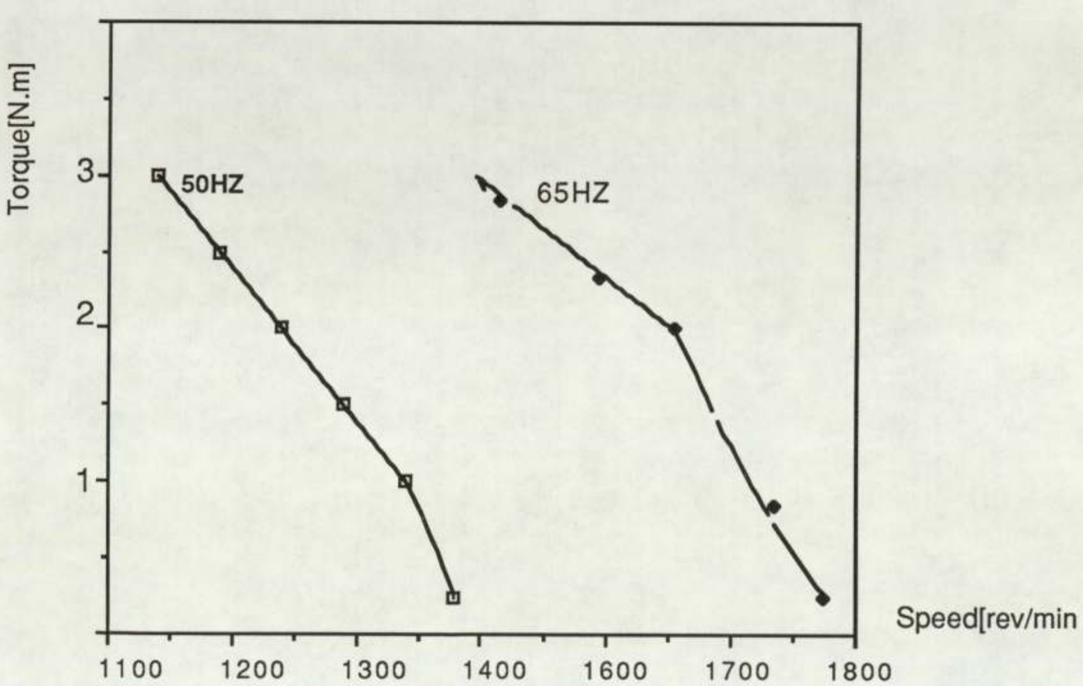


Fig 6.6 Load characteristic at 50Hz

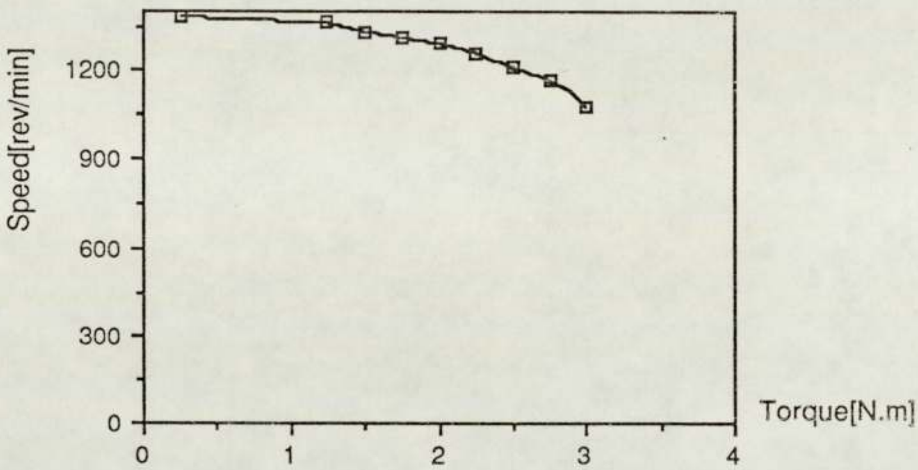
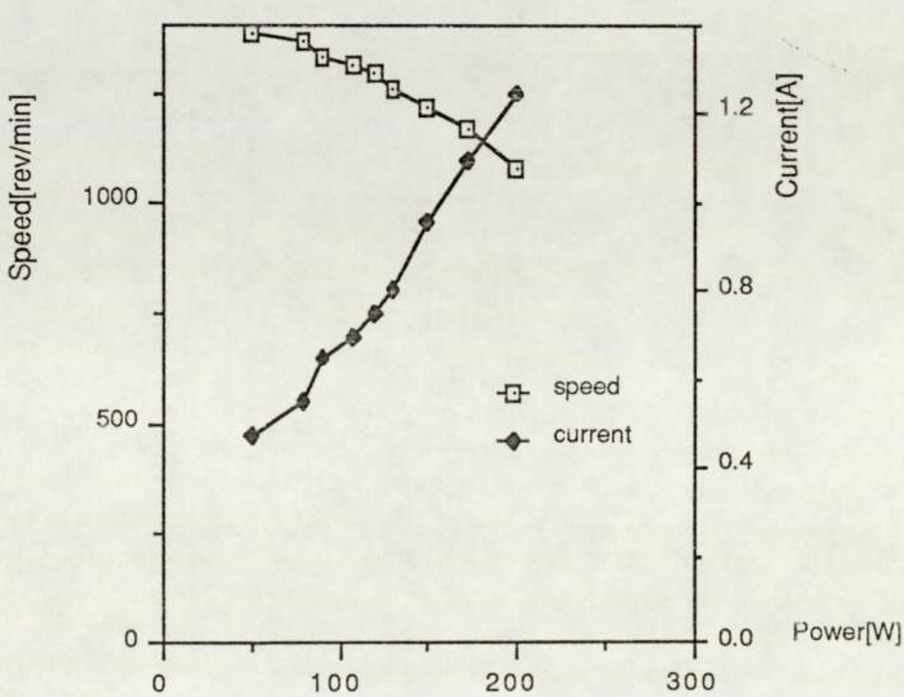


Fig 6.7 Load characteristics at 50Hz



As can be seen from fig (6.6)under "V/Hz control" which is an open loop control the relative speed drop represents 10%.of the nominal speed.

g) **Load test with sinusoidal power supply**

This test has been carried out in order to compare the speed/torque characteristics of the drive system supplied by a non sinusoidal voltage(square wave) and the motor supplied by a sinusoidal voltage

From this test,it can be concluded that,with a voltage wave input(quasi square waves) rich in harmonics,the machine operates as if the applied sinusoidal voltage were decreased.,this can be seen from the test result shown in fig 6.10,where the torque of the sinusoidal input voltage is higher than the torque developed by the machine when supplied by an inverter(non sinusoidal voltage input)of the same magnitude.

This comparison shows that for a better utilization of the motor,it is desirable,therefore,that the fundamental component in the non sinusoidal voltage be kept high.

