Sensorless Control of a
Super Synchronous
Slip Ring Induction Machine

Thesis submitted for the Degree of
Doctor of Philosophy
at the University of Leicester

by

Jonathan K.J. Wong  B.Eng (Leicester), AMIEE
Department of Engineering
University of Leicester

January 1998
To the memory of my father,

*Wong Ang Joo*
Sensorless Control of a Super Synchronous Slip Ring Induction Machine

ABSTRACT

This thesis presents the development and verification of a novel sensorless control scheme for a super synchronous slip ring induction machine (SRIM). The novel scheme overcomes the limitations of the encoder-based system, with a reduced cost.

This research re-utilised a LUED slip power facility, previously funded by SERC in the 80s.

The SRIM uses a conventional current source inverter (CSI) connected between the supply and the rotor windings. Two techniques of slip power recovery, the Kramer and the Scherbius system, are discussed. The Kramer system recovers slip power to the supply. The Scherbius system is capable of four-quadrant operation, with the main advantage of twice the rated shaft output power when operating at twice the rated speed.

The CSI is investigated and its commutation problems studied. For stable operation, the commutation sequence in the CSI is synchronised electronically with the varying rotor e.m.f.'s. When motoring sub-synchronously, or generating super-synchronously, the rotor e.m.f.'s are in the direction to aid the commutation process. However, when motoring super-synchronously or generating sub-synchronously, the rotor e.m.f.'s oppose commutation. These limit the performance of the CSI.

Provided the rotor terminal voltages are isolated from the mains, the rotor speed and phase information can be derived from the measured and filtered rotor terminal voltages. The derived signals can synchronise the CSI to the secondary e.m.f., thereby eliminating the need of an encoder. This novel scheme is different to the techniques known for cage type machines, and is not affected by the inherent machine parameter variations.

The SRIM harmonics are analysed. From which, a novel adaptive filter was developed to remove the rotor speed dependent harmonics and to track the varying e.m.f. with negligible phase delay. The filter and the scheme were verified using an analogue method, although a DSP system may provide better flexibility.

Finally, suggestions have been made for further work particularly for sensorless operation of the SRIM.
MEMORANDUM AND ACKNOWLEDGEMENTS

Sensorless Control of a Super Synchronous Slip Ring Induction Machine

The accompanying thesis submitted for the degree of Ph.D. entitled "Sensorless Control of a Super Synchronous Slip Ring Induction Machine" was based on the work conducted by the author in the Department of Engineering, Leicester University, during the period between January 1995 and October 1997.

All the work recorded in this thesis is original unless otherwise acknowledged in the text or by references. None of the author’s work has been submitted for another degree in this or any other University.

The main contributions the author claims to have made to the subject on ‘Sensorless control of a super synchronous SRIM’ are as follows:

1. Harmonic analysis of the electrical quantities, particularly the rotor terminal voltages, of an inside out (the rotor is connected to the mains, and the stator connected to the power bridge. For this thesis, the analysis treats the SRIM as in the conventional layout) wound rotor machine. Analysis is carried out using real time DSP and FFT. A theory to predict the number and magnitude of the induced voltage harmonics is presented. The theory is further verified by experimental results.

2. A Scherbius analysis of the slip ring machine with a CSI connected to the rotor circuits. The analysis allows the prediction of the performance of the slip ring machine, provided the machine parameters are known.

3. Commutation analysis of the CSI, particularly the different commutation processes during sub and super synchronous speed regions. The effect on commutation by the power factor correction capacitors in the rotor side circuit.
4. Introduction of rotor side capacitors on the SRIM. The capacitors reduce voltage spikes caused by CSI, and allows better observation of the rotor terminal voltage.

5. A theory of a novel sensorless control scheme for the slip ring machine. The theory outlines the pre-requisite of sensorless control for the slip ring machine.

6. Implementation of the author's proposed novel sensorless control scheme. Specific suggestion of DSP implementation is included.

7. The account of the operational difficulties of controlling the SRIM from sub to super synchronous speeds.

The author acknowledges the continual inspiration and teaching from Dr. G.A. Smith. The author also acknowledges the enthusiastic supervision and encouragement from Dr. P.W. Lefley. The author acknowledges the constructive technical challenges and feedback from Dr. J.A.M. Bleijs. Finally, the author is grateful to Dr. J.C. Fothergill for the help and encouragement at the final stage of this research.
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<td>$P_1$</td>
<td>Power input to the primary stator</td>
<td>Watts</td>
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<td>$P_2$</td>
<td>Inverter Electrical Power</td>
<td>Watts</td>
</tr>
<tr>
<td>$P_s$</td>
<td>Rotor Electrical Power</td>
<td>Watts</td>
</tr>
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<td>$P_r$</td>
<td>Power Crossing Air Gap</td>
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<td>Generator Mechanical Power</td>
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<td>Synchronous speed of machine shaft</td>
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<td>$\omega_s$</td>
<td>Synchronous Angular Velocity</td>
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<td>$V_{is}$</td>
<td>a.c. inverter supply voltage</td>
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<td>$V_{dr}$</td>
<td>Maximum rectified output voltage at unity slip</td>
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<td>$I_o$</td>
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<td>$I_2'$</td>
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<td>$I_{2t}$</td>
<td>Total rms current</td>
<td>Amps</td>
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<td>$I_{2f}$</td>
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<td>$i_1$</td>
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<td>$i_2$</td>
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<td>$i_0$</td>
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<td>$i_c$</td>
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<td>$R_1$</td>
<td>Primary winding resistance</td>
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<td>$R_2$</td>
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<td>$R_{ts}$</td>
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<td>ohms</td>
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<td>Primary winding reactance</td>
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<td>$X_{ts}$</td>
<td>Total primary equivalent series reactance</td>
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<td>$L_e$</td>
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<td>$L_f$</td>
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<td>$C_{x,y,z}$</td>
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<td>$\alpha$</td>
<td>Firing angle of controlled rectifier</td>
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<td>Inverter Power factor</td>
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Chapter 1

Sensorless Control of Super-Synchronous Slip Ring Induction Machine

1.1 Introduction

Slip Ring Induction Machines (SRIM) are the workhorses for industry with a considerable history. The SRIM is receiving a renewed interest, due to its doubly fed advantages and in wind energy conversion. With the stator permanently connected to the mains, and power controlled to flow into or out of the rotor windings, excellent torque and speed control can be achieved for variable speed motoring or generating operation. The SRIM has tremendous cost benefits when operating in the super-synchronous mode, when the shaft power can be twice greater than the rated values. Sensorless control of the SRIM in four-quadrant operation is a challenge not commonly attempted.

Variable speed control of SRIM may be achieved by simply dissipating the rotor power into resistor banks [15, 70, 78]. However, the drive efficiency is very low and speed regulation is poor. When the drive requirements include good efficiency, dynamic response and accurate speed or torque regulation, then such open loop control is unsatisfactory. It is necessary to operate the SRIM in a closed loop mode, especially when the dynamic operation of the drive has an important effect on the overall performance of the system.
Usually, a SRIM control scheme relies on a position encoder for closed loop operation. The mechanical encoder is undesirable due to high initial cost, low reliability and poor noise immunity. A control system without the use of any mechanical encoders, i.e. sensorless control, can eliminate the disadvantages associated with the encoders with a reduced initial costs. Moreover, in certain applications, the operating conditions may even prohibit the presence of an encoder. Although many sensorless approaches have been presented for the cage type induction machine, sensorless control of the SRIM remains a field not commonly attempted.

This research is possible after the author's re-commissioning and further improvement work on an encoder based SRIM slip recovery facility at Leicester University. The machine and drive was previously funded by SERC in the 80s.

For systems of low power levels, the costs for the control system may equal or exceed that for the power electronics. However, for large systems, the cost of power electronics and the amount of slip power become very significant. Theoretically, the SRIM in the Scherbius configuration can deliver twice the rated shaft power at twice the synchronous speed. In addition, the cost of the power electronics can be relatively low, to handle the slip power only. Therefore, the SRIM becomes an attractive option where a high output with a limited space and low cost is involved. For example, the SRIM operating as a generator can provide twice the shaft power using the same size power electronic system.
This thesis presents the development and verification of a novel theory required for the sensorless control for a super-synchronous SRIM. An encoder based slip power recovery system using a current fed inverter connected to the rotor of a SRIM was reviewed and studied [74,75]. The system characteristics and experimental observations allowed the development of the novel sensorless control system.

1.2 Slip Power Recovery Systems

When operated from the fixed frequency grid, the SRIM operates at a constant speed. In earlier days, speed control of the SRIM below synchronism was done by dissipating the slip power through the external resistor banks. This type of control has a very poor efficiency considering the large amount of slip power lost.

Instead of wasting the slip power, the electrical energy can be recovered by feeding it back to the supply system. This brings a significant improvement to the drive efficiency. As the secondary e.m.f. voltage and frequency vary with the rotor speed, direct connection between the slip rings and the supply was impossible. Therefore, before the introduction of power electronic inverters, auxiliary machines were used for slip power recovery [63].

Many authors have presented methods for slip recovery from the secondary rotor circuit, which utilised either mechanical or electrical methods [6, 7, 8, 15, 17, 20, 21, 27, 31, 32, 55, 65, 70, 71, 74, 75, 82, 98]. They are generally classified as the Kramer or the Scherbius systems. For the earlier Kramer system, the slip power of the
SRIM was mechanically recovered. This was achieved via the shaft of a d.c. motor.

For the early Scherbius system, power is frequency converted and electrically returned to the mains supply using a second machine. This is shown in Fig. 1.2.

Both the early Kramer and Scherbius systems employed auxiliary machines such as the d.c. or the a.c. commutator motors and synchronous motors. In modern configurations, solid state rectifiers and inverters replaced most or all of the auxiliary
machines. This improved drive efficiencies and significantly reduced costs and maintenance.

Although slip power recovery is generally applied to Kramer or Scherbius schemes using SRIM with a separate external recovery system, it is noted that the a.c. commutator motors running at sub-synchronous speed also feed slip power back into the supply system [63].

1.2.1 **Static Kramer Sub-Synchronous Slip Recovery**

For sub-synchronous operation, a static converter replaces the d.c. machine of the early Kramer drive.

![Fig. 1.3 Static Kramer operation for Induction Machine](image)

This system consists of a three-phase rectifier bridge connected to the rotor, and cascaded with a d.c. link and a mains side converter. The rectified slip power is return to the a.c. supply via the d.c. link and the frequency matching converter. Both the rectifier and the inverter are naturally commutated by the alternating e.m.f. at the secondary output terminals and the a.c. supply, respectively.
The performance characteristics of the Static Kramer system have been analysed in detail by many authors [6, 8]. The torque developed by the system was shown to be proportional to the fundamental component of the secondary current from the d.c. link.

The principal disadvantage of the sub-synchronous drive is the low power factor [73]. Several circuit modifications have been devised to counter this disadvantage. A special type of 'through pass' inverter has been developed to replace the phase controlled inverter. This inverter circuit uses forced commutation and returns slip power supply without drawing large amounts of reactive power.

The static Kramer drive is common in situations where limited sub-synchronous speed range is adequate. The system is particularly advantageous at high power levels because of better overall efficiency.

1.2.2 Static Scherbius Sub & Super-Synchronous Slip Recovery

For sub and super-synchronous operation, a fully controlled inverter replaces the rectifier bridge in the Kramer set-up. The Scherbius configuration allows the SRIM speed to reach beyond synchronism, by allowing power to flow into both the rotor and stator windings [98].
When the current is injected, in or out of phase to the induced rotor e.m.f., the power in the secondary rotor circuit can flow in either directions via the d.c. link. However, difficulties are experienced near synchronism, when the slip frequency e.m.f. vanishes and phase commutation is impossible, if thyristors are used. A special forced commutation method becomes necessary. Such commutation circuits must be controlled according to the secondary induced e.m.f. of the SRIM. Many authors have presented methods to implement such CSI system [75].

Smith [75] described a well-known and reliable technique. He describes an encoder based electronic signal generator with a three-phase square wave output locked in phase with the secondary e.m.f. of the SRIM. The constant amplitude square waves can be used through additional logic for phase adjustments, to control the precise firing points of the current source inverter.
1.3 **Power Electronic Configurations for Super-Synchronous Operation**

To drive the SRIM above synchronism, power has to be controlled to flow into or out of the rotor for either motoring or generating operations. Two types of power electronic configurations have been developed, which do not require the induced e.m.f. for commutation. The two systems are the Cyclo-converter and the Current Source Inverter (CSI).

1.3.1 **The Cyclo-converter**

The naturally commutated cyclo-converter is connected between the secondary windings of a SRIM and the a.c. mains [10, 24, 43]. This configuration allows reversible power flow, and speed control is possible for sub and super-synchronous operations by controlling the injected secondary voltage. Although the power devices can be rated at a fraction of the total power, this configuration requires 18 switching devices as shown in fig. 1.6.

The cyclo-converter is an a.c. to a.c. frequency ‘step-down’ converter. The generated output frequency is lower than the primary supply frequency. A basic three phase to three-phase set-up requires 36 line commutated power electronic switches. The quantity of power electronics may increase the cost of the drive in large systems. The power circuit in each phase consists of positive and negative groups of thyristor switches, which allow bi-directional flow of power into or from the mains supply.
It is difficult to prevent excessive circulating currents between the positive and the negative thyristor groups. For continuous current operation, a suitable reactor is inserted between the groups to limit the circulating current. The continuous circulating current operation has the advantage of reduced output voltage harmonics and increasing the output frequency range. However, in practice, the circulating current mode would only be used when the load current is low. This is to maintain a continuous load current in a better waveform.

The cyclo-converter may be attractive for large power installations with a limited speed range above and below synchronous speeds. A possible advantage of the drive is that regeneration is simple and the system can be easily designed for four-quadrant operation.
1.3.2 The Current Source Inverter

The limitations of thyristors prevented the use of conventional voltage source (VS) inverters for super-synchronous operation (inverter grade thyristors are expensive). The reverse connected diodes across the thyristors can prevent the reversal of the D.C. link voltage, so that any regeneration has to involve current reversal through an additional fully controlled line commutated inverter. This may be a practical difficulty. However, the introduction of IGBTs and the superior advantages of modern power electronics systems, allow the VS inverter configuration to be a possibility [84].

The use of the standard current source inverter (CSI) is well known. The CSI allows reversal of power flow and is simple using conventional non-inverter grade thyristors. The D.C. link voltage is allowed to reverse with a constant current direction. With a reversed voltage, the converter can be used to feed power into or out of the secondary circuit of the induction machine. By a suitable sequence of the CSI, operation in any of the four quadrants is possible [75].

![Fig. 1.7 CSI Slip Recovery System](image_url)

The CSI is inherently reliable as the current source reactor limits any large
current surges during fault conditions. The quasi-square wave current generated is not a particular disadvantage in slip energy recovery scheme. The stator winding is directly connected to the sinusoidal mains, and therefore the machine has a sinusoidal flux. As sub-harmonics are low, cogging is very unlikely.

The CSI drive is suitable for applications where abrupt changes in shaft torque should be avoided. The d.c. link choke prevents sudden current changes, which in turn limits the torque changes. A standard auxiliary commutated current source inverter (CSI) which uses six capacitors and six diodes for commutation is shown as follows.

![Fig 1.5 CSI - Forced Commutated Bridge](image)

The advantage of the auxiliary commutated CSI is its ability to commutate regardless of the a.c. voltages. The stored capacitor voltages provide commutation. The diodes isolate the capacitors from the load, where energy is stored for commutation.
When a thyristor is gated on, the current through the outgoing thyristor is diverted through a pre-charged capacitor. This allows the thyristor to turn off. After a gated thyristor is fully turned on, the commutating capacitor is reverse charged by the direct current from the current source.

The disadvantage of the CSI is its speed limitation due to the commutation mechanism. For the total commutation time to be short for a given current, the capacitor must be sufficiently small for rapid charging and discharging. However, as the capacitor absorbs the stored energy in the machine windings, then a small capacitance causes a large voltage rise above the normal line voltage. The commutation time is also a function of the d.c. link current. Excessive commutation angles or abnormal behaviour may result during light load conditions. If the capacitor is too small, then there may not be sufficient stored energy for commutation [7, 12, 63, 84].

Also, at super-synchronous speed the induced e.m.f. will oppose the commutation process. This problem is noted and presented in chapter 4.

1.4 **Sensorless Control for Slip Ring Machines**

The use of an encoder within a drive system has many disadvantages such as costs, size, reliability, alignments and maintenance [13,56]. A number of sensorless vector control techniques for the squirrel cage type induction machine control have been proposed. Principally, these techniques utilise the rotor position or speed information through measurement and the further processing of the machine terminal
quantities. The sensorless techniques can be mainly classified as follows:

1. Open loop speed control with slip compensation.
2. Closed loop control with speed estimation

The two main methods may be further broadly classified into the following: slip frequency calculation method, speed estimation using state equations, speed estimation by slot harmonic voltages, flux estimation and space flux observer, direct control of torque and flux, observer-based speed sensorless control, model reference adaptive control, Kalman filtering techniques, sensorless control with parameter adaptation, neural network based sensorless control, and etc.

A serious problem with the development of many sensorless schemes, is that the robustness of the control may be adversely affected by the changing machine parameters [35, 44, 45, 54, 60, 64, 72, 77]. Many sensorless schemes rely on the dynamic machine equations from the Generalised Machine Theory. The speed of the machine is calculated from the estimated stator and rotor resistance and inductance. Unfortunately, such quantities vary due to machine thermal effects and result in considerable system complexity and poor robustness. For all machine types, a sensorless scheme that is independent of machine parameters is desired [72].
Unlike the cage type machine, the SRIM has a secondary winding which has a large induced voltage dependant on rotor speed. A controlled current can be injected in or out of phase to this induced voltage, to provide the desired machine operation. As both the angle and the magnitude of the rotor current can be controlled separately, this is a special type of Vector Control [75].

1.4.1. The Proposed Approach

From experimental observations, it was realised that measurement and further processing of rotor terminal voltages can provide rotor position and speed information. When electrical isolation from the mains is provided, the rotor terminal voltages are a good approximation of the induced rotor e.m.f. An analogue adaptive filter system was developed and implemented to track the terminal voltages, with minimum or negligible phase errors.

For sensorless control of the SRIM, two essential parameters are required:

1. The rotor position.
2. The rotor speed.

The common sensorless techniques mentioned for the cage type machines do not suggest any easy or related solution for the estimation of SRIM rotor position and speed information. Furthermore, the SRIM has the following operating characteristics, which present problems in determining both position and speed information.
1. The induced rotor e.m.f. varies in both phase sequence, frequency and magnitude, according to the rotor speed.

2. At synchronous speed, there is zero induced fundamental e.m.f.

3. The induced e.m.f. contains very rich harmonics at all speeds. The dominating harmonics are the 5th and 7th.

4. Due to the machine slots, other harmonics are also generated.

5. At super-synchronous speed, the rotor phase sequence is reversed, i.e., the RBY sequence at sub-synchronous will become RYB sequence.

As direct measurement of electrical quantities is used, the inherently rich machine harmonics result in very poor signal to noise ratio. Furthermore, as the stator is connected to the mains, and quasi-square wave current feeds the rotor, the drive system is extremely noisy. Signal processing of the measured terminal quantities is absolutely necessary.

A novel sensorless control technique for the SRIM is developed and verified. The rotor position/speed information is derived from the analogue measurement and signal processing of the rotor terminal voltages. Although a DSP system may provide a more flexible and complete solution [39, 72, 84], the aforementioned method was adopted to illustrate the theory within the limited cost and short time constraints.
CHAPTER 2

Kramer Slip Recovery System

2.0 The Kramer System

The Kramer sub-synchronous static cascade is a common slip power recovery system found in very large installations that have a limited speed range. Instead of using auxiliary machines, control can be better achieved by using power semiconductors. The power rating of the recovery equipment can be smaller than the slip ring machine’s rated power, and is proportional to the speed range required.

The Kramer power electronics configuration allows higher efficiency, lower maintenance and a more compact system than using auxiliary machines. Theoretically, the speed range covers from zero to near synchronous speed. A sufficient starting torque can be achieved under current control, without any extra starting resistive elements. However, the recovery equipment must then be rated at the full power of the drive for starting. Top speed is restricted below synchronous speed as the secondary e.m.f. will not be sufficient to circulate the required current as the speed approaches synchronism [8,28,31].

