



THE PRACTICAL APPLICATION
OF
OPTIMAL CONTROL TECHNIQUES
TO
SYNCHRONOUS GENERATOR EXCITATION

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ABSTRACT

The small signal behaviour of synchronous generators has been a topic of research for many years. With the current trend in machine design towards larger units with higher per unit reactances it is continuing to attract interest. As the economy of power generation is improved with these units, there are increasing demands on regulators to offset the resulting reduction of stability margins, and so permit operation with satisfactory leading power factors. Previous studies have been largely aimed at determining the maximum theoretical extension of the stability limit into the leading reactive power region. Due to the non linear nature of the alternator, regulator gains which allow operation near this limit produce unacceptable system performance under other loading conditions. Thus, as an alternative, procedures have been developed for choosing regulator gains, which, although reducing this limit, provide acceptable regulation over as wide an operating region as possible.

The non linear single generator tie line (SGTL) system may be described mathematically by a small signal linear model, the parameters of which are dependent on the present operating conditions of the system. When applying optimal control theory to determine the excitation for this model it is seen that theoretical stability limits cease to exist, provided the gains of the controller are made to vary with the operating point. In addition, optimal control theory provides a systematic approach to the evaluation of suitable regulator gains. To determine whether these theoretical advantages exist in practice, an experimental SGTL system was established in the laboratory.

A D.C. motor, with a speed torque characteristic similar to that of a steam turbine, is used as the prime mover for a 5 KVA alternator connected to the infinite bus through adjustable tie line reactors. Transducers were developed through which a small digital computer, programmed as the regulator, monitors the states of the system. The computed control signal is applied to an electronic amplifier exciting the alternator field.

For a given steady state operating point in the complex power plane, the regulator gains satisfy a simple quadratic cost function. To allow for the non linear nature of the SCTL system, these gains are precalculated "off-line" at a matrix of grid-points across the PQ plane, and then stored in the controlling computer. The gains used by the regulator at any particular operating point are those corresponding to the nearest grid-point.

Stable operation of the closed loop experimental system is observed under leading power factor conditions in an extended region of the PQ plane. Comparisons are made between the small disturbance responses and those predicted by the linear mathematical model. Consideration is given to system performance under operating conditions where the mathematical model is obviously invalid so allowing an assessment to be made of any practical deficiencies existing in the theoretically ideal regulator.

STATEMENT

This thesis contains no material which has been accepted for the award of any other degree or diploma in any University, and, to the best of the author's knowledge and belief, contains no material previously published or written by another person, except when due reference is made in the text.

“ John P. Bartlett

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NOTATION

A, B	=	state equation matrices
C	=	output matrix multiplier
C_e	=	electrical torque
C_m	=	prime mover torque
D	=	damping coefficient
E	=	input driving force
F_i	=	optimum regulator
f_s	=	synchronous frequency
G	=	weighting matrix for J
g	=	weighting factor for J
H	=	inertia constant
H'	=	Hamiltonian
i_d, i_q	=	direct and quadrature axis currents
J	=	system cost function
J_i	=	inertia constant in MKS units
K_i	=	open loop system parameters
L	=	Riccati matrix
M	=	$H/\pi f_s$ = inertia constant
N	=	weighting matrix for J
n	=	reference input
P	=	real power output of alternator
P_b	=	base power
p	=	differential operator d/d_t
P_m	=	prime mover power
Q	=	reactive power measured at infinite bus
Q_m	=	reactive power measured at machine terminals
r_a	=	armature resistance

r_e	=	tie line resistance
r	=	$r_a + r_e$
S	=	weighting matrix for J
T	=	$\frac{x_e + x'_d}{x_e + x_d} T'd_o = \text{field time constant}$
$T'd_o$	=	field open circuit time constant
u	=	control vector
v_b	=	infinite bus voltage
v_d, v_q	=	direct and quadrature axis voltages
v_f	=	applied field voltage
v_{ref}	=	reference input voltage
v_t	=	machine terminal voltage
w	=	input noise vector
x	=	state vector
x_d, x_q	=	direct and quadrature axis synchronous reactances
x'_d	=	direct axis transient reactance
x_e	=	tie line reactance
x_p	=	Potier reactance
x_e	=	leakage reactance $\rightarrow x_l$?
y	=	input noise multiplier matrix
z	=	output vector
δ	=	power angle
ω	=	machine speed
ω_s	=	synchronous speed
ψ_d, ψ_q	=	direct and quadrature axis flux linkages
λ	=	Lagrange multiplier

Prefix

Δ implies small perturbation

Subscript

o implies steady state value

All dimensioned quantities are in per unit except where specified otherwise in the text.

CHAPTER 1

INTRODUCTION



1.1 BACKGROUND

For many years researchers have been concerned with the problem of improving dynamic behaviour of the synchronous machine coupled into the power grid. With the current trend in economic generator design towards larger units with higher per unit reactances and lower per unit resistances, the problem is becoming increasingly important as it becomes necessary to offset the inherent reduction in stability margins resulting from these design trends.

Generation of leading reactive power is limited by the ability of regulators to maintain stability and produce otherwise acceptable system performance under these loading conditions. Various forms of closed loop control have been analysed in an attempt to improve this performance.^{1,2,3,4} Feedback signals from system variables such as terminal voltage, machine speed, power angle, flux linkages, line current and power, as well as the derivatives of some of these, have been used in a variety of combinations to achieve improved excitation control. In general the choice is based on physical insight into the system operation.

As varied as the combinations of feedback signals are the techniques used to assess the effects of the desired regulator on the stability of the system. These include the D partition method, the Nyquist and Routh criteria and root locus techniques. The system performance under stable conditions is evaluated using analogue or digital computer models and occasionally scaled laboratory model-power-systems.

More recently modern optimization techniques have been applied to this problem.^{5,6,7} The main difference between this and other approaches is that a mathematical definition of the required closed loop performance is the starting point of the design procedure. This provides the designer with a more methodical approach, as once the system and its goals are mathematically

defined, stabilizing feedback gains can generally be derived which force the system to satisfy a predefined performance criterion. Investigations along this line to date have been largely confined to computer simulated systems operating with a limited range of loading conditions.

The problems associated with the control of synchronous power systems stem from the inherent non linear nature of the attenuator. Due to the variation of the machine parameters with operating conditions, a compromise must be made, when fixed gain regulators are used, between the extension of the stability limit and the system performance obtained in the unity power factor region. Although previous studies have not made note of the fact, by using optimization theory to derive the gains of an adaptive regulator, it is theoretically possible to operate a single generator tie line system (SGTL) well beyond all previously defined stability limits in the leading reactive power region. The range of operation afforded by such a regulator is limited by the accuracy of the mathematical model of the system for which the regulator gains are derived, and whether the required regulator is physically realizable. This thesis describes a practical investigation into the feasibility of such a regulator.

1.2 THE SINGLE MACHINE APPROACH

The rigorous mathematical analysis of an operational power grid is in general prohibitively complex due to the interaction of its contributing parts. In addition, data necessary to allow even a separate analysis of each part is either not available or is difficult to obtain.

To simplify the analytical procedures, major approximations are made regarding the individual machines when it is felt that the problem lies mainly in the system configuration, and when the investigation is centred on an individual machine the remainder of the system is represented simply by an infinite bus, the single machine being connected to this bus by a tie line

or series reactance.

The accuracy of the SGTL representation is dependent on the relationship between the machine under study and the remainder of the system. For a remote plant connected to a tightly coupled power grid, the representation is more accurate than for a machine which dominates the system by virtue of its capacity. However this approach is widely used in the study of regulator performances as it does allow a mathematical analysis to be made on a particular machine.

1.3 OPTIMIZATION THEORY

The existence of a stability limit for the regulated SGTL system operating with a leading power factor is a phenomena which has long been recognized. Park's two axis theory for the synchronous machine^{8,9} provided a basis for the investigation of this phenomena. Much theoretical analysis and some experimentation has been performed to determine the locus of this limit in the complex power plane (PQ plane) and to extend it as far as possible into the leading reactive power region. In general this is achieved by representing the non linear system by a small signal linearized mathematical model and adjusting the gains of the feedback loops until no further extension is possible, as indicated by the failure of the closed loop model to satisfy one of the stability criteria mentioned previously.

In recent years the theory of systems control has been revolutionized by the application of the calculus of variations to control studies, so generating the modern optimal control theory. Using optimal control techniques, for any system satisfying the controllability criterion¹⁰, stabilizing feedback gains may be computed which result in the optimization of some predefined system performance criterion. This criterion is a mathematical description of the required closed loop behaviour of the system. The controller gains so computed are then referred to as the optimum gains. Thus the best set of controller gains

is obtained by direct computation, rather than from the essentially trial and error methods previously employed.

For any linear system with a known mathematical model the optimum feedback gains may be computed and then implemented on the system. However there are many difficulties associated with the application of optimal control theory to complex non linear systems (such as the excitation control of the SGTL system). While approximations to the optimum gains can theoretically be determined, the computation involved may be prohibitive, particularly when on line control is envisaged.^{10,11} As a result of this, a gap exists at present between the theory of optimal control and its application to practical systems. By accepting approximations in the use of the theory however, useful guides to the choice of controller gains can be found. To this end, in this study, the non linear SGTL system is represented by a small signal linear model (as used in the classical approach) and linear optimization theory applied.

It is shown in this thesis that, with a judicious choice of the variables describing the state of the system, the controllability criterion is satisfied by the small signal mathematical model over an extended region of the PQ plane. Therefore stabilizing feedback gains can be computed throughout this region.

The region of the PQ plane investigated is limited by the ratings of the equipment used in the practical implementation and not by theoretical stability considerations. These limits lie well beyond the open loop stability limit of the system studied, and beyond the closed loop limit existing with a conventional voltage regulator whose gains are continuously adjusted for the maximum extension of the limit.

Thus by causing the feedback gains, computed using optimization techniques, to vary with the system operating point (a necessity arising from the non linear nature of the system), the concept of a theoretical stability limit is replaced by the consideration of the practical limitations mentioned

previously. These were: is the small signal linear model a sufficiently accurate representation of the non linear system, bearing in mind that the optimum controller assures insensitivity to plant parameter variations^{7,12}; what can be done to overcome significant errors which may exist in the model; are the required gains and accuracy in state estimation attainable; and how can such a system be implemented? The answers to these questions also determine whether the regulator forces the system to achieve its desired goal when operating in a more stable region of the PQ plane.

The investigation of these queries forms the basis of the feasibility study.

1.4 THE MODEL SYSTEM

A SGTL system was established in the laboratory using a 5 KVA 3 phase alternator driven by a D.C. motor having a speed torque curve similar to that of a steam turbine. The alternator is connected to the infinite bus through adjustable line reactors.

Analogue transducers were developed to monitor terminal voltage and the real and reactive power delivered to the infinite bus. A general purpose interface was produced to permit the transfer of information from these transducers to a small digital computer used as the controller. Machine speed and power angle are measured using digital techniques, and the outputs from these are passed directly to the controlling computer.

The evaluation of the optimal regulator gains involves mathematical procedures of a complexity which prohibits the use of a small digital computer operating on line. Instead the gains are computed off line using a large computer. The system is regarded as a point to point linear one. The parameters of the system are calculated at a matrix of steady state operating points, or grid points, resulting from the intersection of lines of constant real and reactive power dividing the PQ plane. The parameters are then assumed to be constant in the immediate vicinity of each of these points. The gains

which are optimal for operation at each of these grid points are computed off line and then stored in the controlling computer. As the system operating point changes, the controller automatically retrieves from its store the gains which are most appropriate to that operating point. The digital computer is ideally suited to this role as a variable gain controller.

Deviations of terminal voltage, machine speed and electrical torque from their initial steady state values are used as the three states necessary to describe the third order system. These are estimated by the controller from the monitored information, and, using the selected gains, the control signal computed and transmitted via a D/A converter to an electronic amplifier exciting the field of the alternator.

The behaviour of this system is investigated under operating conditions in the leading power factor region. Comparisons are made between the predicted and actual small signal responses. Consideration is given to the effect of large disturbances on the system and other phenomena which invalidate the small signal linear mathematical model. The performance under all conditions is related qualitatively to the choice of the cost function describing the physical goals of the system with the aim of allowing more suitable terms for this function to be chosen.

CHAPTER 2

THE OPTIMAL CONTROL PROBLEM

2.1 INTRODUCTION

The fundamental problem of optimization theory is the search for the control which attains a desired objective while minimizing a defined system performance index or cost function.¹⁰

This problem may be considered in five steps.

- (1) From a knowledge of the plant and its desired behaviour, a set of goals or objectives is defined in physical terms.
- (2) Consideration is given to any physical limitations or constraints on the system whether internal or external.
- (3) The physical plant is characterized by what is considered to be an adequate mathematical model which relates the system states to the input.
- (4) The desired behaviour and system constraints are described mathematically, in a form compatible with the model of the system, to generate a cost function.
- (5) With a knowledge of the current state of the system, the plant inputs are determined which result in the minimization (optimization) of the cost function.

It may be found that the optimal controller derived initially by this process is not satisfactory for some reason. For example the mathematical model may prove to be insufficiently accurate, or alternatively while the inputs to the plant are admissible, the optimal controller required to produce these inputs may not be physically realizable. A re-evaluation of the problem is then required. In general the solution of the idealized problem can be regarded as a guide to developing a practical control system.

2.2 THE POTENTIAL OF OPTIMAL CONTROL

Optimal control theory is such that for any system satisfying the controllability criterion, stabilizing feedback gains may be found which force the system to satisfy a predefined performance index.¹⁰

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This statement, if taken without consideration of the difficulties associated with determining any optimal controller for a practical system, implies that optimization theory can be used to provide an ideal solution to almost any control problem. In fact if optimization techniques are utilized only in accordance with an exact application of the theory, the role that optimal control can play in practice is small.¹³ This arises from several considerations.

At present, analytical solutions exist only for the linear time invariant system and few systems fall into this category; approximation techniques with numerical solutions may be employed in some other instances. The choice of a suitable cost function may not be apparent; it is generally not possible to analytically relate the elements of the cost function to the behaviour of the practical system. This means that a suitable cost function may have to be chosen by trial and error. (Tyler and Tuteur have developed a method for expressing the characteristic polynomial of a linear time invariant optimized system as an explicit function of the elements of a quadratic cost function¹⁴, although the location of the zeros of the transfer functions for each of the system states is not considered.) Once the cost function has been defined, the mathematical procedures for determining the optimal controller may appear prohibitive. This is particularly so if a real time solution is required.¹¹

If the concepts of optimization theory are not interpreted literally, then they can be successfully applied to a wide range of practical systems. Many non linear systems can be mathematically represented by a small signal linear model, and if a prior knowledge of the inputs to the controller system is assumed, (perhaps from a statistical analysis) then the problem is reduced to that of the well defined optimal regulator. The elements of the cost function may be varied until the desired predicted performance is achieved, while any single cost creterion assures insensitivity of

the controlled system to the variation of the actual system parameters from those of the model.¹²

Thus in applying optimization theory to a practical system one cannot hope to obtain an ideal solution, but, by careful consideration of the possible approaches, an acceptable approximation may be found.

The single generator tie line system studied is a non linear system. Difficulties became apparent when attempting to establish an approximate solution by employing optimization techniques developed for direct application to non linear systems. The approach adopted here to determine a suitable controller is to derive a small signal linear mathematical model and optimize performance for deviation of the perturbed terminal voltage from its desired value (see chapter 3). Thus the control problem becomes that of the optimal linear tracking regulator, which is presented in a general form in the following section.

2.3 THE OPTIMAL TRACKING REGULATOR

Consider a linear time varying dynamic system described by the state equation.

$$\dot{x}(t) = A(t) x(t) + B(t) u(t) \quad (2.1)$$

where $x(t)$ is the state vector with elements $x(t) \dots \dots \dots x_n(t)$, $u(t)$ is the control force vector, which forms the input vector to the plant, while $A(t)$ and $B(t)$ are the matrices relating these vectors. The plant output is a vector related to the states by a matrix multiplier;

$$z(t) = C(t) x(t) \quad (2.2)$$

Assume that the control force is not constrained. (If u is constrained a quadratic cost function, with the associated advantages discussed later, cannot be used¹⁰). The goal of the system is now considered in physical terms.

The tracking regulator problem is to find a control which causes the output $z(t)$ to follow a desired output state $n(t)$. Since the control

force $u(t)$ is not constrained, it is possible that it may become exceedingly large. From the point of view of implementation large control forces are not desirable as input limits may be exceeded on the system, or expensive, high gain equipment may be required in the control loop. A quadratic cost function is used in translating these requirements into mathematical terms. This form of cost function has three advantages:

- (1) an analytical solution of the problem is possible;
- (2) it produces a linear closed loop control;
- (3) if the plant satisfies the controllability criterion (see section 2.4) a stable system is always produced.¹³

The quadratic cost function is expressed as;

$$J = \frac{1}{2} // n(t_f) - z(t_f) // \frac{2}{S} + \frac{1}{2} \int_{t_0}^{t_f} // n(t) - z(t) // \frac{2}{N} + // u(t) // \frac{2}{G} dt \quad (2.3)$$

(If a is an m vector and A is an $m \times m$ matrix then $// a // \frac{2}{A}$ is the scalar product $a^T A a$). The matrices S , N and G may be assumed symmetrical with no loss in generality.

Considering briefly the physical significance of each term, the minimization of J implies the following. Large deviations of the output from its required value are penalized by the quadratic nature of $// n(t) - z(t) // \frac{2}{N}$. Excessive control force is penalized by $// u(t) // \frac{2}{G}$, and $\frac{1}{2} // n(t_f) - z(t_f) // \frac{2}{S}$ guarantees that the end point error is small. S may be put equal to zero if the end point error is not particularly important, as the rest of the cost function ensures excessive error will not exist. The matrices S , N and G are weighting factors, allowing emphasis to be placed on the individual elements of the cost function as required by the designer. By computing the second variation of J , it can be shown that a sufficient condition for a minimum to exist is for S , N and G to be at least positive semidefinite.¹⁰ In addition equation (2.6) shows that G must have an inverse so that G must be positive definite.

A solution for u which minimizes J is obtained via Pontryagin's maximum principle.¹⁵ The Hamiltonian canonic equation is obtained as

$$H'(x, u, \lambda, t) = \frac{1}{2} n(t) - C(t) x(t) / \frac{2}{N} + \frac{1}{2} u(t) / \frac{2}{G} + \lambda^T(t) (A(t) x(t) + B(t) u(t)) \quad (2.4)$$

H' is minimized with respect to u , i.e.,

$$\frac{\partial H'}{\partial u} = 0 \quad (2.5)$$

giving

$$u(t) = - G^{-1}(t) B^T(t) \lambda(t) \quad (2.6)$$

and

$$\frac{\partial H'}{\partial x} = - \dot{\lambda} = C^T(t) N \{C(t) x(t) - n(t)\} + A^T(t) \lambda(t) \quad (2.7)$$

plus the end point condition,

$$\lambda(t_f) = C^T(t_f) S \{C(t_f) x(t_f) - n(t_f)\} \quad (2.8)$$

To obtain a closed loop control assume

$$\lambda(t) = L(t) x(t) - E(t) \quad (2.9)$$

This is substituted into the canonic equations to determine a solution. The requirement on $L(t)$ is found to be

$$\begin{aligned} \dot{L} = & -L(t) A(t) - A^T(t) L(t) - C^T(t) N C(t) \\ & + L(t) B(t) G^{-1} B^T(t) L(t) \end{aligned} \quad (2.10)$$

$$L(t_f) = C^T(t_f) S C(t_f) \quad (2.11)$$

and for the driving force $E(t)$;

$$\dot{E} = - \{A(t) - B(t) G^{-1} B^T(t) L(t)\}^T E - C^T(t) N n(t) \quad (2.12)$$

$$E(t_f) = C^T(t_f) S n(t_f) \quad (2.13)$$

Equation (2.10) is referred to as the Riccati Equation.

The linear optimum control law is obtained from (2.6) and (2.9)

as;

$$u(t) = - G^{-1} B^T(t) \{L(t) x(t) - E(t)\} \quad (2.14)$$

This control force is unique for a given cost function.¹⁰

Thus an analytical solution is obtained for the optimal control force, although a knowledge of $n(t)$ for all time t_e (t_0, t_f) is required for

2.3.1 APPROXIMATIONS FOR THE TIME INVARIANT SYSTEM

In considering the application of the above theory to the small signal linearized SCTL system it is noted that the matrices A , B and C in the system equations are independent of time (see Chapter 3). That is, equations (2.1) and (2.2) can be rewritten as

$$\dot{x}(t) = Ax(t) + Bu(t) \quad (2.15)$$

and

$$z(t) = Cx(t) \quad (2.16)$$

If deviation of the output from its desired end point is not specifically penalized (i.e. $S = 0$) and the desired output n is assumed to be a constant then the cost function becomes

$$J = \int_{t_0}^{t_f} \left(\| n - z(t) \|^2_N + \| u(t) \|^2_G \right) dt \quad (2.17)$$

As $t_f \rightarrow \infty$, the matrix $L(t)$ obtained as a solution to the Riccati equation (2.10) tends to the positive definite matrix $\hat{L}$.¹⁵ However if t_f is allowed to tend to infinity, then, in all but a few special cases,^{16,17,20} a minimum of J cannot exist as zero error cannot be maintained by zero control force at the infinite terminal time, i.e. $n - z(t_f)$ and $u(t_f)$ cannot both be zero, and thus J must tend to infinity as t_f tends to infinity. Consequently t_f is assumed to be large but finite, and $L(t)$ is approximated by $\hat{L}$, i.e. $\dot{L}$ becomes zero and $\hat{L}$ may be obtained as a solution of the constant Riccati equation;

$$0 = -LA - A^T L - C^T N C + L B G^{-1} B^T L \quad (2.18)$$

Also with the output reference n constant, E in equation (2.12) is zero and (2.12) may be rewritten as

$$0 = -\{A - B G^{-1} B^T L\}^T E - C^T N n \quad (2.19)$$

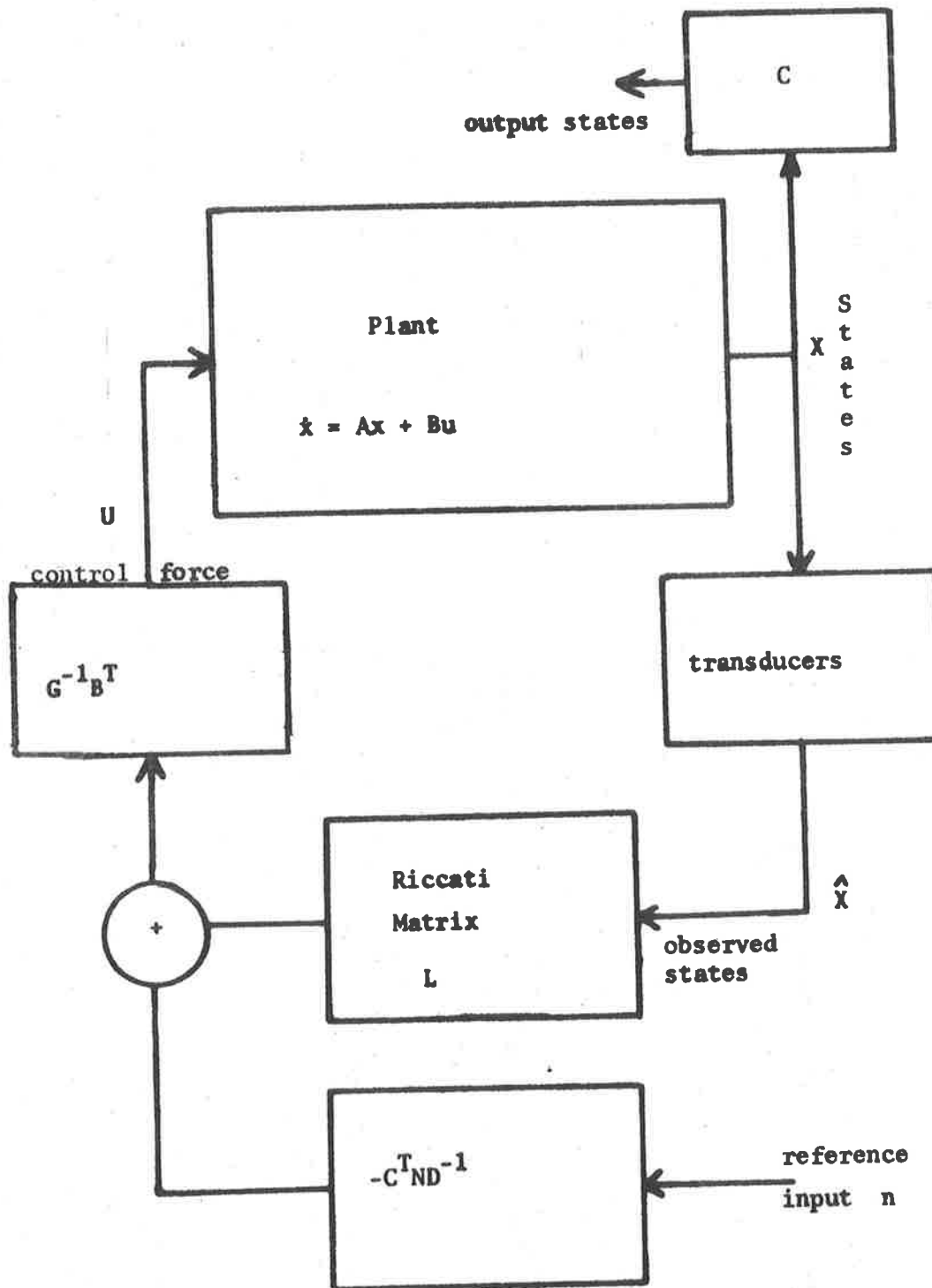


Figure 2.1 The structure of the 'approximate' optimal tracking regulator for large t_f .

