A STATIC SCHERBIUS SYSTEM OF INDUCTION MOTOR SPEED CONTROL

by

GORDON ARTHUR SMITH

B.Sc.,C.Eng.,MIEE.

A thesis submitted to the University of Leicester
in support of an application for the degree of Ph.D.

JUNE 1975

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<u>Contents</u>	<u>Page No.</u>
Memorandum.	i
List of principal symbols.	ii
Abstract.	iv
 <u>1.0. SPEED CONTROL OF INDUCTION MOTORS</u>	 1
1.1. Introduction.	1
1.2. Control of synchronous speed with limited slip.	1
1.3. Control of slip with constant synchronous speed.	4
1.4. Slip energy recovery systems.	5
1.4.1. The Scherbius system.	6
1.4.2. The Kramer system.	6
1.4.3. Subsynchronous static slip recovery systems.	8
1.4.4. Supersynchronous static slip recovery systems.	9
1.5. The cycloconverter slip energy recovery system.	10
 <u>2.0. THE CYCLOCONVERTER</u>	 11
2.1. The three-phase to single-phase cycloconverter.	11
2.1.1. Synchronisation of the thyristor gate pulses.	11
2.2. The three-phase to three-phase cycloconverter.	13
2.2.1. The three-phase fully controlled bridge rectifier.	13
2.2.2. Control requirements of the cycloconverter with neutral connection.	14
2.3. Conclusions	16
 <u>3.0. THE DESIGN OF THE SECONDARY EMF SIGNAL GENERATOR</u>	 18
3.1. General considerations	18
3.2. The pulse-rate generator for the motor shaft.	19
3.3. The pulse-rate generator to represent angular increments of the mains supply cycle.	20
3.4. The counter and decoder.	20

3.4.1.	The up-down counter.	20
3.4.2.	The pulse splitter.	21
3.4.3.	The four phase clock.	22
3.4.4.	The decoder.	23
3.5.	Waveform generation.	23
3.6.	Observations of signal generator operation.	24
4.0.	<u>THE CYCLOCONVERTER-MOTOR SLIP RECOVERY SYSTEM</u>	26
4.1.	Power flow considerations.	26
4.1.1.	Power flow in an induction machine.	26
4.1.2.	Power flow considerations in the cycloconverter.	28
4.2.	The control system.	29
4.3.	Equivalent circuits of the induction motor with a secondary power source.	32
4.3.1.	An equivalent circuit with a voltage source in the secondary.	32
4.3.2.	Effect of the magnitude and phase of the voltage source on the developed torque and the power factor.	33
4.3.3.	An equivalent circuit with a current source in the secondary circuit.	35
4.4.	Consideration of secondary current waveform.	36
4.5.	Summary.	40
5.0.	<u>EXPERIMENTAL RESULTS</u>	41
5.1.	Equipment and instrumentation.	41
5.2.	Standard machine tests.	42
5.3.	Cycloconverter slip recovery tests.	43
5.3.1.	Observation of secondary current waveform.	43
5.3.2.	Open-loop torque-speed curves.	44
5.3.3.	Closed-loop torque-speed curves.	45

5.3.4.	Comparison between secondary resistance and cycloconverter control.	45
5.3.5.	Power factor control.	46
5.3.6.	Cube law power characteristics.	47
5.4.	Summary of results.	48
6.0.	<u>CONCLUSIONS AND SUGGESTIONS FOR FURTHER WORK</u>	49
6.1.	Summary of the static Scherbius system.	49
6.2.	Limitations of the system.	50
6.3.	Application of the system.	51
6.4.	Future work.	52
7.0.	<u>ACKNOWLEDGEMENTS</u>	53
8.0.	<u>REFERENCES</u>	54
9.0.	<u>APPENDIX 1 POWER FLOW IN A SLIP RECOVERY SYSTEM</u>	59
9.1.	Terminology.	59
9.2.	Relationships between torque and power in an induction motor.	59
9.3.	Power flow in the electromechanical slip energy recovery system.	60
9.4.	Power flow in the rotary electric system.	60
9.5.	The frame size rating of the d.c. motor in the electromechanical system.	61
9.6.	Power recovery for loads whose torque varies as the square of speed.	61
10.0.	<u>APPENDIX 2 THE DESIGN OF THE THREE-PHASE TO SINGLE-PHASE CYCLOCONVERTER</u>	63
10.1.	The power circuit.	63

10.1.1. The thyristor power assembly.	63
10.1.2. Voltage transient suppression.	64
10.1.3. Current overload protection.	65
10.1.4. Output rating of the cycloconverter.	67
10.1.5. Thyristor cooling.	68
10.2. The pulse generator.	69
10.2.1. Analysis of the charging circuit.	70
10.2.2. Design considerations.	71
10.3. The pulse amplifier.	76
10.3.1. Operation of the pulse amplifier.	76
10.3.2. Analysis of the blocking oscillator.	77
10.3.3. Pulse amplifier design.	78
10.4. The thyristor gate circuit.	79
10.5. The electronic power supplies.	80
10.6. The control system.	81
10.7. Detailed operation of the control system.	83
10.8. Operation of the cycloconverter.	85
11.0. <u>APPENDIX 3 COMPUTER PROGRAMS FOR MACHINE PERFORMANCE</u>	87
11.1. Computer prediction of machine performance with external secondary resistors.	87
11.2. Computer prediction of machine performance with cycloconverter control.	90
12.0. <u>APPENDIX 4 LOAD EQUIPMENT FOR PERFORMANCE TESTING</u>	95

Memorandum

The accompanying thesis is based on work carried out by the author in the Engineering Laboratories of the University of Leicester between October 1969 and March 1975. All the work recorded in this thesis is original unless otherwise acknowledged in the text or by reference. None of the work has been submitted for another Degree at this University, nor for the award of a Degree or Diploma of any other institution.

The main contributions the author claims to have made to the subject of variable speed control of induction machines using cycloconverters include the following:

1. The operating requirements to enable a cycloconverter to control the flow of power in the secondary circuit of an induction machine have been determined.
2. A novel digital signal generator has been designed and constructed to produce three-phase square waves controlled in phase and synchronised to the secondary induced emf of the induction machine. This generator is necessary for the correct operation of the drive system.
3. A theory has been proposed for prediction of machine performance in which the cycloconverter is represented as a controlled current source in the secondary circuit of the induction machine.
4. A complete three-phase drive system has been designed and successfully operated over a limited range above and below synchronous speed with both braking and overhauling loads.
5. Experimental results have shown smooth control through synchronous speed to be possible.

G.A.SMITH
UNIVERSITY OF LEICESTER
ENGINEERING DEPARTMENT
MARCH 1975

List of principle symbols

P_m	The rated power of the induction motor	watts.
P_r	Power crossing the air gap from primary to secondary	"
P_o	Mechanical power developed at the motor shaft	"
P_{C_1}	Power dissipated in the primary winding	"
P_{C_2}	Power dissipated in the secondary winding	"
P_1	Power input to the stator	
P_2	Electrical power out of the secondary	
T	Torque developed at the motor shaft	N-m
S	Slip in the induction motor	
η	Instantaneous speed of the motor shaft	rev/s
η_s	Synchronous speed of the motor shaft	"
ω	Angular velocity of the motor shaft	rad/s
ω_s	Synchronous angular velocity	"
V_d	Average direct output voltage from a controlled rectifier	volts
V_{d_o}	The value of V_d with zero delay angle	"
α	The delay angle of a controlled rectifier	
e	External voltage source in the secondary circuit of the induction motor	volts
e'	External voltage source referred to the primary side	"
e_p	In phase component of secondary voltage source	"
e_q	Quadrature component of secondary voltage source	"
E_1	Primary emf of the induction motor	"
E_2	Secondary emf of the induction motor	"
I_1	Primary current of the induction motor	amps
I_2	Secondary current of the induction motor	
I_1'	Secondary current referred to the primary side.	

R	Resistance	ohms
X	Reactance	"
N_1	Primary turns per phase	
N_2	Secondary turns per phase	
T_0	Time between alternate polarity half cycles of secondary current	s
T_1	Repetition time of current pulses in the cycloconverter output waveform	s
T_2	Conduction time of current pulses in the secondary circuit	s
T	Time for one quarter period of the secondary cycle	s
I_{rms}	Total rms value of the secondary current	amps
I_p	Peak value of the secondary current waveform	"
I_{fund}	The rms value of the fundamental component of secondary current	amps
f_s	The supply frequency Hz	
β	The turns ratio. N_2/N_1	
θ	The phase angle between secondary current and primary supply voltage.	

Abstract

This work describes a control system in which a cycloconverter is connected between the secondary winding of a three-phase induction machine and the a.c. mains supply to give variable speed sub- and super-synchronous operation.

The operating requirements for a three-phase cycloconverter operating in the secondary circuit of an induction machine have been determined. These considerations show that the cycloconverter control signal must be synchronised to the induced secondary emf of the machine. A novel generator based upon digital circuitry was developed to produce such a control signal.

A theory in which the cycloconverter is represented as a controlled current source in the secondary circuit of the induction machine has been proposed and verified. Computer programs based upon this theory were used to analyse machine performance.

A complete three-phase drive system has been designed and successfully operated over a range above and below synchronous speed with both braking and overhauling loads. Experimental observations have shown that smooth control is possible through synchronous speed.

Chapter 1

Speed Control of Induction Motors

1.1. Introduction

Research into the control of a.c. induction motors has been revitalised during the last two decades by the development of thyristor controlled power sources. Control of induction motors falls into two main categories:

- (a) limited slip operation at a variable synchronous speed, and
- (b) slip control at a constant synchronous speed.

1.2. Control of synchronous speed with limited slip

Most research and development effort has concentrated upon controlling the motor synchronous speed with subsequent operation in a limited slip region. This mode of control permits the use of a machine of the squirrel cage type which, with its basic simplicity, robustness and low manufacturing cost, is so attractive to many industries.

The synchronous speed of an induction motor is proportional to the supply frequency which can be obtained from one of two types of static frequency changer; the d.c. link inverter and the cycloconverter. Many industrial applications such as paper making, printing, extrusion and conveyor lines require a drive capable of delivering full load torque at all speeds. To achieve constant torque without excessive motor current requires a constant magnetic flux level in the machine. The static frequency source must, therefore, be capable of generating a constant voltage to frequency ratio over the operating frequency range. At very low frequencies the voltage frequency ratio may have to be increased to compensate for the stator impedance drop.

The various types of d.c. link inverter are well documented⁽¹⁻⁷⁾. The essential elements of the d.c. link inverter are shown in Fig.1.1a. The a.c. supply is rectified, smoothed and then changed to a variable frequency a.c. output by the operation of a forced commutated thyristor inverter. The output is a function of the thyristor switching sequence and the simplest waveform, a stepped wave or quasi-square wave, is shown in Fig.1.1b. This waveform is generated by allowing each thyristor to conduct for 180° of the output cycle. Although the waveform contains no third or triplen harmonics there are high fifth and seventh harmonics which cause extra heating in the motor. To generate this waveform forced commutation of the thyristors is necessary and many circuits have been devised to achieve this^(8,9). The most efficient circuits require the use of expensive thyristors with fast turn off and high dv/dt capabilities. The amplitude of the output voltage is directly related to the smoothed d.c. supply voltage which can be controlled by a rectifier, series regulator or variable input transformer to give a constant voltage to frequency ratio.

The quasi-square wave output is not suitable for motor operation at low frequencies because of its high harmonic content and saturation effects in the motor. It is adequate, however, for high frequency operation and is used for constant power applications such as machine tool spindle drives at frequencies in excess of 25 Hz⁽¹⁰⁾.

Output voltage control and improved waveform is possible by using pulse width modulation (PWM) techniques in the inverter⁽¹¹⁾. The input to the inverter can be at a fixed voltage and so an uncontrolled diode rectifier may be used. Voltage control is achieved by switching each inverter thyristor "on" and "off" more than once during the inverter cycle. In more refined PWM systems the pulse width is varied throughout the cycle in a sinusoidal manner, Fig.1.2. Good output waveforms are possible at

a) Basic D.C. Link Inverter Circuit

b) Quasi-Square Wave Inverter Output Waveform

FIG. 1.1. D.C. LINK INVERTER

FIG. 1.2. A PULSEWIDTH MODULATED INVERTER OUTPUT WAVEFORM WITH SINUSOIDAL MODULATION OF PULSE WIDTH

low frequencies. At high frequencies, however, commutation energy losses can be considerable unless adaptive ratio modulation⁽¹¹⁾ is used to reduce the number of commutations as the frequency is increased.

The cycloconverter^(12,13) is a direct a.c. to a.c. frequency changer without a d.c. link and uses standard thyristors rather than the expensive inverter grade devices used in the d.c. link inverter. The operation of the cycloconverter may best be understood by considering the three-phase half-wave rectifier connection in Fig.1.3a.

The average direct voltage output is given by

$$V_d = V_{do} \cos \alpha \quad 1.1$$

where α is the delay angle and V_{do} the mean output voltage with zero delay angle. The average voltage V_d can be varied sinusoidally through a complete output cycle by variation of the delay angle. To provide current flow in both directions two similar circuits must be connected in inverse parallel, Fig.1.3b. Voltage and current waveforms with an inductive load are shown for this cycloconverter in Fig.1.3c for conventional circulating current-free operation⁽³⁹⁾. In this mode of operation power output and efficiency are only reasonable with output frequencies from zero to about one-third of the supply frequency depending upon the degree of complexity of the cycloconverter. As the cycloconverter output frequency approaches the supply frequency the harmonic distortion increases because the output waveform consists of fewer part cycles of the supply waveform. Despite this restriction there are applications that particularly suit the cycloconverter drive scheme and these include roller run-out tables in steel mills⁽¹⁴⁾ and control of directly coupled low speed drives as used in the cement industry⁽¹⁵⁾.

Regeneration, i.e. recovery of mechanical energy as electrical energy to the supply system, is often required for those applications with overhauling loads such as hoists, cable pay-off stands, or drives where rapid

a) Three-Phase Half Wave Rectifier

b) Three -Phase to Single Phase Cycloconverter

c) Voltage and Current Waveforms for a Cycloconverter supplying an Inductive Load

FIG. 1.3. THE CYCLOCONVERTER

reversal is necessary as in reversing mills, planing machines, radar scanners etc. Regeneration is inherently possible with the cycloconverter whereas the voltage controlled d.c. link inverter requires an additional controlled rectifier. Recent papers^(16,17) have shown that the current source inverter may be the best solution for an inverter fed system where regeneration is required.

1.3. Control of slip with constant synchronous speed

For constant synchronous speed the primary winding (usually the stator) of the machine is connected to a source of constant frequency.

For squirrel cage machines speed control is possible by using a stator voltage regulator as shown in Fig.1.4a. With closed loop speed control, operation is possible at any speed but the torque per unit of current decreases rapidly as the slip increases, Fig.1.4b. Rating limitations of the motor and of the a.c. regulator restrict this mode of operation to pumps, compressors and ventilation fans where the load torque varies approximately as the square of motor speed. Slip power is dissipated within the machine and the method is inefficient and consequently only suitable for low power applications.

For slipring machines resistors may be added externally to the secondary (usually the rotor) circuit to change the speed torque characteristics as in Fig.1.5. In this case electrical energy proportional to slip, i.e. slip energy, is dissipated as copper loss in the secondary circuit (see Appendix 1 9.2) making the system very inefficient. The effect of adding rotor resistance may be obtained statically by controlling the ratio of on-time to off-time of a thyristor across the output of a rectifier bridge connected to the secondary terminals⁽¹⁸⁾ of the machine as shown in Fig.1.6. Closed loop control is readily possible but once again the method is inefficient as power is dissipated in the rotor circuit.

a) Block Schematic of AC Voltage Regulator

b) Characteristics for Voltage Controlled Induction Motor

FIG. 1-4. AC VOLTAGE REGULATOR CONTROL OF INDUCTION MOTORS

FIG. 1.5. SPEED CONTROL BY VARYING ROTOR RESISTANCE

FIG. 1.6. STATIC CONTROL OF ROTOR RESISTANCE

For the secondary control methods so far described, the electrical power dissipated in the rotor circuit would be considerable, particularly for constant torque applications. In high power drives it is essential that the electrical power in the rotor circuit is not dissipated in the machine and its control equipment but is returned electrically to the supply. Schemes to recover this energy are called 'slip energy recovery systems'.

1.4. Slip energy recovery systems

In slip energy recovery systems electrical power from the secondary (rotor) of a slipring induction motor is recovered either electrically or mechanically after some form of frequency conversion.

For an induction motor to operate and develop a specified torque a certain level of current must be present in the secondary circuit. Normally the induced emf in the secondary circuit, which is proportional to slip, circulates this current through the circuit impedance. If voltage from an external power source is impressed upon, or injected into, the secondary circuit, then the same torque will be developed only if the induced emf changes to maintain the secondary current constant. As the induced emf is proportional to slip the machine will run at a new speed. If the injected voltage is in phase with the induced emf then the motor speed will rise so reducing the induced emf and maintaining the rotor circuit current constant. Similarly an antiphase injected voltage will result in the motor slowing down so increasing the induced emf and maintaining the same current and torque.

It is essential, then, that the external voltage be controlled in magnitude, frequency and phase with respect to the secondary induced emf.

1.4.1. The Scherbius system

In the classical Scherbius⁽¹⁹⁾ system of induction motor speed control a Scherbius machine which is essentially an induction generator coupled to an a.c. commutator motor, Fig.1.7, is used as a source of low frequency injected voltage to the induction motor. The injected voltage and, therefore, the speed and power factor of the motor is controlled by the excitation of the Scherbius machine. The excitation current must have the same frequency and phase sequence as the secondary voltage of the machine which it is to regulate. Series excitation by a winding carrying the main current in the motor secondary ensures that the induced voltage is in phase with and proportional to the secondary current. In this case the Scherbius machine is equivalent to a resistance but the energy is returned to the supply by the induction generator.

Shunt excitation from a transformer connected to the terminals of the Scherbius machine is simple but gives only subsynchronous operation due to the inability of the transformer to provide any excitation at zero frequency. If, however, additional separate excitation is provided from, say, a rotary frequency changer coupled to the main motor, then subsynchronous and supersynchronous operation is possible. Scherbius equipment with all three methods of excitation is shown in Fig.1.8.

1.4.2. The Kramer system

In 1908 Kramer⁽²⁰⁾ proposed a system of speed control for an induction motor in which variable frequency rotor emf is converted to a direct voltage by a rotary converter. The rectified power is then used to supply a d.c. motor. The d.c. motor may be mechanically coupled to the induction motor whereby electrical energy is converted to mechanical energy which supplements the mechanical output of the induction motor. Alternatively, the d.c. motor may drive a separate induction generator thus

FIG.1.7. CONNECTIONS OF AN INDUCTION MOTOR CONTROLLED BY A SCHERBIUS MACHINE

FIG.18. SCHERBIUS SYSTEM WITH METHODS OF EXCITATION

returning the slip power electrically to the mains supply.

In the 'modified' Kramer system the rotating frequency changer has been replaced by a semiconductor rectifier, thus reducing the number of rotating machines and raising the efficiency. Expressions for the power flow in the electromechanical and rotary-electric recovery systems are developed in Appendix 1, 9.3 and 9.4, and are shown in Fig.1.9. Closed loop operation is achieved by amplifier control of the d.c. machine excitation.

The speed range determines the rating of the recovery equipment. Appendix 1, 9.2, shows that for a constant torque application the maximum power that can be recovered is $S.P_m$ where P_m is the rated power of the induction motor and S the per unit slip. In the electromechanical system the d.c. motor must recover this power at the minimum speed of the induction motor and yet be capable of rotating at maximum speed. In this case Appendix 1, 9.5, shows that the effective frame size rating of the d.c. motor must be $S.P_m/(1 - S)$. Table 1 in Fig.1.10 shows a comparison of the d.c. motor rating for the two systems with a constant torque load. The electromechanical system is usually limited to centrifugal fan or pump loads where the speed range is small and the power varies as the cube of the speed. Appendix 1, 9.6, shows that in this case the maximum slip power is about 15% of the rated power of the induction motor and occurs at two-thirds of synchronous speed. Typical efficiencies over a range of speeds for the rotary-electric Kramer system are given in Table 2 of Fig.1.11.