The power factor and the r.m.s. value of the a.c. current are important considerations for the Kramer drive. A simple single-phase model of the drive system is presented for the evaluation of the r.m.s. secondary current waveform and the power factor. The analysis in section 2.2 allows the prediction of the total and the fundamental r.m.s. secondary current, and also the total harmonics generated by the system.
2.1 **Kramer Analysis**

Although much work has been done for the Kramer system, it is reviewed and further analysed in this chapter as a foundation for the development of a novel analysis of the Scherbius system presented in Chapter 3.

The analysis of the machine using a single phase equivalent model and the use of a system equivalent circuit are presented. These allow the prediction of the steady state operation of the Kramer system.
2.1.1 Single Phase Machine Model

Considering only one phase of the SRIM.

The stator winding is represented by the series resistance $R_1$ and reactance $X_1$, with the parallel magnetising impedance $R_m$ and $X_m$. The rotor current, $I_2'$, will be forced by the diode rectifier bridge to be in-phase with the rotor e.m.f., $V_2'$. Therefore, the rotor current will be always in-phase with the rotor e.m.f.

The induced voltage $V_2'$ has a quantity $V_2'=sE_0$, where $E_0$ is the standstill e.m.f. Choosing $V_2$ as the reference phasor, the phasor diagrams in the sub-synchronous motoring and super synchronous generating regions are as follows:
\[ \theta = \text{angle between } I_2, V_2 \]

\[ \theta \text{ is set at } 0 \text{ by power electronics in Kramer} \]

Fig. 2.3 Super Synchronous Generating

Fig. 2.4 Sub Synchronous Motoring

A delta-delta model is assumed in this analysis. From Fig. 2.2, the rotor current \( I_2 \) is set electronically by the power electronics to an angle \( \theta \), referenced to the rotor terminal voltage \( V_2 \). In the Kramer system, the angle \( \theta \) is zero.
The rotor e.m.f. \( E \) is therefore:

\[
E = V_2 + (R_2 + jX_2)(I_2 \cos \theta - jI_2 \sin \theta)
\]

\[
= V_2 + I_2(R_2 \cos \theta + X_2 \sin \theta) + jI_2(X_2 \cos \theta - R_2 \sin \theta)
\]

\[
= V_2 + a + b
\]

\[
\ldots \quad (2.1)
\]

where,

\[
a = I_2(R_2 \cos \theta + X_2 \sin \theta), \quad b = jI_2(X_2 \cos \theta - R_2 \sin \theta)
\]

The stator current \( I_1 \) is:

\[
I_0 = \frac{E}{R_m} + \frac{E}{jX_m}
\]

\[
I_1 = I_2 + I_0
\]

\[
\therefore I_1 = \left( I_2 \cos \theta + \frac{E}{R_m} \right) + j \left( -I_2 \sin \theta - \frac{E}{X_m} \right)
\]

\[
\ldots \quad (2.2)
\]

The stator voltage \( V_1 \) is:

\[
V_1 = E + I_1(R_1 + jX_1)
\]

\[
= E + \frac{R_1}{R_m}I_1 + E\frac{X_1}{X_m} + j\left( \frac{E}{X_m} - \frac{E}{R_m} \right)
\]

\[
+ R_2I_2 \cos \theta + X_1I_2 \sin \theta + j \left( X_1I_2 \cos \theta - R_1I_2 \sin \theta \right)
\]

\[
\ldots \quad (2.3)
\]

\[
V_1 = E \left[ \left( 1 + \frac{R_1}{R_m} + \frac{X_1}{X_m} \right) + j \left( \frac{X_1}{X_m} - \frac{R_1}{R_m} \right) \right] + (R_1I_2 \cos \theta + X_1I_2 \sin \theta) +
\]

\[
\ldots \quad (2.4)
\]

\[
V_1 = E \left[ e + jf \right] + c + jd
\]

where \( e, f, c \) and \( d \) denote the terms in the parentheses.
\[ V_1 = \left( V_2 + a + j \cdot b \right) \left[ e + j \cdot f \right] + c + j \cdot d \]
\[ V_1 = \left( V_2 \cdot e + g \right) + j \left( V_2 \cdot f + h \right) \]

\[ \text{Where,} \]
\[ g = ae - bf + c, \quad h = be + fa + d \]

As the stator voltage \( V_1 \) is known, then \( V_2 \) can be found.

\[ V_1^2 = (V_2 \cdot e + g)^2 + (V_2 \cdot f + h)^2 \]
\[ V_2^2 \left( e^2 + f^2 \right) + V_2^2 \left( g \cdot e + f \cdot h \right) + \left( h^2 + g^2 - V_1^2 \right) = 0 \]
\[ V_2^2 \cdot k + V_2 \cdot L + M = 0 \]
\[ \therefore V_2 = \frac{-L \pm \sqrt{L^2 - 4k \cdot M}}{2k} \]

\[ \text{.... (2.5)} \]

The power crossing through the air gap from the stator to the rotor is \( P_r \). The power does so in a constant rotating field and the torque is:

\[ T = \frac{P_r}{\omega} \text{ Nm} \]

\[ \text{.... (2.6)} \]

Considering only one single phase, the real power \( P_r \) is the in phase product of \( I_1 \) and \( E \), i.e. in phase product of:

\[ I_1 (\cos \theta - j \sin \theta) \left[ (V_2 + a) + j b \right] \]
The real power is expressed as:

\[ P_r = I_2 \cos \theta (V_2 + a) + b \cdot I_2 \sin \theta \]

\[ \therefore P_r = I_2 (V_2 + a) \]

\[ \therefore \theta = 0 \]

\[ \ldots (2.7) \]

The total power crossing the air gap is:

\[ P_r = 3 \left[ I_2 \cos \theta (V_2 + a) + b \cdot I_2 \sin \theta \right] \]

\[ T = \frac{P_r}{\omega_s} \quad \text{Nm} \]

\[ P_o = T \cdot \omega \quad \text{Watts} \]

\[ \ldots (2.8) \]

The electrical power out of the rotor to the supply is

\[ P_o = P_r - 3 I_2^2 R_2 - P_o \]

\[ \ldots (2.9) \]

Consider only one phase of the stator power. Using \( V_2 \) as the reference phasor, the phasor diagram is as follows:

![Fig. 2.5 Phasor Diagram of Power](image)
\( P_1 = V_1 I_1 \cos \phi_1 \)

\[
I_1 = I_2 \cos \theta + \left( V_2 + a + j b \right) \frac{1}{R_m} - j \left( I_2 \sin \theta + \left( V_2 + a + j b \right) \frac{1}{X_m} \right)
\]

\[
= \left( I_2 \cos \theta + \frac{V_2 + a}{R_m} + \frac{b}{X_m} \right) + j \left( \frac{b}{R_m} - I_2 \sin \theta - \frac{V_2 + a}{X_m} \right)
\]

\[ \cdots (2.10) \]

\[
I_i = p + j q
\]

\[
|I_i| = \sqrt{p^2 + q^2}
\]

\[ \cdots (2.11) \]

\[
\therefore \quad V_i = (V_2 \cdot e + g) + j (V_2 \cdot f + h)
\]

\[
\theta_i = \cos^{-1} \frac{p}{I_i} \quad \text{and} \quad \theta_2 = \cos^{-1} \frac{(V_2 e + g)}{V_1}
\]

\[
\cos \phi_1 = \cos (\theta_2 - \theta_i)
\]

Hence \( p \) can be calculated. Assuming negligible losses in the inverter,

\[
\sqrt{3} V_1 \cdot I_1 \cdot \cos \phi_2 = P_{r2}
\]

\[
\therefore \cos \phi_2 = \frac{P_{r2}}{3 \cdot V_1 \cdot I_2}
\]

\[ \cdots (2.12) \]
The power into the stator

\[ P_i = \sqrt{3} V_i I_i \cos \phi_i \]  

\[ \text{.... (2.13)} \]

### 2.1.2 Further Consideration of the Single Phase Model

Referring to the Fig 2.2, and neglecting losses in the machine, the total impedance of the stator winding can be expressed as:

\[ R_s = \frac{R_i \times Z_m^2}{R_i^2 + X_T^2} \]  

\[ \text{.... (2.14)} \]

\[ X_s = \frac{Z_m (R_i^2 + X_i \cdot X_{10})}{R_i^2 + X_T^2} \]  

\[ \text{.... (2.15)} \]

Where,

\[ R_i \] = Stator winding resistance

\[ X_i \] = Stator winding impedance

\[ Z_m \] = Magnetising branch impedance

\[ X_T \] = \( X_i + Z_m \)

The stator e.m.f is therefore,

\[ E_S = |V_i| \cdot \frac{Z_m}{X_T} \]  

\[ \text{.... (2.16)} \]
Let $\beta$ be the machine effective turns ratio and $S$ is the per unit slip. The total impedance of the machine is the sum of the referred equivalent primary impedance and the secondary impedance,

\[
X_e = X_t \cdot \beta^2 + (S \cdot X_2) = 2\pi f \cdot L_e
\]
\[
R_e = R_s \cdot \beta^2 + R_2
\]

Fig. 2.6 Developed Single Phase Model of Kramer System

Assuming low losses in the power devices, the bridge rectifier connected to the secondary is presented as a variable d.c. voltage source $V_{dr}$. $V_{dr}$ is also dependent on the slip voltage at the secondary terminals.
\[ V_{dr} = 1.35 \cdot S \cdot E_{so} \quad \text{.... (2.17)} \]

Assuming the line voltage \( e_{ac} \) connected to the controlled thyristor bridge is sinusoidal and contains no harmonics,

\[ e_{ac} = V_M \sin(\omega t + \phi) \quad \text{.... (2.18)} \]

where,

\[ V_M \] is the maximum supply voltage

\[ \phi \] is the total delay angle \( \left( \phi = \frac{\pi}{3} + \alpha \right) \).

\[ \alpha \] is the firing angle

2.2.2 D.C. Link Current

With reference to the work by Smith and Goodfellow, for one single phase, the differential equation describing the operation of the Kramer system is:

\[ e_{ac} + V_{dr} = 2R_e \cdot I_T + \left(2L_e + L_F\right) \frac{dI_T}{dt} \quad \text{.... (2.19)} \]

where,

\[ 2R_e \] is the equivalent machine resistance referred to d.c. side

\[ 2L_e \] is the equivalent machine inductance referred to d.c. side

\[ L_F \] is the d.c. link inductance

\[ V_{dr} = 1.35s \cdot E_{so} \]
The current in the d.c. link therefore can be determined.

\[ V_d + V_M \sin(\omega t + \phi) = I_T R + L \frac{dI}{dt} \]

\[ \ldots (2.20) \]

where \( V_m \) is the maximum supply voltage. Hence,

\[ V_d + V_M (\sin \omega t \cdot \cos \phi + \cos \omega t \cdot \sin \phi) = I_T R + L \frac{dI}{dt} \]

\[ \ldots (2.21) \]

Taking the Laplace Transform,

\[ V_M \left( \frac{\omega}{p^2 + \omega^2} \cdot \cos \phi + \frac{p}{p^2 + \omega^2} \cdot \sin \phi \right) + \frac{V_d}{p} = R \cdot I_p + L \cdot I_p - L \cdot I_0 \]

\[ \ldots (2.22) \]

therefore, \( I_p \) is expressed as

\[ I_p = \frac{V_M}{L} \left| \frac{\omega \cdot \cos \phi + p \cdot \sin \phi}{p + \frac{R}{L}} \right| \left( \frac{p^2 + \omega^2}{p + \frac{R}{L}} \right) + \frac{V_d}{P \cdot L \cdot (P + \frac{R}{L})} + \frac{I_0}{P + \frac{R}{L}} \]

Let the time constant

\[ \delta = \frac{R}{L} \]

then,

\[ I_p = \frac{V_M}{L} \left( \frac{A}{P + \delta} + \frac{B \cdot P + C}{p^2 + \omega^2} \right) + \frac{V_d}{P \cdot L \cdot (P + \delta)} + \frac{I_0}{P + \delta} \]

where,
A = \frac{\omega \cos \phi - \delta \sin \phi}{(\delta^2 + \omega^2)}

B = -A

C = \sin \phi + \frac{1}{\delta} \left( \frac{\omega \cos \phi - \sigma \sin \phi}{1 + \frac{\omega^2}{\delta^2}} \right)

then,

I_p = \frac{V_m}{L} \frac{A - \frac{AP}{(p + \omega)^2} + \frac{C}{(p^2 + \omega^2)}}{p + \delta} + \frac{V_\omega}{PL(p + \delta)} + \frac{I_o}{(P + \delta)}

\ldots (2.23)
Taking the time domain value, then the d.c. link current equation is

\[
I_t = \frac{V_d}{2R_e} (1 - e^{-8t}) + \left[ \frac{V_m}{2L_e + L_f} \{ A(e^{-8t} - \cos \omega T) - A \cos \omega t \} + I_o \cdot e^{-8t} \right]
\]

Therefore, the current in the d.c. link is

\[
I_T = \frac{V_d}{2R_e} (1 - e^{-\alpha t}) + \left[ \frac{V_m}{2L_e + L_f} \{ A(e^{-\alpha t} - \cos \omega T) + \left( \frac{\sin \phi + A \sigma}{\omega} \right) \sin(\omega t) \} + I_o \cdot e^{-\alpha t} \right]
\]

\[
\ldots \ (2.24)
\]

where,

\[
A = \left( \frac{\omega \cos \phi - \delta \sin \phi}{\omega^2 + \sigma^2} \right), \quad \sigma = \frac{2R_e}{2L_e + L_f}, \quad \phi = \frac{\pi}{3} + \alpha
\]

2.2 Total Root Mean Square and Fundamental Root Mean Square

Secondary Currents

Assuming the secondary per phase line current is a square wave with a pulse width of \(2/3 \pi \text{ (120°)}\) of the secondary e.m.f. period. For a perfectly flat topped current waveform, then the link inductance must be approaching infinity. In actual case, the d.c. link inductance is of limited values.
The analysis of a perfect square waveform using Fourier series taking X-X (in fig. 2.8) as the zero axis indicates that the harmonic components are of the order 6K±1, where K is an integer.

For a perfect square wave the total and fundamental r.m.s. currents are given by:

\[ I_{2T} = \frac{2}{\sqrt{3}} I \]
\[ = 0.816 I \]
\[ = 0.816 I_{\text{MAX}} \]

\[ I_{2F} = \frac{6}{\sqrt{\pi}} I \]
\[ = 0.78 I \]
\[ = 0.78 I_{\text{MAX}} \]

\((I \approx I_{\text{MAX}} \therefore L_F \approx \infty)\)

\[ \ldots (2.25) \]

In practice, the current wave is not flat topped because it contains a ripple component depending upon the inductance and circuit parameters. If the instantaneous peak value of the current is \( I_{\text{MAX}} \), then the total secondary r.m.s. current \( I_{2T} \) is,
\[ I_{2T} = A_1 I_{MAX} \]

and, the fundamental r.m.s. \( I_{2f} \) is

\[ I_{2F} = A_2 I_{MAX} \]

Equation (2.26)

where \( A_1 \) and \( A_2 \) are constant correction factors for the choke and machine inductance. The effect of varying the correction factor and the link inductance has already been discussed by Smith [15].

The presence of harmonics in the d.c. link current, which could be high for small choke values, will allow harmonic currents in the secondary circuit. These will increase secondary losses and so reduce the overall system efficiency. Increasing the link inductance may be impractical and the filtering effect is limited.

The distortion relating to a particular harmonic content in a waveform can be expressed as the relative magnitude of the total r.m.s. harmonic current of order \( n \) to the r.m.s. amplitude of the fundamental. The total harmonic distortion factor is the ratio of the r.m.s. value of all the harmonic components together, to the r.m.s. amplitude of the fundamental. Total harmonic distortion factor is given as follows.

\[ \text{THD} = \frac{\sqrt{I_{2T}^2 - I_{2F}^2}}{I_{2F}} \]

Equation (2.27)
2.3 Developed Torque and the d.c. Link Current Relationship

The relation between the d.c. link current, $I$, and the developed torque, $T_d$, has been studied by many authors [6,38]. The torque developed is independent of the slip of the motor and is directly proportional to the link current.

$$T_d = \frac{P_R}{\omega_s} = \frac{P_R}{2\pi N_s} \text{Nm}$$

where,

$$P_R = \frac{P_s}{s} = \frac{\sqrt{3} s E I \cos \phi}{s} \text{Watts}$$

Assuming the losses in the secondary circuit are low, $P_s = V_{dr} I$, which is the d.c. link power, then,

$$P_R = \frac{1.35 \cdot s \cdot E_s \cdot I}{s} \text{Watts}$$

Hence,

$$T_d \approx 12.89 \frac{E_s}{N_s} I \text{Nm}$$

$$\Rightarrow T_d \approx K_D I \text{Nm}$$

.... (2.30)
To maintain a constant developed torque, the fundamental r.m.s. current must also be constant. As the choke size is reduced, the instantaneous maximum current will be increased.

A high ripple content in the secondary current will generate more harmonic currents in the primary winding by transformer action. These current harmonics will cause additional harmonic torque pulsation and losses, which necessitate some de-rating of the machine. The cost and size of a d.c. choke is considerable. For economical reasons it is necessary to minimise this value.

### 2.4 Power Factor in the Secondary Circuit

The true power factor in the secondary circuit is difficult to analyse due to the inevitable distorted current waveform due to the limited real inductance, distorted voltage due to the SRIM and mains pollution, and also the CSI action. A Voltech power analyser is capable of analysing the true power factor of the secondary circuit. For analysis purpose, the power factor is derived.

Referring to the secondary circuit of the SRIM, if the firing angle $\alpha$ is as defined in section 2.2, then the angle $\phi_i$, which is the angle between the fundamental secondary current and the supply volts is as follows:

$$\cos \alpha = \cos \phi_i$$

The fundamental inverter power transmitted to the supply is
The apparent inverter power is given as

\[ P_{1F} = \sqrt{3} V_1 I_{2F} \cos \phi_1 \]

\[ = \sqrt{3} V_1 I_{2F} \cos \alpha \text{ Watts} \]

\[ \quad \quad \quad \ldots (2.31) \]

The secondary power \( P_s \) is given by

\[ P_s = \sqrt{3} s E_s I_{2F} \cos \phi_2 \text{ Watts} \]

\[ \quad \quad \quad \ldots (2.34) \]

where, \( \phi_2 \) is the angle between the fundamental secondary current and the slip e.m.f.

Neglecting any secondary losses,

\[ P_{1F} = P_s \]

therefore,

\[ \cos \phi_1 = \frac{s E_s}{V_1} \cos \phi_2 \]

and

\[ \cos \alpha = \cos \phi_1 \]

\[ = \frac{s E_s}{V_1} \cos \phi_2 \]
then from previous equations,

\[ \cos \phi_1 = \cos \alpha \]

the effective inverter power is

\[ P.F. = \frac{sE_s \cos \phi_2}{V_I} \]

\[ \ldots (2.35) \]

Including the machine turns ration, \( \beta \), then

\[ P.F. = \left( \frac{\beta_0 V_I}{\beta_1 V_{Is}} \right) \cos \phi_2 \]

\[ = K_0 \cos \cos \phi_2 \]

\[ \ldots (2.37) \]

where,

\( \beta_0 \) is the SRIM turns ration
\( \beta_1 \) is the optional recovery transformer turns ratio
\( V_I \) is the Stator input voltage
\( V_{Is} \) is the Inverter supply voltage
\( K_0 \) is the distortion factor

To maximise the inverter power factor, the machine turns ratio should be chosen as unity. The open circuit secondary voltage should match the supply. Various techniques have been suggested to improve the inverter power factor. The power factor can be compensated by providing a capacitor bank across the supply terminals, which also serve as a short circuit path to current harmonics.
Another method is to reduce the duration of conduction of each thyristor of the controlled bridge. This is achieved by a circuit known as a 'through-pass inverter'. The output power factor can be further improved by using forced commutation with a modified 'through-pass inverter'.

In the Kramer systems, power factor will be less than unity, even if the turns ratio is chosen to be unity. To achieve on the machine power factor of unity, the diode bridge must be replaced by a controlled thyristor bridge in which the firing control to the thyristors is synchronised to the secondary e.m.f.