These results have thus demonstrated:

Fig 6.8 Load current versus Speed

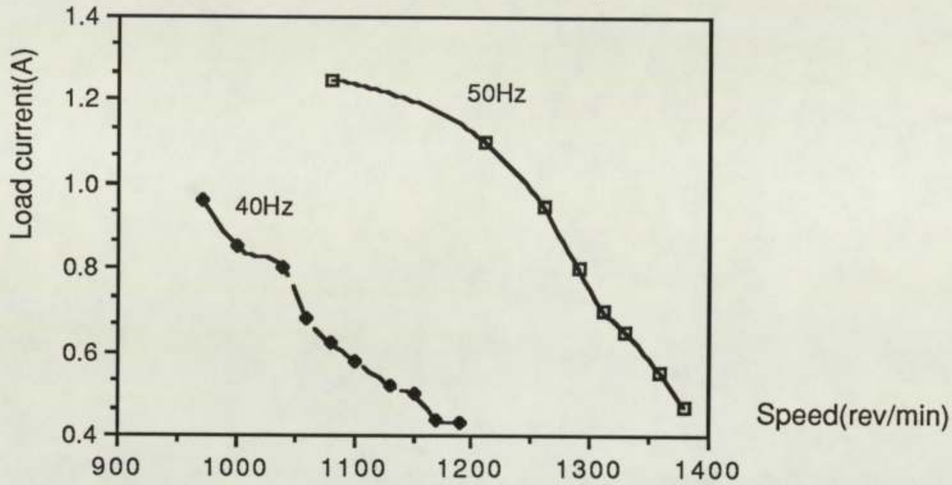
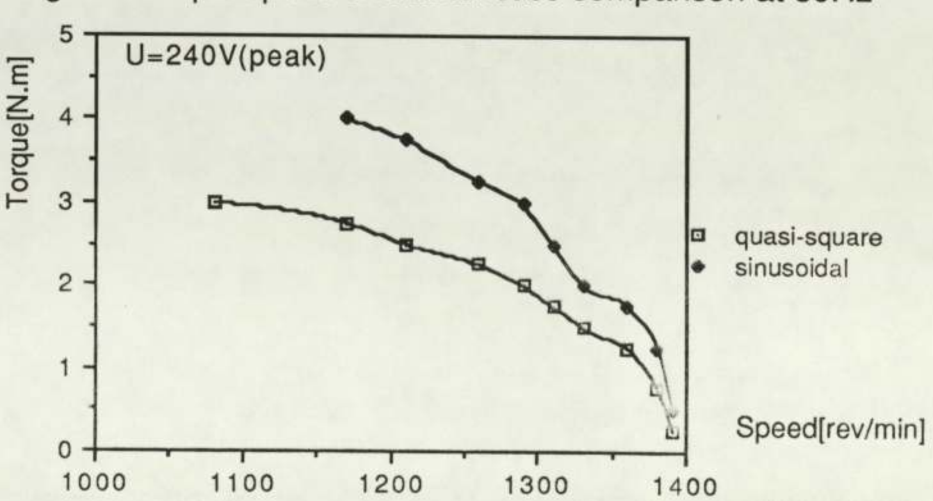


Fig 6.9 Torque/speed characteristics comparison at 50Hz



1) the operating conditions of the chopper coupled to the voltage source inverter(the first part of the work).

2) the drive operation was very stable over a large speed range.The first part of the work constitutes the framework for the second part of the project.

The second part of the project dealt with the implementation of the vector controller,in which the magnetising current component and the torque current component are controlled independantly..

3) this controller differs from the one used in the preliminary tests since the setting of the flux in the motor is no more dependant on the strict ratio of the stator voltage to the stator frequency delivered by the "V/Hz-controller" but set manually, the torque signal is varied separately.

4) this means of control has some resemblance to the conventional DC motor drive system.

5) It is based on the simple and direct relationship that exists between flux,torque and slip frequency of the motor on one hand

and components of the stator current vector on the other.hand:

6) It does not require either measurement or computation of flux and torque angle as in the vector control,which is far more complicated to implement,because:

a)the vector control method requires refined measurement techniques and complex circuitry.

b) the scheme is attractive in comparison with the "V/Hz control" presented earlier in the sense that the speed is controlled smoothly at the lowest range(from 10rev/min),whereas in the V/Hz control ,at low speed (below 100 rev/min)the drive shows instability and high current pulsations

c) the rotor shaft starts moving only at a speed reference of one volt(that is to say a base frequency of about 18Hz),whereas the method used in the final practical work shows that the rotor shaft starts moving even with a very low speed reference (0.2 V) (less than 8Hz)corresponding to a speed of 10rev/min..

6.2 Suggestions for further work

The following areas can be expanded and improved.

1) The Pulse Width Modulation strategy may be used (instead of the combination Chopper-Inverter power converter)to eliminate the low order harmonics in the system which are harmful to the motor(heating,pulsating torque).

2) A microprocessor generating the triggering pulses(P.W.M) allowed the system to run at low speed (with low harmonic content) the latest development in speed control

3) The control strategy used in this project to separetly control the magnetising current component and the torque current component are not satisfactory(Chopper-V.S.I-drive),thus a P.W.M may improve the dynamic behaviour of the drive system.

APPENDICES

Appendix 1

The GTO-Thyristor-construction

The GTO-Thyristor is a four layer pnpn device.Constructed similarly to a conventional thyristor.However a new factor,turn-off gain(Goff),is introduced,and defined as equal to the anode current Ia,divided by the amount of reverse gate current(Ig)required to turn-off the anode current,Goff=Ia/Ig.

Because of design compromises to achieve a fast turn-off capability in GTO-Thyristors,the high regenerative gain inherent in conventional SCRs is reduced in GTO design.GTOs are specially configured so that a negative current flow to the gate can turn-off the device by drawing off current from Ia sufficiently to break the regenerative loop.

1-1 The two-transistors analogy of the GTO.

The approximate equivalent circuit,shows that the device acts as if it contained two transistors,a p-n-p and a n-p-n interconnected to form a regenerative feed-back pair as shown in (fig-A),and a zener diode.

The turn-on process is exactly the same as for ordinary SCRs.

From the fig (A-3),it is evident that the collector of the n-p-n transistor provides base drive for the p-n-p transistor.

$Ib1 = Ic2 + Ig(n)$

Similarly the collector of the p-n-p transistor along with any p-gate current Ig(p) supplies the base drive for the n-p-n transistor.

$Ib2 = Ic1 + Ig(p)$

When both transistors Q1 and Q2 have very small forward bias of the emitter-base junction,the value of the gain $(\alpha1+\alpha2)<<1$,consequently the anode current Ia is also small.The sum $(\alpha1+\alpha2)$ can be made temporarly close to unity by injecting a short duration positive gate current at the P-gate terminal,which is the base of transistor Q2.This causes collector current Ic2 to flow in transistor Q2;as we can see in fig (A -1/2/3),this collector current Ic2 is the base current of Q1 as a result Q1 will be switched-on.At this point each transistor supplies the base current of the other and the action is regenerative.For this reason the gate current is no longer needed.This regenerative situation may be expressed by the following equations:

$Ib1 = (1-\alpha1).Ia - I_{leak1}$

$Ic2 = \alpha2.Ik + Ic + I_{leak2}$

Since the base current Ib1 of Q1 and the collector current Ic2 of Q2 are the same,then:

$$I_{b1} = I_{c2}$$

$$(1-\alpha 1).I_a - I_{leak1} = \alpha 2.I_k + I_{leak2}$$

and since Ia =Ik,

$$(1-\alpha 1).I_a - \alpha 2.I_k = I_{leak1} + I_{leak2}$$

$$(1-\alpha 1-\alpha 2).I_a = I_{leak1} + I_{leak2}$$

$$I_a = I_k = I_{leakt}/[1-(\alpha 1+\alpha 2)]$$

1.2 - Turn-off design criteria

By a consideration of the two transistors analogy the base current Ib2 required to sustain the n-p-n transistor -ON,is $(1-\alpha 2).I_k$,whereas the actual base current Ib1 is$[\alpha 1.I_a + I_{g(n)}]$.Therefore the GTO will turn-off if :

$$\alpha 1.I_a + I_{g(n)} < (1 - \alpha 2).I_k$$

But as $I_a + I_{g(n)} = I_k$,this can be written as:

$$-I_g > I_a[(\alpha 1+\alpha 2 - 1)/\alpha 2]$$

This equation therefore,gives the condition for the minimum negative current which will turn-off the device.It shows that in order to arrive at the desirable low value of Ig,$\alpha 2$ should be large and $\alpha 1$ should be small.