Applications for Kramer systems include centrifugal pumps and fans for mine ventilation. The powers are usually very high and operation is often continuous, requiring a limited speed range with high overall efficiency and the minimum of maintenance. Table 2 in Fig.1.11 shows that the slip recovery d.c. motor and the induction generator make a

a) Electro Mechanical Kramer System

b) Rotary -Electric Kramer System

FIG. 1.9. KRAMER SLIP ENERGY SCHEMES

Table 1

Slip S		0.25	0.5	0.6
D.C. motor rating as a % of the maximum output of the induction motor	Electro-mechanical system	33	100	150
	Rotary-electric system	25	50	60

FIG. 1.10. COMPARISON OF D.C. MOTOR FRAME SIZE FOR THE ELECTRO - MECHANICAL AND ROTARY ELECTRIC KRAMER SYSTEMS

Table 2

Speed	r p m	740	600	450
Induction motor output	%	92.0	89.0	80.0
Induction motor loss	%	5.15	6.35	9.0
Rectifier loss	%	0.25	0.25	0.4
D.C. motor loss	%	1.6	2.6	5.6
Induction generator loss	%	1.0	1.8	5.0
Total loss	%	8.0	11.0	20.0

Losses are expressed as a % of input power

FIG. 1.11. TYPICAL ROTARY ELECTRIC KRAMER PERFORMANCE FOR A HIGH POWER PUMPING APPLICATION

FIG. 1.12. SUB SYNCHRONOUS STATIC SLIP ENERGY RECOVERY SYSTEM

substantial contribution to the system losses and being rotating plant require regular maintenance.

1.4.3. Subsynchronous static slip recovery systems

Recent developments in the Kramer system have replaced the rotating slip recovery equipment by a thyristor bridge operating in an inverting mode⁽²²⁻²⁴⁾. In this system, shown in Fig.1.12, the inverter returns the rectified slip power to the a.c. supply. The rectifier and the inverter are naturally commutated by the alternating emfs at the secondary output terminals and the supply bus-bars. Speed variation is obtained by controlling the inverter delay angle but only subsynchronous operation is possible. Where a limited range of slip is required the supply voltage to the inverter terminals is reduced by a step down transformer so that the most economical inverter can be used operating with the best possible power factor. This minimises the rating of the rectifier to $S.P_m$ for constant torque loads (see Appendix 1, 9.2). A disadvantage of the system is that to ensure satisfactory operation of the inverter a choke is necessary in the d.c. link between the rectifier and the inverter to give continuous current flow.

The electrical machine can operate satisfactorily with a current overload for a time well in excess of the capability of the power semi-conductors. It is necessary, therefore, to operate the inverter in a closed loop control system, Fig.1.13. The inverter delay angle is automatically changed to control the current to a level demanded by the output of the speed servo amplifier, which must have a restricted maximum value.

Satisfactory operation of such systems has been achieved⁽²⁵⁾, but commutation failure may short-circuit the inverter if a momentary mains failure occurs. Research work is continuing to incorporate a system of forced commutation in the event of a supply failure which should overcome this problem.

FIG. 1.13. CONTROL SYSTEM FOR A SUBSYNCHRONOUS SLIP RECOVERY SYSTEM

FIG. 1.14. SUB AND SUPER - SYNCHRONOUS STATIC SLIP ENERGY RECOVERY SYSTEM WITH FORCED COMMUTATION

1.4.4. Supersynchronous static slip recovery systems

Supersynchronous operation of the previously described static slip recovery system is possible if the diode rectifier is replaced by a controlled rectifier operating in an inverting mode as in Fig.1.14. The main difficulty with this system is that near synchronism the slip frequency emf's in the motor secondary are insufficient for natural commutation of the inverter⁽²⁶⁾. Forced commutation circuits are therefore necessary for operation through and near synchronism and these commutation circuits must be controlled in accordance with the secondary induced voltage of the motor⁽²⁷⁾.

At synchronous speed, the secondary emf is zero and a means of producing special control instructions to the commutation circuits must be devised. Ohno and Akamatsu⁽²⁷⁾ developed their control signal from a small slipping motor coupled to the induction motor and excited by a 10 KHz high frequency modulated by a three phase 50 Hz supply. Suitable demodulation of the secondary voltage gave a three-phase voltage of constant amplitude at slip frequency. With a system of this type subsynchronous and supersynchronous operation is possible in both braking and motoring modes.

A similar technique was tried by Erlicki⁽²⁸⁾ who used the secondary terminal voltage to vary the frequency of the inverter output. Erlicki arranged for the inverter voltage to be greater than the secondary induced voltage at standstill and the motor revolved against the rotating field with an inverter frequency greater than that of the supply to the primary. The problem of the disappearance of the control signal near synchronism was not, therefore, encountered but it should be realised that when current flows the secondary terminal voltage is not the induced emf due to the winding impedance drop.

1.5. The cycloconverter slip energy recovery system

The subsynchronous and supersynchronous slip energy recovery system as described by Ohno and Akamatsu, although capable of operation over a wide speed range, requires a forced commutated inverter and a naturally commutated inverter. Both inverters present operational problems of reliable commutation. The same function can be fulfilled by a cycloconverter which is inherently more reliable.

The work described in this thesis considers a static Scherbius system of induction motor control in which a cycloconverter frequency changer is used to control the slip energy in the secondary circuit of a slipring induction motor.

For pumping and ventilating applications considerable mass flow change can be achieved with a limited speed range above and below synchronous speed. The secondary emf would be at a low slip frequency and would, therefore, be suitable for connection through a cycloconverter to the mains supply. The cycloconverter with its capability of controlling power flow both out of, and into, the mains supply would make it possible to operate the slipring induction motor both above and below synchronous speed. To control the cycloconverter through the speed range would require a secondary emf signal generator that, if correctly designed, would make possible the control of motor torque and power factor.

The basic circuit requirements of the proposed system are shown in Fig.1.15. The research work includes the design of a suitable cycloconverter and secondary emf signal generator and the prediction and observation of system performance.

FIG.1.15. A CYCLOCONVERTER SLIP RECOVERY SYSTEM

Chapter 2

The Cycloconverter

2.1. The three-phase to single-phase cycloconverter

The three-phase to single-phase cycloconverter in Fig.2.1 is a very simple connection requiring the conduction of only one thyristor at any instant in time. The operational requirements of this connection have been considered in detail by many authors^(12,13,29). The positive half cycle of output to the load is obtained by firing the 'p' group thyristors at controlled firing angles. The current flows from the supply through the cycloconverter and load and returns to the supply by the neutral link. Similar control of the 'n' group thyristors generates the negative half cycle of output. The main operational problem is the prevention of a short circuit due to simultaneous operation of groups 'p' and 'n' thyristors. A circulating current reactor may be used to limit this current or, alternatively, electronic zero current detection may be used to prevent simultaneous operation of the two thyristor groups⁽³⁰⁾.

For slip energy recovery systems the load not only constitutes an impedance but also a voltage source which is the induced secondary emf of the induction motor. The cycloconverter must control current out of, and into, this voltage source in order to obtain both supersynchronous and subsynchronous operation of the motor (see Chapter 4).

2.1.1. Synchronisation of the thyristor gate pulses

The induced secondary emf superimposed on the supply waveform is shown in Fig.2.2. To control the current during regeneration, when power flows from the load to the supply, it is necessary to retard firing beyond the 0° point. Thyristor gate circuits are normally synchronised by a supply waveform in phase with the voltage across the thyristor being

FIG. 2.1. THREE PHASE TO SINGLE PHASE CYCLOCONVERTER

a) Energy Flow Supply to Load

'n' group operating during the load emf positive half cycle

b) Energy Flow Load to Supply

FIG. 2.2. ALTERNATIVE CONTROL MODES WITH RESISTIVE LOAD INCLUDING AN EMF SOURCE

controlled, but it is evident from Fig.2.2b that the resulting 0° to 180° control range would not give controlled regeneration. For full control in the regenerative region a control range of -60° to $+120^\circ$ must be used and a suitable synchronisation sequence for Figs.2.1 and 2.2 is given in Table 3.

TABLE 3

Synchronisation sequence for -60° to $+120^\circ$ control

when waveform	B-N	is negative thyristor	R_p	can be fired
"	R-N	negative	Y_p	"
"	Y-N	negative	B_p	"
"	B-N	is positive	R_n	"
"	R-N	positive	Y_n	"
"	Y-N	positive	B_n	"

Fig.2.2b shows that for energy flow from the load to the supply, (hereafter called positive energy flow) the 'n' group thyristors are fired during the positive half cycle of secondary induced emf and the 'p' group during the negative half cycle. For energy flow from the supply to the load (hereafter called negative energy flow) the thyristor groups being fired for a given polarity of secondary emf are interchanged.

The requirements for control logic are stated in Table 4.

TABLE 4

Requirements for control logic

Polarity of induced emf	Required direction of power flow	Thyristor Group to be controlled
positive	positive	n
negative	positive	p
positive	negative	p
negative	negative	n

2.2. The three-phase to three-phase cycloconverter

For induction motor control a three-phase to three-phase cycloconverter is required. There are many alternative arrangements of cycloconverter circuits having varying degrees of complexity. For high power applications many thyristors must be used and they are operated in complex configurations giving a high number of effective phases and thereby minimising harmonics.

In this instance, however, the motor to be controlled was quite small and, in addition, economy of equipment was essential. It was decided to use the three-pulse star connected cycloconverter as shown in Fig.2.3. This circuit uses three identical, four-quadrant, three-phase to single-phase cycloconverters as described earlier.

For operation with a neutral connection the controls are quite independent as only one thyristor at any time need be fired, the current return path being via the neutral connection. This, however, would present a problem for slip energy recovery from the rotor of an induction motor as normally the rotor neutral is not available.

Consideration must be given to operation without a neutral connection. To ensure a continuous current path thyristors in more than one single-phase cycloconverter must be fired at the same instant. Thus 'p' group thyristors in cycloconverter 1 would have to be fired at the same instant as 'n' group thyristors in cycloconverter 2, for the (A-B) positive half cycle.

2.2.1. The three-phase fully controlled bridge rectifier

The problem of the simultaneous conduction of more than one thyristor as discussed in the previous section can be considered in more detail by reference to the bridge rectifier shown in Fig.2.4.

The waveforms for the phase control of a fully controlled bridge with a resistive load are shown in Fig.2.5a. The pulse generators can be

Fig 2.3 THREE PULSE STAR CONNECTED CYCLOCONVERTER

Fig 2.4 THREE PHASE FULLY CONTROLLED BRIDGE RECTIFIER

synchronised for the required 0° to 120° operation by the synchronising or timing, waveforms as shown. To complete a current path two thyristors must receive gate pulses at the same instant, Fig.2.5b. This is achieved by arranging for two pulse amplifiers to receive an input from a single pulse generator. The interconnection of all pulse amplifiers and generators is termed a firing 'ring' for which two alternative arrangements are shown in Fig.2.5c. The diode gates allow the output of each pulse generator to be amplified and thereby control two thyristors.

With an inductive load including a source of emf E , the timing requirements are different to those for a resistive load when controlled regeneration is to be used. The control range must be from -60° to $+120^\circ$ as described in Section 2.1.1. and output waveforms with suitable synchronisation, pulse generation and firing-ring alternatives are shown in Fig.2.6. The load has been considered highly inductive so that the d.c. output current in Fig.2.6 is continuous and constant. Regenerative energy flow from the load to the supply is shown as the shaded areas in Fig.2.6b.

2.2.2. Control requirements of the cycloconverter with neutral connection

The fully controlled three-phase bridge rectifier has illustrated how the need to fire two thyristors simultaneously can be met. A similar requirement for the cycloconverter of Fig.2.3 without the neutral connection is shown in Fig.2.7. This shows the conduction of each thyristor to generate one output line to line voltage waveform across a resistive load. The thyristors in cycloconverters 1 and 2 are shown to require simultaneous gate pulses in a prescribed sequence.

The generation of three-phase displaced output sine waves requires even more complex firing circuitry. For example, the $1p$ group thyristors in Fig.2.3 must be simultaneously fired with the $2n$ group thyristors during

Fig. 2.5 SYNCHRONISING AND GATE PULSE SCHEME FOR FULLY CONTROLLED BRIDGE RECTIFIER WITH RESISTIVE LOAD.

thyristor being fired	synchronising waveform scheme A	gate	pulses	synchronising waveform scheme B
Rp	R - B			Y - B
Yp	Y - R			B - R
Bp	B - Y			R - Y
Rn	B - R			B - Y
Yn	R - Y			R - B
Bn	Y - B			Y - R

FIG. 2.6. SYNCHRONISING AND GATE PULSE SCHEMES FOR A FULLY CONTROLLED BRIDGE RECTIFIER WITH INDUCTIVE LOAD AND EMF SOURCE

Output waveform (A-B) Fig.(2.3)

note-R2n means 'n' group thyristor on Red line of cycloconverter '2' (see Fig. 2.3)

Fig 2.7 GENERATION OF ONE LINE-LINE OUTPUT WAVEFORM

the (A-B) positive half cycle and with the 3n group thyristors during the (A-C) positive half cycle.

To generate the output waveforms of Fig.2.8b for the cycloconverter connection of Fig.2.8a a reference frequency generator must be used to control the thyristor firing instants. These reference waveforms would permit gate pulses to be generated to the thyristors during the periods shown in Fig.2.8b. It should be noted that each thyristor group would be controlled from two reference waveforms i.e. group 1p thyristors would be controlled during the positive half-cycle of (A-B) and during the negative half-cycle of (C-A).

This would give the periods shown in Fig.2.8c when both the 'p' group thyristors and 'n' group thyristors of any one cycloconverter would receive gate pulses resulting in a short circuit.

To prevent the short circuit condition, the pulse generation period would have to be restricted electronically according to a pre-determined inhibit sequence. This further restriction on gate pulse generation is illustrated in Figs.2.9,10,11). For example, in Fig.2.9 1p group thyristors are not permitted to operate during the positive half of the (C-A) waveforms. This and similar restrictions to the other groups prevents pulse generation for the first 60° and eliminates the short circuit condition. Three possible inhibit sequences are tabulated in Fig.2.12.

There are, then, two operating requirements for operation without a neutral connection.

- (i) Each thyristor group must operate from two ring of six pulse generation circuits in order that they can be simultaneously fired with either of two other thyristor groups. For example in Fig.2.9 the 1p group must operate for 60° with the 2n group and for 60° with the 3n group.

(a) Power circuit

(b) Thyristors conducting to generate line-line waveforms

[A-B] 1p-2n 1p-2n 1p-2n 1n-2p 1n-2p 1n-2p 1p-2n 1p-2n 1p-2n 1n-2p 1n-2p 1n-4p 1p-2n 1p-2n 1p-2n
 [B-C] 3p 1n 3n 1p 3a 1p 3n 1p 3p 1n 3p 1n 3p 1n 3n 1p 3n 1p 3p 1n 3p 1n 3p 1n 3n 1p 3n 1p
 [C-A] 2n 3p 2n 3p 2p 3n 2p 3n 2p 3n 2p 3n 2p 3n 2p 3n 2p 3n 2p 3n 2p 3n 2p 3n 2p 3n 2p 3n

● Shaded areas show simultaneous operation of 'P' and 'n' group thyristors that would cause cycloconverter short circuit.

(c) Thyristor 'ON' periods

Fig. 2.8 GENERATION OF 3 PHASE OUTPUT WAVEFORM

Fig. 2.9 Inhibit sequence 1 for cycloconverter operation

Fig. 2.10 Inhibit sequence 2 for cycloconverter operation

Fig. 2-11 INHIBIT SEQUENCE 3 FOR CYCLOCONVERTER OPERATION

output WAVEFORM	Thyristors	A-B	A-B	B-C	B-C	C-A	C-A	A-N	A-N	B-N	B-N	C-N	C-N
		+	-	+	-	+	-	+	-	+	-	+	-
inhibit sequence 1	1p					off	on						
	1n					on							
	2p		on										
	2n	on											
	3p			on	on								
	3n			on									
inhibit sequence 2	1p	on											
	1n		on										
	2p			on									
	2n				on								
	3p					on							
	3n						on						
inhibit sequence 3	1p											on	on
	1n											on	
	2p							on	on				
	2n							on		on			
	3p										on		
	3n									on			

shaded area represents period when output waveform must prevent thyristor operation

Fig 2-12 THREE POSSIBLE INHIBIT SEQUENCES FOR THREE PHASE CYCLOCONVERTER OPERATION

- (ii) An inhibit sequence must be used controlled by the output frequency reference waveforms to prevent simultaneous firing of 'p' and 'n' groups of one cycloconverter.

Firing 'ring' connections for the pulse generators and amplifiers can be determined for the power circuit of Fig.2.8a and are tabulated in Fig.2.13.

The considerable complexity of operation in this manner can be seen by reference to Fig.2.14 which shows a thyristor gate pulse scheme for the ring of six alternative B in Fig.2.13 and the inhibit sequence 1 shown in Fig.2.12. In Fig.2.14 there are 18 gate amplifiers - one for each thyristor - and six rings of six pulse generators.

The gate pulse control circuits are further complicated when inductive loads are considered. The inhibit sequence is in itself not sufficient to prevent a short circuit condition as the incoming thyristor group may receive gate pulses before the current in the outgoing group has fallen below the minimum holding current level of the thyristors. Current detection would be required as described in Appendix 2 to further limit the generation of gate pulses.

The operating safety margin for short-circuit prevention can be increased by further restriction of the range of control of any one thyristor group. The resultant waveform for a group conduction period of 120° and the resulting 60° safety angle between the on periods of the 'p' and 'n' group thyristors is shown in Fig.2.15.

2.3. Conclusions

To operate a three-phase to three-phase star-connected cycloconverter without a neutral connection greatly increases the complexity of the gate pulse circuits compared with operation with a neutral connection. Few authors have considered this problem. Langer (45) describes a

SUPPLY	OUTPUT WAVEFORM (A-B) + ve		OUTPUT WAVEFORM (A-B) - ve	
WAVE	THYRISTORS	FIRING RING	THYRISTORS	FIRING RING
R-Y	R1p Y2n	(R-B) → Y2n	R2p Y1n	(R-B) → Y1n
R-B	R1p B2n	(Y-B) → R1p	R2p B1n	(Y-B) → R2p
Y-B	Y1p B2n	(Y-R) → B2n	Y2p B1n	(Y-R) → B1n
Y-R	Y1p R2n	(B-R) → Y1p	Y2p R1n	(B-R) → Y2p
B-R	B1p R2n	(B-Y) → R2n	B2p R1n	(B-Y) → R1n
B-Y	B1p Y2n	(R-Y) → B1p	B2p Y1n	(R-Y) → B2p
		L		M
	WAVEFORM (B-C) + ve		WAVEFORM (B-C) - ve	
R-Y	R2p Y3n	(R-B) → Y3n	R3p Y2n	(R-B) → Y2n
R-B	R2p B3n	(Y-B) → R2p	R3p B2n	(R-B) → R3p
Y-B	Y2p B3n	(Y-R) → B3n	Y3p B2n	(Y-R) → B2n
Y-R	Y2p R3n	(B-R) → Y2p	Y3p R2n	(B-R) → Y3p
B-R	B2p R3n	(B-Y) → R3n	B3p R2n	(B-Y) → R2n
B-Y	B2p Y3n	(R-Y) → B2p	B3p Y2n	(R-Y) → B3p
		N		O
	WAVEFORM (C-A) + ve		WAVEFORM (C-A) - ve	
R-Y	R3p Y1n	(R-B) → Y1n	R1p Y3n	(R-B) → Y3n
R-B	R3p B1n	(Y-B) → R3p	R1p B3n	(Y-B) → R1p
Y-B	Y3p B1n	(Y-R) → B1n	Y1p B3n	(Y-R) → B3n
Y-R	Y3p R1n	(B-R) → Y3p	Y1p R3n	(B-R) → Y1p
B-R	B3p R1n	(B-Y) → R1n	B1p R3n	(B-Y) → R3n
B-Y	B3p Y1n	(R-Y) → B3p	B1p Y3n	(R-Y) → B1p
		P		Q

FIG.2.13. FIRING RING OF SIX GATE PULSE AMPLIFIERS

FIG.2-14 THYRISTOR GATE CONTROL AND TIMING ARRANGEMENT FOR A THREE-PHASE TO THREE-PHASE CYCLOCONVERTER WITH NO NEUTRAL CONNECTION AND DISCONTINUOUS LOAD CURRENT.