A Scherbius system [7, 27, 31, 51, 55, 75] is described in Chapter 3, in which a current source inverter connected to the rotor is used. This system has many advantages over the 'Through-pass Inverter' [49], although both systems require complex control and forced commutation circuits compared with conventional static Kramer equipment.
CHAPTER 3

Scherbius Slip Recovery System

3.1 The Scherbius Slip Recovery System

The Scherbius system permits bi-directional power flow in the secondary rotor circuits, and hence allows operation in both the sub and super synchronous speed regions. A cyclo-converter and a current source inverter were mentioned in Chapter 1, to illustrate the recovery of slip power from the secondary circuit.

The cyclo-converter has the advantage of natural commutation without chokes and commutating capacitors [10,24]. However, the speed range of the system is very limited and dependent on the supply. Also, there are a large number of power devices and control circuits, which have cost implications in high power systems.

The current source inverter (CSI) was used for research to provide sub and super synchronous, motor or generating operation. Theoretically, the induction machine can operate at twice the synchronous speed, without exceeding the rated voltage and current in the secondary circuit, to give twice the rated mechanical output power with the same size power electronics and machine rating [98].
3.2 Scherbius System using CSI

The CSI is suitable for wide speed, four-quadrant operation. It is also preferred for its exceptional robustness and reliability [75]. The CSI consists of a mains side three-phase naturally commutated thyristor current source, coupled via a d.c. choke, to a three phase forced commutated thyristor inverter. The three-phase naturally commutated bridge converter operates in conjunction with the d.c. choke to provide a variable current controlled by a closed loop servo system. The controlled current d.c. current is fed to the forced commutated inverter. With a correct firing sequence to the thyristor gates, the controlled current will generate a three-phase-quasi square wave in the machine windings.
For simplicity, the rotor circuit is redrawn with the same definitions of e.m.f. and current directions. The directions of the currents and the voltages are defined positive in the order shown in the rotor circuit diagram, where \(e_1\), \(e_2\), \(e_3\) are slip dependent e.m.f.'s. The method for controlling the direction of power flow is discussed with reference to this circuit. The motor is represented by slip dependent e.m.f.'s \(e_1\), \(e_2\), \(e_3\).
In the CSI, there are two delta-connected banks of capacitors, which serve to commutate the conducting thyristor. Series diodes in each arm of the inverter are also necessary to ensure that adequate charge is stored in the capacitors and are not affected by varying load conditions.

The CSI controls the direction and magnitude of slip power in the secondary circuit of the SRIM. It will provide both sub- and super-synchronous operation. This requires that the firing sequence applied to the thyristor gates of the inverter bridge be synchronised to the secondary e.m.f. An absolute encoder is used as a signal generator to provide synchronisation. This system has already been discussed by Smith [75].

3.2.1 Encoder Based E.M.F. Signal Generator

The electronic signal generator compares the supply frequency represented by a constant number of pulses per cycle with pulses generated from a slotted disc mounted on the machine shaft. A difference counter is then used to generate a three-phase constant amplitude square-wave locked in phase with the secondary e.m.f. This can then be used through additional logic circuits to control the precise firing points of the CSI.

For the machine to operate through synchronous speed, the direction of power flow must change. Therefore, the direction of secondary current for each half cycle of the secondary e.m.f. must be reversed, and the phase sequence of the signal generator output must be reversed. The output of the signal generator and the associated logic is designed to achieve this.
For example, consider the machine is in a sub-synchronous speed. The control of power flow can be seen by reference to section 3.3. When the slip e.m.f. is positive, thyristors T3 and T4 are fired. The current and e.m.f. will both be in the same direction, and so power will be returned through the regenerative naturally commutated thyristor converter to the a.c. supply, to produce the desired driving torque.

As the machine runs through synchronous speed, the phase rotation of the secondary e.m.f. will change. However, provided that the thyristors previously firing on a positive half cycle of secondary e.m.f. are now fired on a negative half cycle, then the current and slip e.m.f. will be in opposite directions. The e.m.f. source will be absorbing power from the a.c. mains and a driving torque can still be developed.
Fig. 3.5 Super-synchronous Motoring Thyristor Sequence

The signal generator is designed to invert the polarity of the signal voltage for each half cycle of e.m.f., as the machine runs through synchronous speed. In addition, this inversion can be achieved manually, or even automatically, to reverse the direction of power flow at any operating speed so changing the machine between motoring and braking modes. This will then allow the different operating modes to be achieved.

Control of the system is simple, if the three-phase signal output is synchronised to the slip e.m.f. this determines the firing points for the current source inverter. A conventional cascade servo system is then used in which the amplified speed error is the demand signal to a current servo amplifier. The output of the current servo controls the angle of advance of the naturally commutated converter over the range -60° to +120° so that the desired current level can be maintained when both rectifying and regenerating.
To ensure that the rotor current never goes to zero under dynamic transient conditions (this can cause malfunction of the inverter) a minimum current demand level is provided by the control system. With a built-in current limit feature, the machine can be started with any level of demanded speed and will run up, within limit, to that speed.

3.3 Power Flow Considerations in Scherbius System

In the SRIM, the primary input power is defined as $P_i$. The power crossing the air gap and the mechanical power output are defined as $P_a$, $P_o$ respectively. The electrical power appearing at the secondary circuit are both proportional to the slip, $s$. The relationship between these quantities, and the rotor electrical power are:

\[
\text{Mechanical Power} \quad : \quad P_o = (1-s) P_a \\
\text{Rotor Electrical Power} \quad : \quad sP_a
\]

If the rotor electrical power can be made positive or negative then there are four possible modes of operation. The four modes are:

Mode 1: **Sub synchronous Motoring**

(Slip power recovery, power out of the rotor.)

Mode 2: **Super synchronous Motoring**

(Slip power into the rotor.)

Mode 3: **Sub synchronous Regeneration**

(Slip power into the rotor.)
Mode 4: Super synchronous Regeneration

(Slip power out of the rotor.)

Fig. 3.6 Scherbius power flow diagram
3.3.1 **Mode 1: Sub-synchronous Motoring**

This is identical to that of the static Kramer system. The stator input or the power crossing the air gap $P_r$ remains constant and the slip power $sPg$, which is proportional to the slip, is returned back to the mains supply. Therefore, the line supplies the net mechanical power, $P_m$, to the shaft. The slip frequency corresponds to the difference between the two frequencies ($\omega_r = \omega_c - \omega_n$). At true synchronous speed ($S=0$), the inverter supplies d.c. excitation to the rotor and the machine operates like a synchronous motor.

**SLIP : $1 > S > 0$ Positive**

**Shaft : Power out**

**Stator : Power in**

**Rotor : Power out**

**Fig. 3.7 Sub-synchronous Motoring Power Flow**
3.3.2 Mode 2: Super-synchronous Motoring

As the shaft speed increases past synchronism, slip becomes negative and the rotor circuit absorbs the slip power. The slip power supplements the stator power for total mechanical power output. The line supplies slip power in addition to the stator power input. In this condition, the phase sequence of the slip voltage is reversed, so that the slip-frequency-induced rotating field is opposite to that of the stator.

\[
\begin{align*}
\text{Slip} & : S < 0 \quad \text{Negative} \\
\text{Shaft} & : \text{Power out} \\
\text{Stator} & : \text{Power in} \\
\text{Rotor} & : \text{Power in}
\end{align*}
\]

![Diagram of power flow](image)

Mode 2: Supersynchronous Driving \( S < 0 \)

Fig. 3.8 Super-synchronous Motoring Power Flow
3.3.3 **Mode 3: Sub-synchronous Regeneration**

In a regenerative braking condition, the load drives the shaft and the mechanical energy is converted into electrical energy. With constant negative shaft torque, the mechanical power input to the shaft increases with speed. This power equals to the quantity of electrical power fed to the line. In the Sub synchronous speed range, the slip power is fed to the rotor so that the total stator power output is constant. The slip voltage has a positive phase sequence (i.e., the direction of magnetic field is the same as that of the stator field). At synchronous speed, the inverter supplies d.c. excitation current to the rotor and the machine behaves as a synchronous generator.

Slip : \( 1 > S > 0 \) Positive  
Stator : Power out  
Rotor : Power in  

**Mode 3 : Regenerative Braking \( 1 > S > 0 \) Subsynchronous**

*Fig. 3.9 Sub-synchronous Generation*
3.3.4 **Mode 4: Super-synchronous Regeneration**

In this mode, the stator output power remains constant but the additional mechanical power input is reflected as slip power output. The inverter phase sequence is now reversed so that the rotor field rotates in the opposite direction.

Slip : $S < 0$ Negative  
Stator : Power out  
Shaft : Power in  
Rotor : Power out

**Primary Input**

- $P_{c1}$ primary core loss
- $P_e$ primary crossing air gap

**Mechanical Power**

- $P_r$ rotor
- $P = (1 - S)P_r$
- $S P_r$ secondary Electrical Power
- $P_{c2}$ secondary $P_r$ loss

**Mode 4: Induction Generator $S < 0$ Supersynchronous**

*Fig. 3.10 Super-synchronous Generation*
3.4 SCHERBIUS ANALYSIS

3.4.1 Equivalent Circuit of the Slip Ring Machine

![Equivalent Circuit Diagram]

Fig. 3.11 Induction Machine equivalent circuit

The single phase equivalent model (Steinmetz model) of the SRIM is shown in fig. 3.11. Initially, with an open circuit rotor, the e.m.f. $E$ lags the stator voltage $V_1$ by a lagging angle $\theta_1$ that will change with load. The rotor current $I_2$ (usually set at zero) is then electronically synchronised at a pre-set angle $\theta_0$ to this e.m.f. Taking the rotor voltage $V_1$ as the reference phasor, the phase relationships are as shown in fig.3.12.

![Phasor Diagram]

Fig. 3.12 Phasor diagram of per phase equivalent circuit
Consider the diagram in fig.3.11, the magnetising impedance $Z_M$ is given by:

$$Z_m = A + jB$$  

....(3.1)

Where,

$$A = \frac{X_m^2 R_m}{R_m^2 + X_m^2} \quad B = \frac{R_m^2 X_m}{R_m^2 + X_m^2}$$

The secondary terminal voltage is:

$$\bar{E} = |E| \cos \theta_h - j |E| \sin \theta_h$$  

..... (3.2)

The relationship between the magnetising current $I_o$ and $\bar{E}$ is:

$$I_o = \frac{\bar{E}}{Z_m}$$  

$$= \frac{|E| \cos \theta_h - j |E| \sin \theta_h}{A + jB}$$  

.....(3.3)

Therefore, by expanding equation (3.3), $I_o$ can be expressed as:

$$I_o = (C|E| \cos \theta_h - D|E| \sin \theta_h) - j(C|E| \sin \theta_h - D|E| \cos \theta_h)$$  

.....(3.4)

Where,

$$C = \frac{A}{A^2 + B^2} \quad D = \frac{B}{A^2 + B^2}$$

The stator current $I_1$ is the vector sum of $I_o$ and $I_2$. The logic control circuit will initially synchronise $I_2$ to $E$ at an adjustable angle $\theta_0$, which is usually set to zero.
From fig.3.12, the angle relationships are:

\[ \theta = \theta_0 - \theta_1 \]  

(3.5)

The current remains after synchronisation at a constant angle \( \theta \) to the supply.

\( I_1 \) can be expressed in its real and imaginary parts as:

\[
\begin{align*}
\bar{I}_1 &= (|I_2| \cos \theta + C |E| \cos \theta_1 - D|E| \sin \theta_1) \\
&+ j(|I_2| \sin \theta - C|E| \sin \theta_1 - D|E| \cos \theta_1)
\end{align*}
\]

(3.6)

Therefore, \( I_1 \) can be expressed as:

\[ \bar{I}_1 = I_{1p} + jI_{1q} \]

(3.7)

Where \( I_{1p} \) is the in phase component with \( V_1 \) and \( I_{1q} \) is in quadrature.

### 3.4.2 Determination of e.m.f. \( E \) and \( \theta_1 \)

The induced e.m.f. \( E \) and the assumed lagging angle \( \theta_1 \) with reference to \( V_1 \), may be determined by the stator voltage, the synchronised current \( I_2 \), the machine parameters and a pre selected angle \( \theta_0 \).

From fig.3.12, the stator voltage is the sum of the induced e.m.f. and the voltage developed by \( I_2 \) across \( R_1 \) and \( X_1 \):

\[ \bar{V}_1 = \bar{E} + \bar{I}_1 (R_1 + jX_1) \]

(3.8)
Substituting equation (3.7), the stator voltage is:

\[ \bar{V}_1 = \left| E \right| \cos \theta_1 - j\left| E \right| \sin \theta_1 + (I_{Ip} + jI_{Iq}) (R_1 + jX_1) \]

\[ = \left( \left| E \right| \cos \theta_1 + I_{Ip} R_1 - I_{Iq} X_1 \right) + j \left( I_{Iq} R_1 + I_{Ip} X_1 - \left| E \right| \sin \theta_1 \right) \]  

\[ \cdots (3.9) \]

Assume that the stator voltage vector consists of an in phase component \( V_{Ip} \) and a component \( V_{Iq} \) in quadrature. As \( V_{Iq} \) is the reference phasor, so \( V_{Iq} \) in quadrature is equal to zero.

\[ V_{Ip} = \left| E \right| \cos \theta_1 + I_{Ip} R_1 - I_{Iq} X_1 \]  

\[ \cdots (3.10) \]

\[ V_{Iq} = I_{Iq} R_1 + I_{Ip} X_1 - \left| E \right| \sin \theta_1 \]

\[ = 0 \]  

\[ \cdots (3.11) \]

Combining equations (3.6) and (3.10) yield:

\[ V_{Ip} = \left| E \right| \cos \theta_1 + R_1 \left( \left| L_2 \right| \cos \theta_1 + C \left| E \right| \cos \theta_1 - D \left| E \right| \sin \theta_1 \right) \\
- X_1 \left( \left| L_2 \right| \sin \theta_1 - C \left| E \right| \sin \theta_1 - D \left| E \right| \cos \theta_1 \right) \]

\[ \cdots (3.12) \]

Combining equation (3.6) and (3.11) yield:

\[ 0 = R_1 \left( \left| L_2 \right| \sin \theta_1 - C \left| E \right| \sin \theta_1 - D \left| E \right| \cos \theta_1 \right) \\
+ X_1 \left( \left| L_2 \right| \cos \theta_1 + C \left| E \right| \cos \theta_1 - D \left| E \right| \sin \theta_1 \right) - \left| E \right| \sin \theta_1 \]

\[ \cdots (3.13) \]
Rearranging equation (3.12),

\[ 0 = |E| \cos \theta_i (1 + R_iC + X_iD) \]
\[ + |E| \sin \theta_i (C X_i - D R_i) + \]
\[ + (R_i |I_2| \cos \theta - X_i |I_2| \sin \theta - V_{ip}) \]

.....(3.14)

This can be simplified to:

\[ 0 = L|E| \cos \theta_i + M|E| \sin \theta_i + F \]

..... (3.15)

Where,

\[ L = (1 + R_iC + X_iD) \]
\[ M = (X_iC - D R_i) \]
\[ F = R_i |I_2| \cos \theta - X_i |I_2| \sin \theta - V_{ip} \]

Similarly, for the zero quadrature component, since from equation (3.13):

\[ 0 = -|E| \sin \theta_i (1 + R_iC + X_iD) \]
\[ + |E| \cos \theta_i (C X_i - D R_i) \]
\[ + R_i |I_2| \sin \theta + X_i |I_2| \cos \theta \]

\[ 0 = L|E| \sin \theta_i + M|E| \cos \theta_i + J \]

..... (3.16)
Where
\[ J = -(I_2 R_1 \sin \theta + I_2 X_1 \cos \theta) \]

Equation (3.15) & (3.16) give:
\[
\frac{L \cos \theta_1 + M \sin \theta_1}{L \sin \theta_1 + M \cos \theta_1} = \frac{F}{J}
\]

\[ \Rightarrow \tan \theta_i = \frac{JL + FM}{FL - MJ} \]

Equation (3.17)

Therefore, the voltage \( E \) and the assumed lagging angle \( \theta_1 \) between \( E \) and \( V_1 \), can be determined by the equations:
\[ \theta_1 = \tan^{-1} \frac{JL + FM}{FL - MJ} \]

Equation (3.18)
\[ E = \frac{-F}{L \cos \theta_1 + M \sin \theta_1} \]

Equation (3.19)

Where \( F, L, J, M \) are already described.

Provided that the machine parameters \( V_1, R_1, X_1, R_m \) and \( X_m \) are known, then the variables \( A, B, C, D, L, M \) in equations (3.15, 3.16, 3.18, 3.19) can be determined.
The rotor current $I_2$ can be set by the electronics to lead the secondary e.m.f. $E$ by the pre-selected angle $\theta_0$. Usually, this angle is set at zero so that the rotor current is in phase with the secondary e.m.f.

The analysis is based on the constant angle $\theta$. The initial value of $\theta_i$ can be obtained from equation (3.8) in which $I_1$ is given by equation (3.6) with $I_2$ set to zero.

With $\theta$ known, the variables $F$, $J$ in equation (3.15) and (3.16) can be determined. Then for each value of rotor current $I_2$, $\theta_1$ can be calculated from equation (3.18). The induced e.m.f. $E$ can be determined from equation (3.19). Also, $I_{ip}$ and $I_{iq}$ can be found from equation (3.6).

It is noted that, the values $R_1$, $X_1$, $R_m$ and $X_m$ of the machine may vary when the machine is in operation.

3.4.3 Power in the Machine

The electrical power passed to the rotor is defined as $P_r$. $P_r$ can be determined using $I_2$ and $E$. With reference to fig.3.12:

$$ P_r = |E| |I_2| \cos(\theta + \theta_i) $$

.....(3.20)

Where $I_2$ is set by the power electronics.
It is assumed that the losses in the machine and power electronics are low. However, equation (3.20) should agree with the results obtained by subtracting the total losses from the stator input power:

\[ P_r = P_i - \left( |I_2|^2 R_1 + \frac{|E|^2}{R_m} \right) \]  

.....(3.21)

The electrical power input to the stator, \( P_i \), can be determined by the product of the stator voltage and the in phase component of the stator current:

\[ P_i = |V_1| I_{ip} \]  

.....(3.22)

From equation (3.6),

\[ P_i = |V_1| \left( |I_2| \cos \theta + C|E| \cos \theta_1 - D|E| \sin \theta_1 \right) \]  

.....(3.23)

The stator power factor, \( \phi \), can be determined by finding the angle between the real and imaginary components of the stator current \( I_1 \) in equation (3.6).

\[ \phi = \tan^{-1} \frac{I_{iq}}{I_{ip}} \]

\[ = \tan^{-1} \left( \frac{|I_2| \sin \theta - C|E| \sin \theta_1 - D|E| \cos \theta_1}{|I_2| \cos \theta - C|E| \cos \theta_1 - D|E| \sin \theta_1} \right) \]  

.....(3.24)
The stator power factor $\phi$ can be determined from equation (3.24). The magnitude of the stator current $I_1$ may be calculated from the modulus of $I_{ip}$ and $I_{iq}$.

The stator input power $P_I$, may be calculated from equation (3.23). As the analysis is based on one single phase of the machine, the total input power to the SRIM should be three times the per phase power.

The power to the rotor, $P_r$, may be determined from equation (3.20) or by taking the product of $V_1$ and $I_{ip}$. The total rotor power should be again multiplied by a factor of three, as the analysis is for one phase only.

The useful torque produced by the rotor is $T = P_r/\omega_s$, and the mechanical output is $P_o = T.\omega$.

A simple model and analysis has been derived for the slip recovery drive in which the power electronic converters set both the phase angle and magnitude of the rotor current. With the machine parameters known and the rotor current set as an input to the model, the operating characteristics can be determined. The analysis can easily be extended to determine the phasor sum of stator current and the line current recovered from the rectifier in the rotor circuit, to obtain the overall operating power factor of the drive.
A Microsoft Basic program for the Scherbius was developed from a previous Apple Basic Kramer program to incorporate the equations in this chapter, for the prediction of the SRIM operation at various conditions. All four modes of operation are studied. The effects of varying the synchronising angle are also presented. All results are in graphical form. The program for the machine operation is listed in Appendix 3, and the simulated machine characteristics are listed in Appendix 4.
CHAPTER 4

Current Source Inverter for Slip Power Recovery System

4.0 Introduction

A conventional thyristor current source inverter (CSI) connected to the slip rings of an induction machine can provide power flowing into and out of the rotor circuits. Therefore, operation both as a motor and as generator are possible at above or below synchronous speeds. Although the CSI is a popular set up with large systems, problems in the forced commutation circuits are not well documented. This can result in operational difficulties, when the SRIM is operating primarily in the super-synchronous region. The problems are analysed and their effects on drive performance are considered.