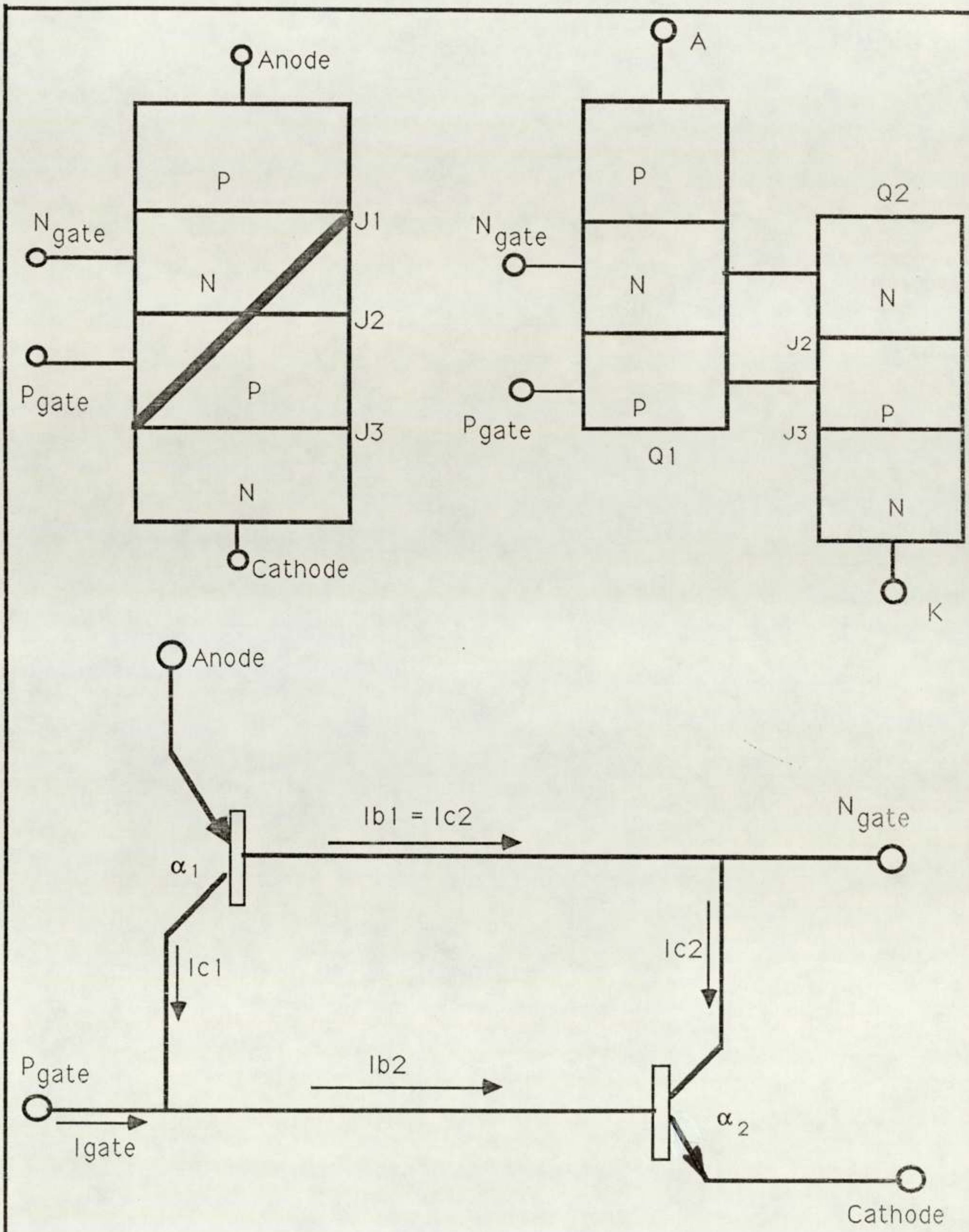


FIG-A TWO-TRANSISTORS ANALOGY OF THE GTO-THYRISTOR

1-3 G.T.O's Inverter snubber design.

1.3.1- dV/dt snubber.

During the G.T.O thyristor turn-off period,it is assumed that the inductive load current remains constant and that it transfers from the G.T.O to the capacitor immediatly the switching -off procedure is initiated.The "worst case" capacitor value is then chosen so that the snubber capacitor voltage does not rise above the allowable anode-cathode voltage when the maximum load current (Ian) is flowing.

The maximum Anode-Cathode voltage was chosen as the maximum rail to rail voltage of the inverter (V_{rr}).

$$\text{i.e} \quad C_V = I_{load} / (dV/dt) = [\, I_{load}/V_{rr} \,]t_f \quad \text{Farads} \qquad \text{a.1}$$

Where t_f is the fall time of the G.T.O .The fall time of the G.T.O used in this project is 0.25 μs and a value of 5amperes was chosen for I_{ANODE}.The maximum rail to rail voltage is 240 volts and so equation (a.1) gives a value for C_V of 0.0052 μF.In practice a value of 0.33 μF was used (to reduce as much as possible the rate of rise of voltage)

The capacitor C_V discharges during the switch-on period of the G.T.O via a resistor R_V.During this discharge, R_V must limit the discharge current to a value I_{dis} that the device can support.

i.e $R_V = V_{rr}/I_{dis}$ (ohms) a.2

The power G.T.O in the inverter used to drive the induction motor can support a load current of at least 5 amperes.I_{dis} was chosen as 5 amperes.Equation a.2 then gives:

$$R_V = V_{rr}/I_{dis} = 240 \text{ volts}/5 \text{ amperes} = 48 \text{ ohms}$$

The power rating of the resistor R_V is then calculated by assuming that C_V discharges its stored energy during each "ON-PERIOD"

The power dissipated (P_v) in the resistor R_V when the G.T.O switches "f_s times " per second is :

$$P_v = 0.5\, C_v\, Vrr^2\, f_s \qquad \text{(watts)} \qquad \text{a.3}$$

Substituting for C_V from equation a1

$$P_V = 0.5\, I_{ANODE}\, V_{rr}\, t_f\, f_s \qquad \text{(watts)}$$

The maximum motor speed of 1400 rpm is achieved with an inverter frequency of 50HZ.Hence for f_s=50HZ,equation a3 gives P_v = 36mW.Practically a resistor of 33 ohms,0.5 watts is chosen for R_v.The time constant R_vC_v is =0.8 ms and so the capacitor C_v can fully discharge in between the G.T.O switch-on operations.

1.3.2 dI/dt limiting snubber.