Fig. 2.15 CYCLOCONVERTER WAVEFORM WITH GROUP CONDUCTION RESTRICTED TO 120°

cycloconverter which is started with a neutral connection subsequently switched out once current has built up in the load. Then, although only one thyristor at any time is fired, there is always another thyristor already conducting to complete the current path.

To maintain simplicity and economy of design, it was decided to operate the cycloconverter with a neutral connection. A consequence of this decision was that with a delta connected rotor on the machine to be used the machine would have to be connected with a direct primary feed to the rotor and the slip energy recovery cycloconverter connected between the star connected stator and the 50 Hz supply.

The cycloconverter would, therefore, be as shown in Fig.2.3 and the detailed design of one three-phase to single-phase unit is given in Appendix 2.

Chapter 3

The design of the secondary emf signal generator

3.1. General considerations

The cycloconverter described in Chapter 2 can be used to control the flow of energy between the induced emf in the motor secondary and the mains supply. The direction of energy flow is determined by whether the 'n' group or the 'p' group thyristors are allowed to conduct during each alternate half cycle of the secondary induced emf. It is essential, therefore, to know whether at any instant the induced emf is positive or negative.

When the motor being controlled is operating with a large slip, the induced emf is considerable and the secondary terminal voltage might be monitored to determine the emf polarity. This technique has limitations:- (i) the terminal voltage is only equal to the induced emf at current zero, and (ii) when the cycloconverter thyristors are conducting the terminal voltage becomes the a.c. supply voltage. Erliki⁽¹⁸⁾ claims success with this method when the machine is operating as a motor with slip values > 1 , i.e. the motor is turning in the opposite direction to the synchronous field, with energy flow into the secondary circuit from an inverter.

The monitoring of the secondary terminal voltage is not practical for motor operation near synchronous speed as the induced emf is small with considerable superimposed slot noise, and it finally becomes zero at synchronous speed. An alternative technique must be developed and as described in Section 1.3.4 Ohno and Akamatsu used a small slip-ring machine to generate a control signal voltage of constant amplitude but in phase with the secondary induced emf.

The method used by Ohno and Akamatsu was restricting in that there was no easy way to change the phase angle between their signal generator

output and the induced secondary emf. This was considered to be important as a means of controlling the motor power factor.

It was decided to develop a secondary induced emf signal generator using a digital technique that would readily permit phase angle control. The generator was developed on the basis that the slip frequency is equal to the difference between the supply frequency and a frequency proportional to motor speed.

The basic elements of the design are as follows:

- (i) A generator is to provide a predetermined number of pulses during each cycle of the sinusoidal supply voltage.
- (ii) A generator is to provide a predetermined number of pulses during each revolution of the induction motor shaft.
- (iii) A counter and decoder is to compare the outputs of the supply and the motor pulse generators and to produce discrete outputs representing angular increments of the slip frequency cycle. A means is to be provided for synchronisation to the supply voltage at a controllable phase angle.
- (iv) An output circuit is to generate three slip frequency constant amplitude square waves displaced by 120° .

3.2. The pulse rate generator for the motor shaft

A pulse rate proportional to motor speed was achieved by optical scanning of a perforated disc mounted on the shaft of the machine being controlled. An optical module type TIL 318 was used which consisted of a Gallium-Arsenide light emitting diode and a phototransistor encapsulated in a pre-focused assembly. The module is T.T.L. compatible and gives a logic 1 output when there is an uninterrupted path between the light source (diode) and the sensor (phototransistor). At slow speeds, a slow moving

disc perforation will cause a gradual change at the sensor output so a Schmitt trigger was used for additional pulse shaping.

The number of perforations in the disc required careful consideration. A compromise solution was required between using a large number of holes to give information at small angular increments and the need to keep the disc size within reasonable proportions. The number of holes and thereby the counts per revolution of the shaft would also be limited by the counter and decoder.

It was decided to use a counter and decoder capacity of 64 counts per cycle of supply frequency. If the disc was mounted on a machine having a synchronous speed of 1500 rpm then one cycle of supply frequency would correspond to half a revolution of the shaft. The number of perforations required was, therefore, 128 with a pulse being generated from the optical module at, nominally, every 5.6° of the supply voltage cycle.

3.3. The pulse rate generator to represent angular movements of the mains supply cycle

Comparison with the pulse rate proportional to motor speed requires exactly 64 counts for each cycle of mains frequency. An oscillator, gated by the mains supply, would produce a number of pulses per cycle which would vary with changes in supply frequency. All possible static techniques considered were complex and it was decided to adopt the simple solution of fitting a disc and optical detector to a small synchronous motor. Thus the number of counts per cycle will be constant irrespective of changes in supply frequency.

3.4. The counter and decoder

3.4.1. The up-down counter

Comparison of the mains supply and motor frequencies generated by the

perforated discs was achieved by using T.T.L. up-down counters of type 74193.

Two SN 74193 up-down binary counters were cascaded as shown in Fig.3.1 which with six output lines would give a total count of 64. With the pulse rate proportional to mains frequency f_m connected to the 'up' clock input and the pulse rate proportional to motor speed f_r to the 'down' clock input, the output count would: (i) increase for subsynchronous motor operation, (ii) decrease for supersynchronous operation and (iii) remain steady at exactly synchronous speed. In addition a phase advance pulse signal at the 'up' clock input and a phase retard pulse signal at the 'down' clock input would be required for phase control.

The SN 74193 counter accepts count-down and count-up inputs and the direction of counting is determined by which count input is pulsed when the other count input is high. A count will be missed if both inputs are pulsed at the same time and this would occur if the mains supply and motor speed pulses and the phase control pulses were directly connected to the clock inputs. To prevent this, the signal inputs were modified in a pulse splitter circuit.

3.4.2. The pulse splitter

The pulse splitter consists of two D-type flip-flops as shown in Fig.3.2 which transfer the information on the D input to the Q output on the rising edge of each clock pulse. The output then remains unchanged until a 'clear' input or a different D input, together with a clock pulse is received.

If, as in Fig.3.2, D_1 is high when a clock pulse arrives, a rising edge at Q_1 is fed to clock 2 which, with D_2 high, gives a zero output from $\overline{Q}_2$ until the end of the clock pulse. Then Clear 2 reverts to the low state so changing $\overline{Q}_2$ back to logic 1. If another clock pulse arrives while D_1 is still high no change occurs at the output $\overline{Q}_2$ as clock 2 needs

6 binary output lines counting 64

Fig 3.1 UP - DOWN COUNTER

(a) Schematic

(b) Logic voltages

Fig 3.2 THE PULSE SPLITTER

a low to high transition to trigger the flip-flop. When D_1 reverts to the low state an input clock pulse will reset Q_1 to zero leaving the output $\overline{Q_2}$ still unchanged. The cycle is repeated when D_1 again becomes high.

Fig.3.2 shows that an output pulse is produced for every input pulse but only for the duration of the clock pulse ϕ_1 .

If identical pulse splitting circuits are used for the other inputs to the up-down counter then providing their clock pulses do not occur at the same time the output pulses to be taken to the up-down counter will never be coincident and a miscount will not occur.

3.4.3. The four-phase clock

To ensure that the pulse splitter circuits used in each of the four signal inputs to the up-down counter are never clocked at the same time two four bit shift registers were used as in Fig.3.3 to produce a four-phase clock.

Input data can be parallel loaded when the mode input is high and this data appears at the outputs after a high to low transition of the clock 2 input. The counter will then shift each bit of information to the left on the receipt of a clock pulse provided the mode input is low.

When first switched on the Schmitt trigger NAND gate 1 output is high to the mode control input and the binary number 10000000 is repeatedly put into the shift register at each clock pulse generated by the master clock Schmitt trigger 2. When capacitor C charges to the trigger level of Schmitt trigger 1 the input to the mode control goes low and the logic 1 at input A starts shifting round the counter at each pulse of the master clock. Alternate outputs as shown in Fig.3.3 are then used as clock inputs to the pulse splitter circuits and as only one clock pulse occurs at any time, correct operation of the up-down counter is ensured as shown in Fig.3.4.

(a) Schematic

(b) Logic voltages

Fig 3.3 THE FOUR PHASE CLOCK

Fig 3·4 COUNT GENERATION PROPORTIONAL TO SLIP FREQUENCY

3.4.4. The decoder

Each SN 74154 decoder has four input lines which are decoded internally to address one of sixteen outputs. The slip frequency as generated at the output of the up-down counter is in the form of a six line binary output count of 64, each count representing 5.6° angular increments of the slip frequency cycle. Four decoders, connected as shown in Fig.3.5, were used to give one exclusive output for each of 64 counts.

All 16 outputs of the decoder are at logic "1" unless both Enable and Data inputs are at logic "0". When Enable and Data inputs are at 0, the SN 74154 operates as a mutually exclusive sixteen line NAND gate decoder of the four input address lines. By using two logic inverters as in Fig.3.5 it is possible to code the two most significant bits of the counter output (E and F) to give gating inputs to the decoders ensuring that only one decoder is operational at a time, Fig.3.5.

The decoder must be synchronised to start at the beginning of a motor secondary voltage half-cycle. To achieve this, the secondary voltage was monitored and all outputs of the up-down counter were held at zero until the start of a positive half-cycle was detected. The decoder outputs representing the beginning and end of each half cycle of all three phases are given in Fig.3.6. Fig.3.6 shows that although decoder outputs can be selected to give 180° periods for each phase the 120° relative displacement can be in error by up to 3.75° .

3.5. Waveform generation

The emf induced in the secondary winding of the induction motor is shown in Fig.3.7. When the slip changes from $S > 0$ to $S < 0$, the polarity and phase sequence of the secondary voltage is changed. Power flow considerations in Section 4.1 of Chapter 4 determine that although the emf changes polarity through synchronous speed the same thyristor group must

count		0—15				16—31				32—47				48—63			
		E	F	$\bar{E}$	$\bar{F}$	E	F	$\bar{E}$	$\bar{F}$	E	F	$\bar{E}$	$\bar{F}$	E	F	$\bar{E}$	$\bar{F}$
decoder 1		○	○	▨	▨	1	○	▨	▨	○	1	▨	▨	1	1	▨	▨
		ON				OFF				OFF				OFF			
decoder 2		▨	○	1	▨	▨	○	○	▨	▨	1	1	▨	▨	1	○	▨
		OFF				ON				OFF				OFF			
decoder 3		○	▨	▨	1	1	▨	▨	1	○	▨	○	1	▨	▨	○	▨
		OFF				OFF				ON				OFF			
decoder 4		▨	▨	1	1	▨	○	1	▨	▨	1	○	▨	▨	○	○	▨
		OFF				OFF				OFF				ON			

▨ inputs not connected

Fig 3-5 THE 64 COUNT DECODER CIRCUIT

Decoder output	Angular degrees	Nearest output to	R phase	Y phase	B phase
0	0°	0°	0	↓	↓
11	61.875°	60°	↓	↓	180
22	123.75°	120°	↓	0	↓
32	180°	180°	180°	↓	↓
43	241.875°	240°	↓	↓	0
54	303.75°	300°	↓	180°	↓
0	360°	360°	0	↓	↓

Fig 3.6 DECODER OUTPUTS REPRESENTING 3 PHASE DISPLACEMENT

Fig 3.7 SECONDARY emf AND THE SIGNAL GENERATOR OUTPUT WAVEFORMS

continue to be controlled. Thus the output of the signal generator must be as shown in Fig.3.7 in order to maintain correct control.

The required waveforms were generated by using six NAND gates each with two inputs as shown with their truth table in Fig.3.8. Diode decoupling of inputs was necessary to prevent short circuiting the decoder outputs.

The three-phase square wave output waveform is shown in Fig.3.9 and meets the requirements of Fig.3.7.

3.6. Observation of signal generator operation

A block diagram of the completed secondary emf signal generator is shown in Fig.3.10 and a detailed circuit diagram is given in Fig.3.11.

The operation of the signal generator was verified with the induction motor driven externally by a d.c. machine (Chapter 5). The emf generated in the open circuited secondary windings of the motor were directly compared with the signal generator output. The output of the signal generator was observed as the rotor was driven slowly through synchronism. The three outputs, as shown in Fig.3.12, are seen to change their phase relationship as the machines are driven through synchronous speed. This meets the operating requirement of Fig.3.7.

A photographic record was taken to show one signal generator output and its associated secondary emf. Fig.3.13 clearly shows the waveforms changing from in-phase to anti-phase as the speed goes supersynchronous. This behaviour is essential for correct operation of the system as will be discussed in Chapter 4. An operational disadvantage of the generator can be seen in Fig.3.14 where the generator output preset to be in phase advance of the secondary emf at subsynchronous speed changes to a phase lag at supersynchronous speed. This would present problems if automatic power factor control was being used.

Fig 3-8 NAND GATE OUTPUT CIRCUIT

Fig 3-9 SIGNAL GENERATOR SQUARE WAVE OUTPUT

Fig 3.10 BLOCK DIAGRAM OF THE SECONDARY EMF SIGNAL GENERATOR

FIG. 3.11. CIRCUIT DIAGRAM OF SECONDARY EMF SIGNAL GENERATOR LOGIC

**Fig 3-12 CHART RECORDING OF THE SECONDARY emf SIGNAL GENERATOR
OPERATING THROUGH SYNCHRONOUS SPEED**

Induced secondary emf and signal generator output in phase

(a) Subsynchronous at 1100 rpm

The induced secondary emf and signal generator output automatically changed to be in antiphase as synchronous speed is exceeded.

(b) Supersynchronous at 1900 r.p.m.

Fig 3.13 SECONDARY EMF AND SIGNAL GENERATOR OUTPUT INITIALLY SET TO BE IN PHASE AT SUBSYNCHRONOUS SPEED

Signal generator in phase advance of the secondary emf

(a) Subsynchronous at 1100 rpm.

The signal generator automatically changes to phase lag the secondary emf as synchronous speed is exceeded

(b) Supersynchronous at 1900 rpm

Fig 3.14 SIGNAL GENERATOR INITIALLY SET TO BE IN PHASE ADVANCE OF SECONDARY
SLIP EMF AT SYNCHRONOUS SPEED

In general the secondary emf signal generator worked well and reliably remained in synchronism with the secondary emf of the induction motor. Considerations in Chapter 4 suggest that the generation of a sine wave output rather than a square wave would be preferable. This could be produced by using a Read Only Memory (ROM) for generating the sine function followed by digital to analogue converters for each output phase of the generator. This could be the subject of further work.

Chapter 4

The cycloconverter - motor slip recovery system

4.1. Power flow considerations

4.1.1. Power flow in an induction machine

An equivalent circuit for one-phase of a three-phase induction machine with sinusoidal supply is shown in Fig.4.1. Of the total power crossing the air gap, P_r , an amount $(1 - S) \cdot P_r$ appears as mechanical power, P_o , developed at the shaft.

If the output torque is T N-m and is developed at an angular velocity ω radians/sec then,

$$P_o = T \cdot \omega \text{ watts.} \quad (4.1)$$

The angular velocity of the shaft is related to the synchronous angular velocity ω_s

$$\omega = \omega_s (1 - S), \quad (4.2)$$

where S is the per-unit slip.

Thus the torque in N-m is given by

$$T = \frac{P_r}{\omega_s} \quad (4.3)$$

The efficiency of the secondary circuit is then given by

$$\eta_s = \frac{P_o}{P_r} = (1 - S). \quad (4.4)$$

Thus the efficiency of the secondary circuit falls linearly as slip increases and the overall efficiency is further reduced by stator copper loss, iron losses and friction losses.

FIG. 4.1. SINGLE PHASE EQUIVALENT CIRCUIT OF AN INDUCTION MOTOR

FIG. 4.2. POWER FLOW IN AN INDUCTION MACHINE

FIG. 4.3. SINGLE PHASE REPRESENTATION OF CYCLOCONVERTER AND MOTOR

The flow of power in the machine may be considered with reference to Fig.4.2. In a slip-ring machine with external secondary resistors the power dissipated in those resistors is represented by P_2 . The directions of power flow shown in Fig.4.2 are considered positive. Provided P_2 can be made positive or negative there are six possible operating modes for the induction machine. There are three cases where mechanical power is out of the shaft when the machine is driving and three cases when mechanical power is into the shaft when the machine is electrically braking or generating.

- Case 1. Conventional Driving. P_0 , P_1 , and P_2 all positive and $1 > S > 0$.
- Case 2. Supersynchronous Driving. P_0 and P_1 are positive and P_2 is negative with $S < 0$, i.e. supersynchronous.
- Case 3. Inverted phase driving. P_0 is positive and P_1 , P_2 are negative with $S > 1$. The machine would be driving in the opposite direction to the synchronous field and the secondary emf would be > 50 Hz.
- Case 4. Regenerative Braking. P_0 , P_1 and P_2 all negative and $1 > S > 0$.
- Case 5. Induction Generator. P_0 and P_1 are negative and P_2 is positive with $S < 0$, i.e. supersynchronous.
- Case 6. Plug Braking. P_0 is negative and P_1 , P_2 are positive with $S > 1$. The machine is running in the opposite direction to the synchronous field with a secondary emf > 50 Hz.

Making use of the bi-directional power flow capability of the cyclo-converter it is possible to investigate cases 1, 2, 4 and 5 over a limited range of slip. The range of slip will be limited by waveform considerations and the number of effective phases of the cycloconverter. Cases 3

FIG. 4.4. VOLTAGE AND CURRENT WAVEFORMS - MOTORING

and 6 are not possible as the cycloconverter cannot operate at the high output frequencies required by the secondary emf.

4.1.2. Power flow considerations in the cycloconverter

A single-phase representation of the cycloconverter and induction motor is shown in Fig.4.3. Voltage sources e_1, e_2, e_3 represent the three supply phases. The secondary current and the changing direction of power flow can be shown by drawing voltage and current waveforms for the basic circuit of Fig.4.3. In drawing these waveforms the secondary emf E_2 is assumed to be a very low frequency compared with the 50 Hz supply and has, therefore, been taken at a constant value over the short time interval considered.

The voltage and current waveforms for the machine operating as a motor are shown in Fig.4.4. Fig.4.4a shows the 'n' group thyristors being fired during a positive half cycle of secondary emf at subsynchronous speeds. All of the current flows from the secondary to the supply during the positive half cycle of the supply. Power is therefore being returned to the supply and P_2 is positive. The machine is operating subsynchronously as a motor. As the slip reduces less of the secondary current flows during the positive half cycle of the supply as indicated by the shaded area in Fig.4.4b and the net power P_2 , although still positive, is reducing.

Provided the same thyristor group is fired, although the polarity of secondary emf has changed as the machine goes supersynchronous, then Figs.4.4c and d show that the same current is still flowing to the supply but increasingly during the negative half cycle of the supply. Thus the power flow has been reversed and P_2 is negative. The machine is now operating supersynchronously as a motor.

The voltage and current waveforms for the machine operating as a

FIG. 4.5. VOLTAGE AND CURRENT WAVEFORMS - BRAKING & GENERATING

brake or generator are shown in Fig.4.5. In this case 'p' group thyristors are being fired during a positive half cycle of secondary emf at subsynchronous speeds. Current flows from the supply to the secondary of the machine and does so, in Fig.4.5a entirely during the positive half cycle of mains supply. The power flow P_2 is, therefore, negative and the machine is braking. Figs. 4.5b - d show the gradual change in the direction of power flow as the machine operates at supersynchronous speed as a generator.