The driving force E can then be expressed as a function of n ;

$$E = -C^T N \{ (A^T - BG^{-1}B^T L)^T \}^{-1} n \quad (2.20)$$

and from equation (2.14) the control force becomes

$$u = -G^{-1}B^T (Lx - E) \quad (2.21)$$

Thus the control force is expressed as a linear weighted sum of the states x of the system and from (2.20), the required output n . Figure 2.1 shows the structure of the optimal controller for the tracking regulator.

2.4 THE CONTROLLABILITY AND OBSERVABILITY CRITERIA

For the linear time invariant system, when t_f is very large, the system must satisfy both the controllability and observability criteria in order than an optimal controller may exist.¹⁰

A state x is defined to be observable if, given any control u , there is a time $t_1 > t_0$ such that a knowledge of u and the output z during this time interval is sufficient to determine x . For the linear time invariant system it is possible to obtain an explicit criterion for observability¹⁶.

For the system described by

$$\dot{x}(t) = Ax(t) + Bu(t) \quad (2.22)$$

$$z(t) = Cx(t) \quad (2.23)$$

where A is an $n \times n$ matrix and C is an $m \times n$ matrix, this criterion is that the partitioned matrix Γ defined by

$$\Gamma = (C^T \mid A^T C^T \mid A^{T2} C^T \mid \dots \mid A^{T(n-1)} C^T) \quad (2.24)$$

is of rank n , i.e. Γ has n linearly independent column vectors.

A system is defined to be state controllable if any initial state vector $x(t_0)$ can be transferred to any final state vector $x(t_f)$, where t_0 and t_f are fixed, by a control force $u(t)$. Using a procedure similar to that for observability, the controllability criterion for the linear time invariant

system becomes: the partitioned matrix Φ defined by

$$\Phi = (B \mid AB \mid A^2B \mid \dots \mid A^{n-1}B) \quad (2.25)$$

must be of rank n .

Choate and Sage¹⁸ have shown that a system which is not controllable must be reducible in order.

2.5 SOLUTION OF THE RICCATI EQUATION

The Riccati equation for the linear time invariant system (equation 2.18) is a non linear equation in the constant matrix L :

$$0 = -LA - A^T L - C^T N C + L B G^{-1} B^T L \quad (2.26)$$

This equation is solved for L using a method developed by Potter¹⁹. The procedure is outlined here for completeness.

The $2n \times 2n$ partitioned matrix M is formed from the coefficients of L in equation (2.26) to give

$$M = \begin{bmatrix} A^T & C^T N C \\ B G^{-1} B^T & -A \end{bmatrix} \quad (2.27)$$

If α_i is an eigenvector of M then α_i may be partitioned into upper and lower halves as

$$\alpha_i = \begin{bmatrix} b_i \\ c_i \end{bmatrix} \quad (2.28)$$

where α_i is a $2n$ column vector and b_i and c_i are n dimensional column vectors.

The solution of (2.26) takes the form

$$L = (b_i, \dots, b_n) (c_i, \dots, c_n)^{-1} \quad (2.29)$$

where the column vectors b_i and c_i are the upper and lower halves of an eigen vector α_i of M , and correspond to eigen values with positive real parts.

Via this procedure the constant Riccati matrix L can be evaluated using the library routines of a large computer installation.

2.6 OPTIMIZATION AGAINST DISTURBANCES

In the analysis of the optimal control system to date, noise or disturbance inputs have been ignored. Disturbances are optimized against by representing them with the vector $w(t)$ and rewriting the state equation (2.15) as

$$\dot{x}(t) = Ax(t) + Bu(t) + Yw(t) \quad (2.30)$$

The output equation (2.16),

$$z(t) = Cx(t) \quad (2.31)$$

is unchanged (assuming that noise is not introduced at this point).

Minimization of the same cost function as before (equation (2.17)),

$$J = \frac{1}{2} \int_{t_0}^{t_f} \left(\|n - z\|_N^2 + \|u\|_G^2 \right) dt \quad (2.32)$$

yields an unchanged Riccati equation (assuming t_f is very large);

$$0 = -LA - A^T L - C^T N C - L B G^{-1} B^T L \quad (2.33)$$

However the equation (2.19), which defines the forcing function E includes a component due to the disturbances:

$$E = - (A - B G^{-1} B^T L)^T E + P Y w(t) - C^T N n \quad (2.34)$$

The control force is still defined as

$$u = -G^{-1} B^T (Lx - E) \quad (2.35)$$

The problem now arises to solve (2.34) for E , and to do this demands a knowledge of $w(t)$ for all time $t \in (t_0, t_f)$ ¹⁶. This is often not known. If an estimate of $w(t)$ is made then the optimal system is as good as the estimate. If $w(t)$ is determined stochastically then the control is optimal only when $w(t)$ is at its average value. Consider now the implications of the $w(t)$ term.

Substituting the equation for u (2.35) into the state equation (2.30) yields

$$\dot{x}(t) = (A - B G^{-1} B^T L) x(t) + B G^{-1} B^T E(t) + Yw(t) \quad (2.36)$$

The eigen values of $-(A - BG^{-1}B^TL)$ from (2.34) are the negative of the eigen values of $(A - BG^{-1}B^TL)$ from (2.36). Therefore the dynamic behaviour of the system is independent of $w(t)$ and the desired output $n(t)$ ¹⁶.

Alternatively, the contribution of the states $(-G^{-1}B^TLx)$ from (2.35) to the control force is independent of E (i.e. $w(t)$ and $n(t)$) as equation (2.33) for L is independent of E . Therefore the same stabilizing feedback gains are derived whether $w(t)$ or n are zero or not. Therefore a system with its input disturbances assumed zero can be stabilized by the feedback loop of the system states (the characteristic polynomial of the transfer function is unchanged), although the control is no longer optimal in the sense that the cost function (2.32) is not minimized.

When the desired output is zero, the optimal tracking regulator reduces identically to the optimal output regulator. Therefore by solving the tracking regulator problem the output regulator is automatically solved. This means that if, at time t_0 , the controlled system is momentarily disturbed from a required output of zero, the states will be returned to zero by an optimal control force.

2.7 SUMMARY

While optimal control theory is an extremely powerful design tool, it must be recognized that in the majority of cases an intimate knowledge of the physical plant is required if its application is to be successful. Generally approximations are made in deriving the mathematical model, and the control is optimal only if these approximations are valid. The cost function is formulated after detailed examination of the physical aims of the controlled system, and the solution of the mathematically defined problem may itself be an approximation.

While these approximations appear to limit the usefulness of optimal control theory, it must be remembered that mathematical models and physical insight form the basis of many design procedures. However once the problem

has been translated into mathematical terms, optimization theory will provide the ideal solution to the idealized model (assuming a solution does exist).

A laboratory model may prove useful at this point to establish whether the controller, determined from the mathematical description of the plant and its requirements, will provide satisfactory when applied to the actual plant. Re-evaluation of the mathematics can then be made on the basis of the laboratory models' performance, without having to go to the actual plant.

CHAPTER 3

OPTIMIZATION OF THE SINGLE GENERATOR

TIE LINE SYSTEM

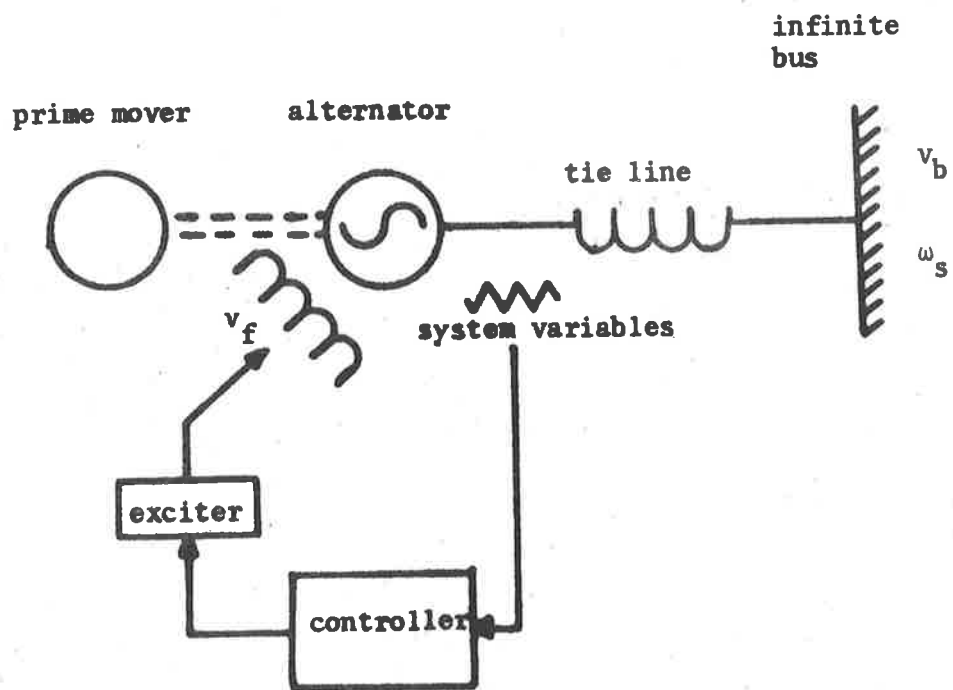


Figure 3.1 The SGTL system, controller and exciter

3.1 INTRODUCTION

In order to calculate the gains of the optimal controller a mathematical model is required for the single generator connected by a transmission line to an infinite bus. A schematic of the SGTL system studied is shown in figure 3.1.

The controller derived in this chapter acts as a voltage regulator which maintains the alternator terminal voltage at, or near, an input reference level while minimizing the defined performance criterion. The controller operates only on the applied field voltage of the alternator. The field exciter is an electronic amplifier with a wide bandwidth (see section 5.11), and time constants associated with it are considered negligible. Thus for the purpose of this study the field exciter is assumed to be simply a unity voltage gain device, and is not considered in the derivation of the optimal controller.

Prime mover power is assumed to be constant during a disturbance. The characteristics of the governing system are not included in the mathematical model, and no attempt is made to optimize performance for perturbations in the required output power.

The time constants of the present thyristor exciters are small (typically 0.03 seconds⁴) in comparison with the field time constant of a large alternator, and could well be neglected in the analysis of a commercial system. If it appeared that the exciter time constant would significantly affect the behaviour of the system, then the procedures adopted here could be extended to include this feature, as well as the characteristics of the prime mover governing system.

3.2 A MATHEMATICAL MODEL OF THE SYSTEM

Optimization theory is applied to the non linear SGTL system by deriving a small signal linear mathematical model of the system from Park's equations.^{8,9} The determination of a suitable controller using optimization techniques is greatly simplified by this approach. In addition this procedure

has been widely used in studies of the small signal behaviour of the SGTL system^{1,2,3,4} so that a comparison can readily be made between the performance of the controller determined using optimization theory, and those resulting from the so called classical approaches, both developed from the same mathematical model.

The linearized equations are first manipulated into a form which allows the model to be structured as a block diagram which is easily related to the physical plant (see section 4.5). The state equation is then derived from this structured form.

The model that is derived ignores the presence of disturbance inputs. In the actual system these may consist of deviations in the infinite bus voltage or frequency or changes in the prime mover power or the Thevenin equivalent of the tie line impedance. As explained in section 2.6 optimization against disturbances is generally only an approximation due to their indeterminate nature. It is considered that any advantage which may be gained in attempting to optimize for these disturbances is offset by the loss of the simple mathematical model otherwise obtained. It is also shown in section 2.6 that a stabilizing controller may be found even if these disturbances are ignored. (Depending on the nature of the disturbance inputs, it may be found that they can only be included in the mathematical representation of the system by substantially altering the composition of the model. This radically changed model may prove to be unstable with the feedback gains derived for the simple model). Thus disturbance inputs have been ignored in the derivation of the mathematical model, although the effects of some of these on the laboratory model are considered in Chapter 8.

3.2.1 THE LINEARIZATION OF PARK'S EQUATIONS

Park's non linear equations describing the synchronous machine may be expressed in the standard operational form;

$$\begin{aligned}
v_d &= -r_a i_d + \frac{p}{\omega_s} \psi_d - \frac{\omega}{\omega_s} \cdot \psi_q \\
v_q &= -r_a i_q + \frac{p}{\omega_s} \cdot \psi_q + \frac{\omega}{\omega_s} \cdot \psi_d \\
\psi_d &= G(p) \cdot U_f = x_d(p) \cdot i_d \\
\psi_q &= -x_q(p) \cdot i_q \\
v_t^2 &= v_d^2 + v_q^2 \\
C_e &= \psi_d \cdot i_q - \psi_q \cdot i_d
\end{aligned} \tag{3.1}$$

A further two equations describe the coupling of the machine to the infinite bus through an external reactance $r_e + jx_e$;

$$\begin{aligned}
v_d &= v_b \cdot \sin \delta + \left(r_e + \frac{px_e}{\omega_s} \right) \cdot i_d - \frac{\omega}{\omega_s} \cdot x_e \cdot i_q \\
v_q &= v_b \cdot \cos \delta + \left(r_e + \frac{px_e}{\omega_s} \right) \cdot i_q + \frac{\omega}{\omega_s} \cdot x_e \cdot i_d
\end{aligned} \tag{3.2}$$

These equations ignore the effect of magnetic saturation. If the direct and quadrature axis subtransient time constants are assumed to be zero, and the machine has no damper windings, or alternatively, if the machine is assumed to be in a "slowly varying steady state" condition, then the functions $G(p)$, $x_d(p)$ and $x_q(p)$ are defined as:

$$\begin{aligned}
G(p) &= \frac{1}{1 + pT'_{do}} \\
x_d(p) &= \frac{x_d + pT'_{do} x'_d}{1 + pT'_{do}} \\
x_q(p) &= x_q
\end{aligned} \tag{3.3}$$

Further simplifying assumptions are made. Perturbations in speed are considered sufficiently small that $\omega/\omega_s = 1$. Under slowly varying conditions, the transformer voltage terms ($p\psi_d$, $p\psi_q$) in equation 3.1 are negligible compared with the speed voltage terms ($\omega\psi_d$, $\omega\psi_q$) so that p/ω_s is assumed zero. The resistances r_a and r_e are assumed negligible in comparison with other impedances in the system, i.e. $r_a = r_e = 0$.

Equations (3.2) and the first two equations in (3.1) simplify to:

$$\begin{aligned} v_d &= v_b \cdot \sin \delta - x_e \cdot i_q \\ v_q &= v_b \cdot \cos \delta + x_e \cdot i_d \\ v_d &= -\psi_q \\ v_q &= \psi_d \end{aligned} \quad (3.4)$$

Small signal linearization of equations (3.3) and the last four equations in (3.1) yield the following small signal linear relationships:

$$\begin{aligned} \Delta v_d &= v_b \cdot \cos \delta_o \cdot \Delta \delta - x_e \Delta i_q \\ \Delta v_q &= v_b \cdot \sin \delta_o \Delta \delta + x_e \Delta i_d \\ \Delta v_d &= -\Delta \psi_q \\ \Delta v_q &= -\Delta \psi_d \\ \Delta \psi_d &= G(p) \cdot \Delta v_f - x_d(p) \cdot \Delta i_d \\ \Delta \psi_q &= -x_q(p) \cdot \Delta i_q \\ \Delta v_t &= \frac{v_{do}}{v_{to}} \cdot \Delta v_d + \frac{v_{qo}}{v_{to}} \cdot \Delta v_q \end{aligned} \quad (3.5)$$

$$\Delta C_e = \psi_{do} \cdot \Delta i_q - \psi_{qo} \Delta i_d + i_{qo} \Delta \psi_d - i_{do} \Delta \psi_q$$

where Δ implies a small perturbation, and the suffix o implies a steady state or initial condition.

Following the procedure of Reggiani and Saccomanno²¹, Δv_d , Δv_q , Δi_d , Δi_q , $\Delta \psi_d$ and $\Delta \psi_q$ are eliminated from (3.4) to derive:

$$\Delta C_e = \frac{(K_1 + pTK')\Delta\delta + K_2\Delta v_f}{1 + pT} \quad (3.5a)$$

$$\Delta v_t = \frac{-(K_3 + pTK')\Delta\delta + K_4\Delta v_f}{1 + pT} \quad (3.5b)$$

$$\text{where } T = \frac{x_e + x'_d}{x_e + x_d} T'_{do} \quad (3.5c)$$

The K parameters are functions of the steady state operating point as defined by,

$$K_1 = \frac{x_e x_q}{(x_e + x_q)v_{to}^2} (Q_{mo} + \frac{v_{to}^2}{x_q}) (\frac{v_o^2}{x_e} - Q_{mo}) - P_o^2 - (x_d - x_q) - \frac{(x_e + x_q)v_{to}^2}{(x_e + x_d)x_q^2} \frac{P_o^2}{P_o^2 + (Q_{mo} + \frac{v_{to}^2}{x_q})^2}$$

$$K_2 = \frac{v_{do}(x_e + x_q)}{x_q(x_e + x_d)} \quad (3.6)$$

$$K_3 = \frac{x_e x_q P_o}{(x_e + x_q)v_{to}} + (x_d - x_q) \frac{x_e v_{to}}{(x_e + x_d)x_q} - \frac{P_o(Q_{mo} + \frac{v_o^2}{x_q})}{P_o^2 + (Q_{mo} + \frac{v_o^2}{x_q})^2}$$

$$K_4 = \frac{v_{qo} x_e}{v_{to} (x_e + x_d)}$$

K'_1 and K'_3 are defined by replacing x_d with x'_d in the expressions for K_1 and K_3 respectively. v_{do} and v_{qo} in the expressions for K_2 and K_4 can be derived as functions of P_o , Q_{mo} , v_b , x_e , x_d and x_q , so that all of the K parameters are functions of the steady state operating point in the complex (PQ) power plane, the infinite bus voltage (v_b), and the machine and tie line reactances (x_d , x_q , x'_d and x_e).

3.2.2 THE MECHANICAL EQUATION OF MOTION

The mechanical torque of the prime mover is balanced by the electrical torque of the alternator and damping and inertia torques:

$$C_m = C_e + C_d + J_i p \omega \quad (3.7)$$

where the damping torque C_d is a function of speed and may be approximated by,

$$C_d = C_{do} + b_1 (\omega - \omega_s) \quad (3.8)$$

C_{do} is a constant at synchronous speed. The mechanical torque C_m can be expressed similarly as a function of speed,

$$C_m = C_{mo} - b_2 (\omega - \omega_s) \quad (3.9)$$

An expression of this form is generally used to represent the speed torque characteristic of the prime mover and is a good approximation for small variations in speed.

From equations (3.7), (3.8) and (3.9) the torque equation becomes:

$$C_{mo} - C_{do} = C_e + B' (\omega - \omega_s) + J_i p \omega \quad (3.10)$$

where $B' = b_1 + b_2$

For small perturbations in speed $C_{mo} - C_{do}$ is constant, and the small signal description is:

$$-\Delta C_e = B' \Delta \omega + J_i p \Delta \omega \quad (3.11)$$

The dimensions of these terms are:

$$\begin{aligned} \Delta C_e &= \text{electrical torque} & \text{Nm} \\ B' &= \text{damping coefficient} & \text{Nm/rad/s} \\ J_i &= \text{moment of inertia} & \text{Nm/rad/s}^2 \text{ (kg m}^2\text{)} \\ \Delta \omega &= \text{rads/s (electrical radians throughout)} \end{aligned}$$

(3.11) may be divided by $P_b / \omega_s = C_b$ where P_b is the volt ampere rating of the alternator (base power), ω_s the synchronous speed, and C_b is the base torque, giving

$$-\Delta C_e = \frac{B' \omega_s}{P_b} \Delta \omega + \frac{J_i \omega_s}{P_b} p \Delta \omega \quad (3.12)$$

ΔC_e is in per unit.

Define the inertia constant

$$\begin{aligned} H &= \frac{\text{stored energy in kWs}}{\text{rating in kVA}} \\ &= \frac{\frac{1}{2} J_i \omega_s^2}{P_b} = \pi f_s \frac{\omega_s J_i}{P_b} \end{aligned} \quad (3.13)$$

where f_s is the rated frequency.

The per unit damping constant is defined as

$$D = \omega_s B' / P_b \quad (3.14)$$

Using these definitions, equation (3.12) is expressed as

$$-\Delta C_e = D \Delta \omega + M p \Delta \omega \quad (3.15)$$

where $M = H / \pi f_s$, and the dimensions of the terms are:

$$\Delta C_e \text{ per unit} = \frac{\text{Nm}}{\text{Nm}}$$

$$D \frac{\text{Nm}}{\text{rad/s}} \cdot \frac{\text{rad/s}}{\text{kW}} = \frac{1}{\text{rad/s}}$$

$$M \frac{\text{Nm}}{\text{rad/s}^2} \cdot \frac{\text{rad/s}}{\text{kW}} = \frac{1}{\text{rad/s}^2}$$

$$\omega \text{ rad/s}$$

For a 50Hz synchronous machine per unit speed is $100\pi \text{ rad/s}$, therefore if ω is expressed as per unit, equation (3.15) must be rewritten as,

$$-\Delta C_e = 100\pi D \Delta\omega + 100\pi M p \Delta\omega \quad (3.16)$$

The dimensions of ΔC_e , M and D are unchanged.

3.2.3 THE STRUCTURE OF THE MODEL

The equations in (3.5) relating perturbations in electrical torque and terminal voltage to field voltage and rotor angle were derived as:

$$\Delta C_e = \frac{(K_1 + pTK') \Delta\delta + K_2 \Delta v_f}{1 + pT} \quad (3.17a)$$

$$\Delta v_t = \frac{(K_3 + pTK') \Delta\delta + K_4 \Delta v_f}{1 + pT} \quad (3.17b)$$

where all quantities are expressed in per unit.

If these are combined with equation (3.16) the mathematical model has the block diagram structure of figure 3.2.

3.2.4 THE STATE EQUATION

The mathematical model is now expressed in the state equation form,

$$\dot{x} = Ax + Bu \quad (3.18)$$

Perturbations in electrical torque, terminal voltage and machine speed were chosen as the three states necessary to

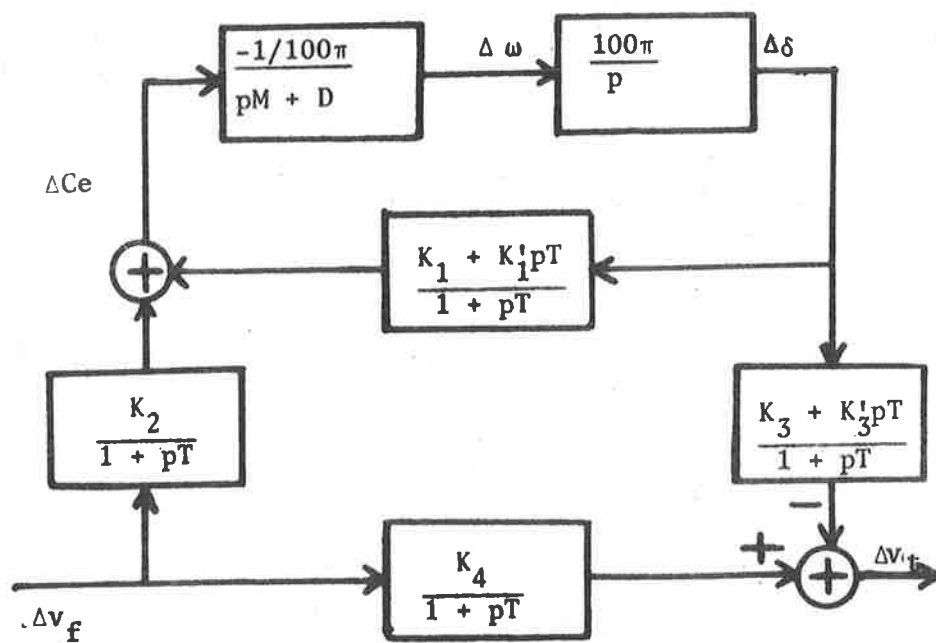


Figure 3.2 The structure of the mathematical model.

describe the condition of the third order system. Perturbation in the applied field voltage, which is the plant input in figure 3.2, becomes the control force u .

Multiplying equations (3.17a) and (3.17b) by $(1 + pT)$ yields:

$$\begin{aligned}\dot{\Delta C_e} &= -\frac{\Delta C_e}{T} + \frac{K_1}{T} \Delta \delta + K'_1 \dot{\Delta \delta} + \frac{K_2}{T} \Delta v_f \\ \dot{\Delta v_t} &= -\frac{\Delta v_t}{T} - \frac{K_3}{T} \Delta \delta - K'_3 \dot{\Delta \delta} + \frac{K_4}{T} \Delta v_f\end{aligned}\quad (3.19)$$

where $x \equiv px$

Using the relationship,

$$K_4(K'_1 - K_1) = K_2(K_3 - K'_3) \quad (3.20)$$

$\Delta \delta$ is eliminated from (3.19) to give

$$\begin{aligned}\dot{\Delta v_t} &= -\frac{1}{T} \left(1 - \frac{K_2 K_3}{DK}\right) \Delta v_t - K'_3 \dot{\Delta \delta} - \frac{1}{T} \left(\frac{K_3 K_4}{DK}\right) \Delta C_e + \frac{K_4}{T} \Delta v_f \\ \dot{\Delta C_e} &= -\frac{1}{T} \left(\frac{K_1 K_2}{DK}\right) \Delta v_t + K'_1 \dot{\Delta \delta} - \frac{1}{T} \left(1 - \frac{K_1 K_4}{DK}\right) \Delta C_e + \frac{K_2}{T} \Delta v_f\end{aligned}\quad (3.21)$$

where $DK = K'_1 K_4 + K'_3 K_2$.