During the G.T.O turn-on period,it is assumed that the load voltage remains constant and that the G.T.O voltage drops immediatly to zero volts.The "worst case " inductance value is then chosen so that during the switching process,the anode current does not rise above a value that the device can support.The maximum voltage that can appear across the inductance "Lt " during the switch operation" is the maximum inverter rail to rail voltage V_{rr}.Hence:

$$Lt = V_{rr}/[dI_{ANODE}/dt] =[V_{rr}/I_{ANODE}]t_r \qquad \text{(Henries)} \qquad \text{a.5}$$

The G.T.O type BTW58-1300R has a switch on time of about 1 μs.Equation (A.5) gives a value of 48 μH

Where Lt is the total inductance for one inverter arm.

With Lt = Li1 +Li2 and Li1 = Li2

Thus Li1 = Li2 =24 μH.Practically a larger value for Li1 and Li2

have been chosen,to further reduce the dI/dt.

Li1 = Li2 50 μH

Appendix 2-Induction motor drives 'modelling and simulation'

2.0 **Modelling and simulation of ac motor drives**

2.1- Introduction

Most variable frequency drive systems are quite complex and it is often difficult to predict their dynamic performance without the aid of a digital computer.

Before the computer era,the limitation of computation facilities restricted the studies to steady-state conditions,often such studies were experimental,accompanied by some qualitative explanations of the results [4.1].In fact the non-linearity of the machine equations added to the discrete-time effect of the converter are of major computational complexity,and therefore simulation on computer become essential.

To avoid unnecessary complications and to properly assess costly systems before proceeding to their construction,it has now become the general trend to simulate and to analyse such systems by computer.The physical system can be modelled mathematically and tested for various conditions and constraints to ensure that the physical model will respond favourably in the prescribed limits.

The aim of this chapter is to highlight and emphasize the need for and the contribution of modelling and simulation on the progress of AC drive systems,and to show how this study can be carried out.

In this chapter,a conventional three axis model of an induction motor will be presented first,before establishing the model in the synchronously rotating frame.This will be followed by the model of the drive system,and finally,how the simulation of this drive can be implemented.

2.2 Mathematical model.

The AC induction machine equations have been derived by many authors [4.2]-[4.4].The transient behaviour of induction machines,are based on the well known two-axis theory,and are solved by numerical integration on digital computers.

Probably the most convenient appproach is to be found in the stationary-axis method so successfully applied by R.H.Park [4.5] to the analysis of salient-pole synchronous machines.

It is therefore,the purpose of the following subsection,to present a conventional three-axis model before introducing the synchronous frame model.

2.2.1 Derivation of the differential equations.

The analysis is based upon the following simplifying assumptions:

1-Balanced rotor windings are assumed for all cases,and the three-phase machine equations are derived upon the additional assumption that the stator windings are also balanced.

2-It is assumed that the coefficient of mutual inductance between any stator windings and any rotor windings is a cosinusoidal function of the electrical angle between the axes of the two windings.

3-It is further assumed that the rotor is smooth and that the self-inductances of all the windings are independant of the rotor position.

4-Finally the effect of saturation,hysterisis,and eddy currents are neglected.

By summing the voltages around the closed circuit of each of the rotor and stator phases,the following fundamental voltage equations may be written (*-Refer to "the nomenclature" for a list of symbols and their definitions).

Stator-voltage-equations

Where, V_{as} (V_{bs},V_{cs}) is the voltage applied to the coil A (B,C) ,this has to overcome the resistance drop and the voltage induced in coil A (B,C) due to the change in the flux linkage (the transformer voltage).Since this coil is stationary in space there cannot be any dynamic generated voltage in it.

Hence:

$$V_{as} = p\Psi_1 + R_s i_{as}$$

$$V_{bs} = p\Psi_2 + R_s i_{bs}$$

$V_{cs} = p\psi_3 + R_s i_{cs}$

Rotor voltage-equations:

For the rotor coils,the situation is different.In addition to the transformer voltage there will also be generated voltages in the rotor conductors when they rotate with respect to the magnetic field.Hence the equation for the applied voltage in a rotor coil will contain the terms:

$V_R = Ri + d\psi/dt + Bp\theta$

The term $Bp\theta$ is the generated voltage,which depends on the speed of the rotor.The rotor equation may be expressed as follows:

$V_{ar} = p\psi_a + R_r i_a$

$V_{br} = p\psi_b + R_r i_b$

$$V_{cr} = p\Psi_c + R_r i_c \qquad \text{b.1}$$

The three-pase stator equations given in (B.1),can also be expressed in matrix form by the following relations:

$$[V_s] = [R_s][I_s] + d[\psi_s]/dt \qquad \text{b.2}$$

Where:$[V_s]$,$[I_s]$,$[\psi_s]$ are instantaneous voltage,current,and flux linkage vectors respectively in the stationary frame.

The rotor equations in matrix form can also be expressed as:

$$[V_r] = [R_r][I_r] + d[\psi_r]/dt. \qquad b.3$$

where:[Vr],[Ir],[y] are instantaneous voltage,current,and flux linkage vectors respectively in the rotor.

All vectors and matrices in (b.2) and (b.3) will be given in detail in appendix (3).

The flux linkage matrix can be expressed as:

$$\psi = \begin{bmatrix} \psi_s \\ \psi_r \end{bmatrix} = \begin{bmatrix} [Ls] & [Msr] \\ [Mrs] & [Lr] \end{bmatrix} = \begin{bmatrix} [Is] \\ [Ir] \end{bmatrix} \qquad b.4$$

detail of submatrices and the [ψ] matrix in its complete form will be given in appendices 3.

From equations (b.2),(b.3),and (b.4) the machine equation in matrix form can be expressed by:

$$\begin{bmatrix} V \end{bmatrix} = \begin{bmatrix} [Vs] \\ [Vr] \end{bmatrix} = \begin{bmatrix} [Rs][Is] + d\{[Ls][Is] \quad [Msr][Ir]\}\,dt \\ [Rr][Ir] + d\{[Msr][Is] \quad [Lr][Ir]\}\,dt \end{bmatrix} \qquad b.5$$

As can be seen from the matrix given in appendix 3,these equations contains numerous trigonometric coefficients, however,by a change of reference-axis,to obtain equations which do not contain these trigonometric function,it is possible to transform equation (3.2) into the synchronously rotating frame.

4.2.2 Synchronously rotating frame model.

As explained in appendix (3) related to axis transformations,this linear transformation into the d-q axis will lead to simpler equations.As shown in the previous subsection,it is possible to express the phase stator equations of the machine in the stationary coordinates as,bs,cs by equation (b.2).Then using the transfomation operator for the stator parameters,the d-q rotating axis transformation,will give the following stator voltage and current vectors as:

$$[V_s] = [P][V_{dq}]$$

$$[I_s] = [P][I_{dq}] \qquad \text{b.6}$$

where:$[V_s]$,$[I_s]$ are instantaneous voltage and current vectors in the conventional stationary frame,and $[V_{dq}]$,$[I_{dq}]$ in the synchronously rotating frame.

[P]-is the required transformation matrix for the three-phase to two phase rotating axis transformation.This matrix is given in equation(b.7),which is also known as the Park's matrix,appendix-3.

$$P = \begin{bmatrix} 1 & 0 & 1 \\ -1/2 & -\sqrt{\frac{3}{2}} & 1 \\ -1/2 & \sqrt{\frac{3}{2}} & 1 \end{bmatrix} \qquad b.7$$

The substitution of $[V_s]$ and $[I_s]$ in equation (b.5),will result in :

$$[P][V_{dq}] = [R_s][P][I_{dq}] + d\{[L_s][P][I_{dq}] + [M_{sr}][I_r]\}/dt.$$

$$[V_r] = [R_r][I_r] + d\{[M_{rs}][P][I_{dq}] + [l_r][I_r]\}/dt \qquad b.8$$

After manipulating (b.8),the final form of the voltage matrix is as given in (b.9).