4.2. The control system

The circuit waveforms given in Figs.4.4 and 4.5 for motoring and braking operation at subsynchronous and supersynchronous speed clearly indicate three requirements of the control system.

- (i) To obtain full control of current over the range of subsynchronous and supersynchronous operation the thyristor firing angle may have to be controlled beyond the 0° point. So, as described in Chapter 2, a firing range of -60° to 120° would be preferable to the more usual 0° to 180° .
- (ii) The voltage and current waveforms clearly show that as the machine runs through synchronous speed, either as a motor or a brake, the thyristor group firing during a positive half cycle of secondary emf at subsynchronous speed must operate during a negative half cycle of secondary emf at supersynchronous speed. The polarity of the secondary emf signal generator will be used to determine which group of thyristors is to conduct and so the change in polarity of the secondary emf when operating through synchronous speed must not be accompanied by a similar change in the signal generator output. The signal generator described in Chapter 3 meets this requirement.

(iii) To change the operating mode of the machine from motoring to braking it is merely necessary to change the thyristor group being controlled by a given polarity of signal generator output.

Lavi and Polge⁽²⁰⁾ carried out induction motor tests using a quasi-square wave secondary current. They showed that the harmonics in the secondary current induced only small currents into the primary winding and produced negligible torque. They suggest that it is sufficient to consider only the torque caused by the fundamental component of secondary current. To minimise wasteful losses in the secondary circuit the control system should ideally generate a sine wave of secondary current.

A reasonably good waveform can be achieved by using the cosine crossing method⁽³⁹⁾ of firing angle control provided the secondary emf is precisely known. Unfortunately the secondary emf signal generator so far developed only produces a fixed amplitude square wave at slip frequency. Accepting this limitation of the signal generator there are two possible methods of operation.

(a) To operate with a firing angle that is constant over each half cycle of secondary emf. This would produce current waveforms shown in Fig.4.6. whose basic shape would change both with slip and with any preset phase angle between the signal generator output and the secondary emf. At small values of slip the secondary induced emf would be very small compared with the supply voltage and the current pulses would be of constant amplitude.

To ensure that the correct torque is developed at any desired operating speed the firing angle can be controlled as a function of speed error by using a speed servo amplifier as shown in Fig. 4.7.

FIG. 4.6. VOLTAGE AND CURRENT WAVEFORMS FOR FIXED FIRING ANGLE CONTROL

FIG. 4.7. SPEED ERROR CONTROL OF FIRING ANGLE

FIG. 4.8. CASCADE CONTROL SYSTEM AND WAVEFORMS

(b) To operate with closed loop current control whereby the firing angle of the thyristors would be continuously changed by the output of the current servo amplifier. With good servo response this would result in a constant amplitude of current during each half cycle of secondary emf. The level of current demanded can be derived from the output of a speed servo amplifier so that the current necessary to produce the required torque at any speed would be obtained. A cascade control system of this type and its output waveform is shown in Fig.4.8.

A control system was designed to enable these control methods to be investigated. Details of the control system are shown in Fig.4.9. The speed demanded is compared with the output of a tachogenerator mounted on the induction machine and the amplified error is produced at the output of amplifier A11. The diodes around amplifier A12 ensure that the current demand to the servo amplifier A1 is of the same polarity irrespective of the polarity of speed error. Details of the operation of amplifiers A1, A2, A3 and their associated interlocks is given in Appendix 10.6.

The changeover switch at the inputs to A2, A3 allows them to operate either from the output of the current servo amplifier or directly from the output of the speed servo amplifier.

Amplifiers A14, A15 were added to operate the changeover relay R in Fig. 10.26 of Appendix 2 whenever the motor speed exceeded the demanded speed. This relay merely changes the thyristor group that is fired during a given polarity of secondary emf signal generator output. By this means the system automatically changes between motoring and braking modes as determined by the polarity of the speed error. Controls around amplifiers, A14, A15 set the sensitivity of the circuit and provide an adjustable dead band range.

FIG. 4.9 THE SERVO CONTROL SYSTEM

4.3. Equivalent Circuits of the Induction Motor with a secondary power source.

4.3.1. An equivalent circuit with a voltage source in the secondary

A single phase representation of the induction motor with a voltage source 'e' in the secondary is shown in Fig.4.10. The voltage source is assumed to be in anti-phase with the secondary induced emf E_2 . The magnetising terms have been omitted.

The voltage and current relationships across the air gap of the machine are

$$E_2 = S. E_1 \left(\frac{N_2}{N_1} \right), \quad (4.5)$$

$$\text{and } I_2 = I_1 \left(\frac{N_1}{N_2} \right) \quad (4.6)$$

An equivalent circuit with all values referred to the primary side may now be drawn as in Fig.4.11. Expressions for the referred impedance and external secondary voltage source can be derived from comparisons between Fig.4.10 and Fig.4.11.

From Fig.4.10

$$\frac{E_2 - e}{I_2} = \sqrt{R_2^2 + (SX_2)^2} \quad (4.7)$$

and from Fig.4.11

$$\frac{E_1 - e'}{I_1} = \sqrt{R^2 + X^2} \quad (4.8)$$

and using equations 4.5, 4.6 and 4.8, gives

$$\frac{E_2 - S. \frac{N_2}{N_1} \cdot e'}{I_2 S \left(\frac{N_2}{N_1} \right)^2} = \sqrt{R^2 + X^2} \quad (4.9)$$

Comparing equations 4.7 and 4.9 gives

FIG. 4.10. EQUIVALENT CIRCUIT OF AN INDUCTION MACHINE WITH A VOLTAGE SOURCE IN THE SECONDARY

FIG. 4.11. A PRIMARY SIDE EQUIVALENT CIRCUIT

FIG. 4.12. EQUIVALENT CIRCUIT WITH REFERRED SECONDARY TERMS

$$e' = \frac{1}{S} \cdot \left(\frac{N_1}{N_2} \right) e, \quad R = \frac{R_2}{S} \cdot \left(\frac{N_1}{N_2} \right)^2 \quad \text{and} \quad X = X_2 \left(\frac{N_1}{N_2} \right)^2$$

The equivalent circuit on the primary side is then as shown in Fig.4.12. This is a very simple representation of the system in which e represents the fundamental value of the cycloconverter voltage and in which the supply impedance have been neglected.

4.3.2. Effect of the magnitude and phase of the voltage source on the developed torque and the power factor.

A phasor diagram is drawn in Fig.4.13 showing the secondary voltage source with component e_p in phase, and component e_q in quadrature, with the secondary induced emf E_2 .

Resolving vertically and horizontally and collecting terms gives,

$$\frac{I_2 \cos \phi_2}{(E_2 + e_p)R_2 + SX_2e_q} = \frac{I_2 \sin \phi_2}{(E_2 + e_p)SX_2 - R_2e_q} = \frac{1}{R_2^2 + S^2X_2^2} \quad (4.10)$$

The power crossing the air gap P_r is given by

$$P_r = E_1 I_1 \cos \phi_2. \quad (4.11)$$

Using equations 4.5 and 4.6 gives

$$P_r = \frac{E_1 I_1}{S} \cos \phi_2 = T_{sw}, \text{ Torque in synchronous watts.} \quad (4.12)$$

So

$$T_{sw} = \frac{E_2}{S} \left[\frac{(E_2 + e_p)R_2 + SX_2e_q}{R_2^2 + S^2X_2^2} \right] \quad (4.13)$$

$$= \frac{E_2 R_2}{S^3 [(R_2/S)^2 + X_2^2]} + \frac{E_2}{S^3} \cdot \left[\frac{e_p R_2 + SX_2e_q}{(R_2/S)^2 + X_2^2} \right] \quad (4.14)$$

If R_1 and X_1 are neglected then $E_1 = V_1$ and so

$$E_2 = E_1 \left(\frac{N_2}{N_1} \right) \times S = V_1 \left(\frac{N_2}{N_1} \right) \cdot S = E'_2 \cdot S \quad (4.15)$$

FIG. 4.13. PHASOR DIAGRAM OF INDUCTION MOTOR WITH VOLTAGE SOURCE IN THE SECONDARY

FIG. 4.14. EFFECT OF SECONDARY SOURCE VOLTAGE ON TORQUE

Now if V_1 is constant, E_2' will be constant and the torque may be written as,

$$T = \frac{V_1^2 \cdot R_2'}{s \left[\left(\frac{R_2'}{s} \right)^2 + (X_2')^2 \right]} + \frac{E_2'}{s^2} \cdot \left[\frac{e_p R_2 + s X_2 e_q}{(R_2/s)^2 + X_2^2} \right] \quad \begin{matrix} \text{synch.} \\ \text{watts.} \end{matrix} \quad (4.16)$$

$$= T_N + T_i$$

where

$$R_2' = R_2 \left(\frac{N_1}{N_2} \right)^2 \quad \text{and} \quad X_2' = X_2 \left(\frac{N_1}{N_2} \right)^2$$

The first term is the well known expression for torque, T_N , for a machine without a secondary voltage source and the second term represents the torque, T_i , due to the presence of the secondary source voltage.

The effect of the additional torque term may be considered as follows:

$$\text{Case 1} \quad e_q = 0 \quad T_i = \frac{E_2 R_2 e_p}{R_2^2 + (s X_2)^2} \quad (4.17)$$

This is shown in Fig.4.14(a) and is a maximum at $s = 0$ so that torque can be developed at synchronous speed.

$$\text{Case 2} \quad e_p = 0 \quad T_i = \frac{E_2 s X_2 \cdot e_q}{R_2^2 + (s X_2)^2} \quad (4.18)$$

This term is shown in Fig.4.14(b) and has the effect of increasing the peak torque above and below synchronism but does not develop torque at synchronous speed.

The effect of the secondary source voltage on the power factor may be seen from equation 4.10 from which

$$\tan \phi_2 = \frac{(E_2 s + e_p) s X_2 - R_2 e_q}{(E_2 s + e_p) R_2 + s X_2 e_q} \quad (4.19)$$

If the secondary source voltage is zero, $e_q = e_p = 0$

and

$$\tan \phi_2 = S X_2 / R_2 \quad (4.20)$$

If $e_q = 0$ but $e_p \neq 0$ then,

$$\tan \phi_2 = S X_2 / R_2$$

The in-phase component of secondary source voltage does not then affect the power factor.

If $e_p = 0$ but $e_q \neq 0$ then

$$\tan \phi_2 = \frac{E_2 S^2 X_2 - R_2 e_q}{S X_2 e_q + E_2 S R_2} \quad (4.21)$$

The quadrature component e_q will reduce the numerator and increase the denominator so reducing $\tan \phi_2$ and improving the Power Factor $\cos \phi_2$.

Thus with control of the magnitude and phase of the secondary source voltage operation is possible through synchronous speed with a controlled power factor.

4.3.3. An equivalent circuit with a current source in the secondary circuit.

The proposed control systems described in Section 4.2 ensure that the secondary current is sufficient to produce the necessary torque to run the machine at any desired speed. Providing the secondary emf signal generator is synchronised to the open-circuit secondary emf prior to start-up, the signal generator will control the phase angle between the primary voltage and the secondary current. The servo system controls the amplitude of the secondary current which is therefore set in magnitude and phase by the cycloconverter rather than by the machine parameters. The cycloconverter may be considered as a current source in the secondary

circuit and a single-phase equivalent circuit is shown in Fig.4.15. For simplicity the magnetising branch has again been omitted.

If the secondary current I_2 is controlled to be at an angle θ with respect to the primary voltage V_1 then,

$$\begin{aligned} \text{Power into the primary } P_1 &= 3 \times V_1 I_1 \cos \theta \\ &= 3 \times V_1 \beta I_2 \cos \theta \quad \text{watts} \end{aligned} \quad (4.22)$$

where β = turns ratio N_2/N_1 .

The total power crossing the air gap from primary to secondary

$$\begin{aligned} P_r &= P_1 - 3 I_1^2 R_1 \\ &= 3[V_1 \beta I_2 \cos \theta - (\beta I_2)^2 R_1] \text{ watts} \\ &= \text{torque in synchronous watts} \end{aligned} \quad (4.23)$$

and the torque in N - m

$$T = \frac{P_r}{\omega_s} \quad (4.3)$$

The mechanical output power $P_o = (1 - S)P_r$ watts

Then the electrical power output external to the secondary winding is,

$$P_2 = S P_r - 3 I_2^2 R_2 \quad \text{watts} \quad (4.24)$$

The power flow diagrams for subsynchronous and supersynchronous operation with forward and reverse currents for motoring and braking are given in Fig.4.16.

4.4. Consideration of secondary current waveform

The current source model described in the preceding section enables the developed torque to be calculated from a knowledge of the secondary current. The model applies to the fundamental component of secondary current. However, the secondary current has already been shown to be

FIG. 4.15. EQUIVALENT CIRCUIT WITH THE CYCLOCONVERTER AS A CURRENT SOURCE

a) Subsynchronous Speed (s positive) - Motoring Load

b) Supersynchronous Speed - Motoring Load

c) Subsynchronous Speed - Braking Load (Current Reversed)

d) Supersynchronous Speed - Generating

FIG. 4.16. POWER FLOW DIAGRAMS FOR THE CURRENT SOURCE MODEL
ARROWS SHOW DIRECTION OF POWER FLOW

far from sinusoidal. In order to obtain a true prediction of torque it is necessary, therefore, to obtain the rms value of the fundamental in the secondary current waveform.

At low levels of current and slip the current waveform will be discontinuous with current pulses of relatively constant shape as shown in Fig.4.17. Furthermore, the repetition rate of the pulses, T_1 , would be small in comparison with the quarter period T of the secondary emf cycle. There will, therefore, be many current pulses during the quarter period T . Fig.4.17 also shows the effect of the current detector, as described in Appendix 2, preventing current for a period T_0 between alternate polarity half cycles of secondary current.

A Fourier analysis is required to obtain the coefficient for the fundamental of this waveform.

The fundamental coefficient is given by

$$a_1 = \frac{1}{\pi} \int_{-\pi}^{+\pi} F(\omega t) \cdot \cos(\omega t) d(\omega t) \quad (4.25)$$

where $F(\omega t)$ represents the waveform being considered.

If there are many identical current pulses in the period T then the value of $\cos \omega t$ can be considered constant during each short time period that the function $F(\omega t)$ exists.

$$\begin{aligned} \text{Then } a_1 &\approx 4 \times \frac{1}{\pi} \times \int \text{area of each current pulse} \times \text{value of } \cos \omega t \text{ at} \\ &\quad \text{each current pulse over the range that the pulses} \\ &\quad \text{exist i.e. } \left(\frac{\pi}{2} - \frac{T_0}{2}\right) \text{ to } 0 \\ &= \frac{4}{\pi} \cdot \left(\text{area of each current pulse} \right) \times \sum_{n=1}^{n=Y} \cos \left(\frac{\pi}{2} - \frac{T_0}{2} - n T_1 \right) \end{aligned} \quad (4.26)$$

where the area of each current pulse must be in ampere radians and the total number of pulses in the period T is given by

$$Y = \frac{T - T_0/2}{T_1} = \frac{1 - 100S \cdot T_0}{200S T_1}, \quad (4.27)$$

as $T = \frac{1}{4Sf_s}$ seconds and $f_s \approx 50$ Hz the supply frequency.

FIG. 4.17. SECONDARY CURRENT WAVEFORM

FIG. 4.18. DISCONTINUOUS CURRENT WAVEFORM

FIG. 4.19. CONTINUOUS CURRENT WAVEFORM

In order to determine the fundamental of secondary current it is therefore necessary to find the area of each current pulse and carry out the above summation.

The total rms value of the secondary current I_{RMS} was measured experimentally by using an electronic integrating meter and the mean value of the current peaks, I_p , was determined by using an oscilloscope.

If the secondary current is assumed to be discontinuous sinusoidal current pulses of time T_2 as shown in Fig.4.18, then the total secondary current is given by

$$I_{RMS} = \sqrt{\frac{Y}{T} \int_0^{T_2} \left(I_p \sin \frac{\pi}{T_2} t \right)^2 dt} \quad (4.28)$$

$$= I_p \sqrt{\frac{Y \cdot T_2}{2 T}} \quad \text{amps.} \quad (4.29)$$

Thus the value of T_2 can be calculated from

$$\begin{aligned} T_2 &= \left(\frac{I_{RMS}}{I_p} \right)^2 \frac{2T}{Y} \\ &= \frac{2 T_1}{(1 - 100S T_o)} \cdot \left(\frac{I_{RMS}}{I_p} \right) \quad \text{secs.} \end{aligned} \quad (4.30)$$

If, from the results taken, the calculated value of T_2 is greater than T_1 then the current must be continuous as shown in Fig.4.19.

To simplify analysis of this waveform it was approximated by a constant current of I_{min} on which is superimposed a sinusoidal current with a half cycle time of T_1 .

For this waveform, as shown in Fig.4.19,

$$I_{RMS} = \sqrt{\frac{Y}{T} \int_0^{T_1} [I_{min} + (I_p - I_{min}) \sin \omega t]^2 dt} \quad \text{amps} \quad (4.31)$$

which when solved gives

$$I_{\min} = -0.6 I_p + \frac{1}{2} \sqrt{(1.2 I_p)^2 - 8.8 \left[I_p^2 - \frac{2 I_{\text{RMS}}^2}{1 - 100S T_o} \right]} \quad \text{amps (4.32)}$$

Thus it is now possible to calculate the area under each current pulse for inclusion in the Fourier fundamental coefficient summation. The area under the discontinuous waveform, Fig.4.18,

$$A_D = \int_0^{T_2} I_p \sin \frac{\pi}{T_2} t \cdot dt = \frac{2 I_p T_2}{\pi} \quad \text{amp seconds.} \quad (4.33)$$

$$\text{and} \quad A_D = 200.S.I_p.T_2 \quad \text{amp radians} \quad (4.34)$$

The area under the continuous waveform, Fig.4.19,

$$A_C = I_{\min} \cdot T_1 + \int_0^{T_1} (I_p - I_{\min}) \sin \frac{\pi}{T_1} t \, dt \quad (4.35)$$

$$= T_1 \cdot \left[I_{\min} + \frac{2}{\pi} (I_p - I_{\min}) \right] \quad \text{amp seconds} \quad (4.36)$$

$$A_C = 100 S \pi T_1 \left[I_{\min} + \frac{2}{\pi} (I_p - I_{\min}) \right] \quad \text{amp radians} \quad (4.37)$$

Thus the appropriate area can be calculated and used in the Fourier summation to give the rms value, $a_1/\sqrt{2} = I_{\text{FUND}}$, of the fundamental component of secondary current.

So

$$I_{\text{FUND}} = 180.S.I_p.T_2 \sum_{n=1}^{n=y} \cos \left(\frac{\pi}{2} - \frac{T_o}{2} - n T_1 \right) \quad (4.38)$$

for discontinuous waveforms or ,

$$I_{\text{FUND}} = 282.8.S.T_1 \cdot \left[I_{\min} + \frac{2}{\pi} (I_p - I_{\min}) \right] \sum_{n=1}^{n=y} \cos \left(\frac{\pi}{2} - \frac{T_o}{2} - n T_1 \right) \quad (4.39)$$

for continuous waveforms when $T_2 > T_1$.

A computer program has been written (Appendix 3) in which the measured values of I_{RMS} , I_p and slip, S , appears as input data. The program calculates whether the waveform was discontinuous or continuous and so uses the appropriate expression for the calculation of the RMS

value of the fundamental of the secondary current.

Using the computed value of I_{FUND} the program then calculates the total torque from

$$T = \frac{3}{50\pi} \times \left[V_1 \cdot I_{\text{FUND}} \cdot \beta \cdot \cos \theta + (I_{\text{FUND}} \cdot \beta)^2 \cdot R_1 \right] \text{ Newtons} \quad (4.40)$$

The program assumes a constant time T_o with slip and calculates the value of $\cos \theta$ from

$$\theta = \frac{T_o}{2} \times 2 \pi f_s \text{ radians,}$$

$$\theta = 50 \pi S \cdot T_o \text{ radians.} \quad (4.41)$$

The program makes a further allowance in the calculation of $\cos \theta$ to include the fact that synchronisation of the secondary emf signal generator on the open circuit secondary emf was at a small angle in phase advance of the primary voltage V_1 . Program details are given in Section 11.2 of Appendix 3.