4.1 Steady State Commutation

The CSI operation generates rich current harmonic currents, which may affect any measurement of the other machine electrical quantities in a sensorless control set-up. The steady state operation of an a.c. drive using a CSI and a three-phase SRIM has been analysed by many authors [28,31,75]. A current source inverter-motor system for slip ring induction motors in which the machine is merely represented by an inductance in series with a secondary e.m.f. has been analysed by Smith [75]. The same analysis is adopted and simplified for sensorless operation based on rotor voltage measurement.
The CSI configuration is shown in fig 4.1. All analysis is based on the figure and the assignments of devices remain the same throughout the analysis. The slip ring machine used has a delta connected stator and a delta connected rotor windings.

The analysis of the CSI connected to the SRIM rotor at various speeds is presented. The operation of the CSI during an inverter period is classified into three stages A, B, C. The voltages and currents during these stages are assigned the subscripts a,b,c.

It is assumed that the resistance and reactance of the machine winding is negligible, the thyristors and diodes are near ideal devices, and the turn-off and turn-on times of the thyristors are short and negligible. The three stages are briefly described as follows with the R-Y as the reference phase as shown in fig 4.1.

Fig 4.1 Current Source Inverter Definition
Stage A: When thyristors T1 and T2 are conducting, stage A begins when T3 is triggered to turn off T1. As soon as T1 turns off, the direct current charges up the capacitor bank and stage A ends when diode D3 becomes forward biased.

Stage B: Stage B begins and continues as the direct current increases in diode D3, and the capacitor bank discharges through D1.

Stage C: Stage C begins when the current in the capacitor bank reaches zero. Stage C lasts until T4 is triggered to turn off T2, the direct current is carried by T3, D3, the machine winding, T2 and D2.

Equivalent circuits for the different stages are used to develop a set of simultaneous algebraic equations relating the time interval of each stage and the peak voltage of the commutation capacitors.

4.1.2 Sub-Synchronous Motoring Commutation Analysis

From fig.4.1, each rotor phase is represented by an induced secondary e.m.f. in series with the total leakage inductance Le. Using the rule of total current in equals total current out, the d.c. current path and the rotor phase current is defined as follows:
where,

$I$ is the total current from the d.c. link

$i_1, i_2, i_3$ are the currents in the rotor phases

Considering at steady state condition, when $T_1$ and $T_2$ are conducting. The slip e.m.f.s at the instant $T_3$ is turned on, are defined as:

\[
e_1 = S\hat{E}_R \sin(s\omega t)
\]

\[
e_2 = S\hat{E}_R \sin(s\omega t - \frac{2\pi}{3})
\]

\[
e_3 = S\hat{E}_R \sin(s\omega t + \frac{2\pi}{3})
\]

\[
\hat{E}_R = \sqrt{2}E_R
\]

\[\ldots (4.1)\]

Let $V_{ini}$ be the initial capacitor voltage, then the voltage on the upper group commutation capacitors are assumed to have the initial values as follows:

\[
V_{CX} = -V_{ini}
\]

\[
V_{CY} = 0
\]

\[
V_{CZ} = +V_{ini}
\]

\[\ldots (4.2)\]

The voltage waveforms are illustrated in the following diagrams.
Fig. 4.2 Commutation Capacitor Currents at sub-synchronous operation

Fig. 4.3 Commutation Capacitor Voltages at sub-synchronous operation
4.2.1 Start of Commutation - Stage A

Fig. 4.4 Before T3 turned on at steady state

Fig 4.4 shows the current path before T3 is on. When T3 is fired, the voltage across T1 will be \(-V_{cx}\). This reverse biases T1, and T1 turns off. The d.c. current which was flowing through T1 now flows through T3 and the capacitor bank. After a charging time \(t_1\) when the voltage across \(C_x\) is equal to the e.m.f. \(e_1\), across R-Y, D3 will start to conduct. If \(V_{cx}\) cannot exceed \(e_1\), then

\[
V_{cx} + e_1 \geq 0
\]

\[
V_{cx} = V_{cx0} + \frac{2I}{3C} t_1
\]

\ ...(4.3)
Then the equation describing the initial voltage is

\[- V_{\text{ini}} + \frac{2I}{3C} t_i + S \hat{E}_R \sin(s \omega t_i) = 0\]

\[t_i = \frac{2I}{3C} [V_{\text{ini}} - S \hat{E}_R \sin(s \omega t_i)]\]

\[\ldots (4.4)\]

The final capacitor voltages at stage A will be as follows:

\[\begin{align*}
V_{\text{CXA}} &= V_{\text{CXO}} + \frac{2I}{3C} t_i \\
V_{\text{CYA}} &= V_{\text{CYO}} - \frac{I}{3C} t_i \\
V_{\text{CZA}} &= V_{\text{CZO}} - \frac{I}{3C} t_i
\end{align*}\]

\[\ldots (4.5)\]

Fig. 4.5 Commutation at Stage A
4.2.2 **Commutation Stage B**

![Diagram](image)

*Fig. 4.6 Commutation at Stage B*

When D3 begins to conduct, the voltage across $C_x$ will be slightly less than or equal to the e.m.f. $e_1$. It is assumed there are three current paths $I_1$, $I_2$ and $I_3$ at this instance for analysis.

**Loop 1 - $I_1$**

Sum of voltages = 0 in loop 1, then:

$$
\frac{1}{X_C} \int_{t_1}^{t_2} dt + L E \left[ \frac{d I_1}{dt} \right] - e_1 - L E \left[ \frac{d I_2}{dt} \right] - \frac{1}{X_C} \int_{t_1}^{t_2} dt = 0
$$

(4.6)
Loop 2 - l2

Also, sum of voltages in loop 2 = 0:

\[-L_e \frac{di_1}{dt} + 3L_e \frac{di_2}{dt} + (e_1 + e_2 + e_3) - L_e \frac{di_3}{dt} = 0\]

.... (4.7)

Loop 3 - l3

For Loop 3, equating the d.c. link current

\[i_3 = I\]

for a D.C. current, then the rate of change with time is zero:

\[\frac{di_3}{dt} = 0\]

\[\int_0^{t_2} i_3 \, dt = (I t)_{t_2}^{t_2} = (1t)_{t_2}^{t_2}\]

.... (4.8)

Since the sum of \(e_1\), \(e_2\) and \(e_3\) must be zero, then from (4.7)

\[\frac{1}{3} L_e \left( \frac{di_1}{dt} \right) = L_e \left( \frac{di_2}{dt} \right)\]

\[L_e \left( \frac{di_1}{dt} \right) + \frac{1}{C} \int_0^{t_2} i_1 \, dt - \left( \frac{t}{c} \right)_{t_2}^{t_2} - \frac{3}{2} e_3 = 0\]

.... (4.9)
The solution for current \( I_1 \) is

\[
i_1 = I \left( 1 - \cos(\omega_0 t) \right) + \frac{3}{2} e \cdot \omega_0 c \cdot \sin(\omega_0 t)
\]

where

\[
\omega_0^2 = \frac{1}{L \cdot C}
\]

Initially, when \( T=0 \), the current in the diode is zero, and as \( I_1 \) flows in to the diode,

\( I_1=I_{d3} \).

\[
t = 0,
\]

\( i_{d3} = 0 \)

\[
t = t_2,
\]

\( i_{d3} = I \left( 1 - \cos(\omega_0 t) \right) + \frac{3}{2} e \cdot \omega_0 c \cdot \sin(\omega_0 t) \)

\[
..... (4.11)
\]

The current flowing in \( D_1 \) is the total current minus \( I_{d3} \).

\[
i_{d1} = \left| I \left( 1 - \cos(\omega_0 t) \right) - \frac{3}{2} e \cdot \omega_0 c \cdot \sin(\omega_0 t) \right|
\]

\[
t = t_2,
\]

\( i_{d1} = \cos(\omega_0 t) - \frac{3}{2} e \cdot \omega_0 c \cdot \sin(\omega_0 t_2) \)

\[
..... (4.12)
\]
Hence, the voltage change across Cx during stage B is

\[
\frac{dV_{CX}}{dt} = \frac{2}{3C} I_d^2 dt
\]

\[
\frac{dV_{CX}}{dt} = \frac{2}{3c} \int_0^{t_2} I_d \cos(\omega_0 r) \, dr - \frac{3}{2} e_0 c \cdot \sin(\omega_0 t_2)
\]

\[
\ldots (4.13)
\]

The final capacitor voltages at stage B will be as follows:

\[
\begin{align*}
V_{CXB} &= V_{CXA} + \Delta V_{CX} \\
V_{CYB} &= V_{CYA} - \frac{1}{2} \Delta V_{CX} \\
V_{CZB} &= V_{CZA} - \frac{1}{2} \Delta V_{CX}
\end{align*}
\]

\[
\ldots (4.15)
\]
4.2.3 **Commutation Stage C**

At this stage, T3 is fully on and D1 is no longer conducting.

\[ i_{at} = 0 \]

\[ \therefore i_c = 0 \]

The capacitor voltages are as follows:

\[
\begin{align*}
V_{cx} &= V_{xb} = V_{xo} \\
V_{cy} &= V_{yb} = V_{yo} \\
V_{cz} &= V_{zb} = V_{zo}
\end{align*}
\]  

\[ \ldots (4.16) \]

where,

\[ V_{xo} = -V_{in} = -\frac{1}{2} \left( \frac{2I}{3C} t_1 + \Delta V_{cx} \right) \]

\[ V_{yo} = 0 \]
\[ V_{CZO} = V_{BI} = \frac{1}{2} \left[ \frac{2}{3C} t_1 + \Delta V_C \right] \] .... (4.17)

### 4.2.4 Waveforms at Stages A,B,C

In summary, the waveforms at the three stages are as follows:

![Waveform Diagram](image)

**Fig. 4.8 Sub-Synchronous sequence and Voltages**
4.3 **Super-Synchronous Motoring Commutation Analysis**

With both the rotor and the stator circuits fed with power from the mains, the machine will accelerate above synchronism. At super-synchronous speed, the phase sequence will change. Using R-Y as the reference, the next appearing phase is not B-Y as in sub-synchronous speed but B-R. Consequently, the firing sequence has to be different from that during sub-synchronism for motoring or generating.

In the super-synchronous speed region, the polarity and phase sequence of the synchronising e.m.f. is different from sub-synchronous speed, the inverter firing sequence is changed accordingly. As such, T5 must be fired to start the commutation process of T1. The above diagram depicts the current path before T5 is turned on.

### 4.3.1 Start of Commutation - Stage A

![Diagram](image)

Fig. 4.9 Before T5 is turned on
The motor phase current is defined as follows:

\[ i_1 = i_2 = \frac{I}{3} \]

\[ i_3 = -\frac{2I}{3} \]

where \( I \) is the d.c. link current

Let \( V_{\text{ini}} \) be the initial capacitor voltage, then the voltage on the upper group of commutation capacitors are assumed to have the initial values as follows:

\[ V_{\text{CX}} = -V_{\text{ini}} \]
\[ V_{\text{CY}} = 0 \]
\[ V_{\text{CZ}} = +V_{\text{ini}} \]

Where \( \text{Cx, Cy and Cz} \) are commutation capacitors connected across the slip rings.

\[ V_{\text{CZa}} + e_i \geq 0 \]
\[ V_{\text{CZo}} - \frac{2I}{3C} t_i + e_i \geq 0 \]

where \( t_i \) is expressed as:

\[ t_i = \frac{3c}{2I(V_x + e_i)} \]
The final capacitor voltages at stage A will be as follows:

\[
\begin{align*}
V_{CXA} &= V_{CXO} + \frac{I}{3C} t_i \\
V_{CYA} &= V_{CYO} + \frac{I}{3C} t_i \\
V_{CZA} &= V_{CZO} - \frac{2I}{3C} t_i \\
\end{align*}
\]

\[\text{... (4.20)}\]
4.3.2 Commutation Stage B

This stage starts when D5 is conducting and ends after the current passing through D1 and the capacitor voltage is zero. Similar to the analysis given for sub-synchronous stages, the diode currents $I_{d1}$ and $I_{d5}$ are found as the following.

$$I_{d1} = \left| I \cos \left( \omega_0 t \right) + \frac{3}{2} e_0 \omega_0 C \sin \left( \omega_0 t \right) \right|$$

$$I_{d5} = \left| I(1 - \cos \left( \omega_0 t \right)) - \frac{2}{3} e_0 \omega_0 C \sin \left( \omega_0 t \right) \right|$$

... (4.21)
The Voltage changes across Cz during stage B is given as:

$$\Delta V_{CZ} = \frac{2}{3C_0} \int_{t_0}^{t_2} I_{DL} \, dt$$

$$= \frac{2}{3C_0} \left[ \frac{1}{2} I \cos(\omega_0 t) + \frac{3}{2} e_i \omega_0 C \sin(\omega_0 t) \right]_{t_0}^{t_2}$$

$$= \frac{2}{3\omega_0 C} \left[ I \sin(\omega_0 t) + \frac{3}{2} e_i \omega_0 C \cos(\omega_0 t) \right]_{t_0}^{t_2}$$

(4.22)

When $t = t_2$, the voltage is:

$$\Delta V_{CZ} = \frac{2}{3\omega_0 C} \left[ I \sin(\omega_0 t_2) + \frac{3}{2} e_i \omega_0 C \cos(\omega_0 t_2) - 1 \right]$$

(4.23)

The final capacitor voltages at stage B will be as follows:

\[
\begin{align*}
V_{CXB} &= V_{CXA} + \frac{1}{2} \Delta V_{CZ} \\
V_{CYB} &= V_{CYA} + \frac{1}{2} \Delta V_{CZ} \\
V_{CZB} &= V_{CZA} - \Delta V_{CZ}
\end{align*}
\]

(4.24)
4.3.3 Commutation Stage C

At this stage, T3 is fully on and D1 is no longer conducting when it is reversed biased.

\[ i_{d1} = 0 \]

The capacitor voltages are as follows:

\[
\begin{align*}
V_{cxc} &= V_{cxb} = V_{czo} \\
V_{cyc} &= V_{cyb} = V_{czo} \\
V_{czc} &= V_{czb} = V_{czo}
\end{align*}
\]

\[ \ldots (4.25) \]
where

\[ V_{\text{CZO}} = -V_{\text{INI}} \]

\[ = -\frac{1}{2} \left[ \frac{2I}{3C} t_i + \Delta V_{\text{CZ}} \right] \]

\[ V_{\text{CYO}} = 0 \]

\[ V_{\text{CZO}} = V_{\text{INI}} = \frac{1}{2} \left[ \frac{2I}{3C} t_i + \Delta V_{\text{CZ}} \right] \quad \ldots \text{(4.26)} \]

Fig. 4.13 Commutation Capacitor Current Super-Synchronous operation

Fig. 4.14 Commutation Capacitor Voltage Super-Synchronous operation
4.3.4 Waveforms at Super Synchronous Stages A,B,C

Fig 4.15 Super-Synchronous Upper Group Capacitor Voltages
4.4 **CSI Commutation Problems**

It is desired to operate the machine in the super-synchronous speed region for higher power handling. Previous work has been done to investigate the suitability of the CSI slip recovery drive for wind energy schemes. It was found that the behaviour of the commutation circuit is affected by the slip e.m.f.'s. Work on the slip power facility shows that the commutation waveforms in the super synchronous region are not conventional. These have detrimental effects on the SRIM performance.

The varying magnitude and phase of the slip e.m.f. affect the commutation behaviour. This result in current and voltage waveform distortions in the commutation circuits, and has adverse effects on the machine operation.

For stable operation of the drive, the commutation sequence in the CSI is synchronised electronically with the rotor e.m.f. A variable phase-angle can be introduced into the synchronising circuit to control the rotor power factor and to overcome some of the commutation problems experienced with the conventional six-step commutation circuit. Such a drive is well reported and has found applications in vertical axis wind turbine applications. Also, the additional charging of the commutation capacitor is well analysed and documented by Smith [33], and will not be repeated in this work.
Fig. 4.16 Sub-Synchronous Phase Advance/Retard

Fig. 4.17 Super-Synchronous Phase Advance/Retard
Experiments indicate that varying phase advance and retard angles can improve the capacitor voltage. The following are taken when the machine at $S=-0.2$ with a line current of approximately 3A. When there is no phase angle adjustment, the capacitor voltage shows extra steps due to additional charging. This effect is obvious when the current is low and the capacitor voltage is low. At higher loads, the capacitor voltage will increase and the effect will not be obvious.

![Fig. 4.18 Rotor Line Current with $\alpha=0$](image)

![Fig. 4.19 Commutating Capacitor voltage with $\alpha=0$](image)
Fig. 4.20 Rotor Line Current with $\alpha=27^\circ$ Right

Fig. 4.21 Commutating Capacitor Voltage with $\alpha=27^\circ$ Right
Fig. 4.22 Rotor Line Current with $\alpha=27^\circ$ Left

Fig. 4.23 Commutating Capacitor Voltage with $\alpha=27^\circ$ Left
The analysis of the commutation circuits in a current source inverter applied to the stator of an induction motor is well known. However the effect of rotor e.m.f. on the commutation behaviour in slip recovery schemes indicate problems. When motoring sub synchronously or generating super synchronously the rotor e.m.f.s are in such a direction to aid the commutation process. This results in conventional waveforms on the commutation capacitors but with reduced commutation time.

In the case of motoring super-synchronously or generating sub-synchronously, the rotor e.m.f.s are in such a direction as to oppose the commutation process. This results in longer commutation times, lower commutation voltages and additional steps in the capacitor waveform due to outgoing blocking diodes turning back on.

The simulation of all SRIM operating modes including generating and motoring, and the effects of varying the synchronising angle from lagging to leading, are listed in appendix 5.
CHAPTER 5

Slip Ring Induction Machine Harmonic Analysis

5.0 **Analysis of SRIM Terminal Quantities**

The SRIM terminal stator currents and voltages, or the rotor terminal currents and voltages, are directly related to the states of the machine. Therefore, the torque and speed of the machine may be derived from the measured terminal quantities, if the machine parameters, such as rotor resistance, are known. Provided all these are possible, then the encoder is no longer required to determine the rotor speed or the e.m.f. The machine information can then be derived from the measured quantities [63].

The study of the SRIM rotor and current harmonics enable the identification of the most appropriate electrical quantities for predicting the SRIM operation, i.e. as a useful signal for the control system. A slip recovery drive and a SRIM were used for harmonic investigation. The SRIM was doubly fed by a current controlled, encoder base drive.

The main advantage of current control of the SRIM is that it allows direct setting of any steady state operation. The torque of the SRIM does not depend on the slip or speed; it is solely determined by a phase or magnitude of the rotor current synchronised to the rotor e.m.f. Dynamic torque control is therefore very easy. This may be considered as a special form of direct torque control [75].
The voltage or current waveforms in the SRIM are seldom purely sinusoidal [63]. The periodic waveform is composed of the superposition of a direct component with a fundamental pure sine wave component, together with harmonic frequencies at multiples of the fundamental. The terminal voltages can be mathematically expressed as follows:

\[
V = V_0 + V_1 \sin(\omega t + \phi_1) + V_2 \sin(2\omega t + \phi_2) + V_3 \sin(3\omega t + \phi_3) + \ldots \nonumber \\
+ V_n \sin(n\omega t + \phi_n)
\]

\[(5.1)\]

where \( V \) is the instantaneous value at any time \( t \)

- \( V_0 \) is the d.c. or mean value
- \( V_1 \) is the maximum fundamental value
- \( V_2 \) is the maximum of the second harmonic component
- \( V_3 \) is the maximum of the third harmonic component
- \( V_n \) is the maximum of the \( n \)th harmonic component
- \( \phi \) is the relative angle reference
- \( \omega = 2 \pi f \)

A Brook Crompton 5.5kW slip ring machine was used for harmonic analysis. The stator windings are wound in Delta, and so are the rotor circuits (Appendix 1). The configuration of the machine windings were identified to correctly relate to the phase quantities. For the machine, the rotor and stator terminal voltages are the phase voltages of the rotor and the stator.
Usually, a delta-star combination is preferred for third harmonic cancellation within the machine. Such configuration is employed in large transformer designs, to reduce undesirable circulating currents due to voltage harmonics. It is noted that the SRIM used is a special ‘inside-out’ machine, made to order for the previous project funded by EPSRC [75].