In the voltage matrix (b.9),the following arrangements are made

1) $d/dt = p$ -is the Laplace operator.

2) $\theta' = p\theta$

$V = r.i + L\{di/dt\} + i.\{dL/dt\}$

$V = r.i + Lpi + p\theta G.i = [r + Lp + \omega_r G].i$ b.9

$\theta' = \omega_r$ -is the angular speed.

B.3 **Induction motor drive modeling**

As mentioned in chapter (two),the converter constitutes an interface between the power supply and the machine.Here in this section a simplified model of the converter-inverter induction motor drive system,will be presented,this model is widely used for stability analysis,assuming that the inverter can generate symmetrical three-phase square-wave,or a P.W.M voltage,and the chopper as a controllable DC-voltage source,as explained in [4.7]. As can be seen in appendix -3,in the synchronously rotating frame the voltage can be derived from:

$$V_{qs}=2/3[V_{AS}.\cos\phi + V_{BS}.\cos(\phi-120) + V_{CS}.\cos(\phi-240)]$$

$$V_{ds}=2/3[V_{AS}.\sin\phi + V_{BS}.\sin(\phi-120) + V_{CS}.SIN(\phi-240)]$$

$$v_{os} = 1/3[V_{AS} + V_{BS} + V_{CS}] \text{(zero sequence)} \qquad b.10$$

or directly from the stationary frame voltage as:

$$V_{qs} = V^s_{qs}.\cos(\omega_r.t) - V^s_{ds}.\sin(\omega_r.t)$$

$$V_{ds} = V^s_{qs}.\sin(\omega_r.t) + V^s_{ds}.\sin(\omega_r.t) \qquad b.11$$

The machine voltage in terms of the DC link voltage are expressed as:

$$V_{qs} = 2/\pi(V_d)G_{qs}$$

$$V_{ds} = 2/\pi(V_d)G_{ds} \qquad b.12$$

Where G_{qs} and G_{ds} are defined as the switching functions.

Assuming that there is no power dissipation through the inverter,which can be considered to be a switching network,the relation between inverter DC link current and machine current can be derived by considering the power balance between the input and output of the inverter.

The instantaneous power relation is given as:

$$V_d I_d = V_{an} I_{as} + V_{bn} I_{bs} + V_{cn} I_{cs} \qquad \text{b.13}$$

Substituting the phase variables (as,bs,cs),in terms of d-q,equation b.13 will be expressed as:

$$V_d I_d = 3/2(V_{ds} I_{ds} + V_{qs} I_{ds}) \qquad \text{b.14}$$

Substituting equation b.12 in equation b.14,the DC link current as a function of machine current will be expressed as:

$$I_d = 3/\pi (I_{ds} G_{ds} + I_{qs} G_{qs}) \qquad \text{b.15}$$

So far it has been asssumed that the inverter,is fed from a fixed DC-voltage,but in general,there will be a filter and the variable DC voltage may be obtained from a controlled rectifier or a chopper in the front end.Hence the DC link equation can be written as :

$$V_{rect.} = L_d\{d(I_{rect.})/dt\} + I_{rect.} R_d + U_d \qquad \text{b.16}$$

-

$I_{rect} - I_d = C_d . \{d(U_d)/dt\}$ b.17

where:

$V_{rect.}$, I_{rect} are the voltage and current at the output the rectifier ,and L_d , R_d , C_d are the filter parameters.

2.4 Induction motor drive simulation

2.4.1 Electric machine simulation

Based on the general theory presented so far and following the assumptions given in b.2,an induction motor can be represented by a set of differential equations,as expressed in (b.9),and can be written in the form:

$$[V] = \{ [R] + \theta'.[G] + p[L] \}.[I]$$ b.18

-

where:

$[V] = [V_{qs}, V_{ds}, V_{qr}, V_{dr}]^T$ is the transposed vector of terminal voltages

[R] = diagonal matrix formed by the resistance terms.

[G] = matrix formed by the cofficients of $\theta' = \omega_r$

[L] = matrix formed by the coefficients of operator $p = d/dt$.

$[I] = [I_{qs}, I_{ds}, I_{qr}, I_{dr}]^T$ is the transposed vector of axis windings currents.

The generalised equation of electromagnetic torque is given by:

$$Te = (P/2).[I]^T.[G].[I]$$ b.19

After further manipulation,equation (b.19) can be written in terms of currents (b.20) or fluxes (b.21).

$$Te=(P/2).\{I_{qs}.(L_{ds}.I_{ds}+L_{md}.I_{dr})-I_{ds}.(L_{qs}.I_{qs}+L_{mq}.I_{qr})\} \qquad b.20$$

$$Te=(P/2).\{I_{qs}.\Psi_{ds}-I_{ds}.\Psi_{qs}\} \qquad b.21$$

The equation of motion is represented as:

$d^2(\theta)/dt^2=\theta''$

-

$$Te-Tl=J.\theta''+D(\theta') \qquad b.22$$

where:

Tl : is the load torque,

J : inertia constant

$D(\theta')$: mechanical viscous damping

ω_r : rotor angular speed

θ : rotor angle

-

The fundamental electric equation of the machine (b.18),with the equation of motion (b.22), makes the complete set of first order differential equations for studying the machine behaviour.

For a numerical solution of the above equation,using a digital computer,equation (b.18) is used in the form:

-

$I'=d[I]/dt$ (derivative of "I")

$$[I'] = [L]^{-1}.\{ [V] - ([R] + \theta'.[G]).[I] \} \qquad b.23$$

which represents a set of first-order differential equations and the second-order differential equation (b.22) is replaced by two first-order equations,one for speed in terms of torque eq. (b.24), and the other for position in terms of speed eq. (b.25).

$$\omega'_r = (1/J).\{ Te - Tl - D(\omega) \} \qquad b.24$$

$$\theta' = d\theta/dt$$

$$\theta' = \omega_r \qquad b.25$$

It is worth noting that eq.(b.18) can also be written in the form of eq.(b.26),for flux linkage in terms of currents and voltages,as explained in [b.6]

–

$$[\psi'] = -[R].[I] + [\omega][\psi] + [V] \qquad b.26$$

where :

[I] , [V], [R] are the same as for eq.(b.18),and

$[\psi] = [\psi_{qs}, \psi_{ds}, \psi_{qr}, \psi_{dr}]^T$ is the transpose of the flux linkages vector.

2.4.2 Choice of the state-variable

In general,eq.(b.22),is used either with eq.(b.18) or with eq (b.26) for studying the machine behaviour,but the choice depends essentially on the aim of the simulation study.In fact this choice automatically implies the choice of state-variable,as explained in

[4.8],[4.9] and [4.10].

Flux vector as state variable.

- Based on the assumption of no saturation effects which means that there is a linear relationship between the magnetizing current and the main flux linkage,hence the behaviour of the machine can be predicted using a simplified method.

- By taking into account the effect of saturation,this implies that there is no linearity between the magnetizing current and the main flux.This method is longer and more complex,but more accurate than the previous one,because an iterative method is used for determining the current vector [I] ,at each step of the integration.