4.5. Summary

Power flow considerations have shown that with the proposed system four operating modes are possible, i.e. sub- and super-synchronous operation with braking or overhauling loads. The simple current source model developed readily shows the effect of varying magnitude and phase angle of the secondary current upon the developed torque. An analytical method has been presented to obtain the rms value of the fundamental component of secondary current for inclusion in calculations based upon the current source model.

Chapter 5

Experimental Results

5.1. Equipment and Instrumentation

The problems associated with the control of three-phase to three-phase cycloconverters have been discussed in Chapter 2. Operation with discontinuous currents has been shown to require complex gate firing arrangements if a neutral link is not used. To minimise the complexity of equipment it was decided to carry out the experimental work using three identical three-phase to single-phase cycloconverters as described in Appendix 2. A neutral connection is therefore necessary, which, because of the usual limitation of three slip rings and a delta rotor winding, means that the primary winding has to be on the stator with the slip recovery cycloconverter connected to the rotor. Only Universal teaching machines were available and their low voltage rotor windings therefore necessitated the use of a high current low voltage transformer in the primary feed.

The induction machine was mechanically connected to a d.c. dynamometer for loading purposes. To enable the dynamometer to present braking and overhauling torques to the induction machine that were constant with speed, a regenerative thyristor loading rig has been developed. This is described in Appendix 4. Current limitations of the primary supply transformer prevented a resistance starting technique from being used. The d.c. load machine and its associated electronic power controller were used to raise the speed of the machine ~~set~~ until the frequency of the secondary emf was within the operating range of the cycloconverter.

The machine set was also fitted with the disc and light sensor required for operation with the secondary emf signal generator. The

output from this disc was monitored by a digital counter to give a precise measurement of machine speed. A d.c. tachogenerator fitted to the machines was used as the speed feedback in the analogue control system.

Dynamometer wattmeters and moving iron ammeters and voltmeters were used on the primary supply and the mains supply side of the cycloconverter. The secondary current and power between the machine and the cycloconverter were measured with an integrating electronic true rms ammeter and an electronic thermal element wattmeter. Torque measurements were recorded from a spring balance operated by the swinging frame of the dynamometer.

A basic diagram of the experimental equipment is shown in Fig.5.1 and photographs of the equipment are shown in Fig.5.2 and 5.3.

5.2. Standard machine tests

Standard locked rotor and no-load machine tests were carried out and from these the equivalent circuit values were derived as shown referred to the low voltage primary side in Fig.5.4.

A series of load tests were made with the machine operating as a motor and a generator. For the motoring tests the loading equipment was operated under closed loop torque control, energy being regeneratively returned to the supply. For the generating tests the induction machine was driven by the loading equipment operating with closed-loop speed control. The load tests were repeated with a range of external rotor resistors.

To assess the validity of the single phase equivalent circuit a computer program was written (see Appendix 3, 11.1) to predict the behaviour of the machine on load. The experimental results and the computer predicted results are compared in Figs.5.5 - 5.8.

The results showed quite good agreement and the model was considered acceptable for analysis with cycloconverter operation.

Fig 5.1 EXPERIMENTAL EQUIPMENT AND INSTRUMENTATION

(a) General view of the laboratory test rig

(b) Close up view of the cycloconverter control panel

(a) Duplex pulse generator and amplifier board

(b) Secondary E.M.F signal generator board

Turns ratio $\frac{\text{secondary}}{\text{primary}} = 6.0$

Fig 5.4 EQUIVALENT CIRCUIT VALUES REFFERED TO THE LOW
VOLTAGE PRIMARY SIDE

Fig 5.5 INDUCTION MOTOR BRAKING LOAD TEST SHORT CIRCUITED ROTOR

Secondary winding resistance 1.5Ω / phase

Fig 5.6 INDUCTION MOTOR BRAKING LOAD TEST WITH
EXTERNAL 8.4Ω / PHASE

Secondary winding resistance $1.5 \Omega / \text{phase}$

Fig 5.7 INDUCTION GENERATOR TEST ROTOR SHORT CIRCUITED

Secondary winding resistance 1.5Ω / phase

Fig 5.8 INDUCTION GENERATOR TEST WITH EXTERNAL 8.4Ω / PHASE

Torque-slip curves are given in Fig.5.9 for a range of external secondary resistance in both motoring and generating modes.

5.3. Cycloconverter slip-recovery tests

Preliminary trials were carried out in order to determine the best method of control of the system. The two methods of control outlined in Section 4.2 were compared and the closed-loop current control scheme showed two problems.

- (i) The discontinuous nature of the current feedback signal caused "ripple instability" which could only be removed by filtering and reducing the servo response.
- (ii) The operation of the current detectors at the end of each half cycle of secondary current caused a period of about 20 mS during which the feedback current signal was zero (see Fig.4.7). During this period the servo amplifier output increased, resulting in a severe current overshoot at the commencement of the next half cycle.

The control scheme in which the output of the speed servo was used to control directly the cycloconverter firing angle gave much better results and was used throughout the test program. The smooth operation of the machine through synchronous speed can be seen in the step response given in Fig.5.10. The speed changes smoothly through synchronous speed, the overshoot being caused by the response of the control system. During deceleration the secondary current is cut off and comes in again at a high frequency as the speed undershoots the demanded lower speed.

5.3.1. Observation of secondary current waveform

The secondary current waveform was discontinuous and a test was carried out to check the waveform analysis of Section 4.4. The induction motor was operated subsynchronously with cycloconverter control and

FIG. 5. 9. MEASURED TORQUE - SLIP CURVES FOR INDUCTION MACHINE WITH ADDED ROTOR RESISTANCE

time scale 10 mm/sec

Fig 5-10 STEP RESPONSE OF SPEED WITH CYCLOCONVERTER CONTROL

photographs were taken of the secondary current, Fig.5.11. The secondary current was monitored by the electronic true rms ammeter and the developed torque noted for three operating conditions as given in Table 5.1. The current zero time T_0 due to the current detectors was observed to be approximately 20 mS.

TABLE 5.1

MEASURED VALUES				CONDUCTION TIME	
Input Voltage	Slip	RMS CURRENT	PEAK CURRENT	Computed	From Photo
36.0	0.016	2.4 A	4.0 A	4.95 mS	4.4 mS
35.5	0.047	3.4 A	5.5 A	5.6 mS	5.2 mS
35	0.068	4.2 A	6.7 A	6.0 mS	5.9 mS

The computer predicted conduction time from equation 4.30 agreed well with photographic observation although the method could at best only be considered approximate. The computer program in Appendix A3, 11.2. was used in the later analysis of cycloconverter results to calculate the fundamental rms current in the secondary and thereby to calculate, using the current source model equation 4.23, the predicted machine torque.

5.3.2. Open-loop torque speed curves

The cycloconverter was operated with the firing angle of the thyristor set at a predetermined value. The torque-speed characteristics for the machine operating with three different firing angles are shown in Fig.5.12.

The machine was not easy to control in this way. With low load torques the machine ran up to a high speed where the developed torque was decreased because of the reduction in the number of mains frequency conduction periods per half cycle at the higher slip frequency.

(a) slip $s = 0.016$

(b) slip $s = 0.047$

(c) slip $s = 0.068$

Scale 2.0 Amperes per cm
5.0 mS per cm

Fig 5.11 PHOTOGRAPHS OF SECONDARY CURRENT WITH
CYCLOCONVERTER CONTROL

FIG. 5. 12. OPEN LOOP TORQUE SPEED CHARACTERISTIC

A similar, but more severe, problem occurred with high load torques at subsynchronous speeds. The increased slip due to the load resulted in less torque being developed because of the reduction in the number of mains frequency conduction periods and the machine would ultimately stall.

The remaining test program was carried out with closed-loop speed control.

5.3.3. Closed loop torque-speed curves

The induction machine was operated with the cycloconverter under closed-loop speed control and gave the torque-speed characteristics of Fig.5.13. Operation with either driving or braking torques at subsynchronous and supersynchronous speeds was possible. The speed range was restricted by the frequency limitations of the cycloconverter. The slope of the characteristics could be changed by altering the gain of the servo amplifier. System stability was not a problem in this instance as controlling the frequency response of the servo amplifier, and including derivative as well as proportional feedback, proved effective.

The computer program in Section 11.2 of Appendix 3 was used to calculate the rms of the fundamental component of secondary current. This was then used to calculate the predicted torque from equation 4.23 of the current source model. A comparison between computer predicted torque and measured torque is shown in Figs. 5.14, 5.15, and shows reasonable agreement considering the approximations being made.

5.3.4. Comparison between secondary resistance and cycloconverter control

The cycloconverter slip recovery system was set up to deliver the same torque at a given slip as had been developed by the induction motor with different secondary resistors. The steady state performance of the two systems is compared graphically in Figs. 5.16 and 5.17.

Fig 5.13 CLOSED LOOP TORQUE SPEED CHARACTERISTIC

FIG. 5. 14. SUBSYNCHRONOUS OPERATION WITH OVERHAULING LOAD

- Experimental results
- x Computer predicted values from equations in appendix 11.2

Fig 5.15 SUBSYNCHRONOUS OPERATION WITH BRAKING LOAD

Fig 5.16 COMPARISON BETWEEN RESISTANCE/CYCLOCONVERTER OPERATION

Comparison between computed and measured torque for cycloconverter control

FIG. 5.17. COMPARISON BETWEEN RESISTANCE / CYCLOCONVERTER OPERATION

The effect of the poor secondary waveform with cycloconverter operation can be clearly seen in Fig.5.17. The total rms current in the secondary circuit with cycloconverter operation is significantly greater than that with resistance control for any given torque level. The computer fundamental component of the cycloconverter current compares well with the rms value of the current when external resistors are added.

The harmonic current waveform in the secondary circuit gives high secondary winding copper losses and the overall efficiency shows little or no improvement over the resistance method. Obviously if the cycloconverter current had been sinusoidal the secondary winding loss in both methods would be the same but most of the external resistance losses would have been recovered by the cycloconverter, thereby showing an overall improved efficiency.

The reduction in power factor when operating with the cycloconverter at high slips is due to the increasing lagging angle of the secondary current with respect to the supply voltage caused by the effect of the current detector operating time, T_o , as shown in Fig.5.18.

The computed fundamental rms current was used to predict developed torque from the current source model and gave reasonable agreement with the measured values as shown in Fig.5.17.

5.35. Power factor control

The secondary emf signal generator as described in Chapter 3 was provided with additional inputs that made it possible to change the phase relationships between the signal generator output and the supply voltage. The effect of the current detector operating time, T_o , had given rise to an increase in the lagging angle of secondary current at high slip values. A test was carried out to manually change the secondary current phase angle with respect to the supply voltage by means of the signal generator controls.

The induction motor was operated at 1453 rpm, a slip of .03, and a

Fig. 5.18 EFFECT OF CURRENT ZERO TIME T_0 ON THE PHASE ANGLE OF THE FUNDAMENTAL OF SECONDARY CURRENT

constant load torque of 5 N-m. The phase relationship between the signal generator output and the supply voltage was changed manually and the effect on secondary current, primary current and primary power factor was observed. The computer program in Section 11.2 of Appendix 3 based upon the fundamental rms analysis and the current source model was used to produce predicted results. Quite good agreement between predicted and measured performance can be seen in Fig.5.19.

The secondary current can be clearly seen in Fig.5.19 to be a minimum when the current is controlled to be in phase with the supply voltage. The larger than normal magnetising component of machine current prevented a dramatic change to leading power factor operation that may occur in a standard slip-ring machine.

5.3.6. Cube law power characteristic

Pumping and ventilating duties would be the most likely application for the supersynchronous cycloconverter slip recovery scheme. In such applications the load has a power requirement proportional to the cube of the operating speed. Such a load characteristic was investigated with the cycloconverter slip recovery system. The equipment operated well over the speed range of ± 150 rpm about the synchronous speed of 1500 rpm i.e. a 20% speed range. The operation could have been further improved by signal generator correction of the increasing lagging angle of secondary current at high values of slip.

The results of the test are shown in Fig.5.20 and of particular interest is the power flow in the secondary circuit as the machine operates through synchronous speed. Power flow from the supply to the machine secondary is necessary for low slip and supersynchronous operation. The copper losses are high due to the high value of total secondary rms compared with the predicted current from the constant current model required to develop the measured torque.

FIG. 5. 19. EFFECT OF CONTROLLING POWER FACTOR BY THE SLIP DETECTOR

Secondary current

Secondary power flow

Fig. 5.20 POWER FLOW FOR A CUBE LAW POWER CHARACTERISTIC

5.4. Summary of results

The slip recovery cycloconverter system operates well over a limited range of slip about synchronous speed. Closed loop speed control is desirable and gives a smooth transition through synchronous speed. Operation in the circulating current free mode using electronic current detectors has the disadvantages of loss of available conduction time and increased lagging power factor at high slip values. The very simple control system whereby speed error determines the thyristor firing angle generates a poor waveform of secondary current. The secondary current and, therefore, the copper loss, is much greater than would be the case with a sinusoidal secondary current. The operating speed range is limited by the cycloconverter power circuit configuration and with the three pulse connection used the operational speed range is limited to approximately $\pm 10\%$ about synchronous speed. The primary power factor of the machine can be controlled by means of the secondary emf signal generator although the high magnetising component of the test machine prevented operation with a substantial leading power factor.

The simple current source model of the system gave quite good prediction of machine torque from a knowledge of secondary current provided only the rms of the fundamental of secondary current was considered.

The slip recovery cycloconverter system seems ideally suited to operation with pumping or ventilating loads where the limited speed can produce a considerable change in mass flow.

Chapter 6

Conclusions and suggestions for further work

6.1. Summary of the static Scherbius system

A static Scherbius system whereby a cycloconverter can be used to control energy in the secondary circuit of a slip-ring induction motor has been designed and satisfactorily operated.

Consideration of three-phase to three-phase cycloconverter operation without a neutral connection but with discontinuous current determined that three identical three-phase to neutral cycloconverters should be used in the experimental equipment. The power and electronic circuits of a suitable cycloconverter were designed and constructed.

A study of power flow requirements in the induction motor, during subsynchronous and supersynchronous operation with overhauling and braking loads, showed that the machine could be correctly controlled provided a signal generator was developed to give a three-phase output locked in phase with the secondary induced emf. Further, that additional means should be provided to change the phase angle between the signal generator output and the secondary emf for system power factor control.

A suitable secondary emf signal generator was developed using digital circuit techniques and proved to be reliable and remain locked in phase to the secondary emf at all times whilst the motor was subjected to sudden changes of load and speed.

Using the signal generator to control the cycloconverter thyristor group to be operated at any instant gave very smooth control of motor speed above, below, and through, synchronous speed.

Experimental tests showed the machine to be easy to control and stable in operation within the operating range of the cycloconverter which corresponded to a slip frequency of approximately ± 5 Hz.

A simple model of the system was developed in which the cycloconverter was represented as a controlled current source in the secondary of the usual single phase model of the induction motor. With cycloconverter control of the secondary current, and of its phase relationship to the primary supply voltage, the prediction of torque from the model was readily possible.

A computer program was written based upon the Fourier series to calculate, from experimental results, the fundamental component of secondary current and to use this to predict the developed torque using the equations from the current source model. Results gave quite good agreement between theory and practice.

The results and performance of the equipment were quite satisfactory and could be the basis for further study.

6.2. Limitations of the system

Operation of the slip energy recovery system was extremely simple under closed loop speed control but a number of limitations were observed.

- (i) A very basic three pulse cycloconverter was used for economic reasons which limited the operation of the motor to a maximum speed range of approximately ± 150 rpm about a synchronous speed of 1500 rpm.
- (ii) The cycloconverter was designed for operation without circulating currents by using electronic current detection. The current detectors produced a period of zero current between each consecutive half cycle of secondary current which, for short circuit prevention, was approximately 20 mS. This had two detrimental effects:
 - (a) The time in each half cycle for which current flow was possible reduced as the machine slip increased

and so caused a further restriction in the operating speed range.

- (b) The fundamental of secondary current was at an increasingly lagging phase angle with respect to secondary emf as the machine slip increased.

The effect of this was to reduce the system power factor and to increase the secondary current for a given load torque.

- (iii) The closed loop speed control system and associated pulse generators and amplifiers did not produce a sinusoidal secondary current. Each half cycle of secondary current was comprised of discontinuous current pulses of essentially constant amplitude. The secondary current therefore had a high harmonic content which gave additional copper losses without substantially contributing to the developed torque.

- (iv) The control system did not include any means for automatically starting the machine and bringing it up to a speed at which the frequency of the secondary emf was within the operating capability of the cycloconverter.

6.3. Application of the system

Subsynchronous slip recovery drives are used in large pumping and ventilating systems where the required operating speedrange is restricted to about 20% below synchronous speed. The deciding factors in drive selection are both initial cost and running costs. Where the drive powers are high and the machines are operating continuously then running costs are of great importance. A very small improvement in efficiency can show a substantial financial saving on the annual running costs. Weiss⁽⁴⁴⁾ considers the possible use of the cycloconverter slip recovery

system for such applications and suggests that operation above and below synchronous speed with some degree of power factor control would be advantageous. Weiss' proposals, however, are based upon a doubly fed machine in which the output of the cycloconverter in the secondary circuit is controlled from an external frequency signal and the machine operates synchronously at all operating speeds. Such a system suffers from the usual stability problems of doubly fed machines.

The cycloconverter slip recovery system investigated here should be ideally suited to these applications, provided the basic operating requirements determined in this work are developed to overcome the present limitations.

6.4. Future work

The present limitations of the system should be further investigated with a view to improved efficiency and extended speed range.

A more complex cycloconverter with a greater number of effective phases should be used to extend the speed range. Improvement of the waveform of the secondary current should be possible by using different control methods. A possible technique would be to use the cosine crossing method⁽³⁹⁾ of phase angle control in which the sinusoidal reference signal be developed by the secondary emf signal generator.

The signal generator could be modified so that the counter output was used as an input to a Read Only Memory (ROM) in which the sine function had been programmed. Linear digital to analogue converters could then be used to generate the sinusoidal waveform to the cosine crossing method control circuits.

The dynamic behaviour of the system should be further investigated both for the start-up condition and for changes in speed and torque with different machine and controller parameters.

Chapter 7.

Acknowledgements

The author wishes to express his thanks to Professor G.D.S. MacLellan, Head of Department of Engineering, for use of the laboratory and computing facilities provided by the University of Leicester. Particular thanks are expressed to Dr. P.H. Mynott who supervised the research work relating to this thesis. The author is grateful to Mr. R.G. Stephens for assistance in taking experimental results and to Dr. P.G. Holmes for his willingness to discuss certain aspects of the work.

Finally thanks to Mrs. H. Sheppard and the Drawing Office for their help in preparation of the thesis.

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Chapter 9

Appendix 1

Power Flow in a Slip Recovery System

9.1. Terminology

P_r	Power crossing the air gap from the primary to the secondary.	watts.
P_o	Mechanical power developed at the motor shaft.	watts.
P_{c_1}	Power dissipated in the secondary winding.	watts.
P_1	Power input to the stator.	watts.
P_2	Electrical power out of the secondary.	watts.
T	Torque developed at the motor shaft.	N-m
η_s	Synchronous speed.	rev/s.
η	Rotor shaft speed	rev/s.
S	Slip = $\frac{\eta_s - \eta}{\eta_s}$	
P_m	The rated power output of the induction motor.	watts.

9.2. Relationships between torque and power in an induction motor

From elementary induction motor theory the usual relationships between torque and power are as follows:-

$$2\pi\eta_s \cdot T = P_r = \frac{P_o}{1 - S} = \frac{P_{c_2} + P_2}{S} \quad (9.1)$$

where $P_{c_2} + P_2$ is the total electrical power in the rotor circuit.