Recalling that when measured in per unit

$$\Delta \delta = 100\pi \Delta \omega \quad (3.22)$$

the state equation is formed from equations (3.21) and (3.16)

as

$$\begin{bmatrix} \dot{\Delta v_t} \\ \dot{\Delta \omega} \\ \dot{\Delta C_e} \end{bmatrix} = \begin{bmatrix} a_{11} & a_{12} & a_{13} \\ 0 & a_{22} & a_{23} \\ a_{31} & a_{32} & a_{33} \end{bmatrix} \begin{bmatrix} \Delta v_t \\ \Delta \omega \\ \Delta C_e \end{bmatrix} + \begin{bmatrix} b_1 \\ 0 \\ b_3 \end{bmatrix} \Delta v_f \quad (3.23)$$

where the elements a_{ij} and b_j are defined as follows:

$$\begin{aligned}
 a_{11} &= -\frac{1}{T} \left(1 - \frac{K_2 K_3}{K'_1 K_4 + K'_3 K_2} \right) \\
 a_{12} &= -100\pi K'_3 \\
 a_{13} &= -\frac{1}{T} \left(\frac{K_3 K_4}{K'_1 K_4 + K'_3 K_2} \right) \\
 a_{21} &= 0 \\
 a_{22} &= -\frac{D}{M} \\
 a_{23} &= -\frac{1}{100\pi M} \\
 a_{31} &= -\frac{1}{T} \left(\frac{K_1 K_2}{K'_1 K_4 + K'_3 K_2} \right) \\
 a_{32} &= 100\pi K'_1 \\
 a_{33} &= -\frac{1}{T} \left(1 - \frac{K_1 K_4}{K'_1 K_4 + K'_3 K_2} \right) \\
 b_1 &= \frac{K_4}{T} \\
 b_2 &= 0 \\
 b_3 &= \frac{K_2}{T}
 \end{aligned} \tag{3.24}$$

The mathematical model of the system is therefore expressed in the standard state equation form.

3.3 THE COST FUNCTION

In Chapter 2 it was shown how the elements of the cost function are chosen after due consideration of the physical requirements of the plant. At present no analytical techniques are available to unambiguously determine a suitable cost function (other than guide lines such as the work of Tyler and Tuteur¹⁴ outlined in section 2.2). Thus the determination of the most

suitable cost function for a particular system involves a considerable amount of research into the physical plant, its requirements, and the way in which these requirements can be translated into mathematical terms. In the case of the alternator regulator, important factors may be: the degree of rotor damping required; the effect of external parameter variations, the effect of disturbance inputs which are not included in the mathematical model; the response to large disturbances; or the rate of change of optimizing gains with operating point. In addition it must be possible to physically realize the required control force in a manner which is economically and technically acceptable.

The elements of the cost function chosen for an operational alternator regulator would depend on the particular circumstances in which the alternator was to be used. That is, there are no physical specifications which can be applied to all systems in general, and each particular installation must be considered separately. It was therefore concluded that a detailed investigation on the choice of a cost function for this particular study was not warranted, and consideration was given mainly to the ease of implementation of the controller. The influence of the cost function on the behaviour of the laboratory model is discussed in conjunction with the presentation of the results in Chapters 7 and 8 and, after a detailed study of the results, greater physical insight into the overall problem is obtained.

The output vector was chosen as

$$z = \begin{bmatrix} \Delta v_t \\ 0 \\ 0 \end{bmatrix} \quad (3.25)$$

that is the matrix C in

$$z = Cx \quad (3.26)$$

is

$$C = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \quad (3.27)$$

In the cost function chosen a penalty is placed only on the control force and the deviation of the perturbed terminal voltage from its required value, so that the optimal controller becomes basically a voltage tracking regulator. The quadratic cost function of section 2.3.1 is used, and in this case takes the simple form;

$$J = \frac{1}{2} \int_{t_0}^{t_f} (\Delta v_{\text{ref}} - \Delta v_t)^2 + \Delta v_f^2 g \, dt. \quad (3.28)$$

Comparing this with the most general form of the quadratic cost function in equation (2.3) the following points arise: the weighting factor S has been assumed zero, i.e. no special penalty has been placed on the deviation of the perturbed terminal voltage from its desired value at the terminal time; the matrix G is reduced to the scalar g ; and the weighting factor N has been put equal to the scalar $d = 1$, as the ratio of the weighting factors d and g determines the relative weighting of the elements, rather than their absolute values. (This arises because $S \equiv 0$, and the minimum value of J is simply multiplied by a constant equal to d if d is varied while the ratio d/g is held constant, i.e.

$$dJ = \frac{d}{2} \int_{t_0}^{t_f} (\Delta v_{\text{ref}} - \Delta v_t)^2 + \Delta v_f^2 g/d \, dt)$$

A value of g for the cost function (3.28) is then chosen such that limiting does not occur in the field exciter for small step changes in reference voltage. In addition to this, a check is made on the resulting

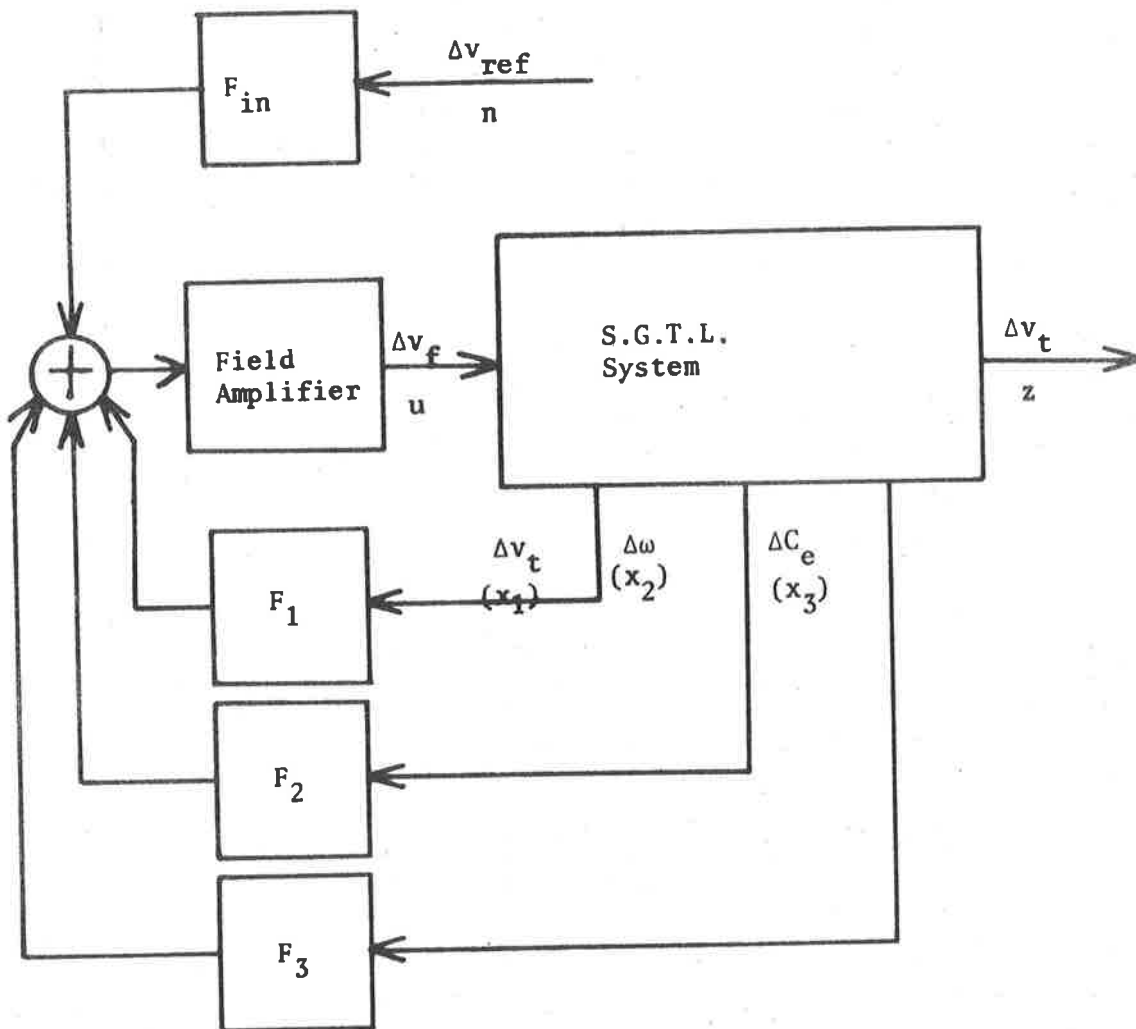


Figure 3.3 The structure of the optimal controller.

controller to ensure that the response of the system states to step changes in reference voltage is such that transducer time constants (see Appendix A) can be ignored in subsequent analysis.

3.4 EVALUATION OF THE OPTIMAL CONTROLLER

The elements of the A and B matrices (equation 3.24) in the state equation are functions of the inertia and damping constants, the K parameters, and the field short circuit time constant. The machine set inertia is constant. The damping coefficient is constant for a given steady state operating point. From equation (3.6) the K parameters are seen to be functions of the PQ operating point, the infinite bus voltage (assumed constant) and constant reactances, so that these are constant for a given steady state operating point. If the magnetic saturation is constant then the field time constant is set. Therefore the elements of the A and B matrices are constant and the state equation linear for small perturbations about a steady state operating point. In general however these elements are non linear functions of P and Q . The gains of the optimal controller are therefore dependent on the steady state operating point.

To facilitate the implementation of the controller (see Chapter 6), the complex power plane (PQ plane) is divided into a grid by lines of real and reactive power. The intersection of these lines yields a set of steady state operating points, or grid points, at which the elements of the A and B matrices are evaluated. The Riccati equation (2.18) is solved using the procedure outlined in section 2.5 and the gains found from equations (2.20) and (2.21) which optimize performance at each of these grid points. For any one of these steady state conditions the optimal controller takes the form of the feedback system shown in figure 3.3.

3.5 CONTROLLABILITY AND OBSERVABILITY

The observability criterion as defined in section 2.4 requires that the states of the system can be determined from the output. The equivalent mathematical statement for the time invariant system is that the partitioned matrix Γ defined in equation (2.24) is of rank n . With the output matrix multiplier

$$C = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \quad (3.29)$$

the matrix

$$\Gamma = (C^T \mid A^T C^T \mid A^{2T} C^T) \quad (3.30)$$

is obviously not of rank 3 and the system therefore fails this definition of the observability criterion. This implies that an optimal controller may not exist¹⁰. The observability criterion however assumes that the states of the system may only be determined from analysis of the control force and the response of the output. In this particular case the voltage and speed states are directly accessible and electrical torque can be estimated from measurements of real power and speed. Thus the system states are all available and an optimal controller may exist even though the states cannot be estimated from the output vector.

The controllability criterion requires that the matrix ϕ defined by equation (2.25) is of rank n . For a third order system this means that

$$\phi = (B \mid AB \mid A^2B) \quad (3.31)$$

must be of rank 3. Considering the elements of the A and B matrices given by equation (3.24), no general conclusion can be drawn concerning the

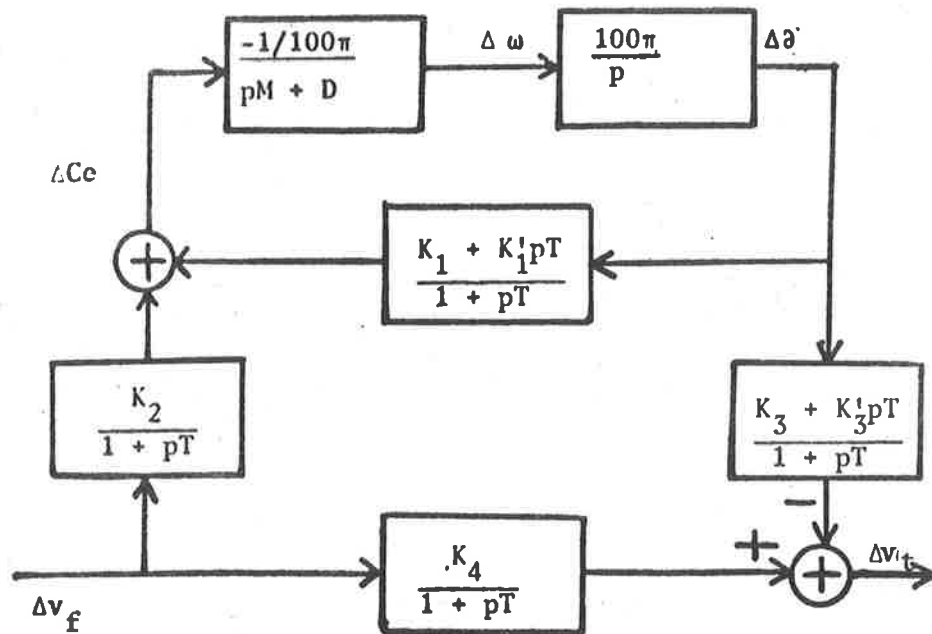


Figure 3.4 The structure of the mathematical model.

controllability of the system. Numerical evaluation of the elements of Φ at each grid point is required before it can be determined whether the system is controllable throughout the complex power plane.

3.5.1 THE CHOICE OF SYSTEM STATES

Considering the structure of the mathematical model in figure 3.2 (shown here as figure 3.4), the obvious choice for the three states of the system would be made from the four variables Δv_t , $\Delta \delta$, $\Delta \omega$ and ΔC_e , or a combination with their derivatives. The derivatives were thought to be unsuitable after considering the accuracy and ease with which they could be measured or estimated.

Initially, perturbations in terminal voltage, machine speed and power angle were selected as system states because these variables have been widely used by other investigators 1,2,22,23. With this choice of state variables the state

equation (3.18) becomes:

$$\begin{bmatrix} \dot{\Delta v} \\ \dot{\Delta \delta} \\ \dot{\Delta \omega} \end{bmatrix} = \begin{bmatrix} -\frac{1}{T} & -\frac{K_3}{T} & -100\pi K'_3 \\ 0 & 0 & 100\pi \\ -\frac{K_2}{100\pi M K_4} & -\frac{1}{100\pi J} (K'_1 + \frac{K'_3 K_2}{K_4}) - \frac{D}{J} \end{bmatrix} \begin{bmatrix} \Delta v \\ \Delta \omega \\ \Delta \delta \end{bmatrix} + \begin{bmatrix} \frac{K_4}{T} \\ 0 \\ 0 \end{bmatrix} \Delta v_f \quad (3.32)$$

The matrix Φ for the controllability criterion is derived from equations (3.31) and (3.32) as

$$\Phi = \begin{bmatrix} \frac{K_4}{T} & -\frac{K_4}{T^2} & \frac{K_4}{T^2} + \frac{K_2 K'_3}{M} \\ 0 & 0 & -\frac{K_2}{M} \\ 0 & \frac{K_2}{100\pi M} & \frac{K_2}{100\pi M} \left(\frac{1}{T} + \frac{D}{J} \right) \end{bmatrix} \quad (3.33)$$

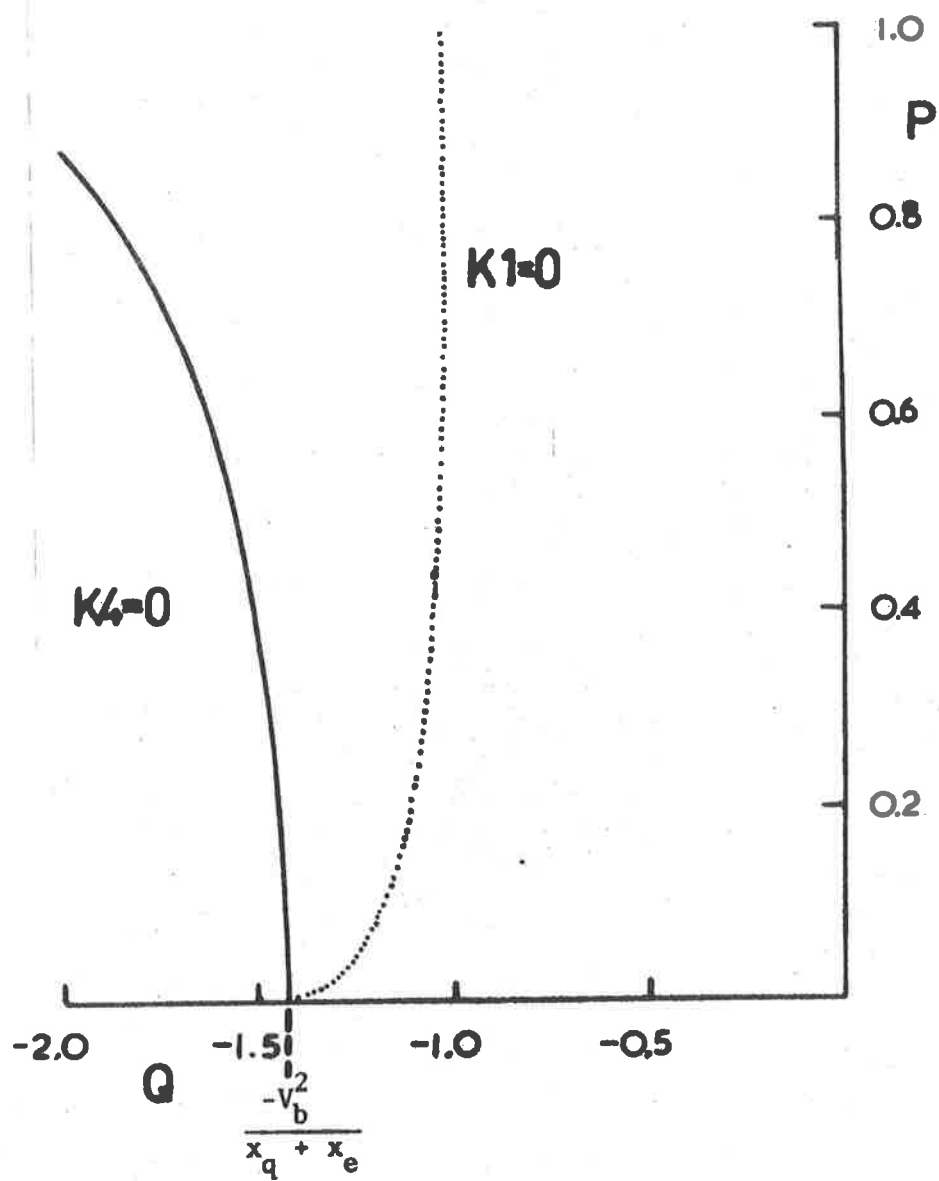


Figure 3.5 The position in the PQ plane of the line $K_1 = 0$ and $K_4 = 0$

When $K_4 = 0$ this reduces to

$$\Phi = \begin{bmatrix} 0 & 0 & \frac{K_2 K'_3}{M} \\ 0 & 0 & \frac{-K_2}{M} \\ 0 & \frac{K_2}{100\pi M} & \frac{K_2}{100\pi M} \left(\frac{1}{T} + \frac{D}{J} \right) \end{bmatrix} \quad (3.34)$$

which is not of rank three. Therefore this system would not be controllable when $K_4 = 0$. The position of this line in the PQ plane is shown in figure 3.5 together with the line $K_1 = 0$. K_4 is equal to zero when the operating point is such that the quadrature axis voltage v_q is zero. $K_1 = 0$ corresponds to the open loop stability limit of the SGTL system. Consider the characteristic equation of the open loop system which is obtained from the state equation (3.23) and the identity (3.20) as

$$p^3 T J + p^2 (T D + J) + p (T K'_1 + D) + K_1 \quad (3.34a)$$

T , J and D are all positive quantities. As the operating point in the PQ plane shifts in the direction of higher leading reactive power, the parameter K_1 changes from positive to negative before a similar sign change in K'_1 . That is, K'_1 is always positive for positive K_1 . Thus, according to the Routh criterion, the point in the PQ plane at which the system becomes open loop unstable occurs when K_1 changes sign.

Using the cost function (3.28) the optimal controller was computed over the complex power plane for this system (as described by state equation 3.32). A plot of the variation of the optimal

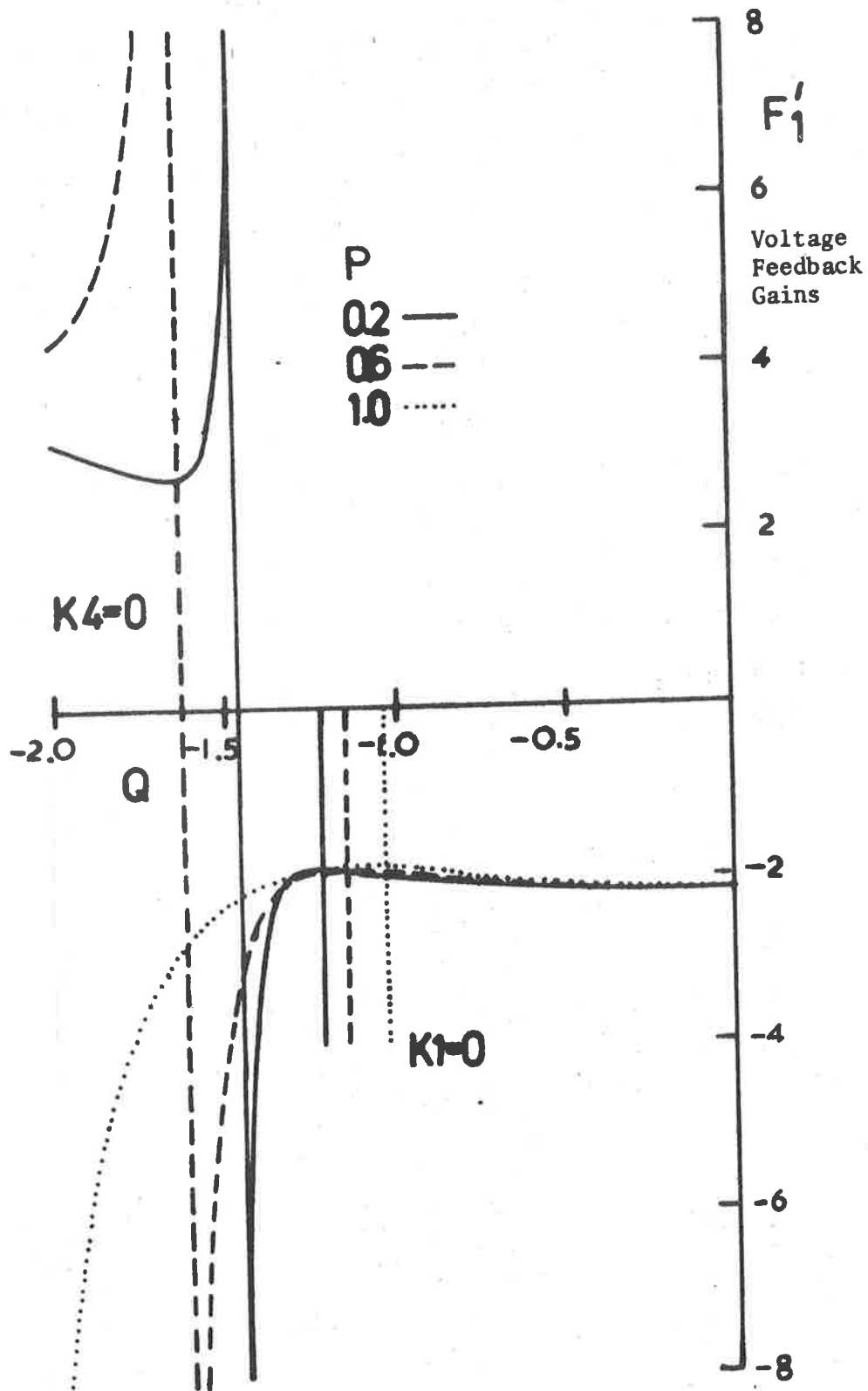


Figure 3.6 The variation of the optimal voltage feedback gain with the steady state operating point. Δv_t , $\Delta \delta$, $\Delta \omega$ as state variables.

voltage feedback gain with the steady state operating point is shown in figure 3.6. When the operating point approaches the region where $K_4 = 0$ the optimal gains tend to plus and minus infinity. The choice of Δv_t , $\Delta \delta$ and $\Delta \omega$ as the state variables therefore proved unsatisfactory.

As mentioned in section 2.4, Choate and Sage¹⁸ have shown that a system which is not controllable must be reducible in order. Equation (3.5b) relating Δv_t , $\Delta \delta$ and Δv_f is

$$\Delta v_t = \frac{-(K_3 + pT K_3') \Delta \delta + K_4 \Delta v_f}{(1 + pT)} \quad (3.35)$$

The K parameters are interdependent according to the equation

$$K_4(K_1 - K_1') = K_2(K_3 - K_3') \quad (3.36)$$

K_2 is not zero coincidentally with K_4 , so that $K_4 = 0$ implies $K_3 = K_3'$. Therefore equation (3.35) reduces to

$$\Delta v_t(t) = -K_3 \Delta \delta(t) \quad (3.37)$$

This means that Δv_t and $\Delta \delta$ can only be controlled along or parallel to the straight line

$$\Delta v_t(t) + K_3 \Delta \delta(t) = 0 \quad (3.38)$$

and the order of the state equation is reduced from three to two. The system is still third order however, and consequently is no longer correctly described by the reduced state equation.