-

Current vector as state-variable

In this method,there is no need to use iterative techniques to define [I] ,because eq.(b.23) is solved directly with step by step procedure.For this reason the choice of [I] (current) as the state variable is more interesting and allows considerable reduction of computing time,as shown by J.FAUCHER et al. [4.11],in their comparative study on the simulation methods of ac motor drive models.

2.5 Simulation results

As an example of the application,the model presented in subsection b.2.2 ,is applied in the simulation of a three phase AC supply.

The induction machine rating and equivalent circuit parameters are given below:

Induction motor: 12.6MVA

Rs=0.00414,Rr=0.0108,Xs=4.0855,Xr=4.0484,Xsr=1.0,F=1.0,U=1.0.

A program was developed to simulate the transient and steady state operation of the induction machine presented previously.

-

2.5.1 Formulation of equations

Equation (b.23) , (b.24) ,and (b.25) are reformulated in a particular way,hence they can be written in the state-space equation form.

-

$$[X'] = [A].[X] + [B].[U] \qquad b.27$$

-

where :

[X '] , [X] state vectors

[U] forcing quantity vector

[A] , [B] are the plant and forcing matrix respectively.

Hence the plant matrix will be expressed as :

-

$$[A] = -[L]^{-1}.\{[R] + \theta'.[G]\} \qquad b.28$$

and the forcing matrix as :

-

$$[B] = [L]^{-1} \qquad \text{b.29}$$

Then ,eq. (b.23) will be reformulated as :

-

$$[I'] = [A].[I] + [B].[V] \qquad 4.30$$

where :

[I] : is the vector formed by the state-variable (currents),and [V] : is the vector formed by the input variables (voltages).

-

2.5.2 Solution of the differential equation.

A step by step numerical solution of the differential equations ,based on the fourth order RUNGE-KUTTA method is used to solve the set of differential equations.For this reason a time increment value (integration-step) has to be defined as a fraction of the shortest time constant in the circuit (typically :1/10) ,then all the variables have to be set to their initial values ,and finally the initial and final times have to be specified .Fig .B.5 shows a general flow diagram for a numerical solution with a digital computer.

2.5.3 Numerical considerations.

Many numerical considerations have to be taken into account

,when developing a simulation program ,to optimise the use of computing time.

Regardless of the simulation aim ,whether the problem is transient or steady state,matrix calculations must be efficiently evaluated.

For a steady state problem,in general the system studied is stable ,so the time increment can be increased without affecting the accuracy of the integration process.In addition to this the motor speed is virtually constant,so the state equations are linear and can be implemented numerically.

For the transient solutions ,the motor speed ω_r is changing and the state matrices are non linear ,so efficient means of linearization must be used.In general for transient calculations,,the state equations can be conveniently linearized by using the fact that the mechanical and electrical transients have significantly different time constants.Therefore ,it is assumed that within an interval "T" which is greater than the solution interval "H" (integration step) ,the rotor speed is constant .Hence the state equations are linearized and have not to be reevaluated at each integration step ,so simplifying considerably the problem.

2.5.4 Program organisation

The flowchart presented in fig B.6 shows in a simplified form ,the

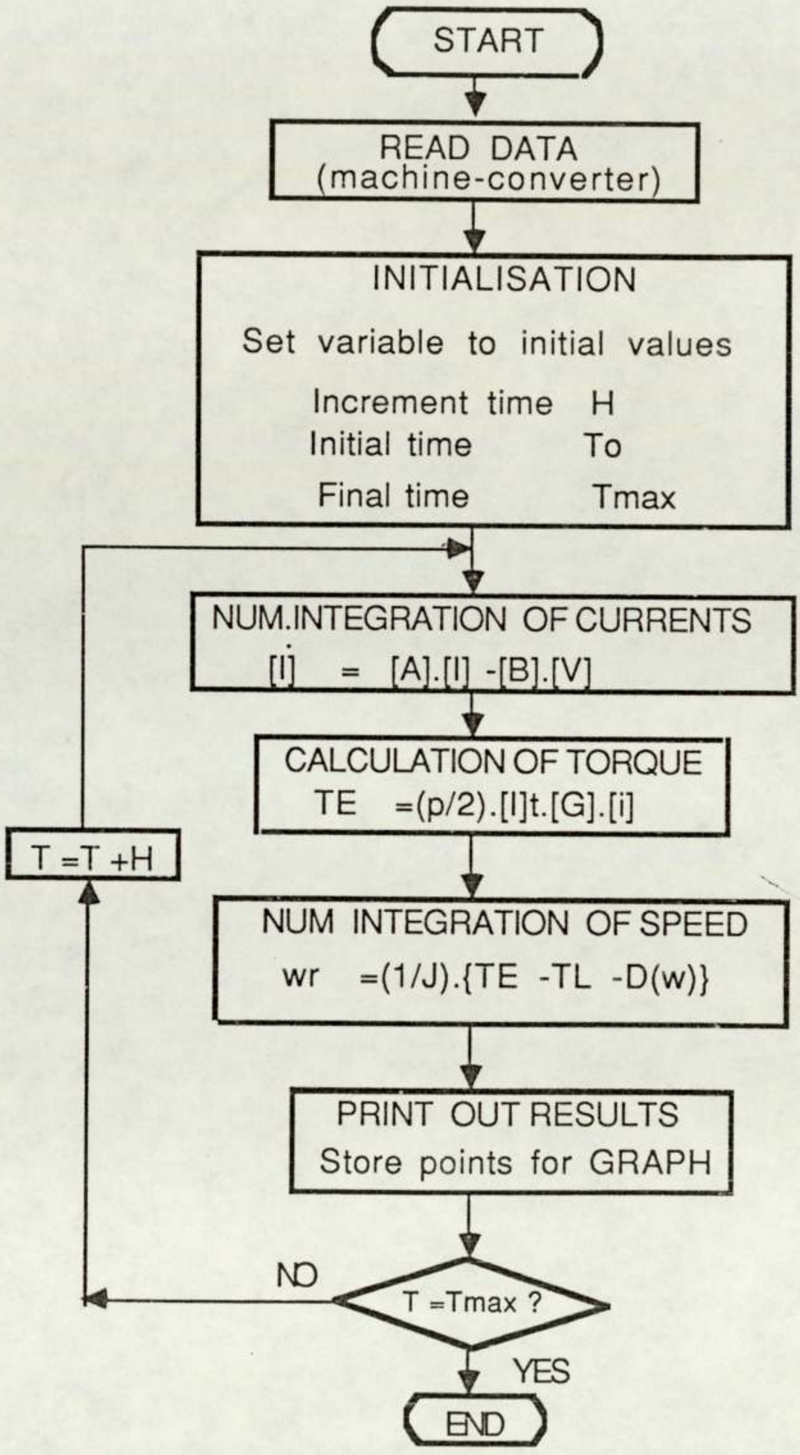


Figure B6 General flow diagram for numerical simulation.

various tasks carried out by this program.

In fact ,the general program was split up into several small sub-programs,for the sake of simplicity and to save computer time.Many NAG routines were used ,to perform all the calculations needed on the matrices (Inverse : F01AAF ,Product : F01CKF ,Sum : F01CDF),and to perform the numerical integration using the RUNGE-KUTTA method.

The NAG routine :D02BBF using the RUNGE-KUTTA-MERSON method ,was used to integrate the set of first-order differential equations ,and was prefered to the routine : D02BAF because it has the advantage of one being able to access to the intermediate values of the computed solution (for example :to print or plot them).

During simulation the program permanently stores all instantaneous values (voltages , currents , torque , speed) in numerical form using output files ,which can be used for graphical purposes.