If iron losses and stator and rotor copper losses are neglected then

$$P_{c_2} = 0 \quad \text{and} \quad P_r = P_1,$$

the stator input power.

$$\text{So} \quad 2\pi\eta_s \cdot T = P_1 = \frac{P_o}{1 - S} = \frac{P_2}{S} \quad (9.2)$$

$$\text{and} \quad P_2 = \frac{S}{1 - S} \cdot P_o \quad (9.3)$$

The maximum power output for a constant torque load at reduced speeds is given by

$$P_o = \frac{\eta}{\eta_s} \cdot P_m, \quad (9.4)$$

when the electrical power out of the secondary terminals, $P_2 = S P_m$.

9.3. Power flow in the electromechanical slip energy recovery system

Let the Total Shaft Power be P . An amount P_o is supplied by the induction motor and an amount

$$P_2 = S P_o / (1 - S) \quad (9.5)$$

is supplied by the coupled d.c. machine.

$$\text{So} \quad P = P_o + \frac{S P_o}{1 - S}$$

$$\text{and so} \quad P_1 = P, \quad (9.6)$$

$$\text{and} \quad P_2 = S P \quad (9.7)$$

These powers are shown for the electromechanical Kramer system in Fig.1.9(a) of Chapter 1.

9.4. Power flow in the rotary electric system

In this case the total shaft power is developed by the induction motor

$$\text{so} \quad P = P_o.$$

$$\text{Therefore} \quad P_1 = \frac{P}{1 - S},$$

$$\text{and} \quad P_2 = \frac{S}{1 - S} \cdot P. \quad (9.8)$$

These powers are shown for the rotary electric Kramer system in Fig.1.9(b) of Chapter 1.

9.5. The frame size rating of the d.c. motor in the electromechanical system

The electrical power from the rotor $P_2 = S P$ but as

$$P = \frac{P_o}{1 - S}$$

then
$$P_2 = \frac{S}{1 - S} \cdot P_o \quad (9.9)$$

For constant torque loads the maximum shaft power at any speed, η is given by,

$$P_o = \frac{\eta}{\eta_s} \cdot P_m$$

where P_m is the rated power of the induction motor at near-synchronous speed.

So
$$P_o = (1 - S)P_m, \quad (9.10)$$

and
$$P_2 = S \cdot P_m. \quad (9.11)$$

The d.c. motor, however, must recover this power at a speed η and so the effective frame size rating of the d.c. machine is

$$S P_m \times \frac{\eta_s}{\eta} = \frac{S}{1 - S} \cdot P_m \quad (9.12)$$

9.6. Power recovery for loads whose torque varies as the square of speed

Let the rated power of the machine $P_m = T_s \cdot \eta_s$ where T_s is the full load torque at a speed very close to the synchronous speed η_s .

At any other speed η

$$T = T_s \left(\frac{\eta}{\eta_s} \right)^2 \quad (9.13)$$

and the shaft output power
$$P_o = \eta \cdot T_s \left(\frac{\eta}{\eta_s} \right)^2$$

In the rotary-electric system

$$P_2 = \frac{s}{1-s} P = \frac{s}{1-s} \cdot P_o$$

So

$$P_2 = \frac{s}{1-s} \cdot \eta T_s \left(\frac{\eta}{\eta_s} \right)^2 = \left(\frac{\eta^2}{\eta_s} - \frac{\eta^3}{2} \right) \cdot T_s \quad (9.14)$$

The maximum value of P_2 can be found by differentiating and equating to zero when the speed at which max. P_2 occurs is given by

$$\eta_{\max} = \frac{2}{3} \cdot \eta_s$$

Substitution gives

$$P_{2_{\max}} = \frac{4}{27} \cdot T_s \eta_s \quad (9.15)$$

i.e. the maximum slip power to be recovered is approximately 15% of the rated power of the induction motor and occurs at about $\frac{2}{3}$ rds. of synchronous speed.

Chapter 10

Appendix 2

The design of the three-phase to single-phase cycloconverter

The cycloconverter converter considerations of Chapter 2 led to a decision to use three identical three-phase to single-phase cycloconverters. The design of the cycloconverter includes all necessary power, electronic, and control circuits.

10.1. The power circuit

The basic elements of the power circuit are shown in Fig.10.1. Six thyristors are required, each protected by its own high speed fuse. The three-phase supply is switched to the thyristors by an electromechanical contactor whose auxilliary contact can be used to clamp the operation of the thyristor pulse generators until a short time after the contactor has been energised. A three-phase line reactor serves the dual purpose of limiting prospective fault current for fusing, and of voltage transient suppression. The phase advance signal to the pulse generators must be controlled to meet the requirements of Table 4 in Chapter 2.

10.1.1. The thyristor power assembly

The thyristor must be carefully chosen so that its performance capability will meet the basic operating requirements of voltage, current and temperature⁽³¹⁾. The thyristor has both a low thermal time constant and low thermal capacity and it is necessary, therefore, to ensure adequate cooling. Thyristors are mounted on heat sinks to form an assembly.

The equipment was designed to operate from a laboratory supply of 240 V 3 phase 4 wire 50 Hz although for slip energy recovery the power circuit was ultimately operated from a low voltage supply. The most suitable thyristors readily available for use in the laboratory were

FIG. 10.1. THE POWER CIRCUIT

Westinghouse CS21Q for which abbreviated data is given in Fig.10.2.

10.1.2. Voltage transient protection

The thyristor must be protected against voltage transients and high rates of rise of forward voltage dv/dt . Excessive dv/dt will cause 'self firing' of the thyristors resulting in possible short circuits in the power circuit.

Local switching transients occur due to operation of the supply contactor. Short duration transients with fast rise times due to contact bounce give rise to problems of stray coupling to thyristor gates. Such transients may be suppressed by capacitors C_1 across each contact of the supply contactor, Fig.10.1. If the contactor closes at the peak of the supply voltage high values of dv/dt can be applied to the thyristors. These values may be limited by suitable selection of R-C components connected across each thyristor which operate in conjunction with the line reactor L , in Fig.10.1.

A simple transient suppression network is shown in Fig.10.3 where the mains supply is replaced by a battery voltage V_b equal to the peak of the mains supply. Merrett⁽³²⁾ analyses this equivalent circuit and his graphical results are given in Fig.10.4 which shows that the minimum of the maximum value of dv/dt occurs with a damping factor $\eta = 0.265$

$$\text{when the output } \frac{dv}{dt} = \left(\frac{dV_o}{dt} \right)_{\max} = 0.81 \cdot V_b \cdot \frac{1}{\sqrt{LC}} \quad (10.1)$$

$$\text{The damping ratio } \eta = \frac{R}{2\omega_\eta L}, \quad (10.2)$$

$$\text{where the resonant frequency } \omega_\eta = \frac{1}{\sqrt{LC}} \text{ radians/sec.}$$

Max repetitive peak reverse voltage VRRM	VRRM	560
Max non repetitive peak reverse voltage	VRSM	700
Max rms on state current	IT (RMS)	19A
Max average on state current at 180° conduction	IT (AV)	16A
Max i^2t for fusing up to 10ms	I^2t	200 A ² s
Peak current during fusing	I	200A
Internal thermal resistance junction to case	R θ_{JC}	1.2 °C/W
Thermal resistance case to heat sink	R θ_{CS}	0.4 °C/W
Maximum junction temperature	T _j MAX	100 °C

FIG. 10.2. DATA FOR THYRISTOR TYPE CS21Q

FIG.10.3. SIMPLE TRANSIENT SUPPRESSION NETWORK

FIG.10.4. VOLTAGE OVERSHOOT AND dV_o/dt DURING SWITCH-ON

At this minimum the voltage overshoot is approximately 50% of the input voltage and so for the CS21Q thyristor the peak repetitive reverse voltage, which is more than twice the crest voltage of the line to neutral supply, may be considered an adequate safety margin.

The line reactor was designed to have a reactive voltage drop of 12 V at 30 A rms 50 Hz which gives an inductance per limb, L , of 1.27 mH. The CS21Q thyristor has a maximum dv/dt capability of 20 V/ μ S which from equation 10.1 requires a value C of 0.05 μ F. For a damping factor $\eta = 0.265$ the required value of R is 85 Ω and from Fig. 10.4 the peak voltage during switch on to a 240 V 3 phase to neutral supply will be 300 V which is well within the capability of the thyristor.

10.1.3. Current overload protection

Thyristors have a very limited overload capacity and a suitable fuse must protect against high prospective fault currents, limit the peak current and limit the thermal energy $\int i^2 dt$ during fault interruption. The factors affecting the design of high speed fuses for protection of semi-conductors are considered by Pearce & Newberry⁽³³⁾.

Current overloads can occur for three reasons:-

- (a) Continuous operation at a load current beyond the normal full load rating.

This is usually prevented by closed loop electronic control of current to limit the current to a preset maximum level. More sophisticated control by monitoring thyristor case temperatures permits occasional long duration overloads.

(b) High rates of change of current, di/dt

Current carriers in the thyristor have a finite propagation time so that high levels of di/dt result in high current densities in the gate region with consequent permanent damage. The line reactor, Fig.10.1, limits the di/dt which for a short circuited load fault will be in this case less than $1A/\mu S$ for the reactor fitted. Where di/dt is a problem at values greater than $30A/\mu S$ driving the thyristor gate with a high peak current pulse having a fast rise time in the order of $1A/\mu S$ is recommended⁽³⁴⁾.

(c) High level short duration overload due to a short circuit.

The most likely short circuit fault in the cycloconverter is a simultaneous operation of group 'p' and group 'n' thyristors. The fault circuit is then line voltage with two reactors of inductance L in series. Each limb of the reactor fitted in this design drops reactively 12 V at 30 A and so the

$$\begin{aligned} \text{prospective fault current} &= \frac{\text{RMS LINE VOLTAGE}}{2 \times \text{line impedance}} \\ &= \frac{240}{2 \times 12/30} = 300 \text{ A} \end{aligned}$$

There will be an error in line impedance due to the iron cored reactor saturating at the prospective fault current but a safety margin is present because of the line source impedance.

Fuse selection

The fuse, then, must carry the continuous rms current, which in this design is to be limited to 15 A rms, operate at a supply voltage of 240 V rms and limit the i^2t below the maximum device capability of $200 \text{ A}^2\text{S}$ for a prospective fault current of 300 A.

A suitable fuse would be an El000-15 having characteristics as in Fig.10.5 which at 300 A prospective fault current and 240 V rms has the following performance compared with the CS21Q thyristor:-

	<u>E1000-15 fuse</u>	<u>CS21Q thyristor</u>
Continuous rms rating	15 A	19 A
Total i^2t	60 A ² /S	200 A ² /S
Maximum peak current	170 A	200 A
Maximum arcing voltage	500 V	$V_{RSM}=700$

The CS21Q thyristor will be adequately protected by the E1000-15 fuse in this application.

10.1.4. Output rating of the cycloconverter

The current rating of the cycloconverter is difficult to determine without precise information of the nature of the load and its affect upon the conduction angle of the thyristors.

For simplicity each thyristor group in Fig.10.6 may be considered as a half-wave three-phase rectifier operating for half of each output cycle. Consider the cycloconverter output as a square wave of rms value I , Fig.10.6, with a half wave on-off cycle for each thyristor group as shown in Fig.10.6(a).

If the load is considered sufficiently inductive to maintain continuous conduction during the 'on' period for each group then, as each thyristor conducts for one third of the time,

the rms current per thyristor = $I/\sqrt{3}$ and
the Average " " " = $I/3$.

When taken over the total period of one output cycle, then,

$$\text{the rms current per thyristor} = I/\sqrt{3} \cdot \sqrt{2} = 0.4 I, \quad (10.3)$$

and the Average " " " = $I/3.2 = 0.166 I$. (10.4)

FIG.10.5. CHARACTERISTICS OF THE E1000 - 15 FUSE

FIG. 10.6. CURRENT RATING WAVEFORMS OF THE CYCLOCONVERTER

If the maximum permitted r_{ms} current for each thyristor is 15 A then the maximum output from the cycloconverter would be 38 A r_{ms} .

10.1.5. Thyristor cooling

The relatively short thermal time constant of the thyristor necessitates the mounting of the device onto a suitable heat sink having a low thermal resistance. This allows the thyristor to run continuously at a given dissipation without exceeding the maximum junction temperature.

Heat sinks cooled by natural convection were used and when a three phase to three phase cycloconverter was considered (Chapter 2) it was found convenient to mount the thyristors in groups of three on a common heat sink.

The heat transfer may be considered from the approximate steady state thermal resistance diagram of Fig.10.7. The heat transfer equation in terms of thermal resistance is given by

$$(T_j - T_a) = P_j \cdot (\theta_{j-c} + \theta_{c-s}) + 3 P_j \theta_{s-a} \quad (10.5)$$

where

P_j is the power developed at each thyristor junction

θ_{j-c} is the thermal resistance junction to case of the thyristor

θ_{c-s} is the thermal resistance case to heat sink

θ_{s-a} is the thermal resistance heat sink to ambient

T_j is the maximum junction temperature

T_a is the maximum ambient temperature

For the CS21Q thyristor,

$$\theta_{j-c} = 1.2 \text{ } ^\circ\text{C/W} ,$$

$$\theta_{c-s} = 0.4 \text{ } ^\circ\text{C/W} ,$$

$$T_j = 125 \text{ } ^\circ\text{C/W} .$$

For a 20 cm length of 60 D heat sink $\theta_{s-a} = 2.4 \text{ } ^\circ\text{C/W} .$

FIG. 10.7. THERMAL RESISTANCE DIAGRAM

FIG. 10.8. POWER DISSIPATION FOR THE CS 21Q THYRISTOR

The maximum power developed at each thyristor junction can now be calculated assuming a maximum ambient temperature of 40°C

$$P_j = \frac{T_j - T_a}{\theta_{j-c} + \theta_{c-s} + 3 \theta_{s-a}} = \frac{85}{8.8} \approx 10 \text{ watts.}$$

The thyristor dissipation curve, Fig.10.8, gives a mean forward current of 7.0 A for a 120° conduction period at 10 watts dissipation. From equation 10.4 the rms output from the cycloconverter is restricted to 42 A by the heat sink, which is in excess of that permitted by the device and fusing considerations.

10.2. The Pulse Generator

When operating as part of a closed loop system the Pulse Generator must provide a pulse at an instant in the mains cycle which is controllable over 180° by a variable d.c. voltage obtained from the output of a control amplifier. Means must be provided to prevent a pulse being generated during the period when the thyristor to be fired is reverse biased by the mains supply. Furthermore, as mentioned in Chapter 2, to prevent short circuit faults in a cycloconverter pulse generation to the incoming thyristor group must be prevented until the outgoing thyristor group has turned off.

The Pulse Generator circuit is shown in Fig.10.9. Capacitor C, charging resistor R_3 , and programmable unijunction transistor (PUT) form a one shot timing circuit.

The charging of C and, therefore, pulse generation, is prevented

- (i) by transistor Q_1 being biased 'on' by R_2 and,
- (ii) by transistor Q_2 being biased 'on' by an input to R_5 whenever thyristors in the other control group are conducting.

For mains synchronisation with a control range of -60° to $+120^{\circ}$ a timing waveform, Fig.10.10, turns off Q_1 by reversing the potential at its

FIG.10. 9. PULSE GENERATOR WITH PHASE ANGLE CONTROL

FIG.10.10. OPERATING WAVEFORMS OF THE PULSE GENERATOR

base with diode D_3 preventing excess reverse voltage. With zero control voltage V_o the capacitor charges via R_3 towards V_s until the trigger level of the PUT is reached at $V_s/2$. When this occurs the anode to cathode path of the PUT changes from a high to a low impedance and discharges C so producing a short duration pulse across R_6 . The anode current supplied by R_3 exceeds the PUT valley current so the PUT remains in a low impedance state until reset by Q_1 saturating at the end of the timing waveform half cycle.

With minimum control voltage V_o resistor R_3 is selected so that the charging time constant CR_3 does not allow the trigger voltage to be reached by the end of the required conduction period and so a pulse is not produced, Fig. 10.10(d). If the control voltage V_o is increased C will charge very rapidly as R_1 is a much lower value than R_3 . When the capacitor voltage exceeds the control voltage diode D_1 will block and C continues to charge at a slower rate via R_3 . The trigger level of the PUT will be reached before the end of the conducting period and a pulse will be generated, Fig. 10.10(e,f). Increasing the control voltage will produce a pulse earlier in the conduction period.

10.2.1. Analysis of the charging circuit

The charging circuit is shown in Fig. 10.11 and can be analysed in two parts.

Period 1 $V_c < (V_o - V_d)$ with diode D_1 conducting with a forward drop of V_d volts.

Assuming an initial condition of $V_c = 0$

$$V_c = \frac{R_1 V_s + R_3 V_1}{(R_1 + R_3)} \cdot (1 - e^{-t/T}) \quad (10.6)$$

where

$$T = \frac{CR_1 R_3}{R_1 + R_3}$$

$$\text{and } V_1 = (V_o - V_d)$$

FIG.10.11. PULSE GENERATOR CHARGING CIRCUIT

FIG.10.12. CONTROL AMPLIFIER INPUT TO THE PULSE GENERATORS

FIG.10.13. SYNCHRONISING ACTION OF THE TIMING CIRCUIT

Period 2 $V_c > (V_o - V_d)$ the diode D_1 is blocking and C is charging from an initial condition of

$$V_c = (V_o - V_d) = V_1$$

when
$$V_c = V_s + (V_1 - V_s)e^{-t/CR_3} \quad (10.7)$$

Time to reach the PUT trigger voltage, V_t

Case 1 $(V_o - V_d) < V_t$. In this case diode D_1 will block before V_t is reached and the capacitor C charges through Period 1 and Period 2.

From eqn.10.6 the time t_1 for V_c to reach $V_1 = (V_o - V_d)$ is

$$t_1 = T \log_e \frac{V_1 R_3 + R_1 V_s}{V_s R_1 + V_1 R_3 - V_1 (R_1 + R_3)} \quad (10.8)$$

From eqn.10.7 the time t_2 for V_c to charge from V_1 to V_t is

$$t_2 = CR_3 \log_e \frac{(V_1 - V_s)}{(V_t - V_s)} \quad (10.9)$$

The total time to trigger the PUT t_T is

$$t_T = \frac{CR_1 R_3}{R_1 + R_3} \log_e \frac{V_1 R_3 + R_1 V_s}{V_s R_1 + V_1 R_3 - V_1 (R_1 + R_3)} + CR_3 \log_e \frac{(V_1 - V_s)}{(V_t - V_s)} \quad (10.10)$$

Case 2 $(V_o - V_d) > V_t$. In this case the capacitor C will charge to the trigger voltage without diode D_1 blocking.

From eqn.10.6

$$t_1 = t_T = \frac{CR_1 R_3}{R_1 + R_3} \log_e \frac{V_1 R_3 + R_1 V_s}{V_s R_1 + V_1 R_3 - V_t (R_1 + R_3)} \quad (10.11)$$

10.2.2. Design considerations

Input Resistance R_1 . The control voltage to the three pulse generators required for each thyristor group was to be developed from an emitter

follower stage at the output of a control amplifier, Fig.10.12. A supply voltage V_s of 24 V was chosen and the output of the control amplifier was clamped at 20.0 volts to avoid possible latch-up problems in the operational amplifier A. To limit the operational amplifier load to about 2.5 mA the minimum gain of transistor Q_1 , Fig.10.12, was assumed to be 10 and the total load resistance at the transistor emitter was limited to approximately 800 Ω . Resistors R and R_1 in Fig.10.12 were selected at 3.3 K Ω .

Synchronising resistors R_2 , R_4

The charging capacitor C in Fig.10.9 is released by transistor Q_1 turning off when the negative current through R_4 is approximately equal to the positive current through R_2 . From Fig.10.13 this occurs when

$$\frac{V_s}{R_2} < \frac{V_T \sin \theta}{R_4} \quad (10.12)$$

In this design values of $V_s = 24$ V, $V_T = 28$ V peak, $R_4 = 2.2$ K Ω and $R_2 = 22$ K Ω were used giving

$$t_3 = 0.25 \text{ ms or } \theta_3 = 4.5^\circ.$$

This represents a restriction on the available 10 ms control range of 15%.