With perturbations in terminal voltage, machine speed and electrical torque as the system state the controllability criterion is satisfied when $K_4 = 0$. Even at the operating point $P = 0$, $Q = -V_b^2 / (x_e + x_q)$ when K_1 and K_4 are simultaneously zero the controllability matrix Φ is given by

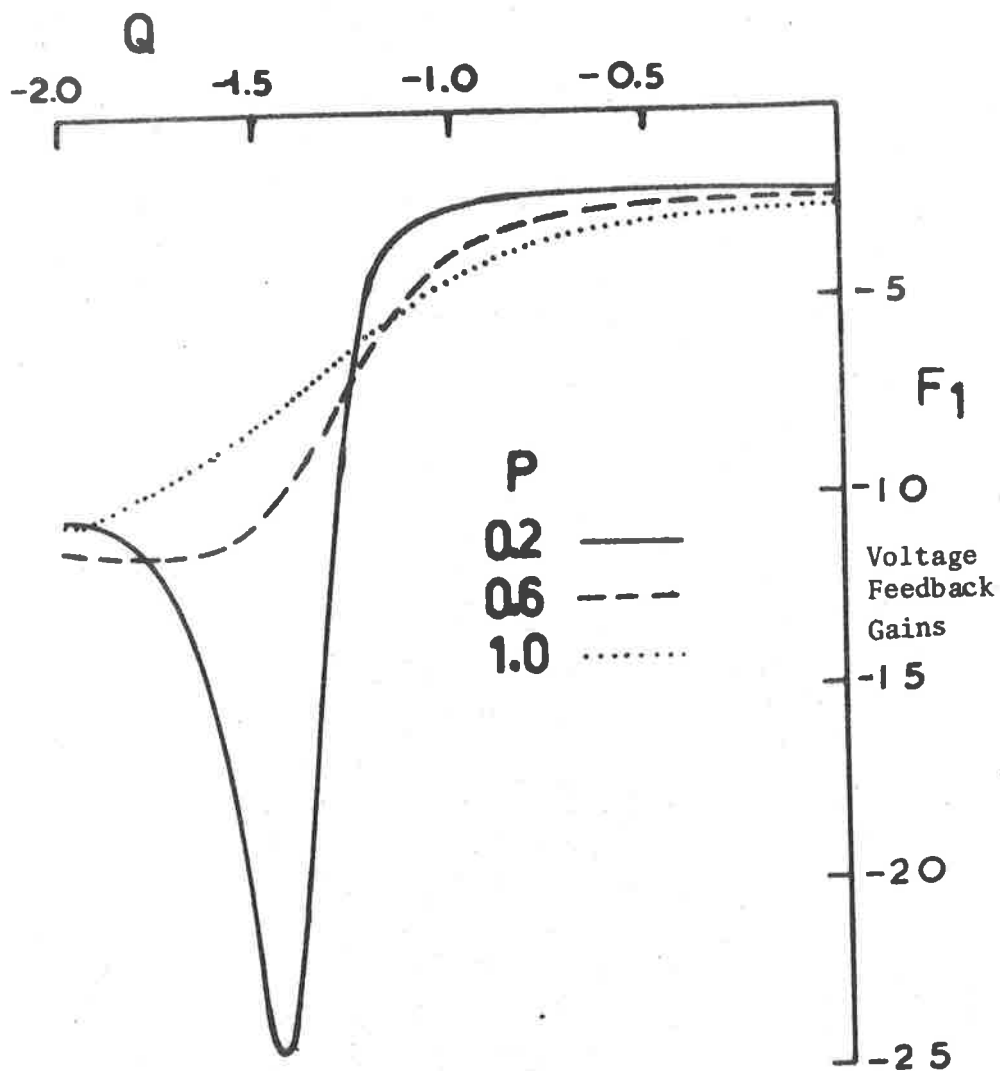


Figure 3.7 Continuous variation of the voltage feedback gain with the PQ operating point when Δv_t , $\Delta \omega$ and ΔC_e are chosen as the system t states.

$$\Phi = \frac{K_2}{T} \begin{bmatrix} 0 & 0 & \frac{K_3}{M} \\ 0 & -\frac{1}{M} & \frac{D}{M^2} + \frac{1}{TM} \\ 1 & -\frac{1}{T} & \frac{-K_1}{T} + \frac{1}{T^2} \end{bmatrix} \quad (3.39)$$

which is of rank 3 so that the system is controllable. If the optimal controller is computed over the operating plane with the system described by these states (Δv_t , $\Delta \omega$, ΔC_e in equation (3.23) then the continuous variation of the voltage feedback gain shown in figure 3.7 is observed. The same cost function (3.28) is used to obtain the curves in figures 3.6 and 3.7.

This simple example demonstrates one of the points which must be considered in choosing system variables as states; that is, if the system parameters vary due to non linearities, is the small signal linear model always controllable? Other points are; can the states be measured or estimated, and if so can they be determined with sufficient accuracy; is it possible to include in the cost function a variable whose deviation is to be penalized, that is, will this variable appear in the output vector z via the identity $z = Cx$? It is apparent from these questions that the choice of suitable state variables may not always be a simple matter of mathematical or practical convenience.

3.6 SUMMARY

The non linear SGTL system is described by a small signal linear model. The following assumptions are made in deriving the model: prime mover power is constant; exciter time constants are negligible; machine speed

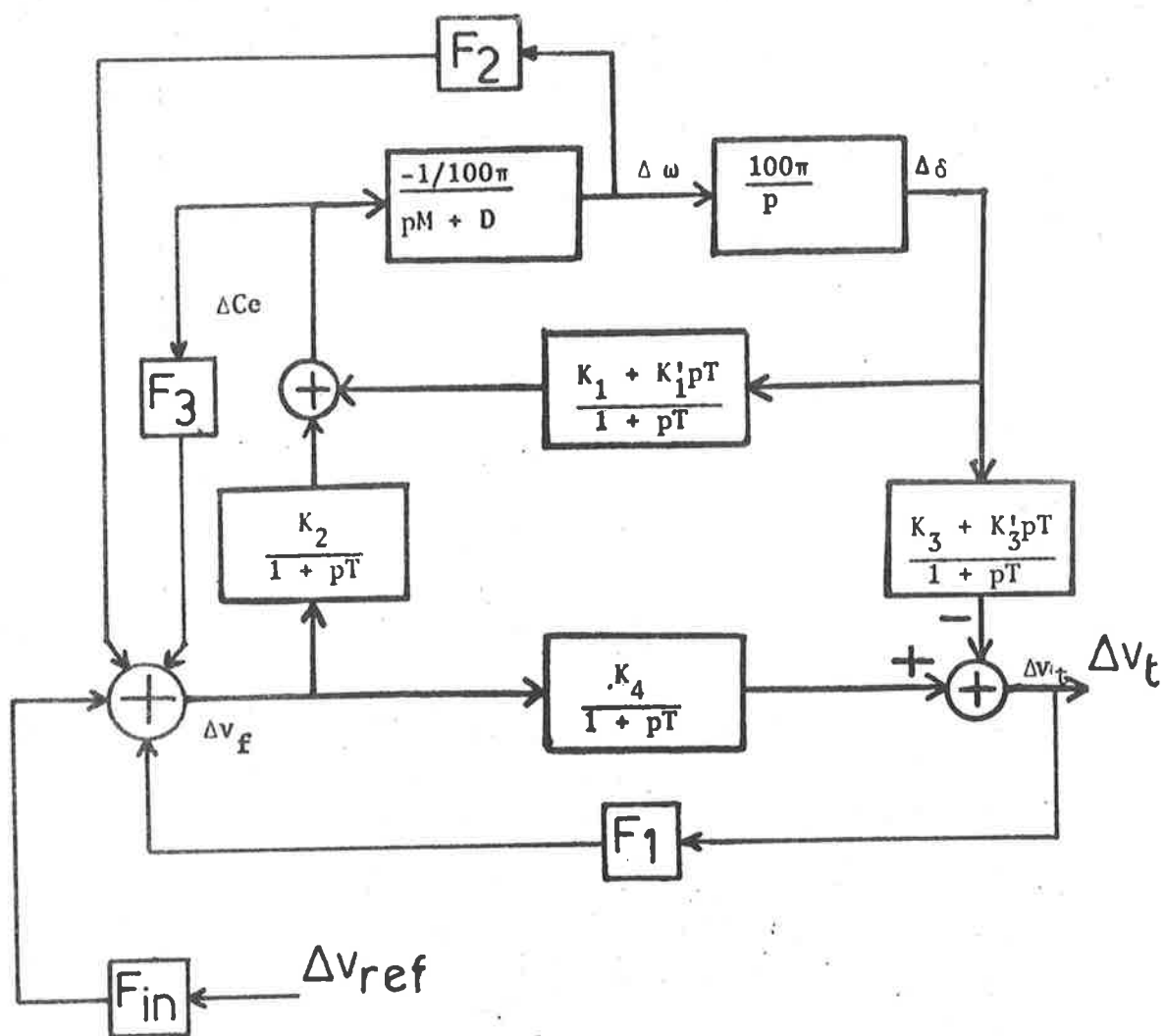


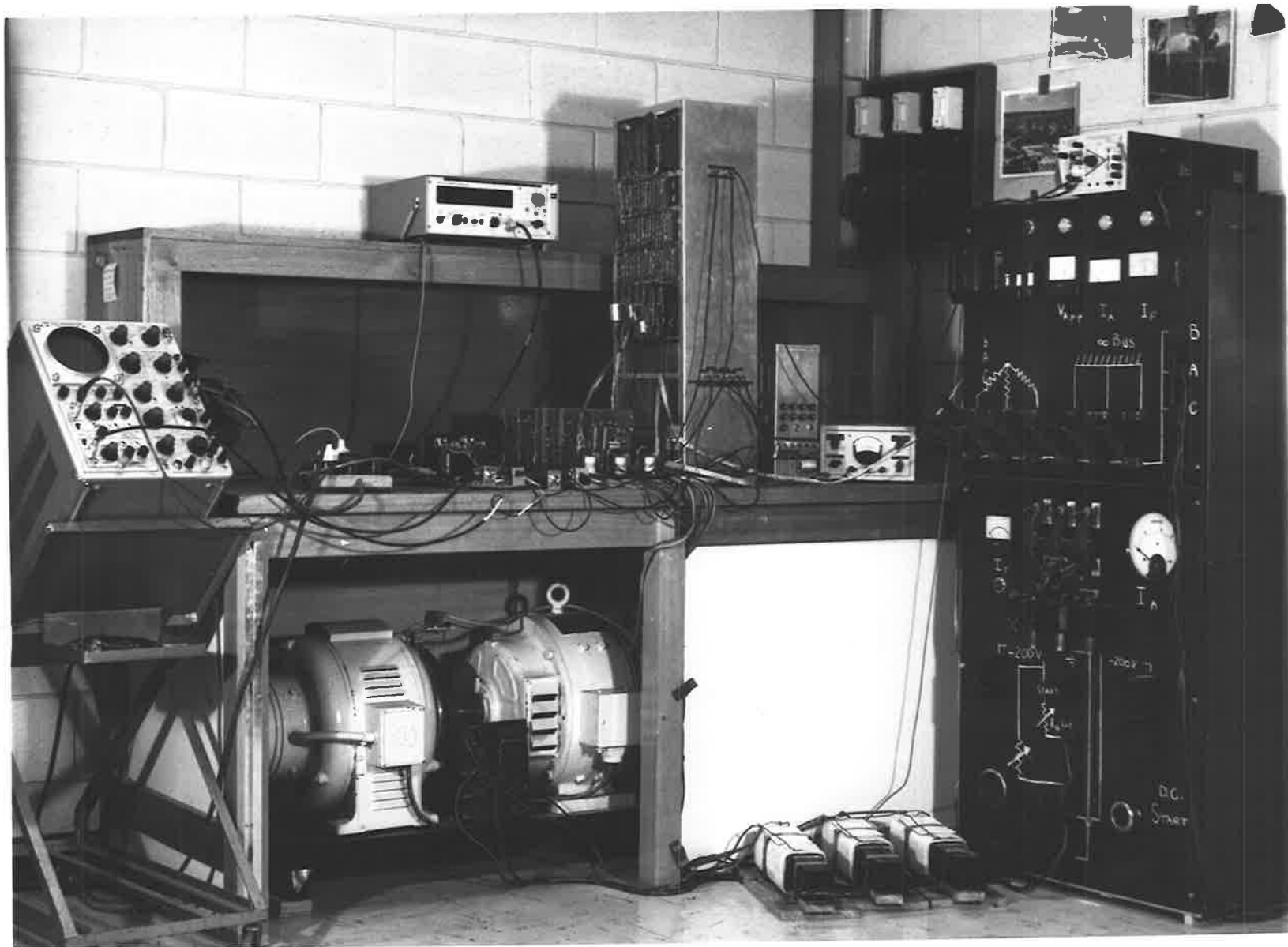
Figure 3.8 The ideal model and optimal controller.

perturbations are small and slow such that $\omega/\omega_s = 1$ and $p/\omega_s = 0$; armature and tie line resistances are zero; the infinite bus voltage and frequency are constant; the tie line reactances are constant; and there is no magnetic saturation. The mathematical model is therefore valid only at a steady state operating point where the above conditions apply, or are, in the case of saturation, allowed for.

The controller which minimizes a simple cost function is computed for the idealized model and the ideal controlled system is shown in figure 3.8.

CHAPTER 4

THE LABORATORY SGT L SYSTEM



4.1 INTRODUCTION

Theoretical analysis of the behaviour of any system is performed using a mathematical model of the system. In this case a number of assumptions are made which simplify and linearize the equations describing the system. To permit these assumptions to be validated and to assess the practical feasibility of applying linear optimal control theory to this non linear system, an experimental model was established in the laboratory. While non linear mathematical models of the SGTL system are sometimes used in the evaluation of a control philosophy,⁵ some basic assumptions must be made in deriving the model. Without checking the model against the actual physical system, its accuracy is open to question. In addition some of the problems associated with the practical implementation of a controller are easily overlooked. These points become evident when analysing the performance of the controlled laboratory model in chapters 7 and 8.

The experimental system is shown in figure 4.1. The alternator and prime mover are positioned under a bench on which is mounted the transducers and an A/D interface (see Chapter 5). The switchboard carries circuit breakers and fuses, meters monitoring system voltages and currents, controlling knife switches and the starter for the D.C. motor prime mover.

In this chapter details of the experimental system are given. The characteristics of the prime mover, and the impedances and time constants of the alternator are determined. The K parameters of the mathematical model are computed and compared with experimentally estimated values to assess the model accuracy. Finally the optimal gains for the linear model are computed to show how they vary with the operating point of the non linear system.

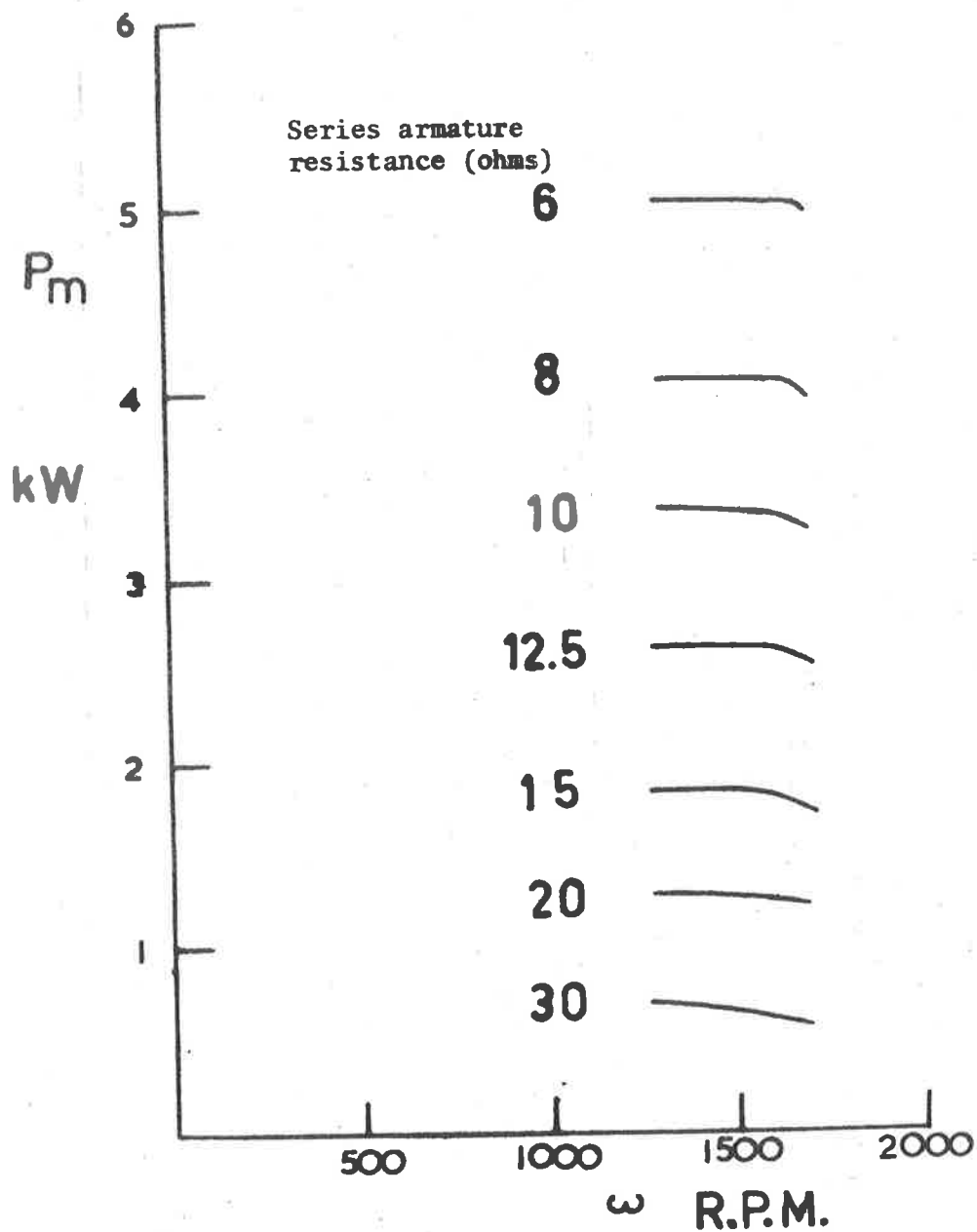


Figure 4.2 Prime mover power against speed for various values of external armature resistance.
Supply voltage 410V, $I_f = 1.0A$

4.2 THE MACHINE SET

The alternator used in the experimental system is 5 KVA 240V, three phase, four pole, salient machine without amortisseur. This is a small general purpose alternator which was available at the time so that the relative magnitudes of its per unit impedances and time constants do not conform with any type of normal commercial machine. No attempt has been made to modify by electronic simulation or other means any of these parameters, the alternator being used as it is. This means that observations on the experimental system cannot be applied directly to any specific supply complex, although this is of little consequence as the procedures and analysis presented are applicable in general.

The alternator is coupled to the supply mains through nonsaturating reactors simulating transmission lines. These can be adjusted from 5 to 12 ohms by means of a sliding core. In determining the optimal controller the tie line reactance is assumed constant at 10 ohms.

The prime mover is a D.C. motor supplied at about two and a half times rated voltage with a large external resistance in series with its armature. The power output is varied by manual adjustment of this resistance.

The equation relating the voltages in the armature circuit of such an arrangement is:

$$V_s = IR + K'w \quad (4.1)$$

where V_s is the supply voltage, I the armature current, R the series resistance, w the machine speed and K' a constant for constant field current (ignoring armature reaction). Neglecting losses, the power output is:

$$P = K'wI \quad (4.2)$$

With a constant supply voltage, a slight increase in speed produces a corresponding reduction on armature current given by:

$$V_T = (I - \Delta I)R + K'(w + \Delta w) \quad (4.3)$$

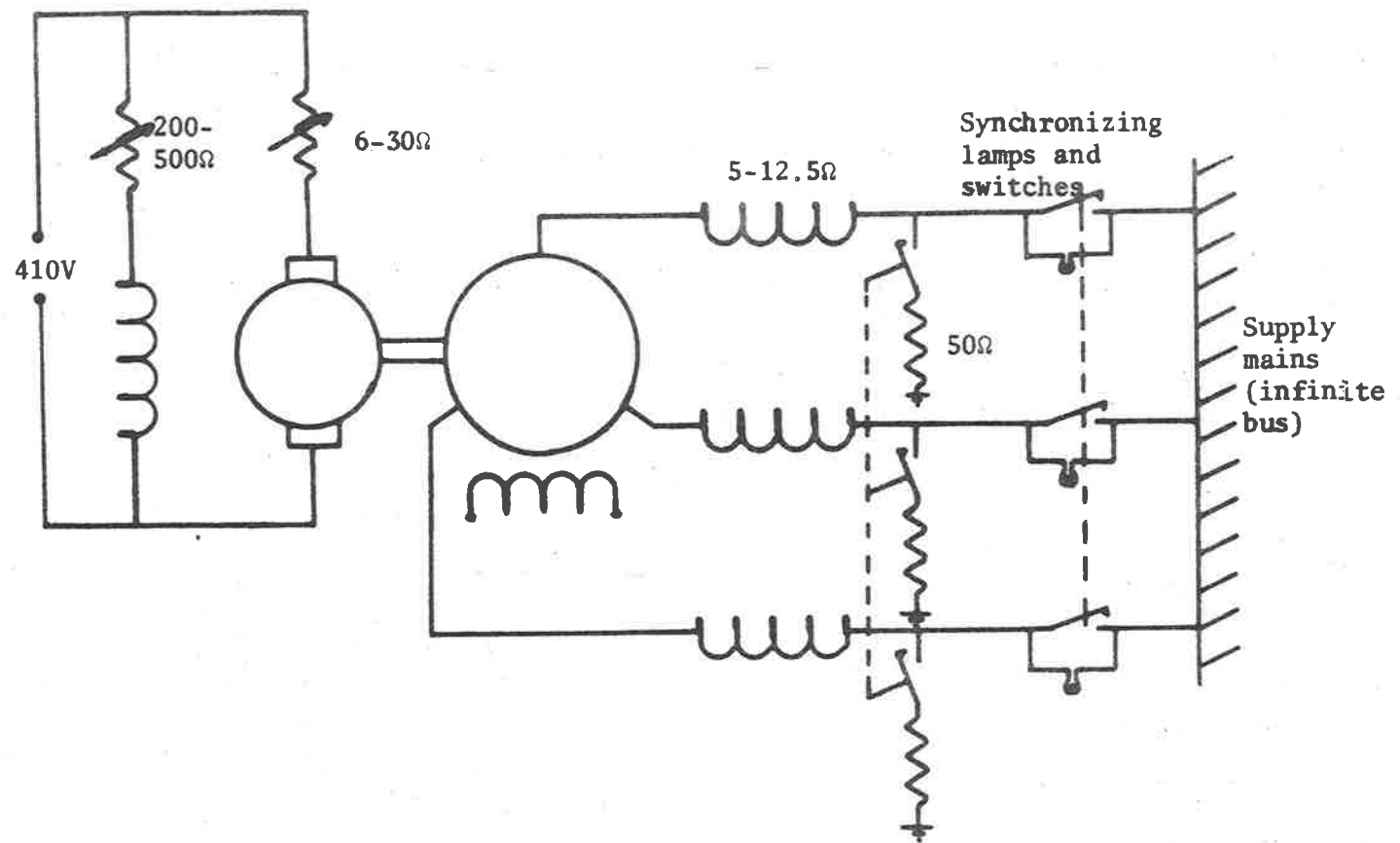


Figure 4.3 The machine set, transmission line reactances and synchronizing system

Thus

$$\Delta I R = K' \Delta w \quad (4.4)$$

or

$$\Delta w = \frac{\Delta I R}{K'}$$

The increase in power output is then given by:

$$\Delta P = K' \Delta w I - K w \Delta I \quad (4.5)$$

This means that the power remains constant provided:

$$I \Delta w = w \Delta I \quad (4.6)$$

Substituting from (4.4) this is true if:

$$K' w = I R = 0.5 V_s \quad (4.7)$$

Therefore with R adjusted so that the power output was 3.3kW, the field current was varied to give equal voltages across the armature and the series resistance. Within the limits of the above assumptions the power then remains constant with speed for all values of R. The power versus speed characteristic is shown in figure 4.2. This plot was obtained with a D.C. supply of 410 volts (approximately 2.5 times rated voltage) and a field current of one amp. These values remained unchanged throughout the experiment.

Figure 4-3 is a schematic of the machine set and tie lines. The 50 ohm resistors are used to load the alternator during synchronization. The D.C. motor armature resistance is adjusted to give approximately the required prime mover power. The 50 ohm resistors are switched in and the machines started. The alternator field is adjusted or the D.C. starter resistance varied until synchronous speed is attained and the synchronizing switch closed when the lamps are dimmed. The 50 ohm resistors may then be switched out. This procedure is necessary because of the way the D.C. motor is supplied. The machine set quickly exceeds synchronous speed if unloaded. Once synchronized the armature resistance of the D.C. motor may be varied as required (provided the stability limit is not exceeded when the alternator field is under manual control).