The start-up phase is illustrated in fig..B.7,which show the transient period.

2.6 <u>Conclusion</u>

This chapter has described the modelling and simulation of an induction motor drive system,and has presented some results

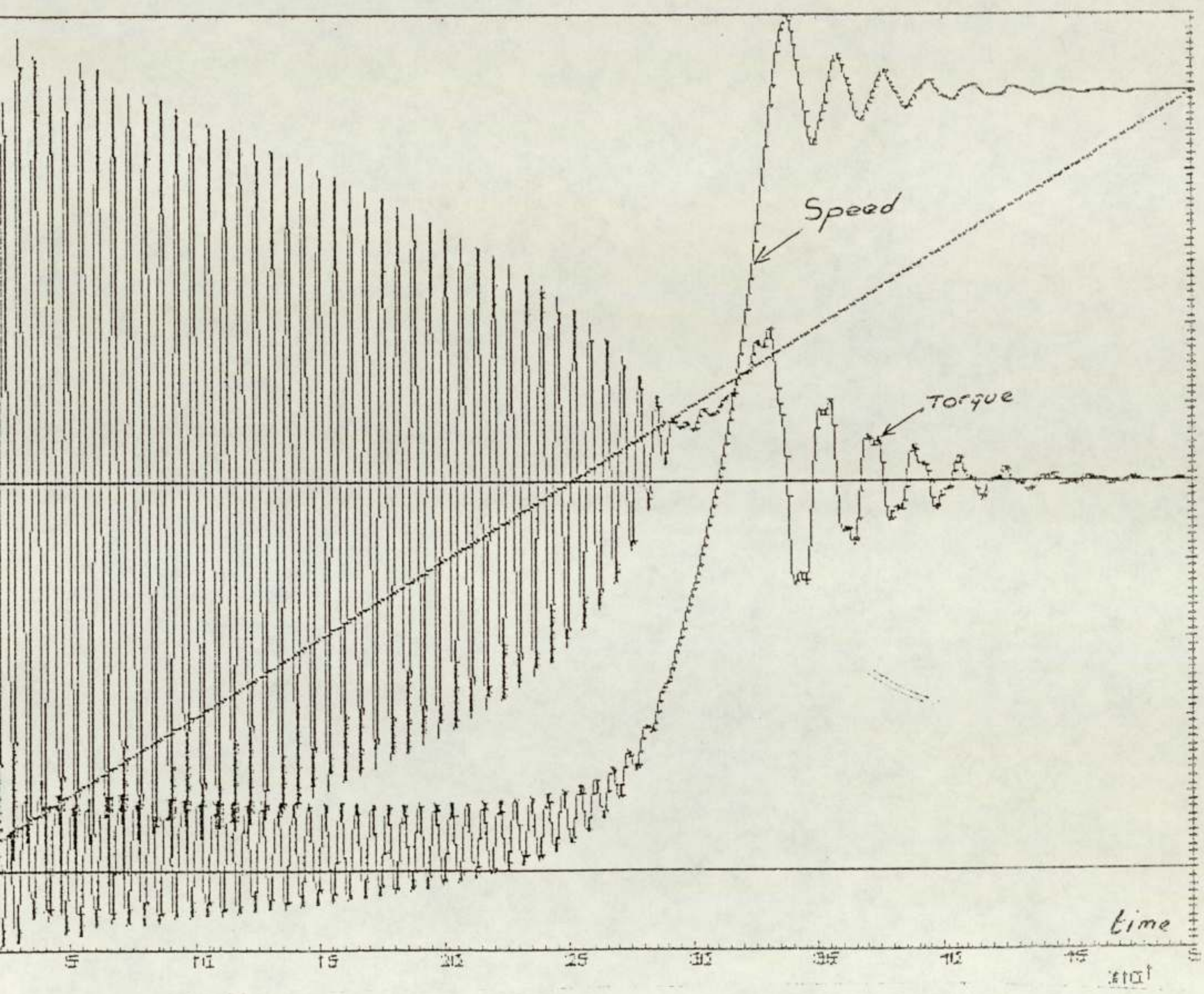


Figure B7 Start-up transient curves (SPEED and TORQUE) of such a simulation.

This study illustrates in a broad way the usefulness of such techniques which are becoming the necessary first step in any design project.

Appendix: A

This appendice contains the details of the vectors and matrices expresse in the general form given in equation (4.4)and (4.5),chapter four.

$$[Vs] = \begin{vmatrix} Va \\ Vb \\ Vc \end{vmatrix} \qquad [Vr] = \begin{vmatrix} 0 \\ 0 \\ 0 \end{vmatrix} \qquad [Is] = \begin{vmatrix} Ias \\ Ibs \\ Ics \end{vmatrix} \qquad [Ir] = \begin{vmatrix} Iar \\ Ibr \\ Icr \end{vmatrix}$$

$$Rs = \begin{vmatrix} Rs & 0 & 0 \\ 0 & Rs & 0 \\ 0 & 0 & Rs \end{vmatrix} \qquad Rr = \begin{vmatrix} Rr & 0 & 0 \\ 0 & Rr & 0 \\ 0 & 0 & Rr \end{vmatrix}$$

$$[Msr]^{t} = [Mrs] \qquad [Ls] = \begin{vmatrix} La & Mab & Mac \\ Mba & Lb & Mbc \\ Mca & Mcb & Lc \end{vmatrix}$$

The mutual inductance between the stator and rotor are the keystone of induction motor performance.The coefficients of mutual inductance vary with rotor position:

$$M(a-A)=\overline{M}\cos\theta + \overline{M}_3\cos 3\theta$$

$$M(b-A)=\overline{M}\cos(\theta+120) + \overline{M}_3\cos 3(\theta+120)$$

$$M(c-A)=\overline{M}\cos(\theta+240) + \overline{M}_3\cos(\theta+240)$$

Appendix C

This appendix contains a summary on axis transformations.

A-Voltage in the stationary frame are expressed as:

$$[Vs]=\begin{vmatrix} Vas \\ Vbs \\ Vcs \end{vmatrix}=\begin{vmatrix} \cos\theta & \sin\theta & 1 \\ \cos(\theta-120) & \sin(\theta-120) & 1 \\ \cos(\theta-240) & \sin(\theta-240) & 1 \end{vmatrix} * \begin{vmatrix} Vqs \\ Vds \\ Vos \end{vmatrix}$$

and the inverse relation is:

$$\begin{vmatrix} Vqs \\ Vds \\ Vos \end{vmatrix}=\sqrt{\frac{2}{3}}\begin{vmatrix} \cos\theta & \cos(\theta-120) & \cos(\theta-240) \\ \sin\theta & \sin(\theta-120) & \sin(\theta-240) \\ 1/2 & 1/2 & 1/2 \end{vmatrix} * \begin{vmatrix} Vas \\ Vbs \\ Vcs \end{vmatrix}$$

Where Vos is the zero sequence component.For a balanced three phase condition,the zero-sequence component does not exist.
It has been considered only to yield the unique transformation relation.
It is convenient to set θ =0,so that the qs axis is coincident with the as axis,and also ignoring the zero-sequence component,the transformati relation can be simplified as:

$$\begin{vmatrix} Vas \\ Vbs \\ Vcs \end{vmatrix} = \begin{vmatrix} 1 & 0 \\ -1/2 & -\sqrt{3/2} \\ -1/2 & \sqrt{3/2} \end{vmatrix} * \begin{vmatrix} Vqs \\ Vds \end{vmatrix}$$

and the inverse relation is

$$\begin{vmatrix} V_{qs} \\ V_{ds} \end{vmatrix} = \begin{vmatrix} 2/3 & -1/3 & -1/3 \\ 0 & -\frac{1}{\sqrt{3}} & \frac{1}{\sqrt{3}} \end{vmatrix} * \begin{vmatrix} V_{as} \\ V_{bs} \\ V_{cs} \end{vmatrix}$$