Charging resistor R_3 .

To allow for biasing the control amplifier of Fig.10.12 the pulse generator was designed to just generate a pulse at approximately -60° when the control voltage was 4.0 volts.

A timing capacitor of $0.16 \mu\text{F} \pm 1\%$ was chosen and it can be seen from Fig.10.13 that this must charge to just reach the PUT trigger voltage V_t in a time $t = 10 \times 10^{-3} - 2 t_3$ seconds.

The value of R_3 was obtained from a Newton Raphson solution of eqn.10.10

$$t_T = \frac{CR_1 R_3}{R_1 + R_3} \log_e \cdot \frac{V_1 R_3 + R_1 V_s}{R_1 V_s + V_1 R_3 - V_1 (R_1 + R_3)} + CR_3 \log_e \frac{(V_1 - V_s)}{(V_t - V_s)} \quad (10.10)$$

where $V_1 = V_o - V_d$, $V_d = 0.6$ (typical diode drop), $V_s = 24.0$, $V_t = 12.0$,
 $V_o = 4.0$, $R_1 = 3.3 \times 10^3$, $C = 0.16 \times 10^{-6}$, $t_T = 9.5 \times 10^3$.

Using the Newton Raphson Method

$$R_3 = R_3 - \frac{f(R_3)}{f'(R_3)}$$

where $f(R_3) = \frac{CR_1 R_3}{R_1 + R_3} \log_e \frac{V_1 R_3 + R_1 V_s}{R_1 (V_s - V_1)} + CR_3 \log_e \frac{(V_1 - V_s)}{(V_t - V_s)} - t_T \quad (10.13)$

$$f'(R_3) = \frac{CR_1^2}{(R_1 + R_3)^2} \log_e \frac{(V_1 R_3 + R_1 V_s)}{R_1 (V_s - V_1)} + C \log_e \frac{(V_1 - V_s)}{(V_t - V_s)} \quad (10.14)$$

Let $C = C$, $A = R_1$, $R = R_3$, $V = V_1$,
 $B = V_s$, $D = V_t$, $T = t_T$.

LIST

```

10 DATA 3.3E+3, 24, 0.16E-6, 12, 9.5E-3, 3.4, 50E+3
20 READ A,B,C,D,T,V,R
25 LET N=R
30 LET Z=(C*A+2)/(A+R)+2
35 LET Y=(V*R+A*B)/(A*(B-V))
40 LET X=(V-B)/(D-B)
45 LET W=C*A*R/(A+R)
50 LET P=W*LOG(Y)+C*R*LOG(X)-T
55 LET Q=Z*LOG(Y)+C*LOG(X)
58 LET R=R-P/Q
60 IF ABS(R-N)<100 GO TO 80
70 GO TO 25
80 PRINT R
85 END
READY
RUN
99164.12

```

The value for R_3 of 99,164 Ω was realised by using a 91 K Ω resistor and a 10 K Ω potentiometer.

Calculation of control range

The equations for the time taken, t_T , for the Pulse Generator to produce a pulse are given in Section 2.3.2.

Case 1 $(V_o - V_d) < V_t$

$$t_T = \frac{CR_1 R_3}{R_1 + R_3} \log_e \frac{V_1 R_3 + R_1 V_s}{R_1 V_s + V_1 R_3 - V_1 (R_1 + R_3)} + CR_3 \log_e \frac{(V_1 - V_s)}{(V_t - V_s)} \quad (10.10)$$

Case 2 $(V_o - V_d) > V_t$

$$t_T = \frac{CR_1 R_3}{R_1 + R_3} \log_e \cdot \frac{V_1 R_3 + R_1 V_s}{V_s R_1 + V_1 R_3 - V_t (R_1 + R_3)} \quad (10.11)$$

These equations were solved for the values given on page 73 for increasing control voltage V_o and the answer expressed in the form of a firing angle in degrees over the range -60° to $+120^\circ$. The computer program below gave the results shown in Fig.10.14 which are compared with the experimentally measured performance.

FIG 10. 14. PULSE GENERATOR CONTROL RANGE

LIST

```

10 DATA 3.3E+3, 24, 0.16E-6, 12, 99164.12
11 PRINT TAB(1)"CONTROL VOLTS";TAB(20)"FIRING ANGLE";TAB(35)"T1";
12 PRINT TAB(45)"T2";TAB(55)"T"
15 READ A,B,C,D,R
17 FOR V=3.4 TO 20.4 STEP 0.5
20 LET Z=C*A*R/(A+R)
22 LET M=Z*LOG((V*R+A*B)/(A*(B-V)))
23 IF V>11.9 GO TO 27
24 LET L=C*R*LOG((V-B)/(D-B))
25 LET T=M+L
26 GOTO 29
27 LET T=Z*LOG((V*R+A*B)/(A*B+R*V-D*(A+R)))
28 LET L=0
29 LET H=INT(M*10000)
30 LET I=INT(L*10000)
32 LET J=INT(T*10000)
34 LET P=T+0.25E-3
35 LET Q=120-P*18E+3
37 LET N=V+0.6
40 PRINT TAB(1)N;TAB(20)Q;TAB(35)H/10;TAB(45)I/10;TAB(55)J/10
45 NEXT V
50 END

```

READY

RUN

CONTROL VOLTS	FIRING ANGLE	T1	T2	T
4	-55.4995	.9	8.5	9.4
4.5	-49.743	.9	8.1	9.1
5	-43.71067	1	7.7	8.8
5.5	-37.41293	1.1	7.3	8.4
6	-30.85464	1.1	6.9	8.1
6.5	-24.0365	1.2	6.5	7.7
7	-16.95597	1.2	6	7.3
7.5	-9.607808	1.3	5.6	6.9
8	-1.984353	1.3	5.1	6.5
8.5	5.924245	1.4	4.6	6
9	14.12999	1.4	4.1	5.6
9.5	22.64703	1.5	3.6	5.1
10	31.49176	1.5	3.1	4.6
10.5	40.68298	1.5	2.5	4.1
11	50.2421	1.6	1.9	3.6
11.5	60.1934	1.6	1.3	3
12	70.56447	1.7	.7	2.4
12.5	81.38659	1.7	.1	1.8
13	89.70825	1.8	0	1.4
13.5	93.83483	1.8	0	1.2
14	96.4996	1.8	0	1
14.5	98.43662	1.9	0	.9
15	99.93874	1.9	0	.8
15.5	101.1526	2	0	.7
16	102.1621	2	0	.7
16.5	103.0196	2.1	0	.6
17	103.7602	2.1	0	.6
17.5	104.4082	2.2	0	.6
18	104.9812	2.2	0	.5
18.5	105.4926	2.3	0	.5
19	105.9524	2.3	0	.5
19.5	106.3686	2.4	0	.5
20	106.7475	2.4	0	.4
20.5	107.0941	2.5	0	.4
21	107.4126	2.6	0	.4

STOP AT LINE 50

READY

10.3. The Pulse Amplifier

The low power pulse developed by the pulse generator must be amplified into a high power pulse of increased duration suitable for driving a thyristor gate. The pulse amplifier requirements are as follows:-

- i) Pulse amplification and electrical isolation between the pulse generator and the thyristor gate
- ii) To convert the short duration pulse from the pulse generator into a long pulse suitable for ensuring thyristor latching in an inductive circuit.
- iii) A fast rise time on the pulse leading edge for minimisation of di/dt problems within the thyristor.
- iv) To accept pulses from two pulse generators when operating as part of a controlled firing 'ring' as in Chapter 2.

There are many possible circuits which could include monostables for pulse length control and optoelectronic isolation. In this case a monostable blocking oscillator^(35,36) was used for pulse length control and a ferrite cored pulse transformer used for electrical isolation as shown in Fig.10.15.

10.3.1. Operation of the pulse amplifier

Operation of the circuit shown in Fig.10.15 is as follows:-

A positive pulse from the pulse generator through D_3 causes transistor Q_1 to start conducting. The collector voltage starts to fall and a positive pulse is applied to the base of Q_1 by the feedback winding of the pulse transformer. The base current through R_1 then drives transistor Q_1 into a fully saturated condition. The

FIG.10.15. CIRCUIT DIAGRAM OF THE PULSE AMPLIFIER

FIG.10.16. WAVEFORM AT THE BLOCKING OSCILLATOR OUTPUT WINDING

voltage across the transformer has built up to the supply voltage and a similar transformed and isolated voltage appears at the output. The magnetic flux builds up proportional to $\int V_s dt$ and the collector current increases until a value $h_{FE} I_b$ is reached. The transistor comes out of saturation and the feedback pulse to the base decreases and Q_1 turns off. As Q_1 turns off the stored energy in the transformer is circulated through D_1 to prevent excessive transient voltage across Q_1 . This energy must be dissipated before the arrival of the next pulse or the magnetic flux in the core will not 'reset' and the next output pulse will be attenuated. The time to dissipate the stored energy may be reduced by making D_1 a number of series diodes.

The waveform across the blocking oscillator output is shown in Fig.10.16.

10.3.2. Analysis of the blocking oscillator

The analysis is considered for the transistor saturated during the pulse period. A simple short circuit model ⁽³⁷⁾ is shown in Fig.10.17

where L_m is the Magnetising Inductance
 R_p is the Primary winding resistance.

This circuit can be analysed to give expressions for primary winding current i_1 and base current i_2 . These currents are then related at the end of the pulse duration time t_d by

$$i_1 = h_{FE} i_2 ,$$

where h_{FE} is the minimum value of large signal current gain for the transistor.

Making the approximations that the base resistance is much greater than the load resistance referred to the primary, which in turn is greater than the primary resistance, gives an expression for t_d as

$$t_d \approx \frac{L_m}{R_p} \cdot \log_e \left[1 + \frac{R_p h_{FE} (n_1)^2}{R_b} \right] \quad (10.15)$$

FIG 10.17. BLOCKING OSCILLATOR SHORT CIRCUIT MODEL

In calculating the pulse duration care must be taken to use the correct value for the relative permeability for the transformer core in determining the value of primary inductance. The effective value⁽³⁸⁾ of the magnetising inductance is

$$L_m = \frac{N^2 A_e \mu_o \mu_p}{l_e} \quad (10.16)$$

where N is the number of primary turns, A_e and l_e are the effective core cross-sectional area and length and μ_p is the pulse permeability defined by

$$\mu_p = \frac{\Delta B}{\Delta H} \quad (10.17)$$

From Faradays Law
$$e = N \cdot \frac{d\phi}{dt}$$

so that
$$\phi = \frac{1}{N} \int e \, dt \quad \text{or} \quad B = \frac{1}{NA} \cdot \int e \, dt$$

So assuming that the core has a maximum rated flux density of ΔB_m and operates on a constant voltage V_s for a time t_d

$$B_m = \frac{V_s \cdot t_d}{N \cdot A_e} \quad (10.18)$$

10.3.3. Pulse Amplifier Design

The amplifier was designed to produce a 6.0 volt secondary pulse of duration up to 800 μ S from a supply voltage of 12.0 volts.

A pulse transformer core type FX2242 was selected having the following data.

Effective permeability 1300

Maximum flux density 0.24 T

Effective magnetic path length $l_e = 5.26 \times 10^{-2}$ m

Effective area of magnetic path $A_e = 2.24 \times 10^{-4}$ m

Fully wound minimum number of turns = 375

Resistance of " " " " = 5 Ω

From equation (10.18) $N = 180$ turns on the primary.

For 6.0 volts output $n_2 = 0.5$ so secondary turns = 90.

Recommended⁽³⁶⁾ value for $n_1 = 0.2$ so feedback turns = 36.

From equation (10.16) $L_m = 220$ mH.

Resistance of primary winding $R_p = 2.4 \Omega$.

The pulse width t_d calculated from equation (10.15) for a range of values of R_b is plotted in Fig.10.18.

The results do not agree with those obtained in practice due, in the main, to the inductance used in the calculation which should be based upon the pulse permeability that varies with pulse width. The results do show that a large range of pulse widths are possible by varying the feedback resistor R_b .

The final output pulse from the transformer secondary, when three series diodes for D_1 in Fig.10.15 are used to reset the core, is shown in Fig.10.19.

10.4. The thyristor gate circuit

Correct triggering of the thyristor requires that the source of the trigger signal should supply adequate gate current and voltage without exceeding gate ratings. Manufacturers provide gate characteristics and specify boundary conditions inside which any device of one type may be reliably triggered.

Gate characteristics for the CS21Q thyristor used in this cyclo-converter are shown in Fig.10.20. The graph shows gate to cathode voltage as a function of positive gate current between minimum and maximum lines for all CS21Q thyristors. The area of recommended triggering is further bounded by lines of peak allowable gate voltage and current, and peak power dissipation. A lower line represents current and voltage values below which no thyristor will trigger.

FIG. 10.18. PULSE WIDTH t_d AS A FUNCTION OF FEEDBACK RESISTOR R_b

FIG.10.19. WAVEFORM OF THE AMPLIFIER OUTPUT PULSE

FIG. 10. 20. CHARACTERISTICS THYRISTOR TYPE CS 21Q

FIG. 10. 21. GATE CIRCUIT FOR THYRISTOR TYPE CS 21Q

The applicable gate power curve is selected on the basis of whether average or peak allowable gate power dissipation is limiting. For continuous d.c. triggering the average maximum allowable gate dissipation is 0.5 watts while for pulse gating the peak power limitation is 5.0 W.

In this design 800 μ S gate pulses were used to prevent latching problems on the inductive motor load and the 6.0 V pulses from the pulse amplifier were connected to the thyristor gate via a 10 Ω series resistor. The load line is drawn in Fig.10.20 and shows a maximum of 0.9 watts. Higher level pulse driving could be used, if necessary, by reducing the series gate resistance but would result in increased power supply ratings.

The final gate circuit is shown in Fig.10.21 where resistor R and capacitor C prevent gate pick-up from unwanted transients. Resistor R lowers the effective gate impedance on very sensitive devices. Capacitor C will degenerate the required fast pulse rise time which in this case is 2.5 μ S. The capacitor C is usually selected to compromise between a fast rise time and adequate suppression of spurious signals. The CR_g time constant is usually about 0.5 μ S. Satisfactory thyristor gating was achieved at all times.

10.5. The electric power supplies

Three separate electronic power supplies were designed for the equipment.

(i) A ± 12.0 V stable supply for the control servo amplifiers in Section 10.6 and pulse generators in Section 10.2. The servo amplifiers are integrated operational amplifiers which in themselves can tolerate power supply voltage variation but are affected by changes in supply to any external bias circuits. The thyristor pulse generators using Programmable Unijunction Transistors can accept reasonable voltage rail regulations as they are within a correcting control loop but a stable supply well suppressed from unwanted transients is preferred to minimise

the chance of spurious pulse generation.

(ii) A 12.0 V supply for the pulse amplifiers in Section 10.3.

These amplifiers generate fast rise step load changes which can greatly disturb the power supply rails and would cause unwanted pulse generation if connected to the pulse generator power supply.

(iii) A 5.5 V supply to the secondary emf signal generator described in Chapter 3. This generator uses TTL logic circuits which require a noise free low voltage supply because of their inherent high frequency response.

Power supply circuits were designed to incorporate modern integrated voltage regulators which were used in all of the three power supplies because of their superior performance. The circuit diagram of a regulator power supply is shown in Fig.10.22 where VR_1 sets the level of output voltage and resistor R determines the level of output current at which the regulator changes to a current limiting mode. The series pass transistor Q_1 is mounted on a suitable heat sink to allow operation up to 700 mA load current. Performance characteristics for the 12.0 V supply are shown in Figs. 10.23 and 10.24.

10.6. The control system

With the equipment so far described the output of the cycloconverter is determined by phase angle control which is a function of the control voltage to the pulse generators. The most usual control technique is the bias shift method or cosine crossing method of phase control⁽³⁹⁾. The method consists of comparing a cosine wave of line frequency with a control voltage and generating, at the instant of zero crossing, a trigger pulse from a one-shot multivibrator. The control voltage is varied sinusoidally by a master frequency reference generator. Continuous variation of the firing angle is obtained with an output waveform as in Fig.1.3.

FIG. 10.22. CIRCUIT DIAGRAM OF THE REGULATOR POWER SUPPLY

FIG. 10.23. OUTPUT CHARACTERISTIC FOR CONSTANT SUPPLY VOLTS

FIG. 10.24. INPUT CHARACTERISTICS 12-0V REGULATOR

For an induction machine with direct mains connection to the primary, but with a cycloconverter controlling current flow between the induced secondary emf and the supply, the secondary current must be at the secondary emf frequency. The reference generator to the cycloconverter must be controlled in phase and at the same frequency as the secondary induced emf. A secondary induced emf signal generator must be used^(26,27) (for details see Chapters 1,3) and the precise form of the generator output waveform will depend upon the generator design. Control of secondary circuit current is further complicated by the motor winding impedance being connected between two voltage sources - the mains supply and the secondary induced emf.

It was decided to operate the cycloconverter in a system where the secondary circuit current would be under closed loop control. A simple block diagram of the system is shown in Fig.10.25. The amplified current error at the output of the current servo amplifier is the control voltage to the pulse generators of both the 'p' and 'n' groups. The output of the secondary emf signal generator indicates the polarity of the induced secondary emf and determines whether the 'p' or 'n' group pulse generator receives the control voltage. An external control instruction to the changeover circuit at the output of the secondary emf signal generator can be used to determine which thyristor group is to operate for a given polarity of slip emf and by so doing control the direction of power flow (Section 4.2, Chapter 4). This enables the requirements of Table 4 of Chapter 2 to be realised. The cycloconverter was to operate in the circulating current-free mode of operation^(40,41,42) and so a current detector prevents the incoming group from receiving the control voltage until the current in the outgoing group is zero. Additional protection is incorporated by a cross interlock which prevents the incoming control amplifier from operating until the outgoing control amplifier has fully returned to minimum.

FIG. 10. 25. BLOCK SCHEMATIC OF THE CONTROL SYSTEM

10.7. Detailed operation of the control system

The required control system was developed using operational amplifiers as detailed in Fig.10.26. The circuit operation can be best understood by consideration of each of the basic elements.

The Current Servo Amplifier A_1

This amplifier compares the single polarity current demand voltage with a rectified output voltage proportional to the cycloconverter current as monitored by a Direct Current Transformer (D.C.T.). Resistor R_3 determines the system gain, and stability adjustment is possible by VR_1 to control the amplifier slew rate and by derivative feedback as set by VR_2 . Capacitor C_3 acts as a ripple filter on unwanted signal noise. The amplifier is biased to zero for zero current demand and develops a negative output as current is demanded.

Control Amplifiers A_2, A_3

The output of these amplifiers is the Control Voltage to the pulse generators which, as described in Section 10.2, require a voltage increasing positively with respect to the negative supply rail. With zero input from amplifier A_1 , amplifiers A_2 and A_3 are biased to a minimum with respect to the negative supply rail. As the output of A_1 increases with current error so the outputs of the Control Amplifiers will increase, producing thyristor gate pulses at 1.0 V (see Fig.10.14) and will reach a 20 V limit as set by VR_3 giving a control angle of approx. $+110^\circ$.

Operation of both the A_2 and A_3 amplifiers would produce simultaneous operation of the 'n' and 'p' group thyristors and cause an output short circuit. The Control Amplifiers are therefore provided with overriding, or clamping, inputs which, when positive, drive the amplifier to

FIG. 10. 26. CIRCUIT DIAGRAM OF THE CONTROL SYSTEM.

minimum output. These clamping inputs are (i) from the secondary emf signal generator which determines which thyristor group is to operate for a given half cycle of secondary induced emf and (ii) from the summing amplifiers A_4, A_5 which operate when the opposite thyristor group amplifier has control and is producing cycloconverter current.