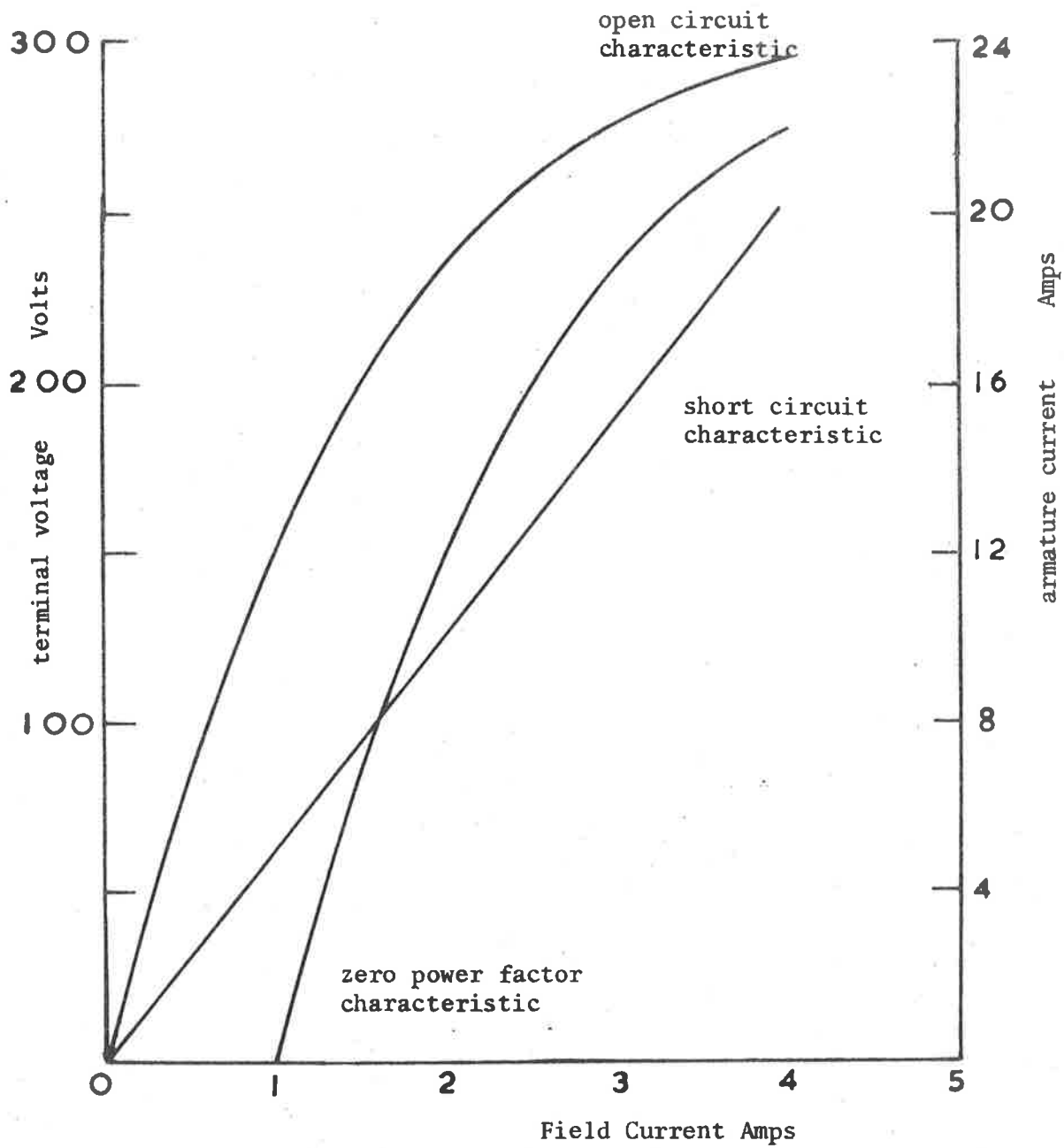


Figure 4.4 The open circuit, zero power factor and short circuit characteristics of the alternator.

Protective switchgear, meters and the D.C. starter are not shown in figure 4.3.

4.2.1 PER UNIT REPRESENTATION

All parameters and variables describing the system are measured in the convenient per unit form. The ratio, expressed as a decimal, of any quantity to a chosen base is the per unit value of that quantity. The base chosen is the VA and V rating of the alternator.

The following quantities then become the one per unit values for the system:

Power	5 KVA	=	P_b	
Voltage	240 V	=	V_b	
Current	6.95 A	=	$I_b = P/V$	
Impedance	34.4 ohms	=	$Z_b = V/I$	
Speed	100π electrad/s	=	ω_s	
Field voltage	20.5V	=	V_{fb}	(That field voltage which produces 1 per unit open circuit terminal voltage at 1 per unit speed)

4.3 THE BASIC MACHINE PARAMETERS

The open circuit and zero power factor characteristics of the synchronous machine are shown in figure 4.4. The zero power factor curve is the same shape as the open circuit characteristic, and the two curves can be overlain by horizontal translation alone. This means that the Poter reactance and hence the leakage reactance, is negligibly small. The leakage reactance is virtually independent of the armature current, and is therefore assumed negligible under all operating conditions, allowing the airgap voltage to be approximated by the terminal voltage. Therefore the effect of saturation on the machine parameters (which was ignored in the derivation of the mathematical model) can be allowed for by plotting their variation with terminal voltage. Knowing (or calculating) the terminal voltage at any particular PQ operating point, the saturated value of the parameters can be determined

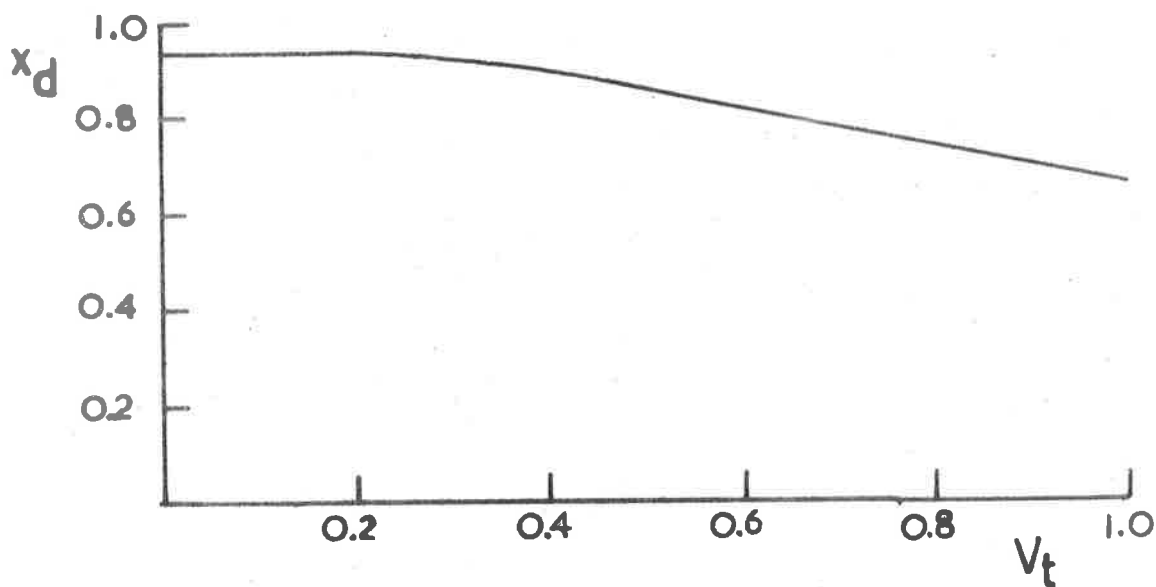


Figure 4.5 The variation of the direct axis synchronous reactance with terminal voltage.

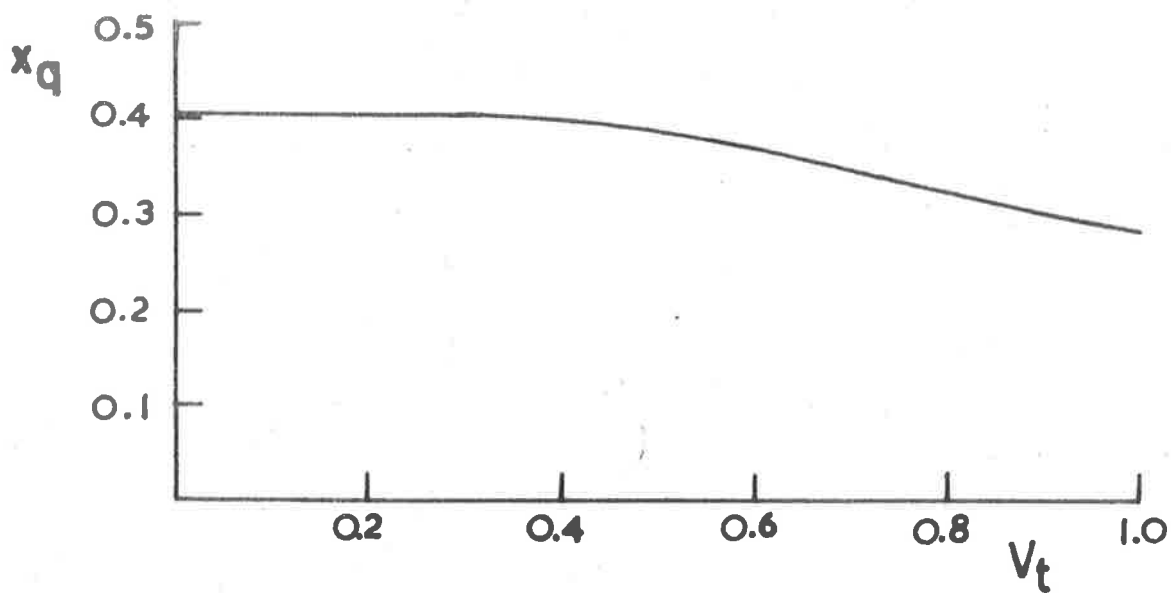


Figure 4.6 The variation of the quadrature axis synchronous reactance with terminal voltage.

from these curves.

4.3.1 THE REACTANCES

Standard tests^{25,27} are used to measure the direct and quadrature axis synchronous reactances, and the direct axis transient reactances.

The direct axis synchronous reactance x_d is determined from the open circuit and short circuit characteristics of figure 4.4. By evaluating this for a range of field currents, figure 4.5 showing the variation of x_d with open circuit terminal voltage is obtained.

The quadrature axis synchronous reactance x_q is determined using the maximum lagging current test. This test is performed with a range of supply voltage to give the plot of x_q against terminal voltage shown in figure 4.6.

Knowing the infinite bus voltage, the tie line reactance and the operating point in the PQ plant, the terminal voltage of the alternator in the single generator tie line system can be computed. Thus for any proposed PQ operating point the appropriate values of x_d and x_q can be found from figures 4.5 and 4.6. Figure 4.7 and 4.8 show how these reactances vary in the PQ operating plane of the SCTL system when the tie line reactance x_e is 0.29 per unit (10 ohms).

The direct axis transient reactance x'_d is found by applying a sudden symmetrical short circuit to the terminals of the alternator. Ignoring the initial rapid decay due to the subtransient reactance, the exponentially decaying waveform of the alternating component of short circuit current is projected to the zero time axis. The direct axis transient reactance is the ratio of the open circuit terminal voltage at $t = 0$ to the projected value of short circuit armature current. For this machine x'_d is measured as 0.12 per unit when the open circuit voltage is 0.4 per unit and the projected short

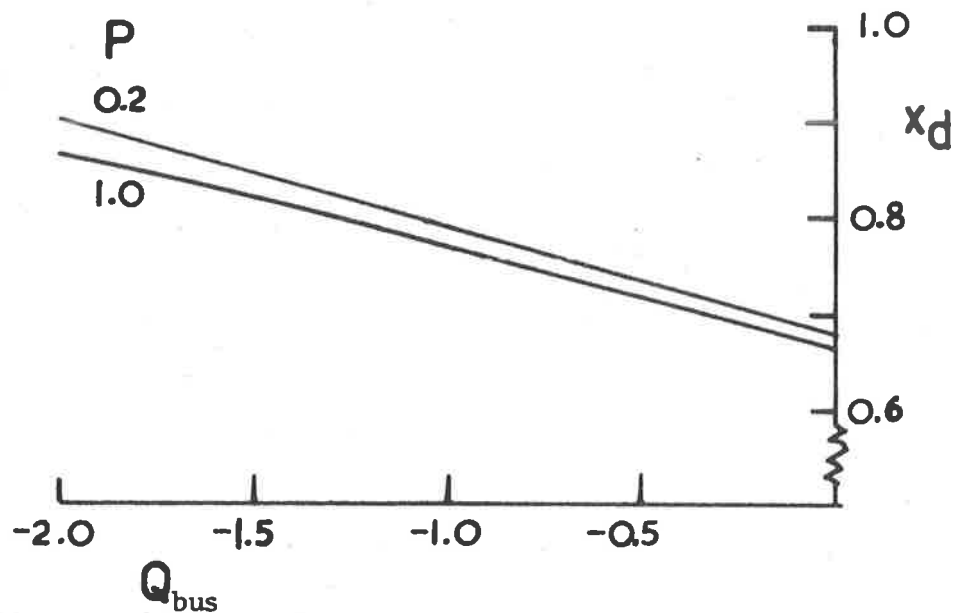


Figure 4.7 Variation of x_d with the PQ operating point of the S.G.T.L. system

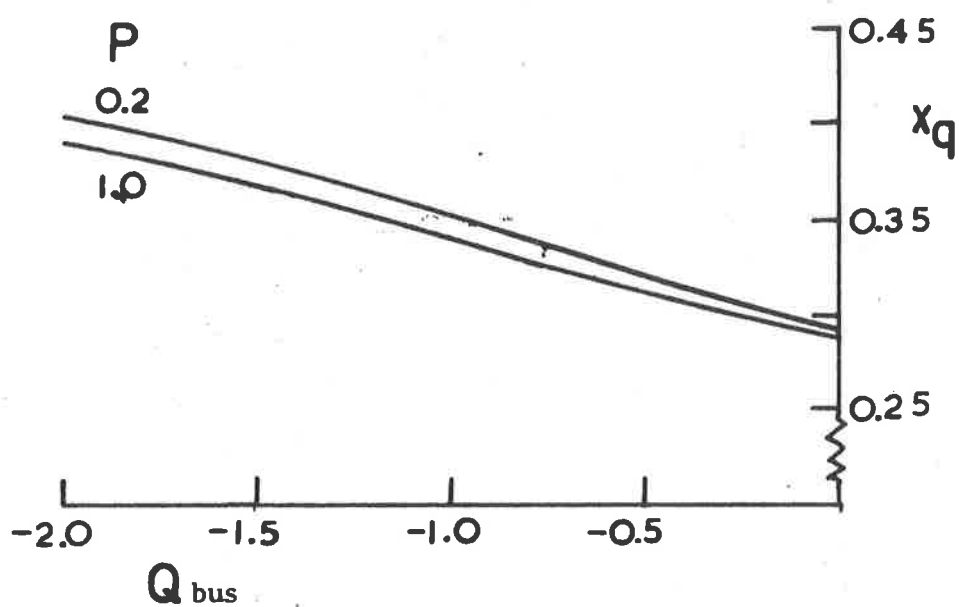


Figure 4.8 Variation of x_q with the PQ operating point of the S.G.T.L. system.

circuit current 3 per unit. Because of the large transient short circuit currents which would result for larger initial terminal voltages, x'_d was not measured as a function of terminal voltage. It may be shown however,²⁶ that for the voltages and currents present in the SGTL system, saturation has a negligible effect on the direct axis transient reactance. Thus x'_d is assumed to be constant at 0.12 per unit.

4.3.2 THE FIELD TIME CONSTANT

While the effect of saturation on the machine reactances is often taken into account variations in the field time constant with the flux levels in the machine are usually ignored.

With the particular alternator used, saturation is seen to have a marked effect on the value of the small signal open circuit field time constant.

Small step deviations in field voltage are applied about a steady state value, and the response of the open circuit terminal voltage recorded and analysed to obtain the curve of T'_{do} against terminal voltage shown in figure 4.9. Once the terminal voltage has been determined for a particular steady state operating point, the corresponding value of field time constant is obtained from this curve. The field short circuit time constant defined by equation 3.5c,

$$T = \frac{x_e + x'_d}{x_e + x_d} T'_{do} \quad (4.8)$$

is used in the mathematical description of the SGTL system. T'_{do} and x_d vary with terminal voltage while x_e and x'_d are constant (0.29 and 0.12 p.u. respectively). Thus T varies with terminal voltage, and therefore with the system operating point. The variation of T with the PQ operating point of the SGTL system is shown in figure 4.10.

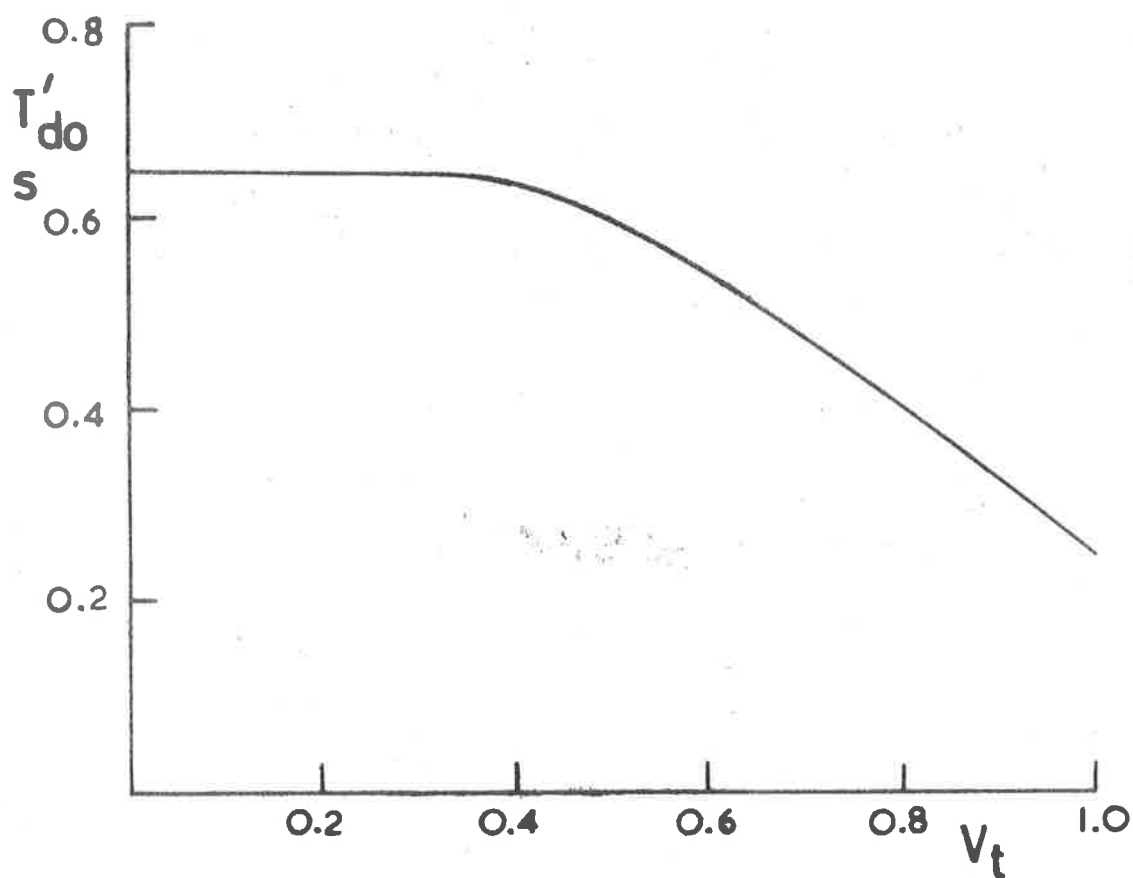


Figure 4.9 The variation of the open circuit field time constant with terminal voltage

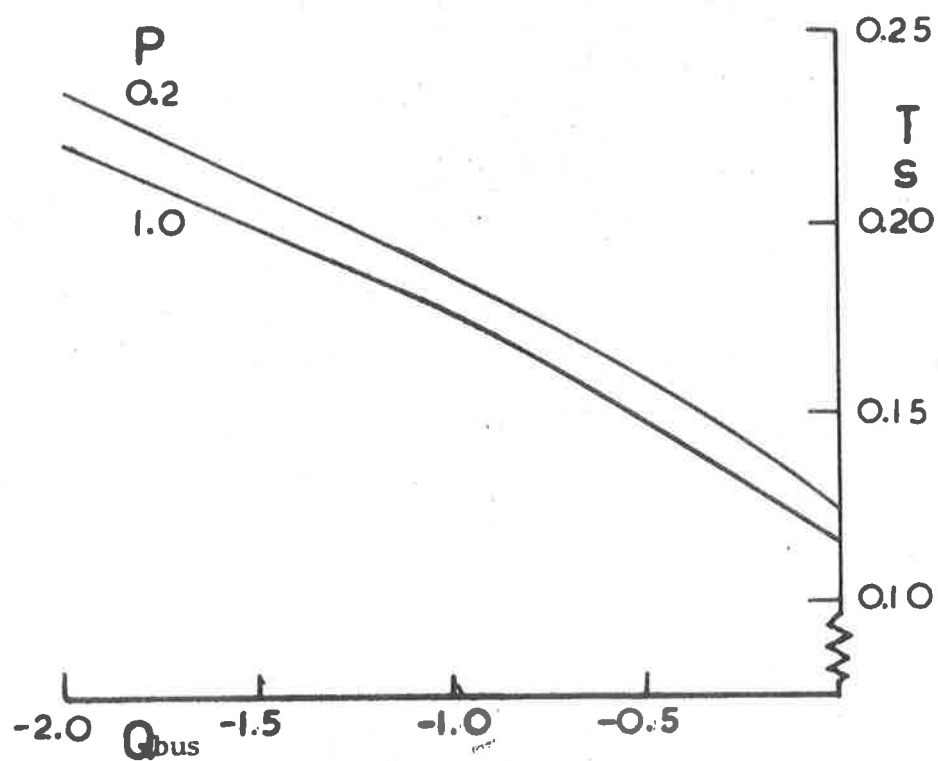


Figure 4.10 Variation of T with the PQ operating point.

4.3.3 THE DAMPING CONSTANT

The mechanical equation of motion for the system under small perturbations is given by equation (3.16):

$$-\Delta C_e = 100 \pi D \Delta \omega + 100 \pi M_p \Delta \omega \quad (4.9)$$

When the alternator is not connected to the infinite bus the effect on the torque of small perturbations in speed may be determined under steady state conditions (implying $p\Delta\omega = 0$). Thus the damping coefficient D is given by

$$D = \frac{-\Delta C_e}{100\pi\Delta\omega} \quad (4.10)$$

Figure 4.2, which is a plot of prime mover power output against speed, was obtained under steady state conditions by varying a resistive load on the alternator. As the alternator has been assumed lossless this shows that the electrical power from the alternator is constant with speed. (Under static or slowly varying conditions). Thus the following equations apply:

$$\begin{aligned} D &= -\Delta C_e / 100\pi \Delta\omega \\ &= \left(\frac{P_1}{\omega_1} - \frac{P_2}{\omega_2} \right) / 100\pi (\omega_1 - \omega_2) \\ &= \frac{-P(\omega_2 - \omega_1)}{\omega_1 \omega_2 100\pi (\omega_1 - \omega_2)} \quad (P_1 = P_2 = P) \\ &= \frac{P}{\omega_1 \omega_2 100 \pi} \end{aligned} \quad (4.11)$$

If speed deviations are small then ω_1 and ω_2 can be assumed equal to 1 per unit and (4.11) reduces to $D = P/100\pi$. If ω is measured in electrical radians per second then the damping coefficient at 1 per unit power would be unity, corresponding to the speed torque characteristic of a turbine. Thus the D.C. motor with a large armature resistance inserted does approximate the characteristic of a turbine at least at one per unit power.

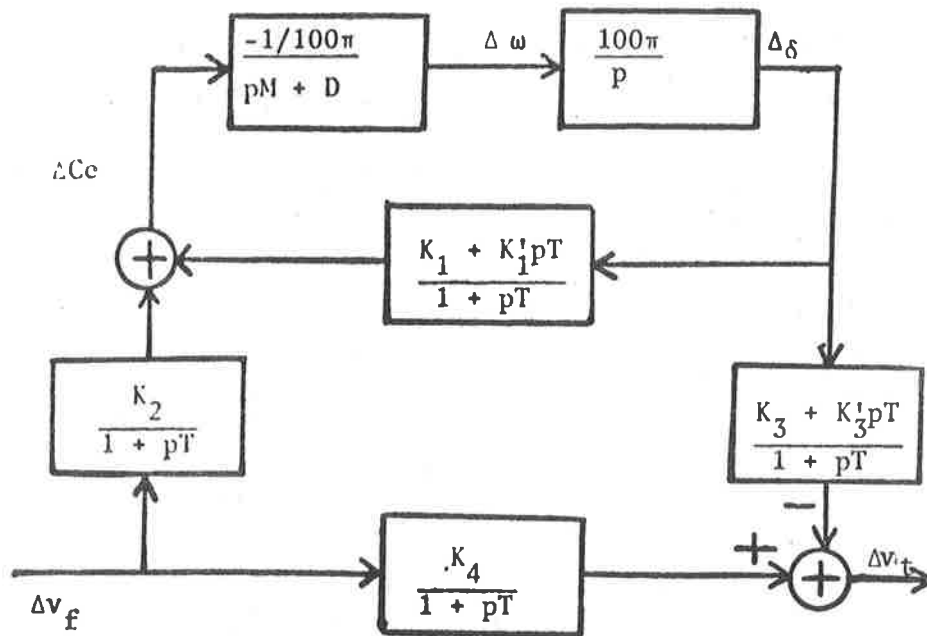


Figure 4.11 The structure of the mathematical model.

Determination of the damping coefficient under steady state conditions as has been done here, ignores the contribution of any dynamic damping torques to the total system damping. As the synchronous machine has no amortisseur circuits, this contribution is considered negligible compared with that arising from the prime mover characteristics.