Normally, the stator is permanently connected to the mains supply. Power is controlled into or out of the rotor to achieve wide speed operation via the CSI. The shaft of the SRIM is rigidly coupled to a separately excited d.c. machine as a simulated load or a prime mover.

The five types of terminal quantities investigated were:

1. d.c. Link current
2. Stator Voltage
   (Largely dominated by mains harmonics.)
3. Stator Current
   (Dominated by mains and machine harmonics.)
4. Rotor Voltage
   (A function of stator voltage and rotor speed.)
5. Rotor Current
   (Set by demand, synchronised and switched according to rotor e.m.f.)

A Texas Instrument TMS320 Starter Kit, a TMS-320C50 DSP starter kit, and a 486 DX2-66 PC were used to record the terminal quantities digitally. The quantities were processed off line by an FFT algorithm to determine and record the
harmonic contents. Also, an Onno-Sokki CF550 real time FFT spectrum analyser was used to help investigate and illustrate the stator and rotor current harmonics under various dynamic conditions.

5.1. D.C. Link Current Harmonics

The SRIM is a noisy system, largely due to the supply and the quasi square current waves from the CSI. The rotor circuits are fed by a servo controlled d.c. current, switched in six steps, in or out of phase with the rotor e.m.f. The rotor current cannot be perfectly flat topped, due to the limited d.c. link choke size. For small inductors, the current ripple or harmonics can be large and undesirable. To minimise such harmonics, it is desired to reduce the current ripples as much as possible within reason. Two methods of reducing the current harmonics in the d.c. link are proposed and implemented:

1. Inserting a $\pi$ filter within the link inductance.
2. Introduce capacitors at the secondary windings.

Fig 5.1 Filter in the D.C. Link to Reduce Current Ripple
The π filter is tuned at below 50Hz and therefore greatly reduces the a.c. component from $V_{do}$ or $V_{dr}$. The output of the filter is a much smoother d.c. current than from a single small choke. The capacitor in the π filter must be a reversible type. The capacitor is an energy storage. This energy is stored and is determined by $Q=CV$. A large capacitance value is avoided to prevent sluggish machine operation.

Capacitors on the slip rings can help reduce voltage spikes on the rotor circuits. A more important function is a reduction in reactive power consumed by the rotor. Experiments show that such capacitors have no direct effect to the CSI operation.

![Fig. 5.2 Link Current Harmonics (2 A) with PI filter](image)

The link current still contains very rich current harmonics with the π filter installed. This is due to the quasi-square waves in the rotor circuits. However, the current waveform becomes more distinct as a square topped waveform.
5.2 **Stator Voltage Harmonics**

As the stator is permanently connected to the mains, the mains largely dominate the stator voltage. The mains is not purely sinusoidal and contains spikes and harmonics, depending on the quality of the utility. When the machine is in operation, there is little or no variation in the stator voltage and the harmonics.

The dominating stator voltage harmonics are the 3\(^{rd}\), 5\(^{th}\) and 7\(^{th}\) (from the supply). As the stator voltage and harmonics generally remain the same at all machine speed or loading conditions, it may not provide useful information for the control system to synchronise the CSI to the rotor e.m.f.'s.

![Stator Voltage Harmonics](image)

**Fig. 5.3 Stator Voltage Harmonics (at all rotor speeds)**

5.3 **Stator Current Harmonics**

The SRIM stator is permanently connected to the mains and provide a constant rotating field. The stator harmonic currents also produce rotating fields in the air gap that rotate at higher speeds than the rotating field produced by the fundamental current. However, the rotor current is non-sinusoidal and a large
number of harmonic currents are present. The harmonics are space and time
harmonics inherent in an a.c. machine.

5.3.1 Stator Time Harmonics

Due to the limited number of slots in the machine, the mmf in the SRIM is
not sinusoidally distributed in the air gap. This produces space harmonics. The space
harmonics produce parasitic torque, which is insignificant compared to the
fundamental torque. For the analysis of time harmonics, a sinusoidally distributed
winding is assumed in the SRIM. The stator phase currents in the 3 phase SRIM are:

\[ i_R = \sum_{h=1}^{\infty} I_{h_{max}} \cos(h \cdot \omega t) \] .... (5.2)

\[ i_B = \sum_{h=1}^{\infty} I_{h_{max}} \cos(h(\omega t - 120^\circ)) \] .... (5.3)

\[ i_V = \sum_{h=1}^{\infty} I_{h_{max}} \cos(h(\omega t + 120^\circ)) \] .... (5.4)

where,

- \( h \) is the number of harmonics.
- \( I_{h_{max}} \) is the amplitude of the \( h \) order harmonics current

Assume that there are \( N \) sinusoidally distributed turns in each phase winding,
then the mmf along \( \theta \) due to the current in phase \( a \) is:

\[ F_R = N I_R \cos \theta \] .... (5.5)
Therefore, the mmf distribution due to phase a is expressed by:

\[ F_r(\theta, t) = \sum_{h=1}^{\infty} N_{h_{\text{max}}} \cosh \omega t \cos \theta \]
\[ = \sum_{h=1}^{\infty} F_{h_{\text{max}}} \cosh \omega t \cos \theta \]
\[ \quad \quad \text{.... (5.6)} \]

Similarly, the mmf contributions from the remaining phases, b and c, are:

\[ F_b(\theta, t) = \sum_{h=1}^{\infty} F_{h_{\text{max}}} \cos (\omega t - 120^\circ) \cos (\theta - 120^\circ) \]
\[ \quad \quad \text{.... (5.7)} \]

\[ F_c(\theta, t) = \sum_{h=1}^{\infty} F_{h_{\text{max}}} \cos (\omega t + 120^\circ) \cos (\theta + 120^\circ) \]
\[ \quad \quad \text{.... (5.8)} \]

By summing the three vector quantities, the resultant is:

\[ F(\theta, t) = \sum_{h=1}^{\infty} F_{h_{\text{max}}} (\cos (\omega t) \cos \theta + \cos (\omega t - 120^\circ) \cos (\theta - 120^\circ) + \cos (\omega t + 120^\circ) \cos (\theta + 120^\circ)) \]
\[ \quad \quad \text{.... (5.9)} \]

The fundamental mmf rotates in the forward direction at an angular speed of \( \omega \) radians per second:

\[ F_1(\theta, t) = \sum_{h=1}^{\infty} F_{h_{\text{max}}} (\cos (\omega t) \cos \theta + \cos (\omega t - 120^\circ) \cos (\theta - 120^\circ) + \cos (\omega t + 120^\circ) \cos (\theta + 120^\circ)) - \frac{3}{2} F_{1_{\text{Max}}} \cos (\theta - \omega t) \]
\[ \quad \quad \text{.... (5.10)} \]
The third harmonic mmf is not present in the machine, as shown in the following equation:

\[ F_3(\theta, t) = \sum_{h=1}^{\infty} F_{h, \text{max}} \left( \cos 3\omega t \left[ \cos (\theta - 120^\circ) + \cos (\theta + 120^\circ) \right] \right) \]

\[ = F_{h, \text{max}} \times 0 \]

\[ = 0 \]

.... (5.11)

The fifth harmonic mmf rotates in the opposite direction, with respect to the fundamental, at five times the speed of the fundamental wave.

\[ F_5(\theta, t) = \frac{3}{2} F_{5, \text{max}} \cdot \cos(\theta + 5\omega t) \]

.... (5.12)

The seventh mmf harmonic rotates at the same direction as the fundamental, but at seven times the angular speed:

\[ F_7(\theta, t) = \frac{3}{2} F_{7, \text{max}} \cdot \cos(\theta - 7\omega t) \]

.... (5.13)

The other mmf harmonic waves are in the order of \( h = 6n \pm 1 \), where \( n \) is an integer. The harmonics are odd numbers and not multiples of 3.

\[ F_h(\theta, t) = \frac{3}{2} F_{h, \text{max}} \cdot \cos(\theta \pm h\omega t) \]

.... (5.14)
In summary, the synchronous angular velocity of the time harmonic mmf waves in the 6-pole SRIM stator are as follows:

<table>
<thead>
<tr>
<th>Current Harmonic</th>
<th>Synchronous Speed (rpm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1000</td>
</tr>
<tr>
<td>3</td>
<td>0</td>
</tr>
<tr>
<td>5</td>
<td>(-5 X 1000) = -5000</td>
</tr>
<tr>
<td>7</td>
<td>(7 X 1000) = +7000</td>
</tr>
</tbody>
</table>

Table 5.1 Synchronous Speed of Time Harmonic mmf for 6-pole SRIM

5.3.2 Stator Space Harmonics

A sinusoidally distributed winding in a SRIM is difficult to achieve as mentioned in 5.3.1. In a practical SRIM, the windings are distributed in a limited number of slots. Therefore, the mmf distribution in space cannot be sinusoidal; it has a stepped waveform.

When a sine wave current flows through the windings, the space harmonic waves rotate at \(1/h\) times the speed of the fundamental. If the harmonic wave is of the order \(h=6m+1\), where \(m\) is an integer, then the direction of rotation is the same as the fundamental. If the order is \(h = 6m-1\), then the direction is opposite.

The synchronous speed of the \(h\)th space harmonic wave is expressed by:

\[
n_{sab} = \frac{n_s}{h} = \frac{120f}{hp} \quad \ldots (5.15)
\]
Table 5.2 Space Harmonic mmf for 6-pole SRIM

<table>
<thead>
<tr>
<th>Space Harmonic</th>
<th>‘Speed’ of Harmonic (rpm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1000</td>
</tr>
<tr>
<td>5</td>
<td>-200</td>
</tr>
<tr>
<td>7</td>
<td>142.86</td>
</tr>
<tr>
<td>11</td>
<td>90.91</td>
</tr>
</tbody>
</table>

5.3.3 Measured Stator Current harmonics

The switching action of the controlled rectifiers inevitably results in non-sinusoidal current being drawn from the a.c. supply system. Experiments indicate that there are ‘signature’ harmonics present in the stator currents that correspond to the rotor speed. The identified harmonics are proportional to the rotor speed in both sub and super-synchronous speed regions.

The stator current contains information that may be sufficient for tracking the rotor speed. Although this may be possible to close the loop for speed feedback, but extracting and tracking the varying harmonics is difficult.

The dominant harmonics in the stator current are the fundamental, 3rd and 5th which remain ‘constant’ throughout the motor speed. There exist a distinct family of harmonics present in the stator current, which varies with the rotor speed. One particular harmonic, which varies linearly with rotor speed from sub to super-synchronous speed, lies above the groups of harmonics close to the fundamental. This harmonic remains observable and distinct, whether the machine is generating or
motoring. The harmonic is shown in Fig. 5.4 at 642Hz, with the rotor at about 800 rpm.

When the SRIM is driven by a d.c. machine and generating, the current harmonic spectrum in the rotor circuit becomes very complex and is difficult to analyse. The stator current harmonic spectrum, however, remains far less complex and the signature harmonics are detectable.

From Fig. 5.6, the signature harmonic has moved to 927Hz with the rotor at about 1200 rpm. This indicates that this signature harmonic varies linearly with rotor speed from sub to super-synchronous range. The linearly varying relationship is indicated in table 5.1.

Although the rotor speed may be observed by the stator current harmonics, there is no direct indication for rotor position. Therefore, the rotor position has to be estimated, or predicted by examining the other terminal quantities.
Fig. 5.4 Stator Current Harmonics Sub-synchronous Motoring 800rpm approx.

Fig. 5.5 Stator Current Harmonics Sub-synchronous Generating 800rpm approx.
The method of current harmonic tracking for SRIM control may not be practical, mainly because of the limitations of the speed of the A to D conversion time. A very fast FFT scheme is necessary to determine the SRIM harmonics and to
identify the signature harmonics. Also, the signature harmonics vary according to rotor speed, which makes the program task complex and difficult.

As the FFT computing time is very significant, implementation in a real time closed loop system requires a very fast computer, fast data acquisition card and complex software. All these will involve prohibitive a cost and long development time. If the hardware and software are compromised, control will be impossible because the time delay will be significant. This harmonic tracking scheme is therefore only suitable for off line signal acquisition and analysis.

5.4 **Rotors Current Harmonics**

Theoretically, the rotor current is a quasi square wave switched from the d.c. link. In practice, the current wave is not flat topped because the d.c. link current contains a ripple component depending upon the inductance and circuit parameters. The presence of harmonics in the link, which could be high for small d.c. link choke values. This will allow harmonic currents in the secondary circuit. Consequently, an increase in secondary loss is inevitable and so reduce the overall system efficiency.

The rotor current observed on the slip recovery rig proved to be complex with rich harmonics in a wide spectrum. Multiple harmonics are expected because the current is a variable frequency six step square wave. Analysis is very difficult as there is no direct indication of the rotor speed or position. The rotor harmonics of the three speed regions are shown in the following figures.
Fig 5.8 Rotor Current Harmonics Sub-synchronous Generating 800rpm approx.

Fig. 5.9 Rotor Current Harmonics at Sub-synchronous Motoring 800rpm approx.
As the rotor current is a controllable vector quantity for the drive, monitoring the rotor current may be relevant for protection purposes or for checking the control circuit operation.

5.5 Rotor Terminal Voltage

The rotor voltage varies in both magnitude and frequency, and is a function of slip and rotor speed. Ignoring losses in the machine, and neglecting both stator and rotor resistance and the leakage inductance, the r.m.s. induced voltage per phase in the rotor winding is:
\[ V_R = \frac{V_s \omega_R}{n \omega_s} \text{ Volts} \]

\[ \text{.... (5.16)} \]

where \( V_s \) is the r.m.s. supply voltage.

The rotor circuit may be modelled by a voltage source, in series with a rotor resistance and rotor reactance. Therefore, the rotor terminal voltage can be expressed by :

\[ V_{RT} = \left( \frac{V_s \omega_R}{n \omega_s} + I_R R_R + L_R \frac{dI_R}{dt} \right) \text{ Volts} \]

\[ \text{.... (5.17)} \]

where

\( V_{RT} \) is the rotor terminal voltage
\( I_R \) is the rotor current
\( R_R \) is the rotor resistance
\( L_R \) is the rotor reactance

As the voltage spikes are suppressed by terminal capacitors, the rotor terminal voltage is approximately equal to the rotor e.m.f.

\[ V_{RT} = \left( \frac{V_s \omega_R}{n \omega_s} + L_R \frac{dI_R}{dt} \right) \text{ Volts} \]

\[ \approx V_R \]

\[ \text{.... (5.18)} \]
By experiment, the rotor terminal voltage is not purely sinusoidal. A large amount of ‘ripple’ was observed. Another serious problem is at near synchronism, the rotor terminal voltage is non zero due to the presence of ‘noise’. It is thought that such ‘noise’ is the result of inverter/converter operation and a limited link inductance.

To investigate the possibility of ripple due to the link inductance, a very large inductor (1H at 3A, air cored inductor) and the proposed π filter in the link were installed. However, the significant voltage ripple at all rotor speeds, and ‘noise’ at near synchronism remained about the same as before.

5.5.1 Rotor Harmonics due to Time and Space mmf Harmonics

Equations (5.16-18) ignored the space and time mmf harmonics in the SRIM as discussed in sections 5.3.2 and 5.3.3. When the rotor is at standstill or locked, the induced voltage magnitude is largest and the voltage harmonics due to the time and space mmf harmonics are comparatively much smaller and therefore not observable. When the rotor speed increases from stand still, the induced e.m.f. diminishes; in this situation the harmonics become much more significant. At near synchronism, the induced e.m.f. is zero but the voltage harmonics are significant. When the rotor speed is above synchronism, the fundamental voltage reappears and the harmonics slowly become less prominent as the speed increase. At near synchronism, between 950 to 1050 rpm, the fundamental induced voltage is difficult to observe due to the large amount of harmonics.
The voltage harmonics appearing in the rotor varies due to the relative speed of the rotor speed to the stator mmf. From equation 5.10, the mmf 'seen' by the rotor is modified as follows:

$$ F_{R1}(\theta, t) = \sum_{h=1}^{\infty} F_{h \text{max}} \left( \cos \left( (h \omega - \omega_R)t \right) \cos \theta + \cos \left( (h \omega - \omega_R)t - 120^\circ \right) \cos (\theta - 120^\circ) + \cos \left( (h \omega - \omega_R)t + 120^\circ \right) \cos (\theta + 120^\circ) - \frac{3}{2} F_{1\text{max}} \cos (\theta - (\omega - \omega_R)t) \right) $$

$$ \ldots \text{(5.19)} $$

where,

$\omega$ is the rotational speed of the mmf in radians per second.

$\omega_R$ is the rotor speed in radians per second.

The third harmonic mmf is absent:

$$ F_{R3}(\theta, t) = \sum_{h=3}^{\infty} F_{h \text{max}} \left( \cos \left( 3 \omega \right) \pm \omega_R \right) t \left[ \cos(\theta - 120^\circ) + \cos(\theta + 120^\circ) \right] $$

$$ = F_{h \text{max}} \times 0 $$

$$ = 0 $$

$$ \ldots \text{(5.20)} $$

The fifth harmonic mmf rotates in the opposite direction, with respect to the fundamental, at five times the speed of the fundamental stator mmf wave. However, with respect to the rotor, the speed of this mmf is not merely five times the speed of the fundamental. The total speed, by superposition, is the sum of the 5th harmonic added to the rotor speed:
\[ F_{R5}(\theta, t) = \frac{3}{2} F_{5\text{max}} \cos (\theta + (5\omega + \omega_R) t) \]

\[ \text{.... (5.21)} \]

The seventh mmf harmonic is rotating in the same direction as the fundamental, and at seven times the angular speed. With reference to the rotor, the net speed is the 7th harmonic speed minus the rotor angular speed:

\[ F_{R7}(\theta, t) = \frac{3}{2} F_{7\text{max}} \cos(\theta - (7\omega - \omega_R) t) \]

\[ \text{.... (5.22)} \]

The other mmf harmonics are in the order of \( h = 6n +\pm 11 \), where \( n \) is an integer. The harmonics are odd numbers and not multiples of 3.

\[ F_h(\theta, t) = \frac{3}{2} F_{h\text{max}} \cos(\theta \pm (h\omega \pm \omega_R) t) \]

\[ \text{.... (5.23)} \]

At synchronism, the induced fundamental voltage is zero. From equations 5.20, the resultant angular speed for the 5th voltage harmonic appearing in the rotor at the synchronous speed of 1000 rpm is \( = 1000 + 5000 = 6000 \) rpm. The resultant angular speed for the 7th voltage harmonic is \( = 7000 - 1000 = 6000 \) rpm. The two harmonics combine to give the 6th time harmonic. With reference to the 50Hz mains fundamental, this is at a frequency of 300Hz.
In the sub-synchronous speed region, say, 900 rpm, the induced fundamental is non zero. The harmonics at 5^{th}, 7^{th} and above are superimposed on to this fundamental voltage. From equations 5.20, the resultant angular speed for the 5^{th} harmonic appearing in the rotor at 900 rpm is = 900 + 5000 = 5900 rpm. The resultant angular speed for the 7^{th} harmonic is = 7000 - 900 = 6100 rpm. These two harmonics are now distinctly different at 5.9 or 6.1 times the fundamental 50Hz mains frequency. Therefore, the frequencies are 295Hz and 305Hz.

In the super synchronous speed region, say, 1100 rpm, the induced fundamental is also non zero and superimposed with the odd order harmonics. Although the phase sequence changed, the directions of angular velocities for the fundamental and harmonics remain the same. From equations 5.20, the net resultant angular speed for the 5^{th} harmonic appearing in the rotor at 1100 rpm is = 1100 + 5000 = 6100 rpm. The resultant angular speed for the 7^{th} voltage harmonic is = 7000 - 1100 = 5900 rpm. The two harmonics are distinctly different at 6.1 and 5.9 times the fundamental 50Hz mains frequency. Therefore, the frequencies are 305Hz and 295Hz. In comparison with the 900 rpm instance, the 5^{th} and 7^{th} harmonics have now swapped positions.

Higher voltage harmonics, the 11^{th}, 13^{th}, etc, are ignored due to their negligible magnitudes.