The Secondary emf Polarity Detector A_{10}, A_{11}

The output from the signal generator (see Chapter 3) is amplified by A_{10} to generate a square wave at the slip emf frequency but of opposite polarity. When the output of A_{10} is positive Control Amplifier A_2 is held at minimum whilst the sign change amplifier A_{11} releases A_3 for control of the 'p' group thyristors. The half cycle of slip emf for which each thyristor group operates can be changed by energising relay R which has a 10 ms operating time.

The Current Detector A_9, A_8

The cycloconverter current is monitored by a high gain amplifier A_9 whose output is positive, when the 'p' group thyristors are controlling current, and negative when the 'n' group thyristors are controlling current. Thus when the 'n' group thyristors are conducting the negative output of A_9 is directed by D_3 to produce a positive output at A_5 and so force Control Amplifier A_3 to a minimum. Thus even if the secondary emf signal generator instructs A_3 to operate it is prevented from doing so until the 'n' group current has ceased and the output of A_9 returned to zero. The capacitor C_4 on A_9 allows the current detector to differentiate between the normal discontinuous current pulses in the cycloconverter output and a true current zero.

Cross Interlock Amplifiers A_6, A_7

These are high gain amplifiers set to be driven to the '0' volt rail, and clamped there by diodes D_7, D_8 , whenever the Control Amplifiers A_3 and A_2 are at a minimum. In this condition they do not produce an output at A_4 but should the output of Control Amplifier A_3 increase from minimum then the Cross Interlock Amplifier A_6 switches to a negative output producing a positive output at A_4 and so forcing Control Amplifier A_2 to a minimum.

Summing Amplifiers A_4, A_5

These amplifiers have a positive output, thereby forcing their associated control amplifier to a minimum whenever the inputs from the brass Interlock Amplifier or from the Current Detector amplifier are negative.

10.8. Operation of the cycloconverter

The three-phase to single-phase cycloconverter was operated into both resistive and inductive loads with simulated secondary emf signal generator inputs from a variable frequency three-phase source. Tests were also taken with an induction motor secondary winding as the load with the machine being driven externally at near synchronous speed. The load was therefore inductive and included a source of emf. Fig.10.27 shows the output current and voltage for this load. The effect of inductance and of the electronic current detector can be clearly seen.

A series of photographs were taken to observe the operation of the control system of Fig.10.26. These are shown in Fig.10.28 and were taken for a very low current of approximately 0.5 A rms. The capacitor across the current detector A_9 maintains the detector output for discontinuous currents and its output only falls to allow the operation

current

voltage

a) 25 ms / cm

current

voltage

b) 5.0 ms / cm

FIG. 10. 27 CYCLOCONVECTOR OUTPUT WAVEFORMS

slip detector
A10

current detector
A9

slip detector

control amplifier
A2

slip detector

output current

25 ms / cm

FIG. 10.28. OPERATION OF THE CONTROL SYSTEM

of the control amplifier after the last output current pulse has been detected. The effect of this and of the response of the control amplifier gives a delay of about 20 mS between changeover of the secondary emf signal generator output and the commencement of output current.

Chapter 11

Appendix 3

Computer Programs for Machine Performance

11.1. Computer prediction of machine performance with external secondary resistors.

The equivalent circuit model is shown in Fig.11.1 and analysis gives the following expressions:-

(i) The secondary phase current referred to the primary side

$$|I_1'| = \frac{V_1 \sqrt{(R_m X_m R_2'/S)^2 + (R_m X_m X_2')^2}}{\sqrt{\left[\frac{R_2'}{S}(L) - X_2'(M)\right]^2 + \left[\frac{R_2'}{S}(M) + X_2'(L)\right]^2}} \text{ amperes (11.1)}$$

where

$$\begin{aligned} L &= R_1(A) - X_1(B) - R_m X_m X_2' \\ M &= X_1(A) - R_1(B) + R_m X_m R_2'/S \\ A &= R_2' R_m / S - X_2' X_m \\ B &= X_2' R_m + R_m X_m + R_2' X_m / S \end{aligned}$$

(ii) The total power crossing the air gap P_r

$$= 3 \times |I_1'|^2 \cdot \frac{R_2'}{S} \text{ watts (11.2)}$$

(iii) The total torque developed at the shaft T

$$= \frac{30 \cdot P_r}{N_s \cdot \pi} \text{ N-m (11.3)}$$

(iv) The mechanical output power P_o

$$= (1 - S) \cdot P_r \text{ watts (11.4)}$$

FIG. 11. 1. EQUIVALENT CIRCUIT OF THE INDUCTION MACHINE

(v) The electric power dissipated externally to the secondary P_2

$$= S.P_r - 3 \cdot |I_1'|^2 \cdot r_2' \quad \text{watts} \quad (11.15)$$

where r_2' is the referred value of the secondary winding resistance.

(vi) The input current to the primary winding

$$I_1 = \frac{V_1 \times [(AL + BM) + j(BL - AM)]}{L^2 + M^2} \quad (11.16)$$

where A, B, L, M, are as previously defined.

(vii) The magnitude of the primary current

$$|I_1| = \frac{V_1 \times \sqrt{A^2 + B^2}}{\sqrt{L^2 + M^2}} \quad \text{amps} \quad (11.17)$$

(viii) The total power into the primary winding

$$\begin{aligned} P_1 &= 3 \times V_1 \times \text{inphase component of } I_1 \\ &= 3 \times V_1^2 \times \frac{(AL + BM)}{(L^2 + M^2)} \quad \text{watts} \end{aligned} \quad (11.18)$$

(ix) The input power factor

$$\cos \phi_1 = \frac{AL + BM}{\sqrt{(BL - AM)^2 + (AL + BM)^2}} \quad (11.19)$$

(x) The overall efficiency

$$\eta = \frac{P_o}{P_1} \times 100 \% . \quad (11.10)$$

In the program

$$V = V_1, A = R_m, B = X_m, Y = r_2, E = X_1, F = X_2'.$$

$$U = R_1 + r_2', Z = \text{turns ratio}, N = \text{speed rpm.}$$

$$R = (R_2 - r_2) \text{ external secondary resistance per phase.}$$

LIST

```

1 DATA 35,32,3,1.5,.13,.13,.27,8.4,6,1500
3 READ V,A,B,Y,E,F,U,R,Z,N
5 PRINT TAB(1)"R2";TAB(7)"S";TAB(12)"I2";TAB(18)"PR";
6 PRINT TAB(24)"T";TAB(30)"PO";TAB(36)"P2";
7 PRINT TAB(41)"I";TAB(47)"W";
8 PRINT TAB(53)"COS";TAB(59)"EFF";
9 PRINT TAB(64)"SP"
10 LET X=1
15 FOR S=-0.1 TO 0.1 STEP 0.01
16 IF INT(S*S*100000)<1 GO TO 70
17 LET P=X*Y
20 LET C=(P+R)/(Z*Z)
22 LET D=U-P/(Z*Z)
24 LET G=A*C/S-F*B
26 LET H=F*A+A*B+C*B/S
28 LET L=D*G-E*H-A*B*F
29 LET M=E*G+D*H+A*B*C/S
31 LET J=SQR((A*B*C/S)*(A*B*C/S)+(A*B*F)*(A*B*F))
33 LET K=SQR((ABS(C*L/S-F*M))^2+(ABS(C*M/S+F*L))^2)
35 LET I2=INT(10*V*J/K)
37 LET O=INT(30*(V*J/K)+2*C/S)
39 LET T=INT(30*O/(N*3.142))
41 LET Q=(1-S)*O/10
43 LET A1=SQR(G*G+H*H)
45 LET A2=SQR(L*L+M*M)
47 LET I=INT(10*V*A1/A2)
49 LET W=INT(30*V+2*(G*L+H*M)/(L*L+M*M))
51 LET Q1=INT(100*(G*L+H*M)/SQR((ABS(H*L-G*M))^2+(ABS(G*L+H*M))^2))
53 LET P2=S*O/10-3*(V*J/K)+2*P/(Z*Z)
54 IFS<0 GOTO 57
55 LET E2=Q*1000/W
56 GOTO 58
57 LET E2=W*10/O
58 LET N1=(1-S)*N
60 PRINT TAB(1)P;TAB(7)S;TAB(12)I2/10;TAB(18)INT(O/10);
61 PRINT TAB(24)T/10;TAB(30)INT(Q);TAB(36)INT(P2);
62 PRINT TAB(41)I/10;TAB(47)INT(W/10);
63 PRINT TAB(53)Q1/100;TAB(59)INT(E2);
64 PRINT TAB(64)INT(N1)
70 NEXT S
80 END
READY

```

11.2. Computer prediction of machine performance with cycloconverter control

(i) To calculate the fundamental rms of secondary current I_{FUND}

Input data. Total rms value of secondary current I_{rms} amps
 Peak value of secondary current I_p amps
 Operating slip S
 Current Detector zero current time T_o seconds
 Current Pulse repetition time T_1 seconds

T_1 is set by the 3 pulse cycle converter at 20/3 mS and T_o is approximately constant at 20 mS. The values of I_{rms} , I_p , S are measured for each test.

Procedure

Calculate the total number of current pulses 'Y' in a quarter cycle of the secondary output frequency from

$$Y = \frac{1 - 100 S T_o}{200 S T_1} \quad (4.27)$$

Calculate the conduction time T_2 for each current pulse

$$T_2 = \frac{2 \cdot T_1}{(1 - 100 S T_o)} \cdot \left(\frac{I_{rms}}{I_p} \right) \text{ secs}$$

If $T_2 < T_1$ calculate the fundamental rms of secondary current

$$I_{FUND} = 180 \cdot S \cdot I_p \cdot T_2 \cdot \sum_{n=1}^{n=Y} \cos \left(\frac{\pi}{2} - \frac{T_o}{2} - n T_1 \right) \quad (4.38)$$

If $T_2 > 1$ calculate the minimum instantaneous current I_{min}

$$I_{min} = -0.6 I_p + \frac{1}{2} \sqrt{(1.2 I_p)^2 - 8.8 \left[I_p^2 - \frac{2 \cdot I_{rms}^2}{1 - 100 S T_o} \right]} \quad (4.32)$$

and so calculate the fundamental rms of secondary current

$$I_{FUND} = 282.8 \cdot S \cdot T_1 \cdot \left[I_{min} + \frac{2}{\pi} (I_p - I_{min}) \right] \cdot \sum_{n=1}^{n=Y} \cos \left(\frac{\pi}{2} - \frac{T_o}{2} - n T_1 \right) \quad (4.39)$$

(ii) To calculate the developed torque from the fundamental rms of secondary current.

The developed torque can be predicted from the current source model (Section 4.3.3).

The total power crossing the air gap

$$P_r = 3.[V_1 \beta I_2 \cos \theta - (\beta I_2)^2 R_1] \quad (4.23)$$

This equation gives the power crossing from primary to secondary for a braking load i.e. a motoring torque. When operating with an overhauling load the secondary emf signal generator changes I_2 through 180° so

$$P_r = -3(V_1 \beta I_2 \cos \theta + (\beta I_2)^2 R_1)$$

or

$$P_r = 3(V_1 \beta I_2 \cos \theta + (\beta I_2)^2 R_1) \quad (11.21)$$

where P_r now represents power crossing the air gap from secondary to primary. These power flows are shown in Fig.4.16.

The developed torque can be obtained from

$$T = \frac{P_r}{\omega} \text{ N-m} \quad (4.3)$$

<u>Input data</u>	Primary voltage per phase	V_1	volts
	Turns ratio of machine winding	β	
	Primary resistance per phase	R_1	ohms

The angle between secondary current and primary voltage θ degrees.

The values of R_1 and β , were determined by machine tests at 0.23 Ω per phase and 6.0 respectively. The primary voltage V_1 must be measured for each test. The angle θ between the secondary current and the primary voltage must be calculated for each test.

Procedure

With the signal generator synchronised to the open circuited secondary voltage its output can be calculated, from Fig.11.1, to be leading the primary voltage V_1 by approximately 0.06 radians.

For any slip S and zero current time T_0 due to the current detectors the angle that the fundamental of secondary current lags the primary voltage is given by $(50 \pi S T_0 - 0.06)$ rads. This is equation 4.41 corrected to allow for the initial synchronising load angle.

The signal generator may be manually operated to give an additional phase shift θ_d which would be input data to the program for each test.

The total lagging angle between secondary current and primary voltage is

$$\theta = (50 \pi S T_0 - 0.06 + \theta_d) \text{ radians} \quad (11.22)$$

Equations 4.3 can now be used to calculate the predicted torque using the calculated value of the rms value of the fundamental of secondary current for I_2 . The value of P_r can be calculated from either 4.23 or 11.21 depending upon whether the machine is developing driving or braking torques.

Alternatively, the measured torque can be presented as input data and equations 4.3, 4.23, 11.21 and 11.22 can be used to calculate the expected value of fundamental secondary current.

(iii) To calculate the primary input current and power factor.

The predicted input current and power factor can be calculated from the rms value of the fundamental of secondary current and its calculated phase angle with respect to the primary voltage, added vectorally to the magnetising and iron loss current. The parallel branch of the equivalent circuit representing the magnetising and iron loss terms was assumed to be directly across the primary voltage.

(iv) The computer program was written to process the results for driving and braking modes at both subsynchronous and supersynchronous speeds.

In the program the symbols used are:-

S = slip, I2 = total rms value of secondary current,
 I3 = peak value of secondary current, V = primary voltage per phase,
 Q5 = manually set phase shift on the secondary emf signal generator,
 H1 = the developed torque (negative for overhauling loads), R3 = R_m the
 iron loss term in the equivalent circuit, X3 = X_m the magnetising
 reactance in the equivalent circuit.

LIST

```

1 DATA.03,2.64,4.8,35.7,-.486,5,32,3
10 READ S,I2,I3,V,Q5,H1,R3,X3
11 PRINTTAB(1)"I-FUND.RMS";TAB(14)"TORQUE N-M";
12 PRINTTAB(25)"PHASE ANGLE";TAB(40)"POWER FACTOR"
14 LET T=0.02
16 LET T1=0.02/3
18 LET T2=(2*T1*(I2/I3)^2)/(1-100*S*T)
19 LET I6=V/(R3*X3)
20 LET Q=50*S*3.142*T
22 LET Q1=Q+Q5-.06
23 LET Q2=INT(Q1*1800/3.142)
24 LET Q3=COS(Q1)
25 LET Q4=INT(1000*Q3)
26 LET Y=(1-100*S*T)/(200*S*T1)
28 LET A=0
30 FOR N=1 TO Y STEP 1
32 LET A=A+COS(1.571-314.2*S*(.5*T +N*T1))
34 NEXT N
36 IF T2>T1 THEN GOTO 39
37 LET E=180*I3*T2*S*A
38 GOTO 48
39 LET C=2.2*(I3^2)-4.4*(I2^2)/(1-100*S*T)
40 LET I1=-.6*I3+.5*SQR((1.2*I3)^2-4*C)
41 LET E=282.8*S*(I1*T1+2*T1*(I3-I1)/3.142)*A
43 IF H1<0 GO TO 46
44 LET D=(-(18*V*E*Q3-24.8*E^2)/(50*3.142)
45 GOTO 53
46 LET D=-((18*V*E*Q3+24.8*E^2)/(50*3.142)
47 GOTO 56
48 IF H1<0 GO TO 51
49 LET D=(18*V*E*Q3-24.8*E^2)/(50*3.142)
50 GOTO 53
51 LET D=(18*V*E*Q3+24.8*E^2)/(50*3.142)
52 GOTO 56
53 LET B1=(I6*R3+E*6*(SIN(Q1)))
54 LET B2=(I6*X3+E*6*Q3)
55 GOTO 62
56 LET B1=(I6*R3-E*6*(SIN(Q1)))
57 LET B2=(I6*X3-E*6*Q3)
58 GOTO 62
62 LET H=INT(100*D)
63 LET F=INT(100*E)
65 PRINT TAB(1)F/100;TAB(14)H/100;TAB(25)Q2/10;TAB(40)Q4/1000
72 LET G=.726*V*Q3
74 LET I5=.5*G-.5*SQR(G*G-25.32*H1)
75 LET B1=(I6*R3+F*6*(SIN(Q1))/100)
76 LET B2=(I6*X3+F*6*Q3/100)
77 LET I7=1.732*(SQR(B1*B1+B2*B2))
78 LET Q6=B2*1.732/I7
79 PRINT TAB(1)"I-PRIDICT";TAB(14)"PF-INPUT";TAB(25)"I IN-LINE"
80 PRINT TAB(1)(INT(I5*10))/10;TAB(14)(INT(Q6*100))/100;
81 PRINT TAB(25)(INT(I7*10))/10
90 END
READY
RUN
I-FUND.RMS      TORQUE N-M  PHASE ANGLE    POWER FACTOR
  1.75          5.97      -25.9          .899
I-PRIDICT      PF-INPUT    I IN-LINE
  1.4           .82       22.2

```

STOP AT LINE 90
READY

Chapter 12

Appendix 4

Load Equipment for Performance Testing

The slip-ring induction motor being controlled by the cycloconverter was coupled to a d.c. swinging frame dynamometer as shown in Fig.5.1. In order to load the induction motor it was necessary to control the armature current of the dynamometer. The usual method is to have a switched load bank of high power resistors connected to the d.c. dynamometer terminals. Resistors can then be selected to give the required load current at a given operating speed. This method is inefficient and, as the armature current for a given load resistor is a function of the generated emf, the load current and, therefore, load torque will change for variations in machine speed.

The test programme for the slip energy recovery control of the induction motor also required an investigation into both braking and motoring modes which require a loading system capable of developing braking or overhauling load torques. To achieve this the dynamometer could be connected to a Motor Generator set to form a Ward Leonard control system.

It was decided to modify the design of the basic three-phase to single phase cycloconverter as described in Appendix 2 to provide a regenerative d.c. controller to the d.c. dynamometer. The schematic diagram for this controller is shown in Fig.12.1. The d.c. dynamometer can be supplied from either the p group or n group thyristors an interlock circuit preventing simultaneous operation.

The direction of load current can be selected to be either out of the positive terminal of the dynamometer when a braking torque is obtained, or into the positive terminal when an overhauling torque can be produced.

FIG. 12.1. SCHEMATIC DIAGRAM OF STATIC LOADING EQUIPMENT

FIG. 12.2. OUTPUT WAVEFORMS FOR GENERATION OF BRAKING TORQUES

a) light loads - discontinuous current

b) heavy loads - continuous current

FIG.12.3. OUTPUT WAVEFORMS FOR GENERATION OF OVERHAULING TORQUES

The circuit waveforms for these two modes of operation are shown in Figs. 12.2 and 12.3. The level of current can be infinitely adjustable up to the full load value of the equipment and a closed loop servo amplifier A_2 maintains this current at the required value irrespective of the speed of the dynamometer. Thus the load torque can be maintained constant during transient speed changes of the test machine. Load torque can also be maintained down to standstill and as the system is regenerative the overall efficiency is high.

Commercial implementation of the cycloconverter slip recovery system would require a resistor starter on the secondary of the slip-ring induction motor. This would be used to bring the motor into the control range of the cycloconverter and would then be switched out of circuit. For the laboratory test programme the static load equipment was modified so that the d.c. dynamometer could be operated under closed loop speed control to bring the induction motor up to a speed where the secondary emf frequency was within the operating capability of the cycloconverter. To achieve this an additional amplifier and mode selector were incorporated, and a d.c. tachogenerator coupled to the dynamometer was used as a speed feedback signal.

Abstract

This work describes a control system in which a cycloconverter is connected between the secondary winding of a three-phase induction machine and the a.c. mains supply to give variable speed sub- and super-synchronous operation.

The operating requirements for a three-phase cycloconverter operating in the secondary circuit of an induction machine have been determined. These considerations show that the cycloconverter control signal must be synchronised to the induced secondary emf of the machine. A novel generator based upon digital circuitry was developed to produce such a control signal.

A theory in which the cycloconverter is represented as a controlled current source in the secondary circuit of the induction machine has been proposed and verified. Computer programs based upon this theory were used to analyse machine performance.

A complete three-phase drive system has been designed and successfully operated over a range above and below synchronous speed with both braking and overhauling loads. Experimental observations have shown that smooth control is possible through synchronous speed.