4.3.4 THE INERTIA CONSTANT

Retardation tests are used to determine the polar moment of inertia J_i of the machine set as 0.45 kg M^2 . The per unit inertia constant H is defined as:

$$\begin{aligned} H &= \frac{\text{stored energy in kWs}}{\text{rating in kVA}} \\ &= \frac{\frac{1}{2} J_i \omega_s^2}{P_b} \\ &= 1.1 \end{aligned} \tag{4.12}$$

Thus the inertia constant M of equation (3.16) is $M = H/\pi f_s = 0.007$

4.4 EVALUATION OF THE K PARAMETERS

The K parameters are defined by equation 3.6 as functions of the steady state operating point of the SGTL system in the PQ plane, the machine and tie line reactances applicable at that point, and the infinite bus voltage. The mathematical model of the system is expressed in the K parameter form of figure 3.4 (figure 4.11) rather than proceeding directly from the linearized Park's equations to a state space form, because the K parameters can be estimated from direct measurement of the system variables v_f , C_e , δ and v_t . Thus the accuracy of the mathematical model is checked by comparing the K values computed from equation (3.6) with the measurements made under open loop conditions.

From figure 4.11 (or equations (3.5a) and (3.5b) it is seen that under steady state conditions at synchronous speed the K parameters may be defined as follows:

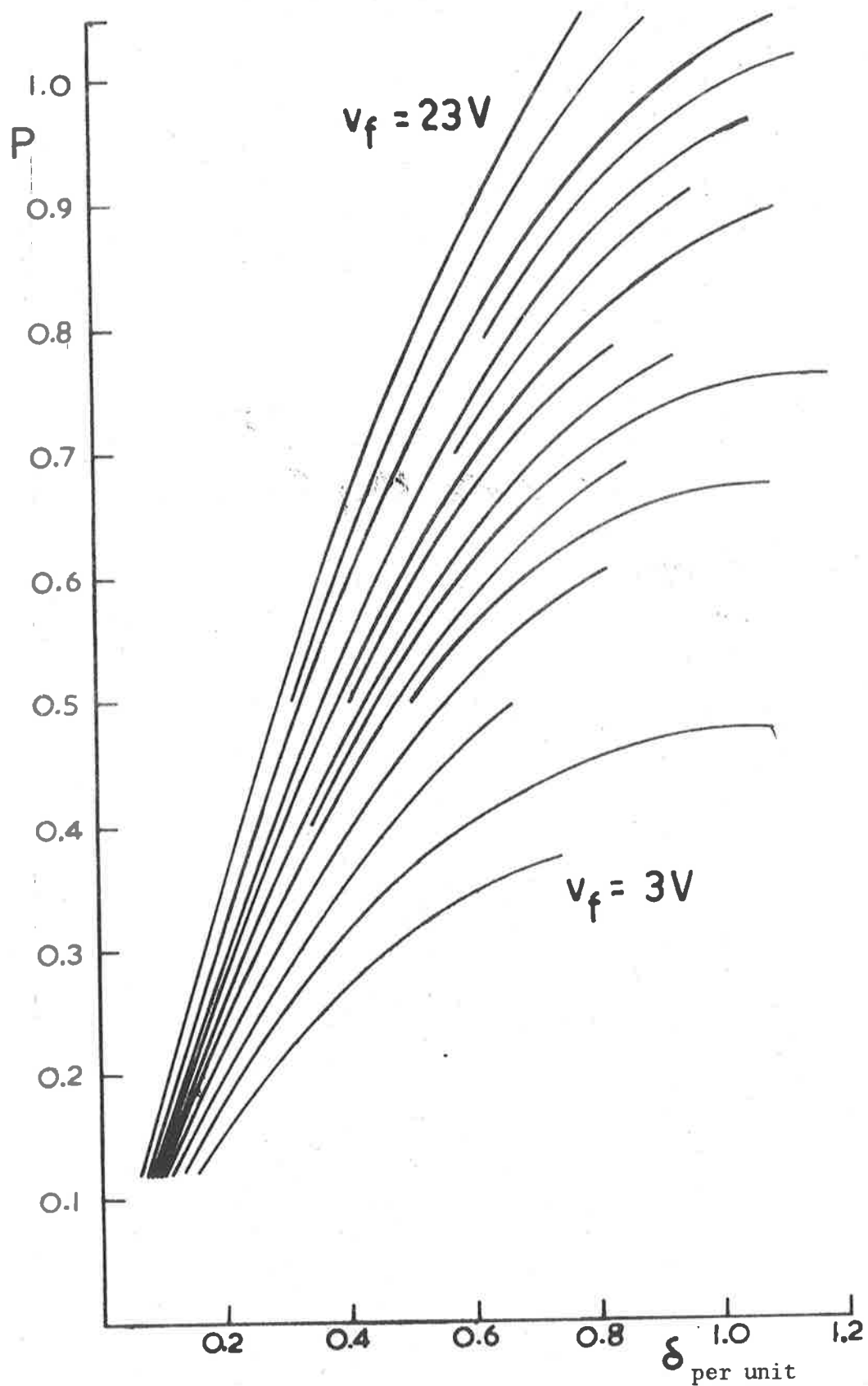


Figure 4.12 The static curve power against rotor angle for a range of constant field voltages at synchronous speed.

$$\begin{aligned}
K_1 &= \frac{\Delta C_e}{\Delta \delta} / \Delta v_f = 0 & = \frac{\Delta P_e}{\Delta \delta} / \Delta v_f &= 0 \\
K_2 &= \frac{\Delta C_c}{\Delta v_f} / \Delta \delta = 0 & = \frac{\Delta P_e}{\Delta \delta} / \Delta \delta &= 0 \\
K_3 &= \frac{\Delta v_t}{\Delta \delta} / \Delta v_f = 0 \\
K_4 &= \frac{\Delta v_t}{\Delta v_f} / \Delta \delta = 0
\end{aligned} \tag{4.13}$$

Under steady state conditions at one per unit speed, $\Delta \omega$ is zero and torque perturbations are equal to power perturbations, i.e. $\Delta C_e = \Delta P_e$ where P_e is the generated power.

Thus the K parameters are the slopes of the following static characteristics:

- K_1 = the slope of (P, δ) at constant field voltage
- K_2 = the slope of (P, v_f) at constant power angle
- K_3 = the slope of (v_t, δ) at constant field voltage
- K_4 = the slope of (v_t, v_f) at constant power angle

These parameters are estimated as follows.

With the SGTL system under manual excitation control the real power P is set to a value P_1 . The field voltage v_f is varied through a range of discrete values v_{f1} to v_{f19} and the values of P , v_f , terminal voltage v_t , rotor angle δ and the reactive power delivered to the infinite bus (Q) are recorded. This procedure is repeated for a range of P of 0.12 p.u. (the minimum value possible with the armature resistance of the D.C. motor at 30 ohms), 0.2 to 1.0 p.u. in steps of 0.1 and at 1.05 p.u. The values of v_{f1} to v_{f19} are 3 to 19 volts in 1 volt increments and 21 and 23 volts. Records of P against δ and v_t against δ for constant v_f are shown in figures 4.12 and 4.13 respectively.

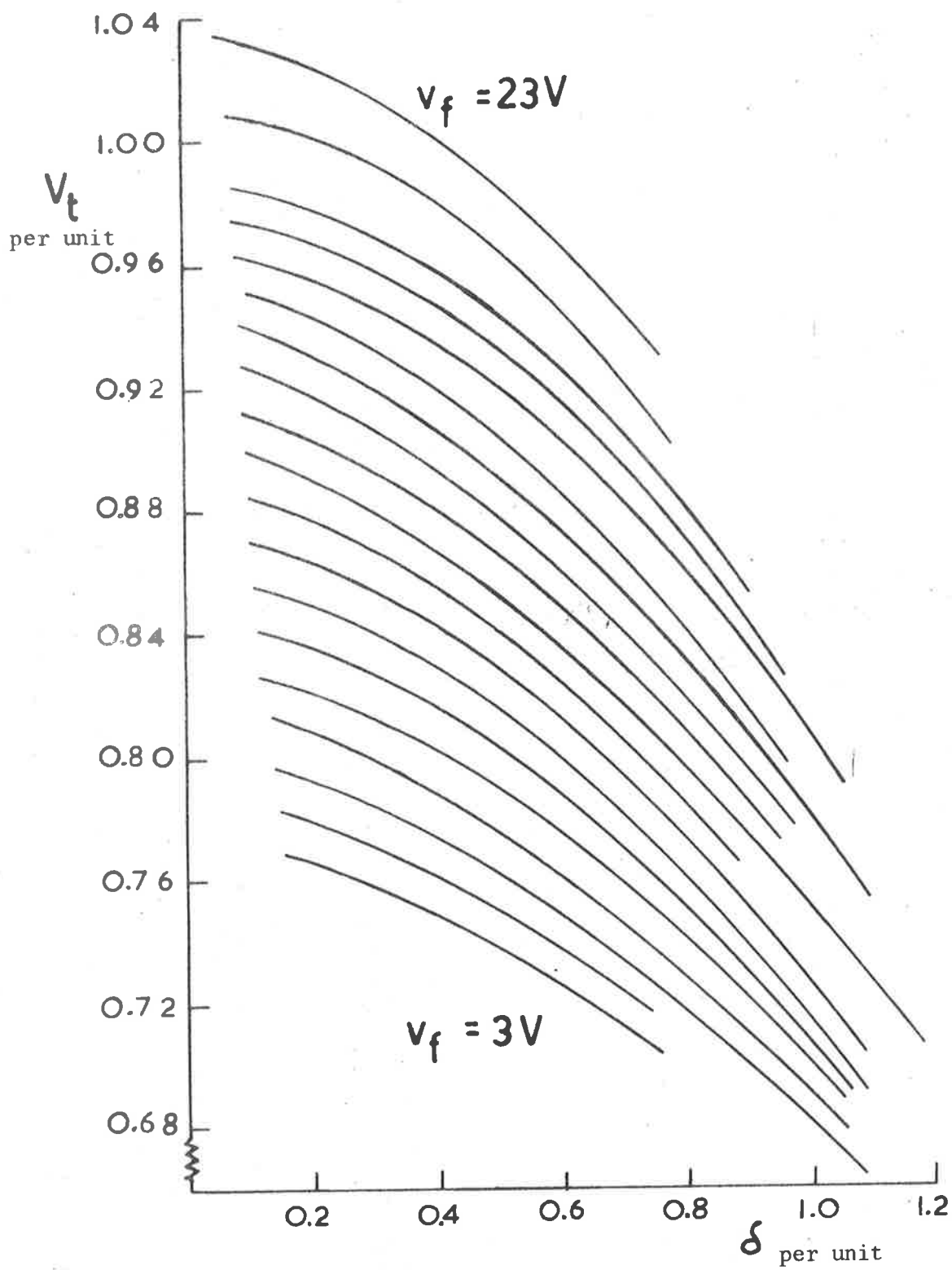


Figure 4.13 The static curve of terminal voltage against rotor angle for a range of constant field voltages.

The recorded data is transferred to a large computer and the coefficients of best fit cubic polynomials found which minimize the squared error between the data and the polynomials. The following functions are formed; $P(\delta)$, $\delta(P)$ and $v_t(\delta)$ for each of the 19 values of constant v_f .

From the series of polynomials $P(\delta)$ at constant v_f best fit cubics $P(v_f)$ for a range of constant δ are obtained. The functions $v_t(v_f)$ at constant δ are found similarly.

From the recorded data a final series of polynomials $Q(\delta)$ at constant P is obtained.

The K parameters at any particular operating point are found from these curves. For example if K_1 at $P = P_1$ is required the procedure is:

From $\delta(P)$ at constant v_f find $\delta_1 = \delta(P_1)$ at $v_f = v_{f1}$. From $P(\delta)$, $K_1 = \frac{\partial P}{\partial \delta} / v_f = v_{f1}$. From $Q(\delta)$ at $P = P_1$ find the appropriate value of $Q_1 = Q(\delta_1)$. Thus the value K_1 at (P_1, Q_1) is found. By using the curves $P(\delta)$ and $\delta(P)$ for different values of constant field voltage the variation of K_1 with Q at $P = P_1$ is found.

K_2 , K_3 and K_4 are determined similarly.

It would be possible to estimate the parameters directly from the plotted curves. For example from figure 4.12 the slope of the curve $P \text{ Vs } \delta (=K_1)$ is found for some value of v_f . From a plot of Q against δ for constant P the appropriate value of Q is found. However the accuracy achieved when determining the slope of a plotted curve is doubtful and the procedure adopted was considered necessary to allow a reliable estimate to be made.

In section 3.2.1 the K parameters were defined as a function of the system operating point in the PQ plant and the basic machine reactances (equation 3.6), these definitions arising from the linearization of Park's equations. These parameters can therefore be evaluated from these basic equations or by using the direct estimation procedure detailed above. In calculating the K values using equation 3.6, the terminal voltage is computed

for each operating point and the values for the direct and quadrature axis reactances found from the curves in figures 4.5 and 4.6. This allows for the effects of saturation on these parameters.

Figures 4.14, 4.15, 4.16 and 4.17 show the variation of these parameters with the operating point in the PQ plane, calculated over a region of leading reactive power to -2.0 per unit and real power to 1.0 p.u. Also shown are values measured using the techniques described above, these being obtainable only up to the open loop stability limit of the system (see figure 3.5), as measurements were made with the system under manual excitation control. Close agreement is observed between the calculated and experimentally derived values of each of these parameters.

K'_1 and K'_3 can be estimated as:

$$\begin{aligned} K'_1 &= \frac{\Delta P}{\Delta \delta} / \psi_f = \text{const.} \\ K'_3 &= \frac{\Delta v_t}{\Delta \delta} / \psi_f = \text{const.} \end{aligned} \tag{4.14}$$

However, because the expressions for K'_1 and K'_3 differ from K_1 and K_3 only in the use of x'_d for x_d , and computed variations in K'_1 and K'_3 with x'_d are slight, experimental determination of the K'_1 and K'_3 parameters was not considered necessary.

Since good agreement between the experimentally derived and calculated values of the K parameters is obtained, the mathematical model is considered to be a satisfactory representation of the open loop system under steady state conditions.

4.5 THE OPTIMAL GAINS

The feedback gains F_1 , F_2 and F_3 and the forward multiplier F_{in} of figure 4.18 are evaluated off line at a series of steady state operating points using the 6400 computer complex at the University of Adelaide. The gains are computed using the procedure in section 3.4 and besides being printed for reference are punched onto paper tape in a format which allows

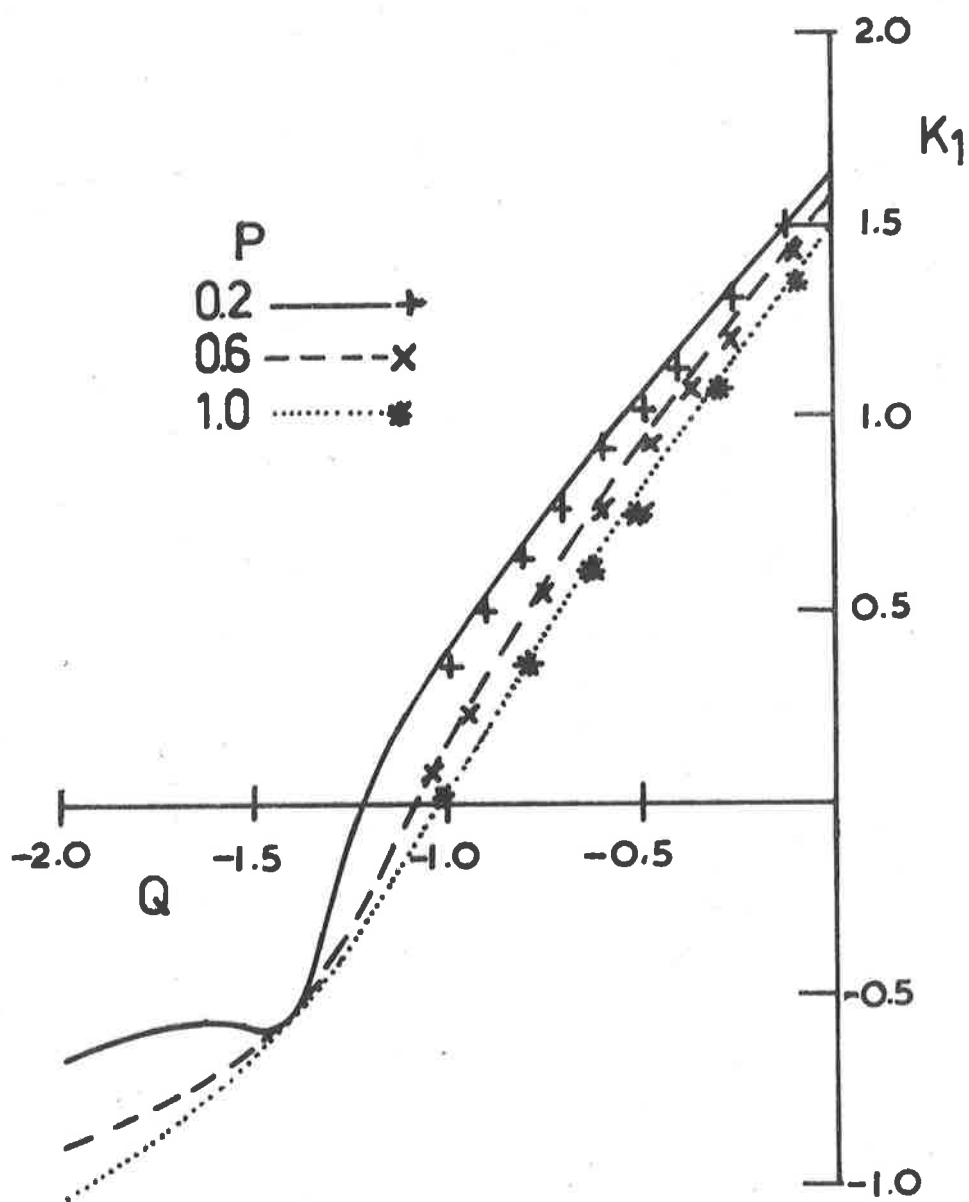


Figure 4.14 The variation of the parameter K_1 with steady state operating point in the PQ power plane. Lines are computed values, discrete points experimentally estimated.

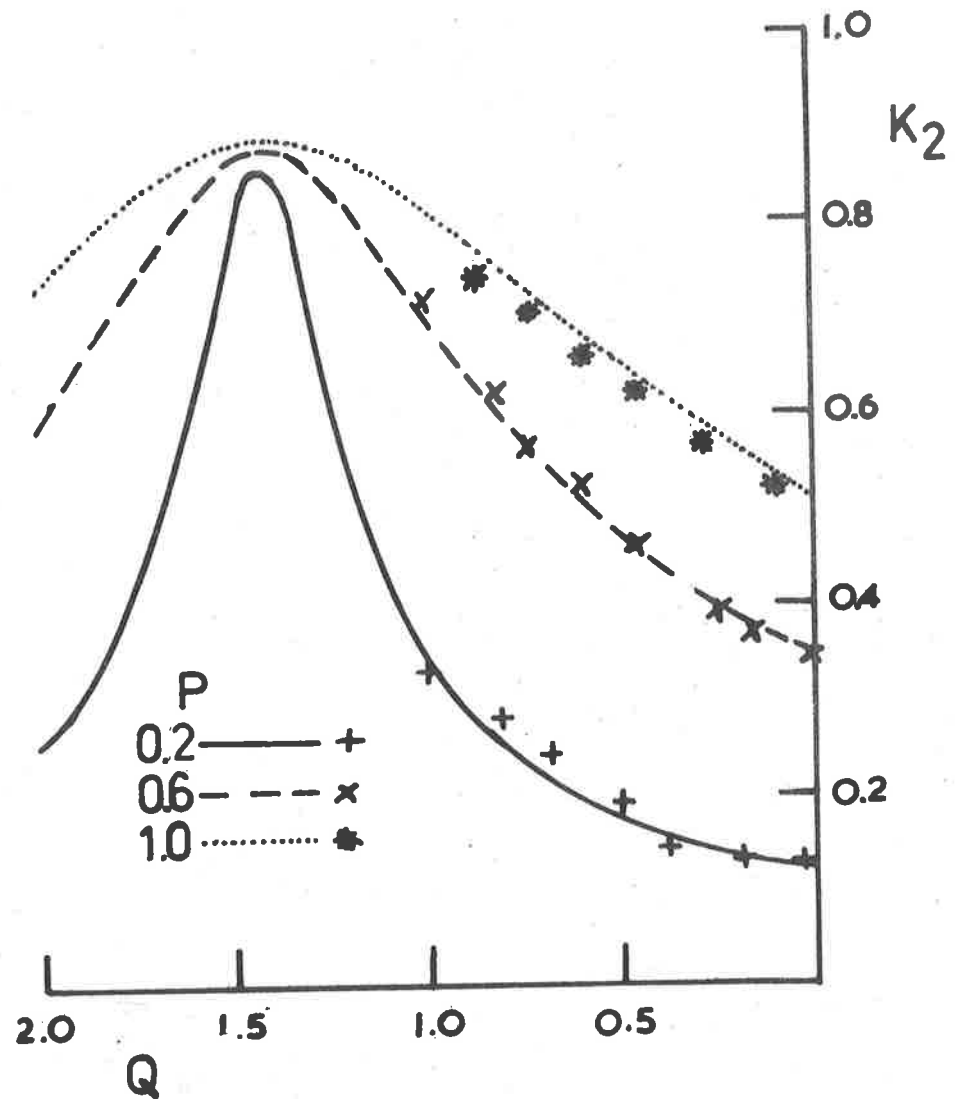


Figure 4.15 The variation of the parameter K_2 with steady state operating point in the PQ power plane. Lines are computed values, discrete points experimentally estimated.

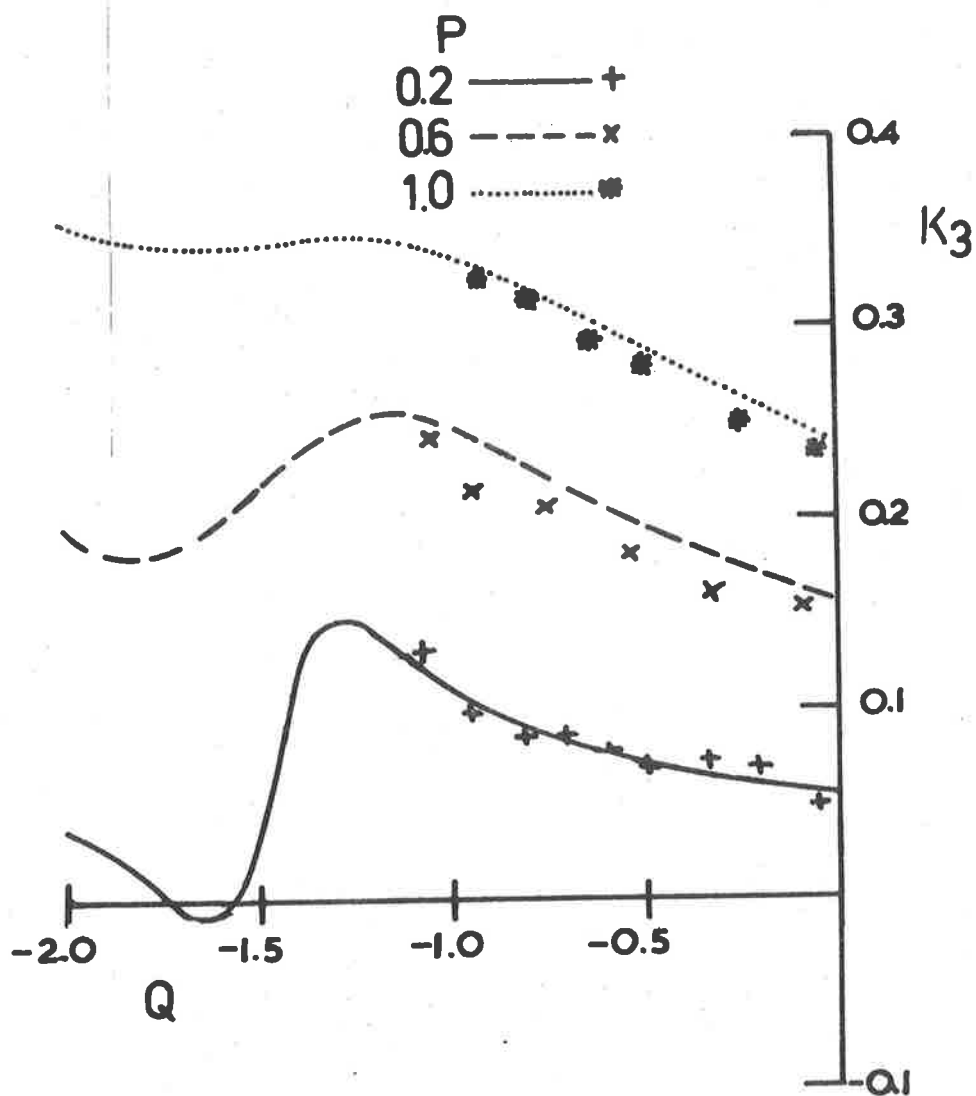


Figure 4.16 The variation of the parameter K_3 with steady state operating point in the PQ power plane. Lines are computed values, discrete points experimentally estimated.