The space harmonics, like the time harmonics, are also altered due to the rotor speeds. The synchronous speed of the hth space harmonic wave is modified from equation 5.14 as:
\[ n_{sh} = \frac{n_s}{h} \]
\[ = \frac{120f - \omega_R}{hp} \]

\[ \ldots (5.24) \]

5.5.2 **Measured Rotor Terminal Voltage Harmonics**

To investigate the induced rotor harmonic content, the machine was isolated from the possibility of interference caused by the CSI or the converter bridge. The SRIM was disconnected from the drive and the stator windings were excited directly from the mains supply via a variac. The rotor windings were open circuited and the terminal voltages were measured, with and without power factor correction capacitors connected in parallel with the windings. The SRIM was mechanically driven by a controllable, separately excited d.c. machine from sub to supersynchronous speeds.

A spectrum of the voltage harmonics of the rotor terminal voltage is shown in fig. 5.11 to fig. 5.17. From fig. 5.11, the measurement at standstill shows the fundamental at 50Hz and two distinct harmonics, the 5th, and the 7th. Although there are higher harmonics, they are disregarded due to their negligible magnitude.

As the rotor began to rotate, the fundamental frequency in the rotor began to decrease as well as the voltage magnitude. As soon as the rotor began to rotate, the distinct groups of harmonics can be seen in the regions of around 100Hz, 200Hz, and 300Hz. If the mains frequency is taken as the reference, then these are the
second, fourth and sixth harmonics. There is no direct explanation of the second and fourth harmonics. However, one possible explanation could be an imbalance in the windings of the machine.

The terminal voltages are non sinusoidal and appear to be much distorted as rotor speed increases. The distortion is due to the higher content of harmonics as the fundamental voltage decrease in both magnitude and frequency. The harmonics are certainly not caused by the switching action of the CSI. At synchronism, the terminal voltage is non zero and contain large amount of 'noise', or frequencies at 100Hz, 150Hz and particularly, the 6th harmonic at 300Hz.
From the theoretical predictions, the two distinct harmonics in the 300Hz region are the result of the rotor speed, and the 5th and 7th harmonic. The two harmonics combine at synchronism to form one single harmonic at 300Hz, which also agrees with the theoretical prediction.

Fig 5.11 Rotor Terminal Voltage Harmonics $\omega = 0$ rpm
Fig. 5.12 Rotor Terminal Voltage Harmonics $\omega = 600$ rpm

Fig. 5.13 Rotor Terminal Voltage Harmonics $\omega = 800$ rpm
Fig. 5.14 Rotor Terminal Voltage Harmonics $\omega = 900$ rpm

Fig. 5.15 Rotor Terminal Voltage Harmonics $\omega = 1000$ rpm
Fig. 5.16 Rotor Terminal Voltage Harmonics $\omega = 1100$ rpm

Fig. 5.17 Rotor Terminal Voltage Harmonics $\omega = 1200$ rpm
5.6 Summary

The investigation of all SRIM terminal voltages and currents have been presented. The investigation show that for the CSI fed SRIM, the phase and magnitude of the rotor terminal voltage is observable despite the presence of significant machine noise.

The stator voltage is largely determined by the mains and do not provide useful information for the control system to synchronise the CSI to the varying rotor e.m.f.'s. The stator current contain harmonics that vary linearly with rotor speed. However, the current harmonics are difficult to extract and track in real time. The hardware and software to implement a real time FFT system is not practical.

The rotor current magnitude is set by demand and its phase sequence is electronically synchronised to the varying rotor e.m.f.'s by power electronics device. The rotor current is observed for determining the torque, protection purposes, and to verify the operation of the CSI.

The rotor terminal voltage may provide information directly related to the rotor e.m.f.'s. The signal to noise ratio is good and the e.m.f.s may be extracted with simple circuitry. However, the terminal voltage contains significant spatial and time mmf harmonics.
Chapter 6

Development and Verification of a Novel Sensorless Control Technique for Slip Ring Induction Machine

6.1 Introduction

Control of electronic drives are becoming increasingly more sophisticated, especially when it is desired that conventional sensors such as a tacho-generator and incremental shaft encoders are to be avoided on grounds of cost, reliability, and adverse environmental factors in which the machine may operate. A brief review of the current techniques (1995-1996) is considered for their performance, and their relevance to the SRIM running in super synchronous mode. Despite the numerous sensorless schemes proposed by many authors, none offer a unified theory to handle different induction machine types, to provide a low cost and robust sensorless system for slip recovery. In principle, the voltage and current quantities from the machine terminals contain the necessary information for deriving the states of the machine. The derivation is done by substituting the V/I quantities into an equivalent machine circuit.

Numerous observer-based and adaptive schemes have been proposed which estimate the rotor slip from the back e.m.f, but these are not suitable for SRIM because of their inherent parameter dependency and failure when the terminal electrical quantity is near zero. These are techniques developed for cage machines. For a doubly fed machine, while operating at synchronous speed, the rotor terminal voltage at the slip rings is superimposed with slot noise and the induced e.m.f. is zero at synchronous speed.
6.1.1. **Rotor Speed Estimation Using An Adaptive System**

The traditional approach to sensorless vector control employs field orientation relative to the rotor flux linkage vector or time derivatives. These methods have difficulty in identifying the instantaneous orientation of the relevant vector. A method to identify the field-orientated reference frame is using a model-reference adaptive system (MRAS). Schauder [13] proposed an MRAS for the estimation of induction motor speed from the measured terminal voltages and currents. The motor voltages and currents are measured in a stationary reference frame. From the complete motor parameters, the instantaneous speed of the rotor can be calculated directly from the measured voltages and currents. To obtain a dynamic representation of the motor speed, the calculations are based on the coupled circuit's equations in the stationary frame. The proposed method does not cover position estimation.

6.1.2. **Rotor Speed Estimation using Current Harmonic Spectral Estimation**

Many sensorless techniques using current harmonic analysis are dependent on specific machine structural parameters. The machine slot harmonic magnitude and frequencies depend on unknown parameters, such as the number of rotor slots. Therefore, the algorithm requires user input such as slot harmonic equations, which may vary from machine to machine.

Habetler [33] indicates that speed-related harmonics that arise from rotor slotting and eccentricity can be analysed using digital signal processing (DSP), as analogue filtering cannot derive accurate speed information from the low bandwidth
harmonics. These current harmonics exist at any non-zero speed and are independent of time varying parameters. The DSP can be used to extract speed-related current harmonics, and the processed data is fed to a computer algorithm. The algorithm samples just one phase of the stator current. After filtering, an aliasing technique combines multiple slot harmonics to increase the signal to noise ratio. A spectral estimation method is applied to determine the slot harmonic frequency. The low-bandwidth harmonics are employed to enhance detection and frequency resolution. The above forms the basis of a speed observer in a closed looped speed control system.

Beguenane [64] proposed that the stator current is sufficient to provide information to determine with great accuracy the steady state rotor speed or slip regardless of the type of machine supply. The rotor time constant identification is achieved from the stator current and voltage, together with the determination of the stator resistance. Slot harmonics are extracted from the stator current using a DSP. However, the procedure is limited to a steady state operation and fail in transient state due to the time required for current acquisition and FFT calculations. Also, there is no proposal for detection of rotor position.

6.1.3. Field Orientation Concepts Employing Saturation-Induced Saliencies in Induction Machines

Jansen and Lorenz [60] proposed a method of continuously tracking the position and magnitude of flux using a heterodyning technique and a closed loop tracking filter, thereby providing a scheme to estimate both rotor speed and position.
Traceable magnetic saliency and high frequency signal injection are the two main features for the accurate and robust estimation, independent of speed and load.

To obtain the rotor speed information, Lorenz [60] observed the stator transient inductance as the basic parameter of the induction machine, instead of the rotor quantity derived from the leakage and magnetising inductance. The stator transient inductance containing a saturation-induced saliency, can be expressed in the synchronous reference frame. The stator transient inductance can also be expressed in the stationary frame, which contains the angle of the rotating saliency.

By injecting a high frequency signal, the stator current will contain useful position information in the q and d axis current. The spatial information contained in the stator current is extracted by heterodyning which provides the linear position error information. The accuracy of estimation is independent of the actual inductance and applied signal voltage magnitudes, if the applied polyphase signals are balanced. The high frequency signal is incorporated in the PWM inverter used.

6.1.4. Encoder-based Control System in LUED Slip Recovery Rig

The Slip recovery facility at Leicester University Engineering Department relies on an encoder to provide rotor speed and position information. The line-commutated inverter must be synchronised to the supply frequency to provide four-quadrant operation and must fire the current source inverter bridge during each half cycle of the rotor e.m.f. A novel rotor e.m.f. signal generator was developed which generates a three phase output locked in phase with the rotor e.m.f., and does not change polarity when the machine goes through synchronous speed. The converter is
controlled by the current servo to maintain the current during rectification or regeneration. The speed servo controls the inverter bridge by comparing speed errors to maintain the desired speed. The system provides inherent current limiting, which is an additional advantage.

To replace the present system for sensorless control, the slip ring machine behaviour was studied in chapter 5 to identify the most suitable observable parameter as a control signal. It is desired to use the machine itself as the encoder, i.e., the electrical quantities to provide rotor speed and position information.

6.1.4.1. Rotor e.m.f. Signal Generator

The present slip power recovery control system relies on an optical mechanical shaft encoder to provide the speed, phase and rotor position information of the machine. The rotor E.M.F. generator generates three phase square waves, which are locked in phase with the rotor e.m.f. Pulses proportional to the supply frequency, the motor speed (divided into 64 counts per frequency cycle of the supply) and the rotor e.m.f., are compared in an up/down counter. Pulses are generated from an encoder on the shaft of the SRIM. The direction of count for the up/down counter is determined by which input is high when the other count input is high. A miscount will occur if both inputs (reference at supply frequency and shaft encoder signals) are pulsed simultaneously. To synchronise the signals and to allow for phase advance control, the pulses are passed through D type flip-flops to ensure that pulses are passed only during the rising edge of the clock timing. The four-phase clock ensures that pulses to the counter never occur simultaneously.
The decoder outputs corresponding to the three-phase displacement are used to operate six AND gates to generate the square waves as shown in Fig. 6.2.

Fig. 6.2 Rotor e.m.f. Waveform
The decoder must be synchronised to start at the beginning of a motor rotor e.m.f. half cycle by enabling the circuit at the instance the positive half cycle starts. The square waves are synchronised to the standstill rotor e.m.f. and will lock to an angle $\theta$ with respect to the motor primary voltage (the angle $\theta$ is usually zero). As the reference count is from the primary side, the rotor e.m.f. generator will remain synchronised at an angle $\theta$ with reference to the primary. The signals, after processing by logic circuits, fire the thyristors in the ASCI.

As shown in the above analysis, the current is injected by examining the phases of the rotor e.m.f.'s. Both the rotor speed and position detection/estimation are required from the intended sensorless control system to replace the existing shaft encoder. It is also desired that the new control system has a lower cost and a more robust performance than the traditional method.

6.2 Rotor Terminal Voltage of SRIM

The control of the slip ring machine in four modes of operation is achieved by injecting the current from the current source into the rotor circuit. The starting of the six step current is synchronised, either in or out of phase, to the start of the rotor e.m.f. To obtain better power factor, the six-step current may be phase advanced or retarded to the e.m.f. The introduction of phase advance or phase retard is done by manual adjustment.
The following diagrams show the theoretical firing sequence for one phase of the rotor circuit in sub/super-synchronous, motor or generating modes.

By transformer action, the rotor e.m.f. is induced in the rotor circuit from the stator. When the rotor is rotating, the magnitude and frequency of the rotor e.m.f. varies according to the slip. From the approximate model of the rotor circuit, the terminal voltage is identified to be a close approximation of the rotor e.m.f. The following are the assumptions, which apply to support the approximation:

1. The series resistance of the rotor circuit is negligible.
2. The series inductance of the rotor circuit (varying under load conditions) is small under all conditions.
3. The rotor parameters do not vary substantially when the machine is hot.
4. The current source inductance is large, and much larger than the self-inductance in the rotor winding.
5. The mains voltages are sinusoidal and harmonics/voltage spikes are low.

6.2.1 **Rotor Terminal Voltage as a Control Signal**

The initial observation of the rotor terminal voltage indicates that the voltage waveform magnitude has a much better signal to noise ratio than observing the current or voltage harmonics as a control signal. The 'ripples' that appear in the rotor voltage is a result of the inverter bridge and harmonics discussed in Chapter 5. Increasing the inductance by 30% from the original value of 1 Henry (5A) help suppress some 'noise'. At super-synchronous speed as shown below, the ripple in the terminal voltage is substantial. It is noted that encoder system does not always guarantee a phase angle of $\alpha=0^\circ$ between the rotor e.m.f. and the rotor current, although the electronics is to ensure no phase difference.

![Fig. 6.4 Terminal Voltage and Rotor Current with low Link Inductance at Slip=-0.5](image-url)
The noise voltage was found to be varying in the region between 300-400Hz, and has been discussed in Chapter 5. The L-C type filter may be the only suitable circuit to effectively remove this unwanted 'noise' from the terminal voltages, with negligible phase delay. Other filters using R-C introduce severe phase delay.

![Fig. 6.5 Ripple/Noise in Terminal Voltage at s = -0.5](image)

By examining the encoder based system characteristics, experimental results confirm that the terminal voltage may be a close approximation of the rotor e.m.f.:

a. With the rotor locked, the terminal voltage reflects the stator voltages as shown in fig. 6.6. The two quantities are in phase.

b. The terminal voltage has an insignificant magnitude variation under a varying line current as shown in fig. 6.7.

c. The terminal voltage has an insignificant small phase variation under a large varying line current as shown in fig. 6.8.
d. The rotor circuit parameters of the Brook Crompton machine was found to be low (Appendix 1).

e. The link inductance was increased by 30%. This helped reduce the current ripple and improve the torque performance. However, extra measures were necessary to suppress the increased $dv/dt$ conditions in the d.c. link.

![Fig. 6.6. Locked Rotor Terminal Voltage](image)

Further investigations of the rotor terminal voltage and the rotor current are presented in Appendix 2. The simulation of a single-phase rotor equivalent circuit showed that the phase variation between the terminal voltage and the rotor e.m.f. is insignificant under a large current. The measured voltages and currents under various rotor speed are also presented Appendix 2. Experiments showed that the existing encoder system has its limitations. The current waveforms were at times leading or lagging the rotor terminal voltages under steady state operations.
Fig. 6.7. Shows a step increase in rotor load torque of approximately 2 Nm

Fig. 6.8. Shows a step decrease in rotor torque of approximately 10 Nm
6.3 Novel Sensorless Control for Slip Ring Induction Machine

6.3.1. Novel Sensorless Scheme

A novel sensorless control method was developed by employing adaptive filtering of the rotor terminal voltage, and further signal processing, to provide the synchronisation signals for the CSI. The majority of the existing encoder based control system does not warrant redesign. Module such as the current servo, power electronics interface and power electronics remain the same. The adaptive filtering is achieved using analogue circuits, instead of using a DSP system, due to time and cost constraints.

Usually, sensorless control schemes for induction machines involve complex DSP and computation power, as shown by Vas [83]. The current trend for sensorless control certainly involve any or all of the following:

1. Digital Signal Processing Theory
2. Machine/ Machine characteristic sensitive parameter estimation
3. Complex machine parameter identification
4. Digital Control Hardware

Due to the time constraints on a three-year research programme, DSP hardware and software systems for the sensorless control of this slip recovery drive involve long development times. (More than 1½ years’ worth of time and effort was spent on digital data acquisition and processing on machine Current and Voltage analysis.) A TMS320 starter kit and a TMS320C5x evaluation module were used
for data acquisition and to generate an F.F.T. function. As there are many
parameters of the machine to observe and many functions to perform, the sequential
nature of the PC makes the computer speed critical. As such, the DSP approach
requires complex hardware and software design. A DSP system was not attempted
in this work. However, a DSP system can be implemented for future work, based on
the theory and verification presented.

6.3.2 Novel Sensorless Control Theory

Work presented in Chapter 5 shows that the rotor terminal voltages allow the
prediction or estimation of the secondary e.m.f.. A novel sensorless scheme is
proposed based on the observation of the terminal voltages, which can be sensed
and further processed to estimate the e.m.f. using analogue circuits or a DSP.

Although it is desired that the link inductance has to be large to provide
isolation from the mains, the adaptive filter is sufficient to extract the required
signals in real time. The terminal voltages are converted into control signals using
simple analogue implementation. When greater flexibility of the control system is
required, a DSP can be implemented, but at cost.

Direct observation of the rotor terminal voltage with adaptive filtering can
provide both the rotor speed and rotor position information. The majority of the
existing slip power recovery rig hardware is retained, in order to compare the
performance with the proposed novel sensorless control scheme. The line side
converter and the current servo can remain the same, as the sensorless scheme
concerns the synchronisation of the CSI with the rotor e.m.f. only.
The new sensorless scheme may control the CSI feeding the rotor circuit in both current and speed control modes. In the current mode, the amount of current into the rotor is set by a current demand. In the speed control mode, the current is regulated such that the rotor speed is made constant.

As the control signal in this sensorless scheme is derived from constantly monitored machine rotor terminal voltages, robustness is expected to be much higher than any other schemes that rely on voltage prediction or parameter estimation. The simpler circuitry also reduces system complexity and costs.
6.3.3 System Description

The following figure shows the system diagram of the proposed novel system. The current servo and the power electronics of this system are similar to that of the existing encoder system.

Fig. 6.9. Novel Sensorless Control System Diagram
The system uses three fast response Hall effect voltage transducers (LEM), connected to the rotor terminals for voltage sensing. The transducers are configured such that their maximum range is the maximum voltage appearing at the rotor terminals. This is to provide the maximum resolution.

The sensed voltage waveform is fed to an adaptive filter, which extracts the fundamental e.m.f. Due to the stringent phase requirement, i.e., zero phase shift, the usual filter designs using R and C, or even the conventional state variable op-amp filters become impractical, due to the large and variable amount of phase shift. It is concluded that practical RC filters cannot meet the brick wall type pass band response and zero phase-shift. Attempts using parallel harmonic traps to remove the noise also proved impractical, due to the limited attenuation and significant phase delay due to R-C effects. In practical circuits, harmonic traps cannot meet the zero phase shift requirements. Filtering using integrators by Bin-Wu [5] was also investigated, but was found to be not suitable due to the phase delay and the problem of integrators with very low frequency signals.

A filter with a fixed cut-off frequency cannot provide the required attenuation for the zero to fifty Hertz frequency range. Two filters, each set at different cut-off frequency, are used to recover the rotor e.m.f. from the measured terminal voltages. The design employs adaptive L-C filters, which have virtually zero phase shifts in a very simple realisation. When the machine e.m.f. is higher that 25Hz, the cut-off frequency is set at 20Hz. This provides excellent filtering with very little phase delay. When the e.m.f. becomes lower, the cut-off frequency is
automatically switched to a low frequency of 1Hz. This ensures zero phase shift at all machine speeds and provides adequate filtering. During the change over from one L-C filter to another, the output from both filters are approximately similar. Another R-C filter stage is used after the L-C filters. This ensures smooth transition when the two L-C filters are switched from one to another.

The filtered signal is fed through to a zero crossing detector with adjustable hysteresis to ensure noise immunity. At every zero crossing, a short pulse of 10ms is generated using an XOR logic function integrated circuit. The zero crossing pulse is coupled to AND gates to give precisely the phase information of each rotor e.m.f. The circuit exhibits good dynamic performance, and the verification is shown in Appendix 5. A limited phase advance/retard of the zero crossings is possible by adding further level shifters to the zero crossing detectors. However, this facility has not been implemented due to time constraints.

From the output of the filter, the rectified mean voltage magnitude is directly proportional to the machine rotor speed, and is used as a speed signal. The adaptive filter also uses this quantity to determine the instant where the L-C filter cut off frequency should change.

The pulses from the zero crossing detectors clock a six step ring counter, which in turns clocks an EPROM that has the firing pattern of the CSI. The pulses from the logic circuits, with additional logic gates, also determine whether the machine is in sub/super-synchronous speed region. This allows the proper page selection of the EPROM stored data, according to the machine status.
In summary, the sensorless control system generates the following control signals to replace the pulses from the encoder:

1. Sub / Super-synchronous speed
2. Motoring/ generating mode
3. Red-Yellow phase voltage zero crossing synchronising pulse
4. Yellow-Blue phase voltage zero crossing synchronising pulse
5. Blue-Red phase voltage zero crossing synchronising pulse

### 6.3.3.1 Adaptive L-C Filter Circuit

![Diagram of Adaptive L-C Filter Circuit](image)

**Fig. 6.10 Novel Adaptive Zero Phase Error Filter**

(Note: The R-C filter after the L-C filters absorbs spikes and ensures a smooth transition)
The practical realisation of the analogue adaptive filter is shown in Fig. 6.10. The filters are constructed using simple L-C filters driven by op-amp buffers. The cut off frequency changes according to the average magnitude of the rotor e.m.f.'s.