B- Then the voltage in the stationary ds-qs frame can be converted to the synchronously rotating d-q frame as follows:

$$V_q = V_{qs}.\cos\theta - V_{ds}.\sin\theta$$

$$V_d = V_{qs}.\sin\theta + V_{ds}.\cos\theta$$

These relations can also be inverted to define relations of stationary frame variables in terms of rotating variables as follows:

$$V_{qs} = V_q.\cos\theta + V_d.\sin\theta$$

$$V_{ds} = -V_q.\sin\theta + V_d.\cos\theta$$

References

2.1 TAKEUCHI.J:"Theory of SCR circuits and application to motor control "Elect.Eng Jap,vol86,n10,pp7-11,1966

2.2 SHEPERD .W:"Thyristor control of AC circuit",IEEE,Tran. on power and systems,vol.PAS.90,n2,pp126-131.

2.3 SHEPERD.W and J.STANWAY:"The polyphase induction motor controlled by firing angle adjustment of silicon controlled rectifier".1964,IEEE Int.Con.Rec,vol12,pt4,pp135-154.

2.4 F.J NOLA:"Power factor control system for AC induction motor "IEEE-PAS,Ann.Meeting,1980,pp194-198.

2.5 BOWLER.P:"The speed control of induction motor using static frequency converters".IEE Proc,vol13,1975,pp432-438.

2.6 SHEPHERD.W:"Slip power recovery in an induction motor by the use of a thyristor inverter".IEEE.Tran.on.IA,pp237-246,1977.

2.7 T.A.LIPO:"The analysis of induction motor with voltage control by symmetrically triggered thyristors"IEEE.Tran.on power and systems,vol.PAS.90,n2,515-523.

2.8...CHAUPRADE.R(JeumontSchneider,France),Abondanti.A(Westinghouse,USA)."Variable-Speed-Drives-Modern concepts and Approaches",IEEE-IAS,1982,CH 1682-4-82,pp20-36.

2.9 GOLDHAMMER.B.A:"Modern KRAMER systems"IEEE Tran.on Ind.App.pp657-663,1978.

2.10 DRURY.W,B.L JONES:"Application of controlled fly wheeling to the recovery bridge of a Static KRAMER drive."IEEE Tran.on IA,pp324-327,1974.

2.11 KHALIL.A.Q.:"Capacitve compensation of thyristor controlled slip recovery system."IEEE-Tran.on.IA,pp426-432,1976.

2.12 GRAF.K:"AC pump drives with static converter slip recovery for the LAKE CONSTANCE" Brown-Boveri Review n6,pp96-102,1975.

2.13 R.J.POLGE:"Speed control of a wound rotor induction motor with static inverter in the rotor circuit."IEEE Tran.on.IA,pp238-

243,1975.

2.18..M.MURRAY.W:"The theory and design of cycloconverter" .Cambrige,MA,MIT Press 1972.

2.19 F.C.SCHWARTZ:"A double sided cycloconverter",IEEE.Power Electronic.PESC Record,1979,pp437-47."

2.20...J.M.D Murphy:"Thyristor control of AC motors "New-York Pergamon Press,1972,p118

2.21 BOSE.B,K:"Power electronics and AC drives",Prentice Hall,New Jersey,U.S.A,1986.

2.22 COSSIE.A:"Traction Application ,French Railway Review vol1,n2,June 1983,pp106-115."

2.23 COSSIE.A:"Traction Apllication,Part 2,French Railway Rev.,vol1,n2,June 1983,pp116-126."

2.24 JACOVIDES,J.J:"Cycloconverter-Application in traction drives.IEEE.Trans.on.I.A.vol.IA-17,n4,JUL-AUG-1981,pp127-132."

2.25 PENKOWSKI.L.J,and PRUZINSKY.K.E:"Fundamentals of a P.W.M power circuits."IEEE Conf Rec of the 5th Annual IGA Meeting,pp669-678.1970.

2.27 TURNBULL,F.G:"Selected Harmonic Reduction in DC/AC inverter."IEEE Tran.CE-83,pp374-378,1964

2.28 PATEL.H.S and HOFT.R.F:"Generalised technique of harmonic elimination."IEEE Tran on Ind.Appl,volIA-10,pp 666-673,Sep/Oct 1974.

2.29 JAYACHANDRAN.P,and KOEPSEL:"Harmonic reduction in thyristor inverter."IEEE Tran on IA,pp245-250,1985.

5.1 Mc Murray,W:'Optimum snubber for semiconductor",IEEE-Power semi-conductor Application,vol-1.1972,pp33-41.

5.2 Hiromichi Ohashi,"Snubber circuit for GTO-Thyristor",IEEE Tran on IA,pp655-664,1983.

5.3 Williams.B.M,and P.R.Palmer:'Drive and snubber techniques for G.T.O's,and Power transistors particularly for inverter bridge'IEE.Conf.Rec.Pub,1983.

5.4 Hall.J.K:'The protection of thyristors and diodes in high power circuit forced commutation',IEEE.Trans.Mag.7,pp297-304.

5.5 Lander,C.W:'Power electronics',Mc Graw-Hill Publishing Cie.LTD,1981.

5.7 Brian,R.Pelly:'Power semiconductor devices-A status Review',IEEE/IAS.INT.1982,pp1-19.

5.8 The GTO-Thyristor ,M.E.D.L-POWER Division-Lincoln

5.9 Sakurada.S et Al:'Recent trends of G.T.O',Hitachi Rev.,vol-30,n-4,1981.

5.10 SCR-Manual.Fifth Edition,General Electric,1972.

5.11 Wilson,P.S:"TheG.T.O.and.requirements"Int.macroelectr.Con .Proc,pp89-97,1982.

5.12 Akituru,Ueda:'G.T.O application for traction motor drives', Hitachi.Rev,1982,vol-31,pp185-188.

5.13 Power MOSFET-Siemens-AG.-Review,1983.

5.14 A new high power MOS.-Electronic design,June-7,1979.

5.15 HP MOSFET,Press information,Hewlett Packard,October-24,1984.

5.16 MOSPOWER FET-design catalogue.Siliconix Manual,1982.

5.17 H.Fukui,Y.Matsuda:'Switching characteristics of a G.T.O thyristor'P.C.I,march-1982,Proc,pp268-277.

5.18 Mitsuyku Honbu,Yasuo.M:'G.T.O inverter for large AC motor drives'IEEE.Trans.on IA,vol-IA,n2,1983,pp198-205.

5.19 Yasuo.Matsura:'Development of inverter employing G.T.O's, IEEE.Trans.on IA,vol-IA,n3,1983.

5.20 T.Ogawa:"A high power,low forward drop G.T.O thyristor", Conf.Rec.1978.Ann.Meeting.IEEE.Idus.Applic.Soc,pp1003-1006.

5.21 J.Karvinen:'Three-phase AC traction design and service experience'IEE.Proc.vol-134,PtB,n3,1987,pp