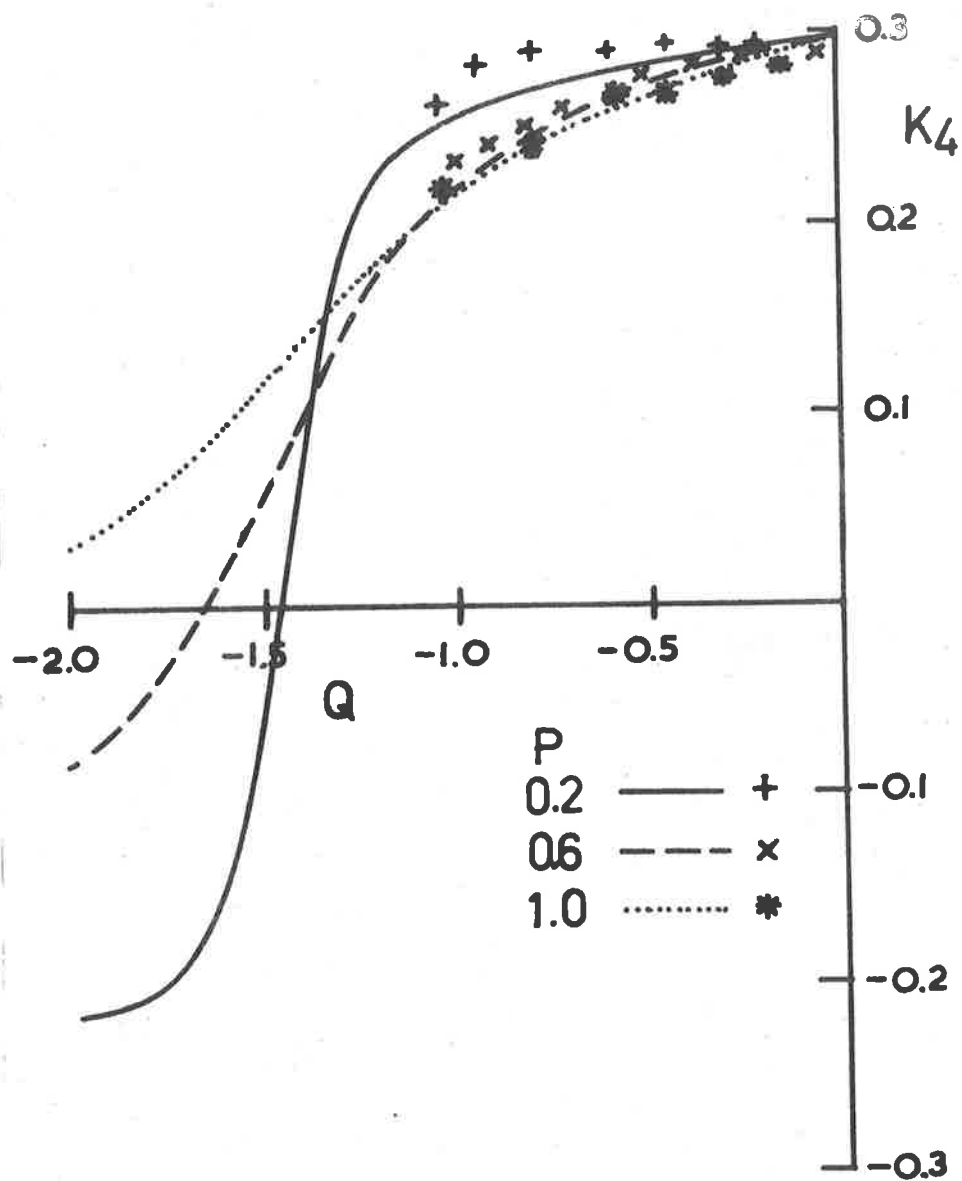


Figure 4.17 The variation of the parameter K_4 with steady state operating point in the PQ plane. Lines are computed values, discrete points experimentally estimated.

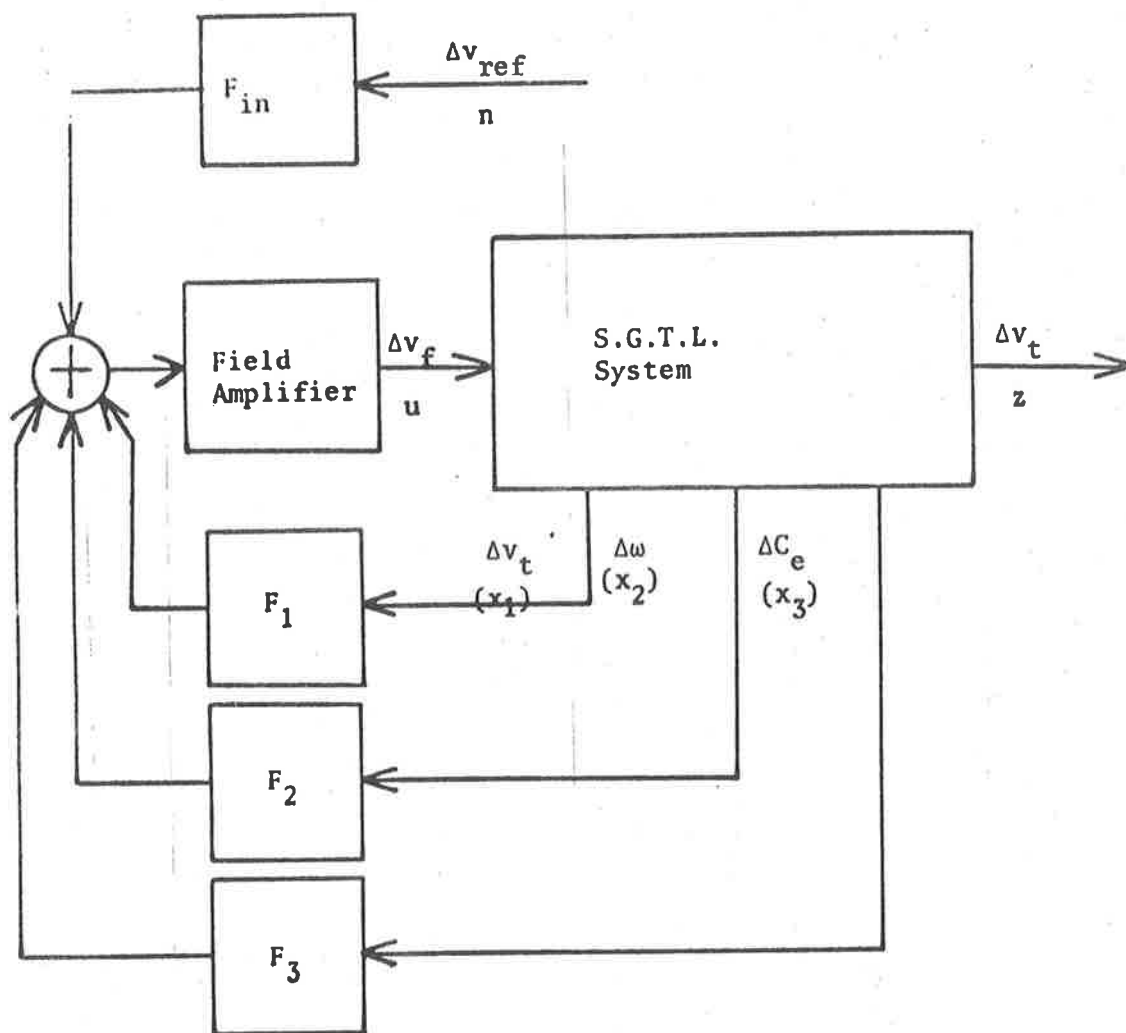
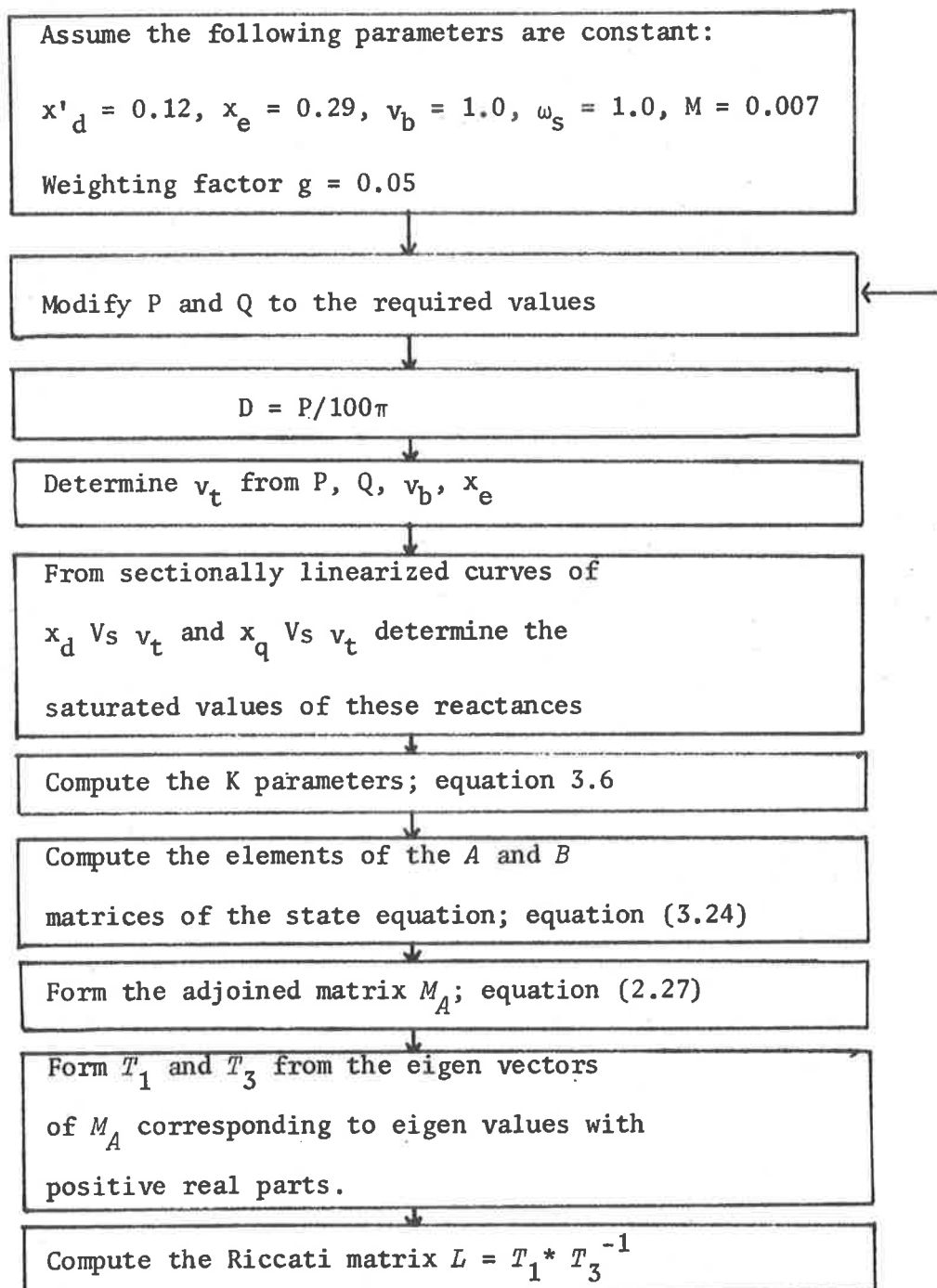


Figure 4.18 The closed loop controller for the SGT.L system.

them to be read directly into the NOVA controller (see section 6.3.1).

In addition to the optimal gains, the coefficients and the pole-zero locations of the transfer function of perturbed reference voltage to terminal voltage deviation are computed.

The outline of this program is as follows:



XXXXXXXXXX PWR= .60 QBUS= -.80 XXXXXXXXXXXX

K1= 4.94E-01 K1D= 1.06E+00 K2= 5.84E-01
K3= 2.28E-01 K3D= -1.25E-02 K4= 2.49E-01

TERMINAL VOLTAGE= .764 LINE CURRENT= 1.020

ADJOINED MATRIX M

-2.79E+00	0.	-6.56E+00	1.00E+00	0.	0.
3.93E+00	-3.15E-01	3.33E+02	0.	0.	0.
-1.28E+00	-4.55E-01	-3.02E+00	0.	0.	0.
4.17E+01	0.	9.80E+01	2.79E+00	-3.93E+00	1.28E+00
0.	0.	0.	-0.	3.15E-01	4.55E-01
9.80E+01	0.	2.31E+02	6.56E+00	-3.33E+02	3.02E+00

RICATTI MATRIX

1.17E-01	-2.44E-01	-6.39E-03
-2.44E-01	1.63E+00	7.62E-03
-6.39E-03	7.62E-03	1.52E-03

FEEDBACK GAINS

F1=-2.940E+00 F2= 6.524E+00 F3= 8.168E-02 ETA= 4.106E+00
MODIFIED ETA= 4.871E+00

TRANSFER FUNCTION COEFFICIENTS

N2= 7.034E+00 N1= 2.215E+00 N0= 1.034E+03
D3= 1.000E+00 D2= 1.009E+01 D1= 1.644E+02 D0= 1.034E+03

ZEROS

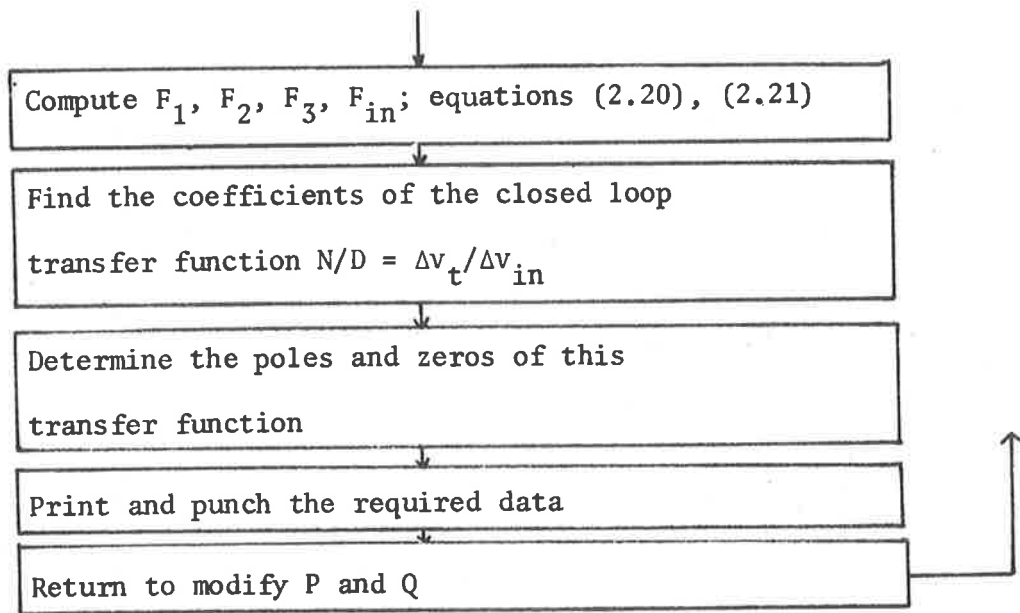
REAL PART	IMAGINARY PART
-.157429	12.124181
-.157429	-12.124181

POLES

REAL PART	IMAGINARY PART
-7.203060	0.000000
-1.444539	11.894508
-1.444539	-11.894508

STEADY STATE VOUT (UNIT STEP VREF, UNMODIFIED ETA)= .84276
UNITY OUTPUT WITH MODIFIED ETA

Figure 4.19 A typical page of computer print
out generated by the optimal gains
program.



A typical page of output from this program is shown in figure 4.19. The quantities ETA and modified ETA refer to the input multiplier F_{in} . The actual value used for F_{in} is the modified ETA. The cost function used did not specifically penalize deviations of the output from its required value at the terminal time, i.e. the weighting factor was put equal to zero, simplifying the minimization procedure. It was found that under these conditions the resulting optimal controller produced a small but significant steady state error which varied in magnitude with the steady state operating point of the system. To obtain uniformity of the steady state value of $\Delta v_t / \Delta v_{in}$ over the PQ plane, ETA was modified to force this ratio to unity throughout the region. This means that the control is no longer optimal in the sense that the chosen cost function is not minimized, although the dynamic behaviour of the system is not effected. This variation to the derived optimal controller may be regarded simply as an additional multiplier being inserted at the input prior to the optimal gain ETA. The product of these input gains is the value used for F_{in} in the implementation of the controller. For purposes of identification the computed gains will still be referred to as the optimal gains although the chosen cost function is not minimized.

Evaluation of the gains for the linear model at a series of steady state operating points over the complex power plane shows that these gains vary appreciably. This variation is shown in figures 4.20, 4.21, 4.22 and 4.23 where the gains are computed for various values of constant real power and a range of reactive power.

4.6 SUMMARY

While the effect of saturation is ignored in the derivation of the equations for the mathematical model, by plotting the reactances and time constants against terminal voltage it can be allowed for in computing the parameters of the steady state model.

The K parameters and the optimal gains are computed for the linear model at a series of steady state operating points, and the variation of these quantities demonstrates the non linear nature of the SGT system.

The accuracy of the open loop model is checked in part by computing the K parameters and comparing them with experimentally derived values. Close correlation is observed implying that the evaluated parameters and the mathematical model provide a satisfactory representation of the system.

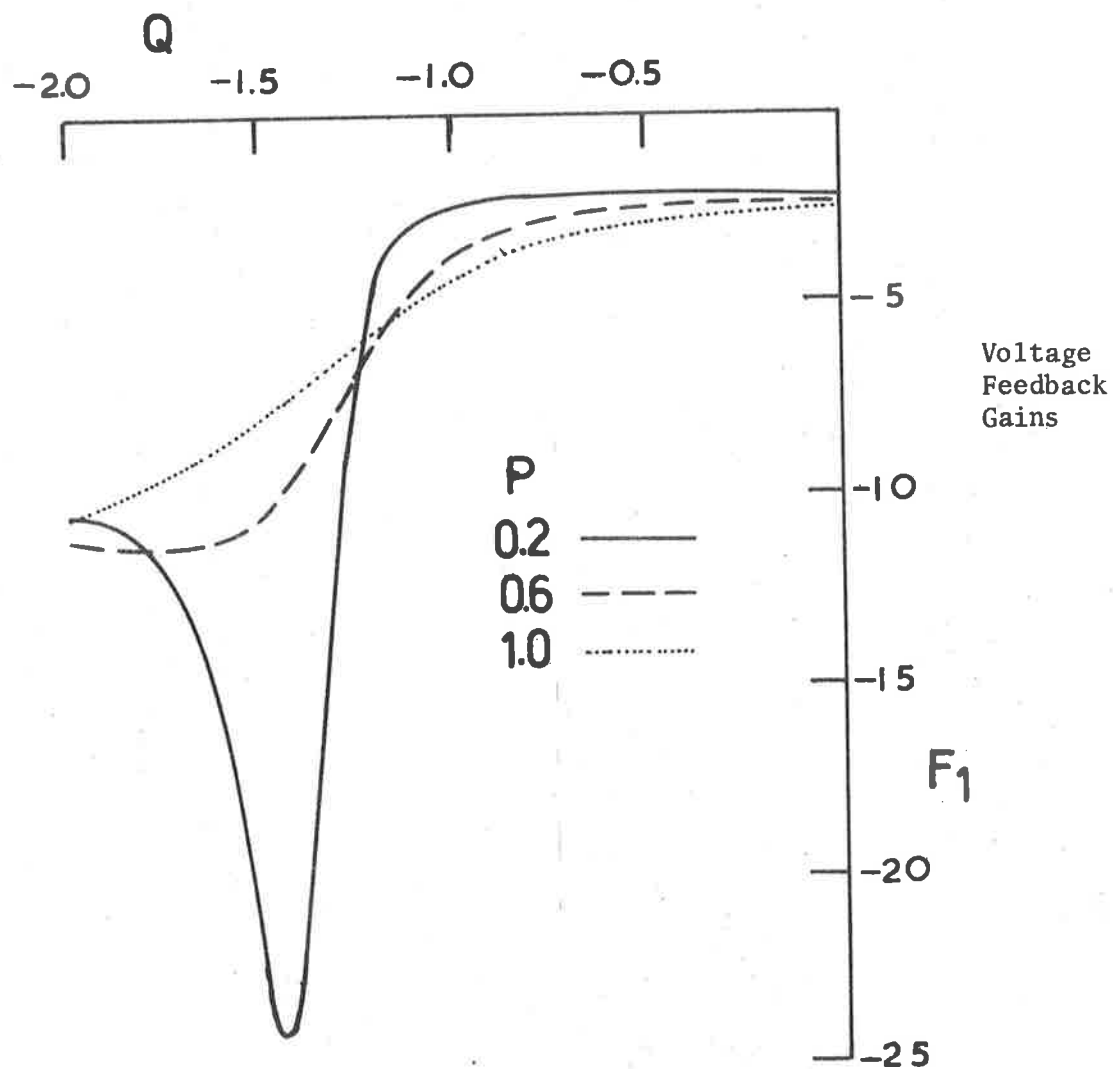


Figure 4.20 The variation of the optimal feedback gain F_1 with steady state operating point.

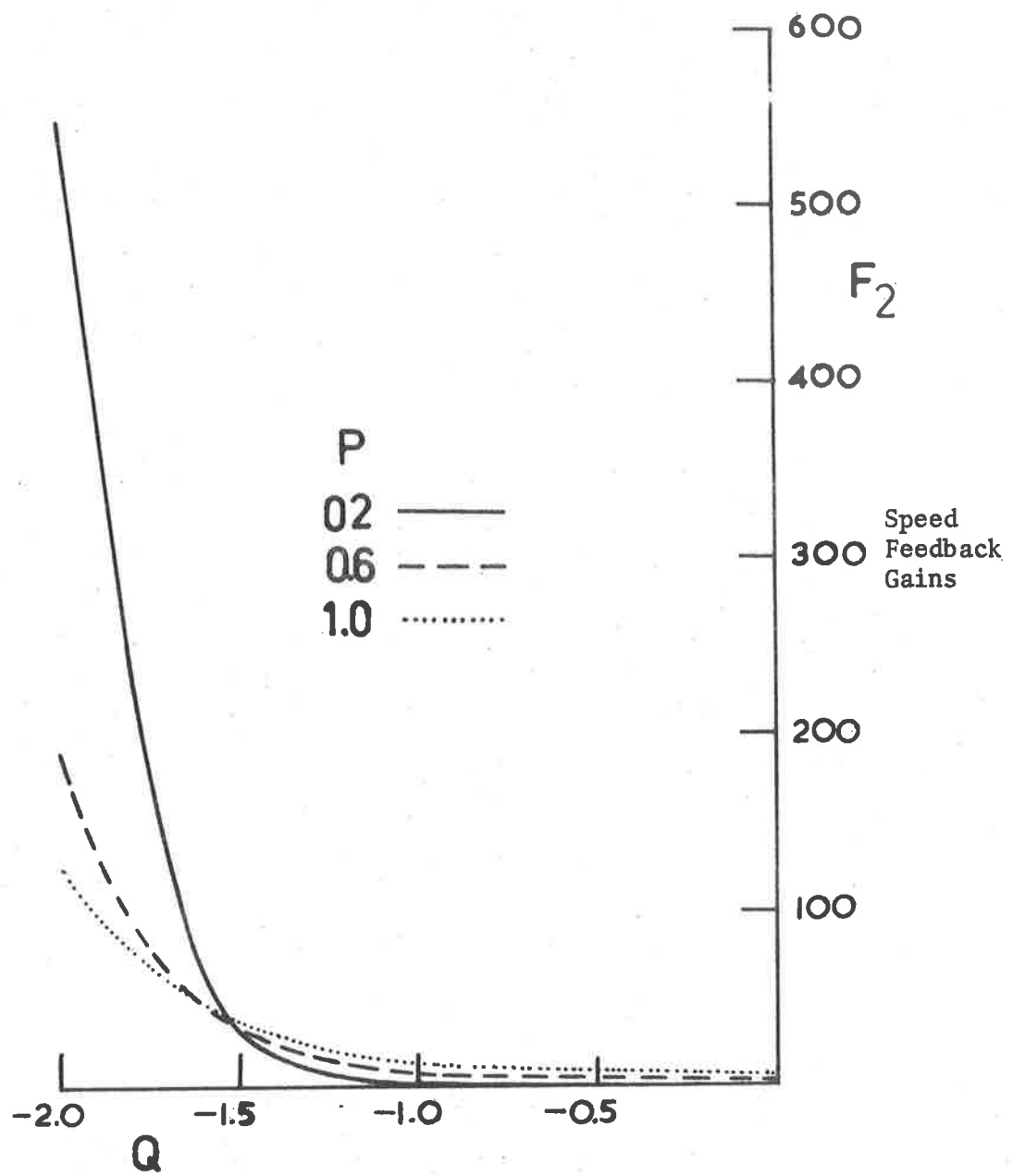


Figure 4.21 The variation of the optimal feedback gain F_2 with steady state operating point.