### 6.3.3.2 Six Step Ring Counter

![Six-Step Ring Counter Diagram](image)

The filtered voltage approximates the rotor e.m.f., and is passed to logic circuits to generate pulses locked to the phase sequences. The zero crossings of the three phase rotor voltages correspond to each change of the six step current waveform. The zero crossing instances of the three phase voltages are identified as R+, R-, B+, B-, Y+ and Y-. When the machine is in the super-synchronous region, the phase sequence reverses.
6.3.3.3 Sub/ Super Synchronous Speed Detection

Examining the phase sequence of the three phase voltages allows the identification of the rotor speed region. At sub-synchronous speeds, the R phase cycle and B+ pulse, together with the AND gate, will yield a positive pulse edge. However, at super-synchronous speed, the same combination will not give any output. This property is used to identify the speed region, which facilitates the change of firing sequence as the machine goes through synchronism. A D-type flip-flop is used. The AND gate delays the EPROM clock to ensure a change of state.

Fig. 6.13 Sub/ Super-Synchronous Speed Detection
6.3.3.4 EPROM Data

The control of the CSI is possible and simple with the generated pulses. The pulses clock the stored waveform in the EPROM according to the variation of the e.m.f., and the triggering signals are fed directly to the existing interface to the thyristor drivers.

![Simplified EPROM Circuit](image)

Fig 6.14 Simplified EPROM Circuit

With reference to the R-Y phase, the EPROM is programmed as in the following tables to provide the appropriate firing sequence.

### Sub Motoring

<table>
<thead>
<tr>
<th>Period °</th>
<th>0-30 °</th>
<th>30-60 °</th>
<th>60-90 °</th>
<th>90-120 °</th>
<th>120-150 °</th>
<th>150-180 °</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sub/Sup</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>Mot/Gen</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>R-Y</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>Y-B</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>B-R</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>Thyristors RY</td>
<td>3-2</td>
<td>3-4</td>
<td>5-4</td>
<td>5-6</td>
<td>1-6</td>
<td>1-2</td>
</tr>
</tbody>
</table>

### Sub Generating

<table>
<thead>
<tr>
<th>Period °</th>
<th>0-30 °</th>
<th>30-60 °</th>
<th>60-90 °</th>
<th>90-120 °</th>
<th>120-150 °</th>
<th>150-180 °</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sub/Sup</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>Mot/Gen</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>R-Y</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>Y-B</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>B-R</td>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>Thyristors RY</td>
<td>5-6</td>
<td>1-6</td>
<td>1-2</td>
<td>3-2</td>
<td>3-4</td>
<td>5-4</td>
</tr>
</tbody>
</table>
Each state of the rotor is represented by a unique address and there is no conflict. However, care must be taken not to short the CSI when changing mode. For example, to switch from sub-motoring to sub-generating, the thyristor firing sequence would suddenly change from T2 to T3 conducting to T5 and T6. Although the commutation capacitors should help turn off the previously conducting devices, it is not certain that the commutation will be complete.

6.4. Verification of Novel Control Circuit

The filter circuit provides good stability in tracking the e.m.f. and provides a similar firing sequence to the existing encoder based system. A verification of the control circuit is presented in Appendix 5.

The plots indicate that from about 500-rpm sub-synchronous speed to 1500 rpm super-synchronous speed, the novel electronics is able to track the zero crossings of the terminal voltage with negligible phase delay.
Difficulties may be encountered to bring the machine from sub-synchronous speed to super-synchronous speed, or to bring the machine down from super synchronous speed to sub synchronous speed. The main reason is that at synchronism there is no induced rotor e.m.f. and therefore there is no signal. The six step ring counter is halted, until a pulsed signal reappears. However, this signal can be from either the sub-synchronous or super-synchronous region, depending on whether the machine is accelerating or decelerating.

The six step counter has a reset pulse timed to occur at the zero crossing of the R-Y's e.m.f.. This ensures that the six step counter always starts at the new cycle of the e.m.f., regardless of which instant it was halted previously in which the e.m.f. diminished to zero.

Due to the fixed sequence in the pages of an EPROM, the 'reappeared' signal only clocks the counter to the end of the steps, and will wait until the new e.m.f. cycle appears. (i.e., the electronics 'wait' for the rotor to spin to the correct position before the CSI is started again). This means that at certain short instances, the injected current may be out of phase with the e.m.f.. The stability of the machine during this period is not affected, because the e.m.f. is almost zero and the angular instant is short.

Comparison with the encoder system signal output indicates good agreement with the output of the encoderless system, as shown in Appendix 5.
Fig. 6.15 Interior of the Slip Power Recovery Rig

Fig. 6.16 Slip Ring Machine Showing Encoder and P.F. Correction Capacitors
Fig. 6.17 Sensorless Control Circuit – VTs and Filters
Chapter 7

Conclusion and Further Work

7.1 Slip Energy Recovery Scheme

The Scherbius system has a distinct advantage than the conventional Kramer system. It enables the secondary circuit power to flow in both directions, and therefore enables sub- and super-synchronous speed control of the SRIM. This provides the advantage of producing up to twice the shaft power when the machine is operating in the super-synchronous speed region. The encoder, which may be looked upon as a mechanical signal generator operating with additional control logic circuits, is conventionally used to initiate the firing instants of each thyristor in the inverter bridge.

Forced commutation of the inverter is necessary near the synchronous speed when the rotor e.m.f. tends to zero. The choice of inverter to be used is severely restricted by the requirement for reversal of power flow for sub- and super-synchronous operation. The CSI inherently provides power reversal, although it has its limitations. Experiments showed that the SRIM could be controlled to provide stable operation from standstill to approaching twice the synchronous speed. The current source inverter controls the secondary current by synchronising to the secondary e.m.f..

A simple model for the Kramer slip recovery system was used for Kramer analysis. A comprehensive Scherbius analysis is also provided to study the machine
behaviour. The test results for both the Kramer and Scherbius systems for constant torque operation over a wide speed range show a good overall efficiency.

The advantage of running the Scherbius system above the synchronous speed is that nearly twice the mechanical power can be developed for the same frame size of the machine and power electronics. The results are shown in Appendix 4.

7.2 **Slip Energy Recovery Operation**

The encoder system does not have the stability problems normally associated with doubly fed machines. A Basic Program was used to simulate the machine operations from sub- to super-synchronous speeds. The tabulated results in Appendix 4, which considers the power flow and VA ratings, show that there is a definite advantage in working over an equal range of speed. There is a trade-off between the VA rating of the rotor controller and the range of on-line power recovery from the SRIM to the grid system.

To maximise the Kramer system inverter power factor, the machine stator to rotor ratio should be unity (the open circuit secondary voltage should match the supply). The power factor can be slightly compensated by providing a capacitor bank across the supply terminals, which also serve as a short circuit path for current harmonics.
In the Kramer system, power factor will be less than unity, even if the turns ratio is chosen to be unity. To achieve a secondary power factor of unity, the diode bridge must be replaced by a controlled thyristor bridge in which the firing control to the thyristors is synchronised to the secondary e.m.f.

A Scherbius system has many advantages over the 'through-pass inverter', although this requires complex control and forced commutation circuits compared with conventional static Kramer equipment. The overall efficiency and power factor of the Scherbius scheme is nearly constant over the entire range of the power recovered.

A Scherbius analysis has been derived for the slip recovery drive in which the power electronic converters set both the phase angle and magnitude of the rotor current. With the machine parameters known and the rotor current set as an input to the model, the operating characteristics can be determined. The analysis can easily be extended to determine the phasor sum of the stator current and the line current recovered from the rectifier in the rotor circuit, to obtain the overall operating power factor of the drive. A Microsoft Basic program can be developed to incorporate the equations in the Scherbius analysis to predict the SRIM operation at various conditions. A program listing is shown in Appendix 3.

The model of the Scherbius system is necessary to maximise the system efficiency in a SRIM Wind Energy Converter system. Usually, the wind turbine has non-linear characteristics. It is desired to control the torque (current) of the SRIM through the Scherbius system, to match the turbine power curve to obtain maximum
power transfer. With a suitable machine model, then by knowing just the rotor speed and torque (current), the stator power, rotor power, machine losses, machine efficiency can all be known. By adjusting the firing angle (advance or retard) in the CSI, the best efficiency of the system can be obtained.

7.3 **CSI Commutation**

The effect of rotor e.m.f. on the commutation behaviour in slip recovery schemes indicate problems. When motoring sub synchronously or generating supersynchronously the rotor e.m.f.'s are in such a direction to aid the commutation process. This results in conventional waveforms on the commutation capacitors but with short commutation time.

In the case of motoring super synchronously or generating sub-synchronously, the rotor e.m.f.'s are in such a direction as to oppose the commutation process. These results in larger commutation times, lower commutation voltages and additional steps in the capacitor waveform due to outgoing blocking diodes turning back on.

The shortcomings of the CSI prove that future work is required to improve system performance. The CSI should be replaced in the Scherbius set up, if the super-synchronous region is the primary region of operation. The presence of the link inductance also increases the size of the system. The future reversible rectifier may replace the CSI for better efficiency.
7.4 **SRIM Machine Harmonics**

For the CSI fed SRIM, the phase and magnitude of the rotor terminal voltage is observable despite the presence of significant machine noise. A study of the SRIM electrical quantities shows that the machine harmonics are speed dependent.

The stator voltage is determined by the mains, and may not clearly provide useful information for the control system, to synchronise the CSI to the varying rotor e.m.f.'s. The stator current contains harmonics that vary linearly with rotor speed. However, the current harmonics are difficult to extract and track in real time.

The rotor current magnitude is set by demand and its phase sequence is electronically synchronised to the varying rotor e.m.f.'s by the power electronics. The rotor current is observed for determining the torque, over current protection purposes and for verification of the operation of the CSI.

The rotor terminal voltage is a good approximation of the rotor e.m.f. despite the noise of the system.

7.5 **Novel Sensorless Control Scheme**

A novel sensorless scheme using the direct observation of rotor terminal voltage is proposed. The analogue system is capable of tracking the rotor e.m.f. to produce synchronising signals for the CSI.

However, the analogue system has many shortcomings, requiring considerable setting up and adjustment. The use of large and bulky inductors as
filters is undesirable. Also, the implementation has to be suited to the individual machine. The filter can be improved by using a DSP implementation.

7.6 **Future work - DSP controlled Sensorless Control System**

The DSP system was attempted but was found to be complex and difficult to use in a practical implementation for the SRIM drive. As such, an analogue system was used and proved adequate for sensorless control within the time and budget constraints. For a digital filter, the statistical approach or filter design by modelling is suggested to meet the zero phase delay requirements.

Due to the extensive signal processing, a fast DSP is necessary, despite the relatively slow machine time constant. The quantities that require observation are the three phase rotor voltages and the Link Current. For the three rotor voltages, the DSP has to perform a zero or low phase shift filtering and also to derive the zero crossing instance. Due to the inductance in the rotor, the current in the rotor circuit may be taken into consideration. This is to adjust for the phase delay between the rotor e.m.f. and terminal voltage.

As the machine parameters can vary due to loading conditions and heat, the DSP system can be adopted in future work for performing the tuning of machine parameter estimation for a higher efficiency, if necessary.

The use of the DSP can also allow more modern and efficient switching schemes, for better performance. The CSI should be phased out as discussed in the previous section. Preferably, a sine wave interface should be considered.
REFERENCES & BIBLIOGRAPHY

1. ACARNLEY, P. P. & FINCH, J.W.,

2. ALBRECHT, P. & VOLLSTEDT, W,

3. BASSI, E., BENZI, F., BOLOGNANI, S. & BUJA, G.S.,

4. BELLININ, A., FIGALLI, G., & ULIVI, G.,

5. BIN-WU & GORDON SLEMON,

6. BLAND, R.J., HANCOCK, N.N., & WHITEHEAD, R.W.,

7. BOSE, B.K.,

8. BROWN, J.E., DRURY, W, JONES, B.L., & VAS, P.,

9. CANAY, L.M.,

10. CHATTOPADHYAY, A.K.,
11. CHATTOPADHYAY, A.K.,  

12. CHIN, T.H., & TOMITA, H.,  

13. COLIN SCHAUDER,  
Adaptive Speed Identification for Vector Control of Induction Motors without Rotational Transducers, IEEE Trans., Vol 28 No.5, 1992

14. COMPARISON OF SPEED-SENSORLESS DTC INDUCTION MOTOR DRIVES,  
Damino A, Vas P & Stronach AF.  
European Intelligent Motion Conference 1997.

15. CROWDER, R.M., & SMITH, G.A.,  

16. DEWAN, S.B., & DUNFORD, W.G.,  

17. DRURY, W, JONES, B.L., & BROWN, J.E,  

18. DRURY, W., FARRER, W, & JONES, B.L.,  

19. E BOGALECKA,  

20. ELGER, H., & WEISS, M.,  
21. ERLICKI, M.S.,

22. FANG-ZHENG, PENG & TADASHI FUKAO,

23. HARMS, K. & LEONHARD, W.,

24. HASSE, K.,
Control of cycloconverters for feeding asynchronous machines, IFAC Conference on Control in Power Electronics & Drives, Dusseldorf, pp 537-46. 1977.

25. HINDMARSH, J.,
Electrical machines & their applications, Pergamon Press, Oxford.

26. HO, E.Y.Y., & SEN, P.C.S.,

27. HORI, T., & HIRO, Y.,

28. JOETTEN, R., & ZIMMERMANN, P.,

29. KAWAMURA, A. AMD HOFT, R.,

30. KAZMIERKOWSKI, M.P. & KOPCKE, H.J.,
Comparison of dynamic behaviour of frequency converter fed induction machine drives, IFAC Symposium on Control in Power Electronics and Drives, Lausanne, p 313-20, 1983.
31. KAZUNO, H.,
A wide-range speed control of an induction motor with static Scherbius and Kramer

32. KELKAR, S.S., & PILLAI, S.K.,
A modified rotor chopper for speed control of slip ring induction motors, Proc. 2nd
IFAC Symp. on Control in Power Electronics and Electrical Drives, 1977, pp. 567-
574.

33. KEVIN HURST & THOMAS G HABETLER
Sensorless Speed Measurement Using Current Harmonic Spectral Estimation in

34. KLI MAN, G.B. & PLUNKETT, A.B.,
Development of a modulation strategy for a PWM inverter drive, , IEEE Ind. Trans.
Appl. IA-15, 1979, pp.72-79.

35. KRISHNAN, R., & DORAN, F.C.,
Study of parameter sensitivity in high-performance inverter fed induction motor

36. KRISHNAN, R., & DORAN, F.C.,
Study of a parameter sensitivity in high-performance inverter-fed induction motor

37. KUSKO, A., & SOMUAH, C.B.,
Speed control of a single-frame cascade induction motor with slip-power pump-

38. LAVI, A., & POLGE, R.J.,
Induction motor speed control with static inverter in the rotor, IEEE Trans. Power

39. LELAND B. JACKEN,

40. LEONHARD, W.,
Field oriented control of a variable speed alternator connected to the constant
frequency line, IEEE Conf on Control in Power Systems, 1979, pp. 149-153.

41. LIENAU, W.,
Torque oscillations in traction drives with current fed asynchronous machines, IEE
42. LIPO, T.A., & WALKER, L.H.,

43. LONG, W.F., & SCHMILLZ, N.L.,

44. LORENZ, R.D. & PATTEN, K.V.,

45. M.T. EL-HAGRY & M.N. ESKANDER

46. MAYER, C.B.,

47. MELLO, F.P. DE, & RIBEIRO, J.R.,

48. MEYER, M.,

49. MILJANIC, P.N.,

50. NABAE, A., SHIMAMURA, T., & KUROSAWA, R.,

51. NODA, J., HIRO, Y., & HORI, T.,
52. NOVOTNY, D.W. & LORENZ, R.D.,
Introduction to field orientation and high performance a.c. drives. Tutorial Course.

53. OHNISHI, K., SUZUKI, H., MIYACHIU, K. & TERASHIMA, M.,
Decoupling control of secondary flux and secondary current in induction motor drive with controlled voltage source and its comparison with volts/hertz control,

54. OHNISHI, K., UEDA, Y., & MIYACHI, K.,

55. OHNO, E., & AKAMATSU, M.,

56. OKUYAMA, T., FUJIMOTO, N., MATSUI, Y., & KUNOTA, Y.,

57. OLIVIER, G., STEFANOVIC, V.R., & APRIL, G.E.,

58. OSMAN, R.H.,

59. PAICE, D.A.,

60. PATRICK L. JENSON, & LORENZ,

61. PLUNKETT, A.B., KLIMAN, G.B. & BOYLE, M.J.,
62. POLMANN, A.,

63. POWER ELECTRONICS CONTROL OF AC MACHINE,
JMD Murphy & FG Turnbull, Pergamon Press.

64. R BEGUENANE, C GHYSELEN & H SCHOORNENS,
A Proposed Induction Motor Speed sensor without contact from slot harmonics application to rotoric time constant identification, Power Electronics and Drives conference, IEE, 1995.

65. RAO, N.N., DUBEY, G.K., & PRABHU, S.S.,

66. ROWAN, T.M. KERKMAN, R.J. & LIPO, T.A.,

67. SAI TO, K., K AMIYAMA, K., & SUKEGAWA, Y.,

68. SATHIKUMAR, S. & VITHAYATHIL, J.,

69. SCHAUDER, C.D., CHOO, F.H., & ROBERTS, M.T.,

70. SEN, P.C., & MA, K.H.J.,

71. SEN, P.C., & MA, K.H.J.,

72. SENSORLESS CONTROL WITH KALMAN FILTER ON TMS320 FIXED POINT DSP,
73. SHEPHERD, W., & KHALIL, A.Q.,

74. SHEPHERD, W., & STANWAY, J.,

75. SMITH, G.A.,

76. STEFANOVIC, V.R.,

77. TAMAI, S., SUGIMOTO, H., & YANO, M.,

78. TANIGUCHI, K., & MORI, H.,

79. TSUCHIYA, T.,

80. TSUGUTOSHI OHTANI & KOUJI TANAKA,

81. VAGATI, A., VILLATA, F.,

82. VAN WYK, J.D.,

83. VAS, P., BROWN, J.E. & HALLENIUS, K.E.,
84. VECTOR CONTROL OF AC MACHINE,
Peter Vas, Oxford Science Publications.

85. WAKABAYASHI, T., HORI, T., SHIMIZU, K., & YOSHIOKA, T.,
Commotor-less Kramer control system for large-capacity induction motors for
dring water service pumps, Conf Rec. IEEE Ind. A pp. Soc. Annual Meeting,

86. WALLACH, Y., ERLICKI, M.S., & BEN-URI, J.,

87. WANG, C., NOVOTNY, D.W., & LIPO, T.A.,
An automated rotor time constant measurement system for indirect field orientated

88. WARD, E.E.,
Inverter suitable for operation over a range of frequency, Proc. IEE, 111, 8, 1964,
pp.1423-1434.

89. WEISS, H.W.,
Adjustable speed ac drive systems for pump and compressor applications, IEEE

90. WILLIAMSON, S., & CANN, R.G.,
A comparison of PWM switching strategies on the basis of drive system efficiency,

91. WOLFINGER, J.F., & LIPO, T.A.,
Stability improvement of inverter driven induction motors by use of feedback, IFAC

92. WU, Z.K. & STRANGAS, E.G.,
Feed forward field orientation control of an induction motor using a PWM voltage
source inverter and standardised single-board computers, IEEE Transactions on

93. XU, X., DE DONCKER, R., & NOVOTNY, D.W.,
Stator flux orientation control of induction machines in the field weakening region,

94. XU, X., DE DONCKER, R., & NOVOTNY, D.W.,
A stator flux orientated induction machine drive, IEEE Power Electronics
95. ZACH, F.C., BERTHOLD, R.J. & KAISER, K.H.,

96. ZHANG, J., & BARTON, T.H.,

97. ZHANG, J., THIAGARANJAN, V., GRANT., & BARTON, T.H.,

98. ZIMMERMANN, P.,

99. ZUBEK, J.,
Appendix 1

Motor and System Data

Name Plate details
Brook Crompton Motor
5.5kw at 950 rev/min
Rating: 415V 11.3A 50Hz

Locked Rotor Test
$I_1 = 10A$
$I_2 = 16A$
Turns ratio = 1.6
$T = 2.2 \text{Nm}$
$V_1 = 94V$
$W_1 = 600W$

The machine is $\Delta$ connected.

$I_{ph} = \frac{10}{\sqrt{3}} = 5.8A$
$R_t = 6\Omega$
$X_t = 15\Omega$
Assume $X_1 = 7.5\Omega$, $X_2 = 7.5\Omega$

Open Circuit test at Synchronous Speed

$V_1 = 415V$
$I_1 = 5.6A$
$W = 300w$

Pull Out Test

$S_m = 0.2$

$S_m = \frac{R_2}{\sqrt{R_1^2 + X_T^2}} \approx \frac{R_2}{X_T}$

Stall Speed = 800 rpm
## Steel Construction

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<th>Type</th>
<th>Wound Rotor</th>
<th>Supply</th>
<th>Insulation</th>
<th>Class F</th>
</tr>
</thead>
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<tr>
<td>(External Rings)</td>
<td></td>
<td>(up to 2236kW)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Enclosure</td>
<td></td>
<td>3PH 50Hz 230-600V</td>
<td></td>
<td></td>
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<tr>
<td>Rating</td>
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<td>3PH 50Hz 380-600V</td>
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<tr>
<td>Maximum Continuous (MC)</td>
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### Performance Data

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<th>100</th>
<th>150</th>
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<th>200</th>
<th>250</th>
<th>300</th>
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<th>400</th>
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<tbody>
<tr>
<td>kW</td>
<td>4.4</td>
<td>10.0</td>
<td>15.0</td>
<td>18.0</td>
<td>20.0</td>
<td>25.0</td>
<td>30.0</td>
<td>35.0</td>
<td>40.0</td>
</tr>
<tr>
<td>Hz</td>
<td>50</td>
<td>50</td>
<td>50</td>
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### Approximate Shipping Specifications

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<tr>
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<th>Bearing Type</th>
<th>Bearing Material</th>
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<td>05</td>
<td>25</td>
<td>35</td>
<td>45</td>
</tr>
<tr>
<td>04</td>
<td>24</td>
<td>34</td>
<td>44</td>
</tr>
<tr>
<td>03</td>
<td>23</td>
<td>33</td>
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</tr>
<tr>
<td>02</td>
<td>22</td>
<td>32</td>
<td>42</td>
</tr>
</tbody>
</table>

### Bearings

<table>
<thead>
<tr>
<th>Bearing Type</th>
<th>Material</th>
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<tbody>
<tr>
<td>Ball</td>
<td>Steel</td>
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<tr>
<td>Needle</td>
<td>Steel</td>
</tr>
<tr>
<td>Taper</td>
<td>Bronze</td>
</tr>
</tbody>
</table>

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Brook Crompton 6 pole 5.5 kW machine data
Appendix 2

Rotor Terminal Voltage and Current

Modelling and Simulation Results

Simulation using the PSPICE package helps investigate and compare the phases and voltages at the rotor terminals. Several models for the simulation have been developed. The presented model shows three ideal current sources producing a six step current waveform into the rotor circuit. The rotor e.m.f. is represented by a voltage source in series with the rotor resistance and reactance. Assuming the current source has negligible ripple and the thyristor switches are ideal, then the per phase equivalent circuit can be expressed in the following PSPICE model.

The current source, which is the combination of a line side converter and a large inductance, is assumed to supply a low ripple current. The six stepped current waveform switched into the rotor is represented by I1, I2 and I3, forming the 180 degree six stepped pulses. There is a commutation capacitance C3 before D1, D2 and
across the rotor circuit. The p.f. correction capacitor C4, added across the rotor terminal also absorbs any voltage spikes. The value for C4 is chosen not to affect commutation. The true secondary e.m.f. is represented by V1, which is a pure sine wave, assuming the mains is near sinusoidal. A resistor and inductor represent the rotor parameters, which is the machine characteristics obtained by machine tests. An additional RC low pass filter (R4, C2) is also included to investigate the option of observing a filtered terminal voltage.

The simulation results are shown below. The plot shows the difference between the rotor e.m.f. and the terminal voltage under the effect of the six stepped currents, provided by the current source inverter.

The rotor e.m.f. and terminal voltage are in good agreement when the rotor is almost locked, at 240V 50Hz. However, the range of the rotor e.m.f. and the current magnitude in the circuit has to be considered, and confirm that the terminal voltages are indeed a good approximation of the true e.m.f. and easily detectable.
The terminal voltage is an observable quantity that has good phase representation of the e.m.f., provided the voltage ripple from the inverter bridge can be removed or suppressed. The terminal voltage is useful for generating control signals using zero crossing to fire the inverter bridge. Despite the ripple present, this approach is very much better than extracting rotor phase/position information using a DSP operation on the stator line current harmonics, which is an almost impossible task because the harmonic content is at a very low level (-60db). The limitation of a real time DSP approach is the inherently random wide spectrum, and very noisy environment.

**DIRECT MEASUREMENT OF ROTOR VOLTAGE AND CURRENT**

The terminal voltage was tapped directly, without any filtering or manipulation, from one phase of the secondary e.m.f., under various speeds. The phase chosen was R-B and the line current measured is the B line current. The results for sub synchronous speed, synchronous speed and super synchronous speeds are illustrated in the following diagrams.

A Voltech power analyser is also used to analyse the power in the rotor circuit. The Voltech is able to compute the true power and displacement factor by considering the ratio of W/VA, even if the waveform is distorted. It was found that the power factor was poor. This is partly due to the control circuit’s shortcoming, that current was not injected at exactly unity p.f. with respect to the rotor e.m.f.’s.
For all diagrams, 1 = Rotor terminal voltage and 2 = rotor line current. The synchronous speed is at 1000 r.p.m.
Speed = 900rpm
V link = -20V
I stator = 6.1A
I link = 1.2A
p.f. = 0.17

Speed = Synchronous
V link = 0V
I stator = 6.2A
I link = 1.2A
p.f. = 0.005

Speed = 1100rpm
V link = +20V
I stator = 6.1A
I link = 1.4A
p.f. = 0.22

Speed = 1200rpm
V link = +50V
I stator = 6.1A
I link = 1.5A
p.f. = 0.36

Speed = 1300rpm
V link = +62V
I stator = 6.2A
I link = 1.8A
p.f. = 0.53

1 = Rotor terminal voltage and 2 = rotor line current.
As shown in the diagrams, the terminal voltages are easily observable, even without filtering or manipulation. This provides good information to estimate the rotor e.m.f. with better accuracy than the encoder. As the rotor e.m.f. can be directly predicted, a more simple and robust control system can be developed.

Experiments show that the encoder control system has limitations, that of poor power factor and phase errors. Comparing the first diagram at 500 rpm to the last diagram at 1500 rpm, the currents are not injected at the same instance.

The per phase terminal voltage can be used to observe/estimate the true rotor e.m.f. Although the phase and magnitude of the terminal voltage is different from the true rotor e.m.f., the difference or error is negligible and can be corrected using electronic phase adjustment if required. This can be achieved using adaptive control.
or other schemes that constantly monitor the machine rotor resistance and inductance variations. As shown in the previous section, the error is almost negligible for the SRIM and therefore the error correction may be omitted.

A sensorless control scheme is developed based on this principle, by observing the terminal voltage to generate the synchronising signals for the inverter bridge. Developing from the existing encoder based system, new hardware required is minimal. The existing slip power system is modified to employ the sensorless scheme. This new sensorless control scheme is a form of Direct Torque Control, as the torque can be directly regulated using a current servo. This scheme is not machine parameter variation sensitive, and can be used for any slip ring machine of the same rating and meets the assumptions listed.
Appendix 3

Basic Program for Scherbius Analysis

10 REM SCHREBIUS ANALYSIS
12 REM GAS VER 1.3
22 REM JKW VER 2.1 1997 FOR MS Q basic
25 CLS
33 PI = 3.141592653#
42 READ VOLTS, FQY, POLES
52 DATA 415,50,6
62 READ ILC
72 DATA 11.3
82 READ R1, X1
92 DATA 3, 7.5
102 READ RM, XM
112 DATA 2000, 130
122 READ R2, X2
132 DATA 3, 7.5
142 ANG = 0
152 READ N
162 DATA 1.6
172 REM RATIO PRI/SEC VOLTS
182 INPUT "SELECT MOTOR(M) OR GENERATE(G)", C5$
192 IF C5$ = "M" THEN C5 = 1
202 IF C5$ = "G" THEN C5 = 0
204 PRINT "R1=", R1, "X1=", X1
205 PRINT "R2=", R2, "X2=", X2
206 PRINT "RM=", RM, "XM=", XM
207 PRINT "TURNS RATIO=", N
222 REM
232 INPUT "TORQUE CONTROL? Y OR N", C6$
242 IF C6$ = "Y" THEN C6 = 1
252 IF C6$ = "N" THEN C6 = 0
262 IF C6 = 0 THEN GOTO 302
271 REM
272 INPUT "TORQUE", TD
273 IF C5 = 0 THEN TD = -TD
282 TRQ = TD
292 IDC = 1.25 * ILC * N
302 INPUT "MACHINE SPEED (RPM) =", SPD
304 REM
314 WS = (2 * PI * FQY) * 2 / POLES
316 W = (SPD / 60) * 2 * PI
318 S = (WS - W) / WS
322 IF C5 = 1 THEN 352
324 REM ANG IN DEGREES
325 IF C5 = 0 THEN ANG = ANG + PI
332 IF S > 0 THEN ANG = ANG + PI
342 GOTO 362
352 IF S < 0 THEN ANG = ANG + PI
362 IF C6 = 1 THEN GOTO 432
372 INPUT "DC CURRENT =", IDC
382 I20 = IDC * 1.4142 / 3
392 REM I20=ROTOR WDG CURRENT /PHASE (DELTA)
402 I2 = I20 / N
412 REM
422 IF C6 = 0 THEN GOTO 442
432 IDC = IDC * TD / TRQ
442 I20 = IDC * 1.4142 / 3
452 REM I20=ROTOR WINDING CURRENT / PHASE (DELTA)
462 I2 = I20 / N
472 REM
481 PRINT TAB(1); "SPD"; TAB(5); "I1"; TAB(12); "PF1";
482 PRINT TAB(19); "STA P"; TAB(26); "ROT P"; TAB(33); "TOT P";
483 PRINT TAB(40); "SHF P"; TAB(47); "PF2"; TAB(54); "PFT";
484 PRINT TAB(60); "T"; TAB(65); "E%"; TAB(69); "IDC";
485 PRINT TAB(74); "I"
502 Z1 = COS(ANG)
512 Z2 = SIN(ANG)
522 REM
532 A = I2 * (R2 * Z1 + X2 * Z2)
542 B = I2 * (X2 * Z1 - R2 * Z2)
552 C = I2 * (R1 * Z1 + X1 * Z2)
562 D = I2 * (X1 * Z1 - R1 * Z2)
572 E = (1 + R1 / RM + X1 / XM)
582 F = (X1 / RM - R1 / XM)
592 G = A * E - B * F + C
602 H = B * E + F * A + D
612 K = (E * E + F * F)
622 L = 2 * (G * E + F * H)
632 M = (H * H + G * G - VOLTS * VOLTS)
642 ROOT = SQR(L * L - 4 * K * M)
652 V2 = (-L + ROOT) / (2 * K)
662 PR = 3 * (I2 * Z1 * (V2 + A) + I2 * B * Z2)
672 TRQ = PR / WS
682 IF C6 = 0 GOTO 712
692 IF ABS(TRQ) < .99 * ABS(TD) GOTO 432
702 IF ABS(TRQ) > 1.01 * ABS(TD) GOTO 432
712 PO = TRQ * W
722 PR2 = PR - 3 * I2 * I2 * R2 - PO
732 IF C5 = 1 GOTO 762
742 IF PR2 < 0 THEN PRINT "CANNOT GENERATE AT "; SPD
743 IF PR2 < 0 THEN GOTO 1032
752 GOTO 772
762 IF PR2 < 0 THEN PRINT 'CANNOT MOTOR AT"; SPD
763 IF PR2 < 0 THEN GOTO 1032
772 P = I2 * Z1 + (V2 + A) / RM + (B / XM)
782 Q = B / RM - I2 * Z2 - (V2 + A) / XM
792 I1 = SQR(P * P + Q * Q)
802 TEMP1 = SQR(I1 * I1 - P * P)
803 ANG1 = ATN(TEMP1 / P)
812 TEMP2 = SQR((VOLTS * VOLTS) - (V2 * E + G) * (V2 * E + G))
813 ANG2 = ATN(TEMP2 / (V2 * E + G))
822 PF1 = COS(ANG2 - ANG1)
832 P1 = 3 * VOLTS * I1 * PF1
842 TRQ = INT(TRQ * 100) / 100
852 PO = INT(PO * 10) / 10
862 PR2 = INT(PR2 * 10) / 10
872 I1 = INT(1.732 * I1 * 100) / 100
882 P1 = INT(P1 * 10) / 10
892 PT = P1 - PR2
902 IF (PO < 0) THEN EFFY = INT(PT * 100 / PO)
912 IF (PO > 0) THEN EFFY = INT(PO * 100 / PT)
922 PF1 = INT(PF1 * 100) / 100
932 PF2 = PR2 / (3 * VOLTS * I20)
942 PF2 = INT(PF2 * 100) / 100
952 ILP = I1 * PF1 - I20 * 1.732 * PF2
962 SA1 = SQR(1! - (PF1 * PF1))
972 SA2 = SQR(1! - (PF2 * PF2))
982 ILQ = I1 * SA1 * I20 + 1.732 * SA2
992 ILT = SQR((ILP * ILP) + (ILQ * ILQ))
1002 PFT = INT(ILP * 100 / ILT) / 100
1012 ILT = INT(ILT * 10) / 10
1022 PRINT SPD; TAB(6); I1; TAB(11); PF1;
1032 PRINT TAB(17); P1; TAB(26); PR2; TAB(33); PT;
1042 PRINT TAB(40); PO; TAB(47); PF2; TAB(54); PFT;
1052 PRINT TAB(60); TRQ; TAB(65); EFFY; TAB(69);
1062 PRINT INT(IDC * 10) / 10; TAB(74); ILT
1032 ANG = 0
1034 IF C6 = 1 THEN GOTO 271
Appendix 4

SRIM Simulated Results

The SRIM characteristics were simulated under the conditions of varying the firing angle of the CSI with respect to the rotor e.m.f., from lagging to leading. Results for motoring and generating are presented, showing the shaft power, power factor and machine efficiency.
MOTORING - SHAFT POWER

MOTORING POWER FACTOR
MOTORING EFFICIENCY

GENERATING SHAFT POWER
GENERATING POWER FACTOR

GENERATING EFFICIENCY
APPENDIX 5

Verification of Adaptive Filtering/Signal Generating System

The dynamic responses of the adaptive filter and the synchronising signal generator are shown from sub to super-synchronous speeds.

![Graph showing synchronising signal versus rotor terminal voltage at different speeds](image)

**Fig. 5.1 Synchronising Signal versus Rotor Terminal Voltage at \( \omega = 200 \text{rpm} \)**

**Fig. 5.2 Synchronising Signal versus Rotor Terminal Voltage at \( \omega = 300 \text{rpm} \)**
Fig. 5.3 Synchronising Signal versus Rotor Terminal Voltage at $\omega = 400$rpm

Fig. 5.4 Synchronising Signal versus Rotor Terminal Voltage at $\omega = 500$rpm
Fig. 5.6 Synchronising Signal versus Rotor Terminal Voltage at $\omega = 600\text{rpm}$

Fig. 5.7 Synchronising Signal versus Rotor Terminal Voltage at $\omega = 700\text{rpm}$
Fig. 5.8 Synchronising Signal versus Rotor Terminal Voltage at \( \omega = 800 \text{rpm} \)

Fig. 5.9 Synchronising Signal versus Rotor Terminal Voltage at \( \omega = 900 \text{rpm} \)
Fig. 5.10 Synchronising Signal versus Rotor Terminal Voltage at $\omega = 1000$ rpm

Fig. 5.11 Synchronising Signal versus Rotor Terminal Voltage at $\omega = 1050$ rpm
Fig. 5.12 Synchronising Signal versus Rotor Terminal Voltage at $\omega = 1200\text{rpm}$

Fig. 5.13 Synchronising Signal versus Rotor Terminal Voltage at $\omega = 1300\text{rpm}$
Fig. 5.14 Synchronising Signal versus Rotor Terminal Voltage at \( \omega = 1500 \text{rpm} \)

Fig. 5.15 Synchronising Signal versus Rotor Terminal Voltage at \( \omega = 1600 \text{rpm} \)
Fig. 5. 16 Dynamic Response of Synchronising Signal Generator

Note: 1 = Diminishing rotor terminal voltage
      2 = Control synchronising signal
Appendix 6

Closed Loop Operations (Current Control) of the Novel Sensorless System vs Encoder

The system was operated, from sub- to super-synchronous speeds, using only current control to compare the operations between the sensorless system and the encoder system. The dynamic responses of the system were determined by the link current servo.

The sensorless system appeared to track the rotor voltage with a slight phase advance than the encoder system. It is noted that the accuracy of the encoder system may be affected by its alignment to the machine. The Sensorless system examines the real time rotor terminal voltage and is free from alignment problems. The following plots show the thyristor control signals under closed loop operations. The top traces were taken from the sensorless system, and the lower traces were taken from the encoder system. The CSI was operating from the signals generated by the sensorless system.

![Graph of Sensorless Control Signals vs Encoder Control Signals at ω= 500rpm](image)

Fig. 6.1 Sensorless Control Signals vs Encoder Control Signals at ω= 500rpm
Fig. 6.2 Sensorless Control Signals vs Encoder Control Signals at ω = 600rpm

Fig. 6.3 Sensorless Control Signals vs Encoder Control Signals at ω = 700rpm

Note: 1 = Sensorless (TTL Output) 2 = Encoder (CMOS Output)
Fig. 6.4 Sensorless Control Signals vs Encoder Control Signals at $\omega = 800$ rpm

Fig. 6.5 Sensorless Control Signals vs Encoder Control Signals at $\omega = 900$ rpm

Note: 1 = Sensorless (TTL Output) 2 = Encoder (CMOS Output)
Fig. 6.6 Sensorless Control Signals vs Encoder Control Signals at $\omega = 1010\text{rpm}$

Note: 1 = Sensorless (TTL Output) 2 = Encoder (CMOS Output)

Fig. 6.7 Sensorless Control Signals vs Encoder Control Signals at $\omega = 1100\text{rpm}$
Fig. 6.8 Sensorless Control Signals vs Encoder Control Signals at ω = 1200rpm

Fig. 6.9 Sensorless Control Signals vs Encoder Control Signals at ω = 1300rpm

Note: 1 = Sensorless (TTL Output) 2 = Encoder (CMOS Output)
Fig. 6.10 Sensorless Control Signals vs Encoder Control Signals at \( \omega = 1400 \text{rpm} \)

Fig. 6.11 Sensorless Control Signals vs Encoder Control Signals at \( \omega = 1500 \text{rpm} \)
Fig. 6.12 Sensorless Control Signals vs Encoder Control Signals at \( \omega = 1600 \text{rpm} \)

Fig. 6.13 Sensorless Control Signals vs Encoder Control Signals at \( \omega = 1700 \text{rpm} \)

Note: 1 = Sensorless (TTL Output) 2 = Encoder (CMOS Output)
List of Papers Published by the Author

REVIEW OF THE LATEST SENSORLESS TECHNIQUES FOR INDUCTION MOTOR CONTROL


SCHERBIUS ANALYSIS USING A CURRENT SOURCE INVERTER AND SLIP RING INDUCTION MACHINE


A DSP CONTROLLED BOOST RECTIFIER FOR SYNCHRONOUS GENERATOR

J.K.J. Wong and J.A.M. Bleijs, UPEC 1997 UMIST.

NOVEL SENSORLESS CONTROL OF A SUPER SYNCHRONOUS SLIP RING INDUCTION MACHINE
(Paper submission)