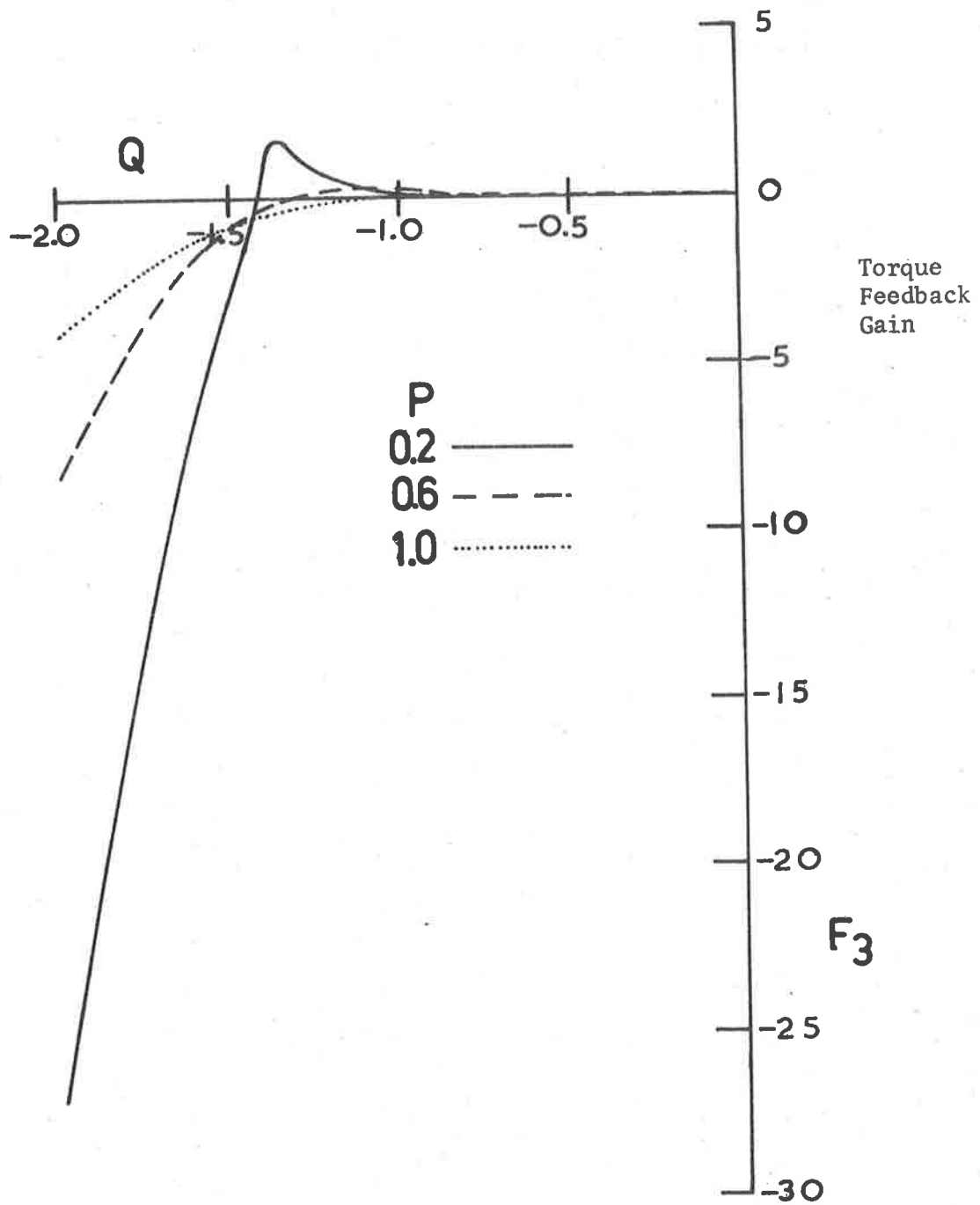


Figure 4.22 The variation of the optimal feedback gain F_3 with steady state operating point.

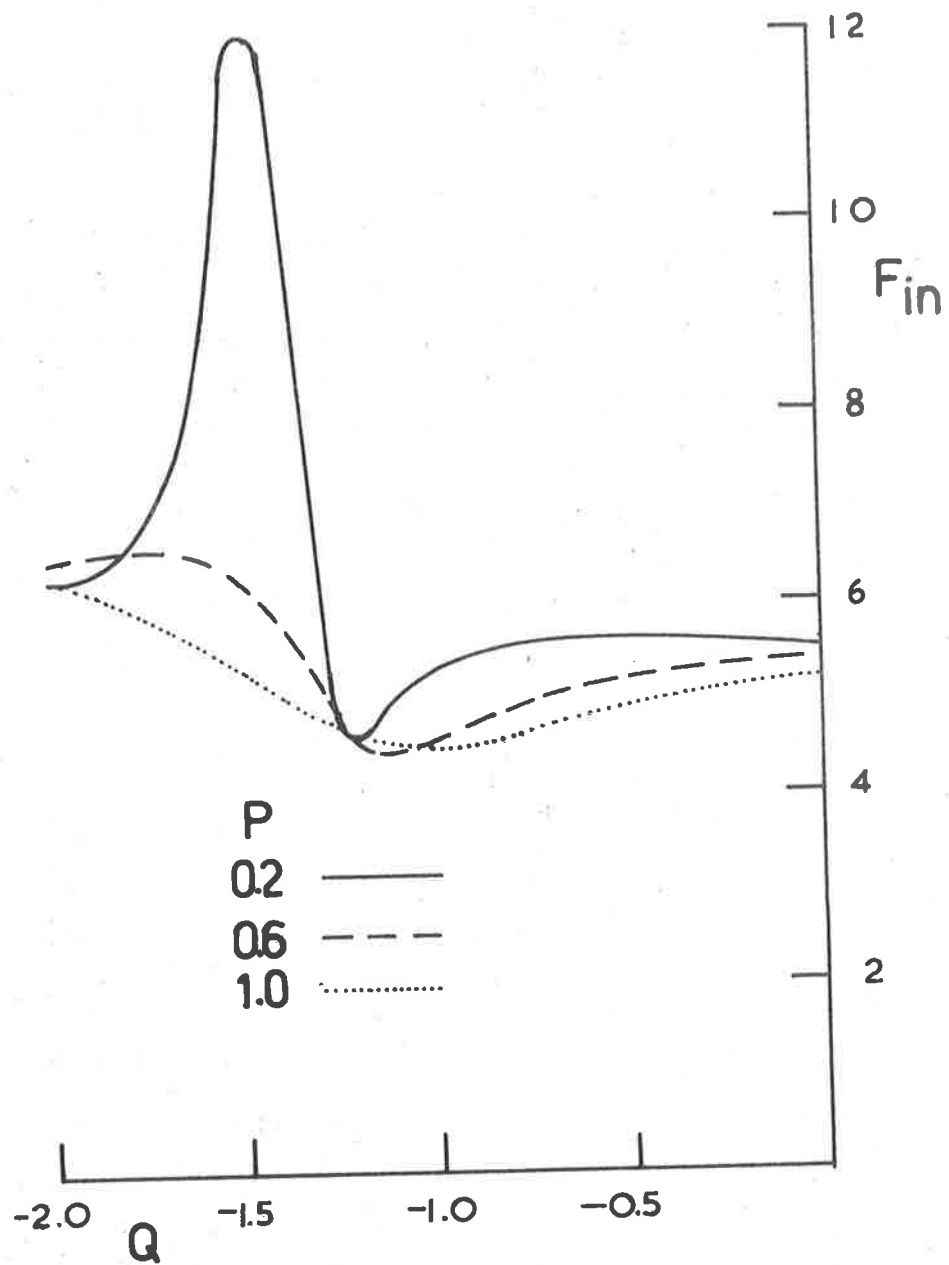


Figure 4.23 The variation of the modified input multiplier with the steady state operating point.

CHAPTER 5

THE CONTROLLER, TRANSDUCERS AND ANALOGUE INTERFACE

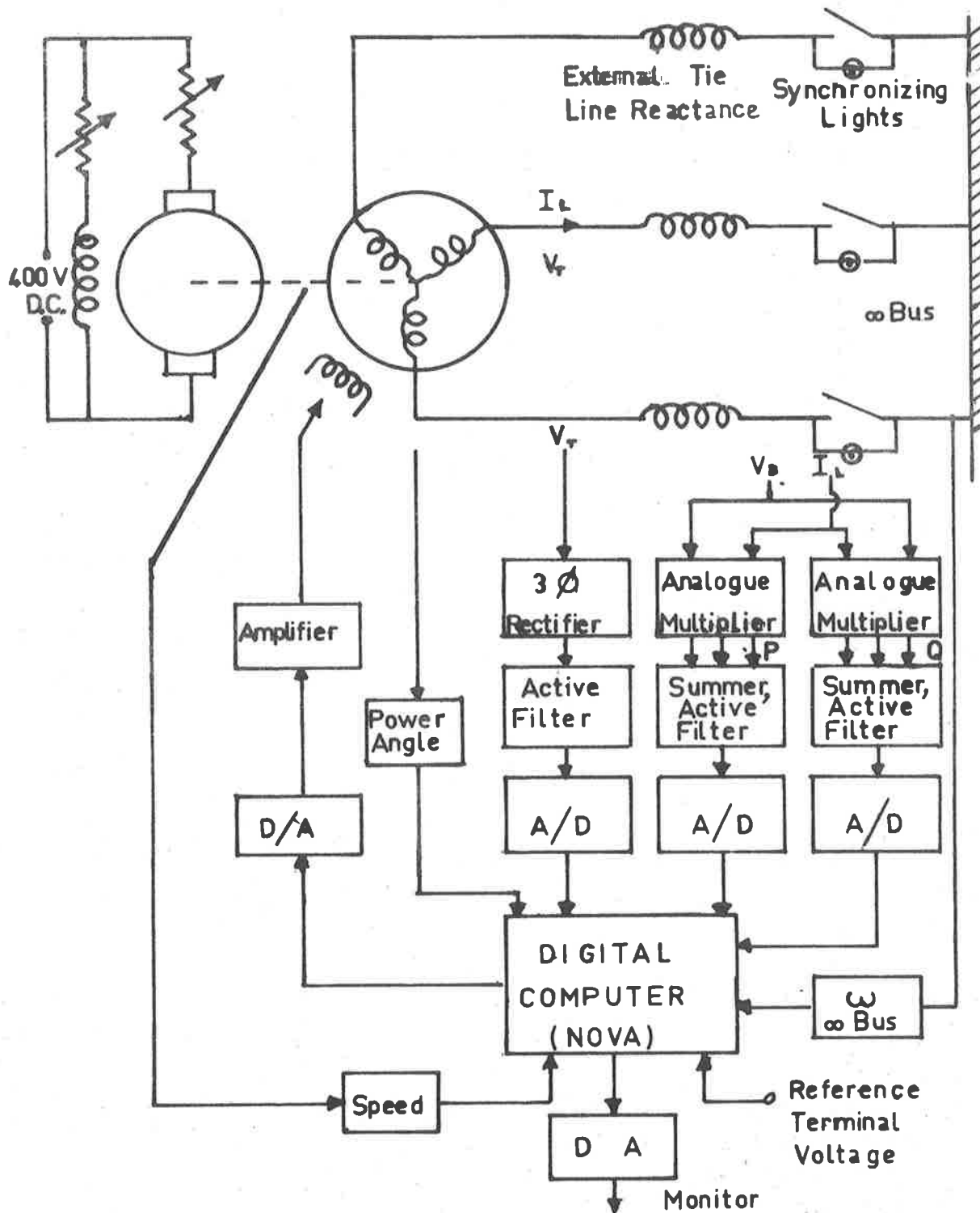


Figure 5.1 The SGTL system with the NOVA controller and transducers.

5.1 INTRODUCTION

Whenever a digital computer is used to monitor or control an analogue system, an interface must be provided for the transfer of information between each section. Transducers may be used which produce either digital or analogue outputs. Digital outputs can be passed directly to the computer, while analogue signals must undergo a conversion to an equivalent digital value before processing. In this study a digital computer is used as the controller for the SCTL system. Various transducers and an A/D interface were developed to form the required link. This chapter outlines the equipment produced with detailed design considerations appearing in Appendix A.

Figure 5.1 is a schematic of the experimental model, showing the SCTL and the controller with its transducers and interface. The real and reactive power transfer at the infinite bus is measured by analogue transducers. The alternator terminal voltage is converted to a proportional D.C. value by a three phase rectifier and an active filter. The outputs of these transducers are sampled by the computer via the A/D interface. The computed control force is applied through a D/A converter to an electronic amplifier exciting the alternator field.

Digital transducers are used to monitor the machine speed and system power angle δ . Although it is not used in the control of the system, δ is measured to permit the direct estimation of the K parameters as shown in section 4.4.

The transducers described in this chapter were developed to perform a specific function, although the analogue to digital interface was designed as a general purpose unit. Although not shown in figure 5.1, all units directly coupled to the computer are under its program's control.

5.2 THE NOVA COMPUTER

A NOVA mini computer is used as the controller. This is a 16 bit word length computer engineered towards on line applications. The central processor has four accumulators (or arithmetic registers) two of which are

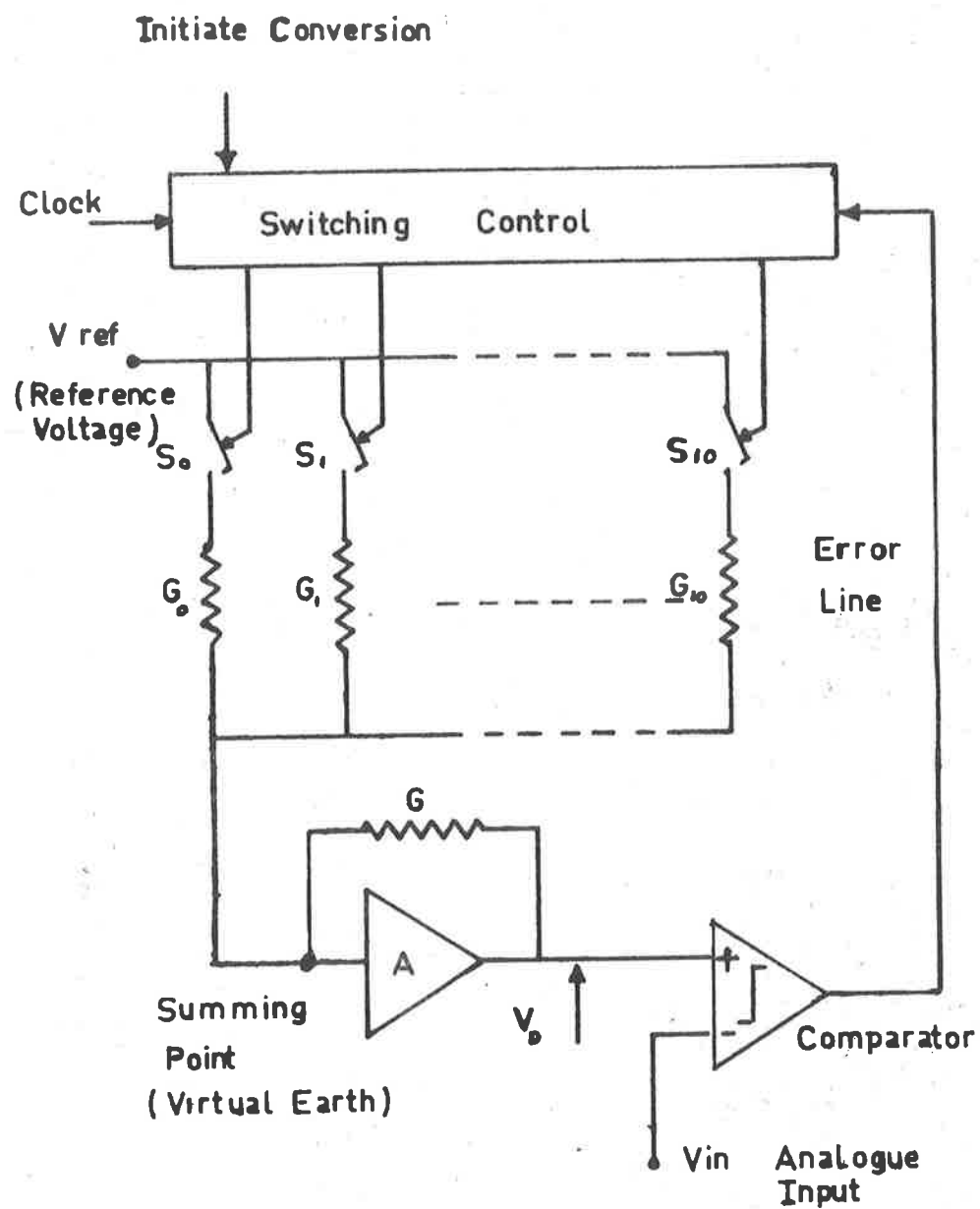


Figure 5.2 Schematic of a successive approximation A/D converter.

available as addressing index registers. The particular installation used has 12K of read-write store and a real time clock. Peripheral equipment includes tele-typewriter, high speed paper tape reader and punch, a graphic display and a hardware multiply-divide unit. All peripheral are connected to the central processor by a bidirectional input/output bus. All programming for on-line applications is in NOVA assembly language. Full details of the NOVA system and its operation are contained in the "How to use NOVA" handbook.²⁴

5.3 THE ANALOGUE TO DIGITAL CONVERTERS

In nearly all A/D interfaces the basic unit is the analogue to digital voltage converter, since most analogue systems or their transducers produce a voltage as their output, or can have their output simply converted to a proportional voltage.

Various types of A/D converters exist, and the type implemented is determined by its particular application. Because of its method of operation the successive approximation type is widely used. This particular system has a conversion time, which unlike some others is independent of the analogue input signal and is therefore more useful when samples are taken at definite time intervals, or where the input voltages cannot be regarded as slowly varying D.C.

Figure 5.2 is a schematic of a successive approximation A/D converter. Briefly, its operation is as follows. Once initiated, the switching control section sequentially closes the switches S_0 to S_{10} coincident with a master clock pulse. Once switch S_0 is closed, a voltage V_d appears at the output of the summing amplifier A, proportional to the magnitude of the conductance G_0 . If V_d is greater than V_{in} , the analogue input signal, the error signal from the comparator causes switch S_0 to be opened, otherwise S_0 is left closed. Similarly each of the binary weighted conductances G_0 to G_{10} are tried in turn,

the voltage V_d at any stage being proportional to the sum of the conductances which have their switch closed. At the completion of the conversion the setting of the switches represents in binary form (to some scale) the analogue input signal, and V_d is equal to V_{in} to within the least significant binary bit.

The switching control register of figure 5.2 must sequentially control the switches and obey commands from the comparator. A digital register was developed to perform this function, and named the rollup register because of the way it achieves the required control. A paper on this register was presented at the 12th International Radio and Electronic Conference.³³

From considerations of switching speeds and junction temperature coefficients, bilateral transistors were chosen as the switches S_0 to S_{10} . As explained in Appendix A, the transistors appear in a unique circuit configuration which overcomes the effect base-emitter currents have on the output of the summing amplifier.

As the digital register switches in the conductances a staircase waveform appears at V_d , the largest step occurring when the largest conductance G_0 is switched in. High conversion rates can only be achieved if these stepped voltages settle rapidly to their steady state value, as a comparison between V_d and V_{in} must be made before the next switch is closed. This requirement proved to be the limiting factor in the conversion rate of the units. A detailed investigation of integrated, hybrid and discrete component operational amplifiers was carried out, considering cost and performance, before a satisfactory integrated circuit was chosen.

Little difficulty was experienced in obtaining a suitable comparator.

A multichannel system is provided by six analogue to digital converters, each having the following specifications:

- (1) Bipolar input range ± 1 volt.
- (2) Digital output 10 bits + sign in two's complement.
- (3) Conversion time 11 μ s (clocking frequency 1MHz).

- (4) Resolution $\pm 1\text{mV}$ ($\pm 0.05\%$ F.S.D.)
- (5) Repeatable conversion accuracy in an operational environment
 $\pm 2\text{mV}$ ($\pm 0.1\%$ F.S.D.)
- (6) Automatic hold of converted input
- (7) Conversion restart by single pulse
- (8) Input impedance $10\text{K}\Omega$

5.4 THE DIGITAL TO ANALOGUE CONVERTERS

The D/A converters are used to transfer computed information to the analogue system. These are similar in configuration to the A/D converters, except that the comparator with its error output signal is not required. The digital input is clocked into a simple digital holding register which operates the appropriate conductance switches. These are brought together at the summing point and the amplifier output is the analogue representation of the switch settings (as occurs at the end of a conversion cycle in the A/D converters).

Four D/A converters were built with the following specifications:

- (1) Digital input 10 bits + sign in two's complement.
- (2) Bipolar output range ± 1 volt.
- (3) Resolution $\pm 1\text{mV}$ ($\pm 0.05\%$ F.S.D.)
- (4) Repeatable conversion accuracy $\pm 1\text{mV}$ ($\pm 0.05\%$ F.S.D.)
- (5) Zero order digital hold.
- (6) Dynamic output impedance 0.5Ω
- (7) Maximum output current 300mA

5.5 COMPUTER CONTROL OF THE CONVERTERS

As mentioned in section 5.2 the NOVA computer has a 16 bit bidirectional input/output (I/O) bus for data transfer between its central processor and peripherals. Connection of particular peripheral to the I/O bus is controlled by a six bit address bus. Each device has a selection net, and this is activated only when the device's code is requested by the

and subsequently applied to the address bus

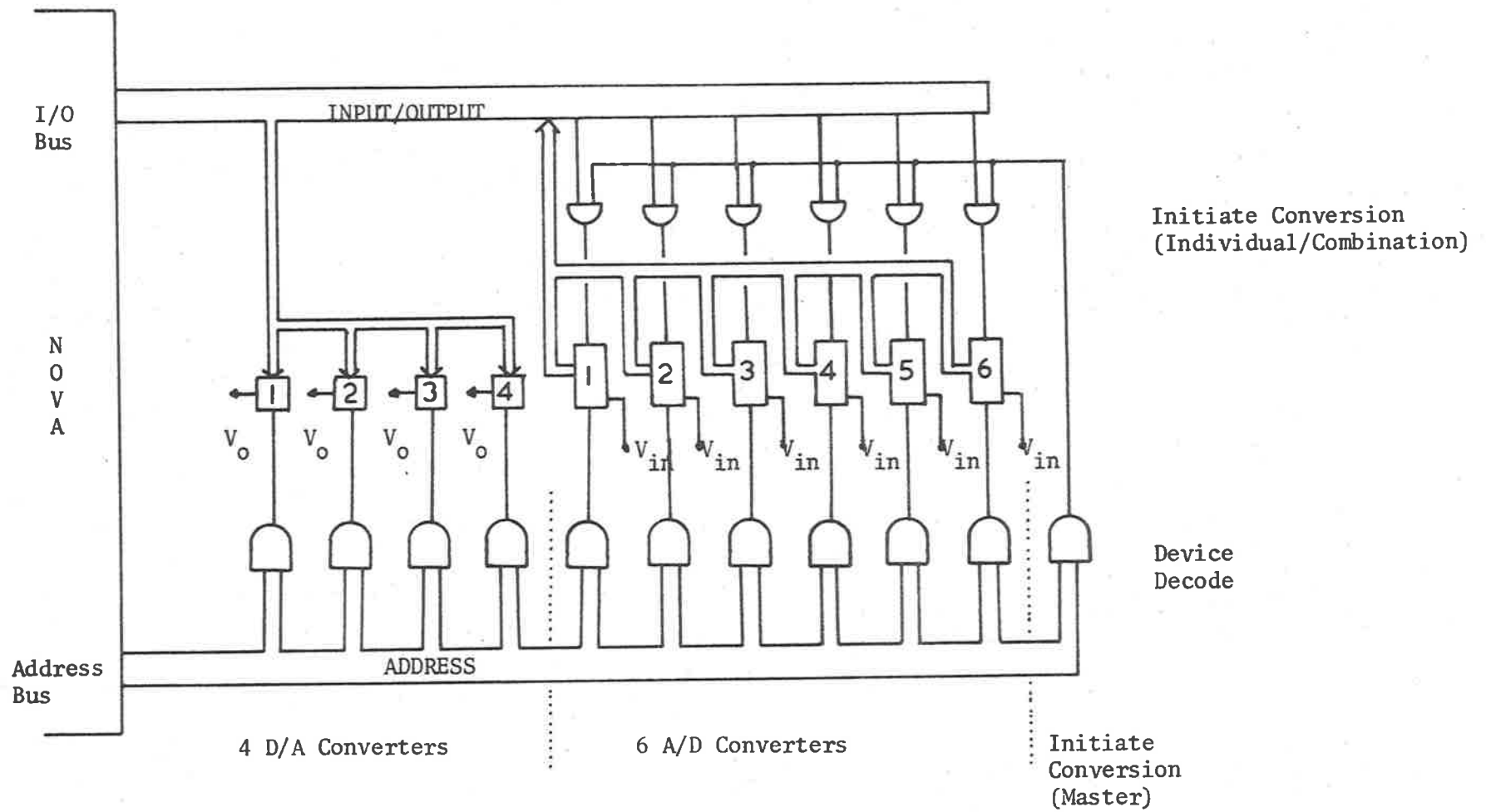


Figure 5.3 NOVA control of the converters.

In an output sequence from the NOVA, the required digital information, held in one of its accumulators, is applied to the I/O bus by the central processor under program control. The appropriate device code is then applied to the address bus. With the D/A converters the decoded address of the required unit is used to clock the digital information into its holding register (figure 5.3). An analogue output voltage representative of the binary data on the I/O bus then appears as the output of the D/A converter.

For the A/D converters an output instruction is used to initiate a conversion. Information supplied to the I/O bus by the program determines which converter is to operate. The hardware decoding this information is such that any individual converter or combination of converters may be initiated as required at any point in the program. This adds to the versatility of the system. Once the converter has completed its operation the digital information held in its register is available to the program as required. An input instruction causes the device decode network of the particular converter to place this data on the I/O bus. From there it is clocked into an accumulator for processing (figure 5.3).

5.6 THE MACHINE SPEED TRANSDUCER

Perturbations in machine speed, together with electrical torque and terminal voltage were chosen as the system states (section 3.5.1). As detailed in appendix A4, the perturbations in torque and voltage which result from an 0.04 p.u. step change in reference voltage can be measured with a resolution of 1 part in 80 of their peak response. To achieve a similar resolution in the measurement of speed perturbations, the transducer must be capable of resolving 1 part in 50,000 of one per unit speed. For transducer time constants to be neglected in system analysis, a sampling time of not more than 10ms was considered necessary. For a 50Hz 4 pole synchronous machine this means that the rotor speed must be resolved to one part in 50,000 in one quarter of a mechanical revolution. A transducer to meet these specifications was developed.

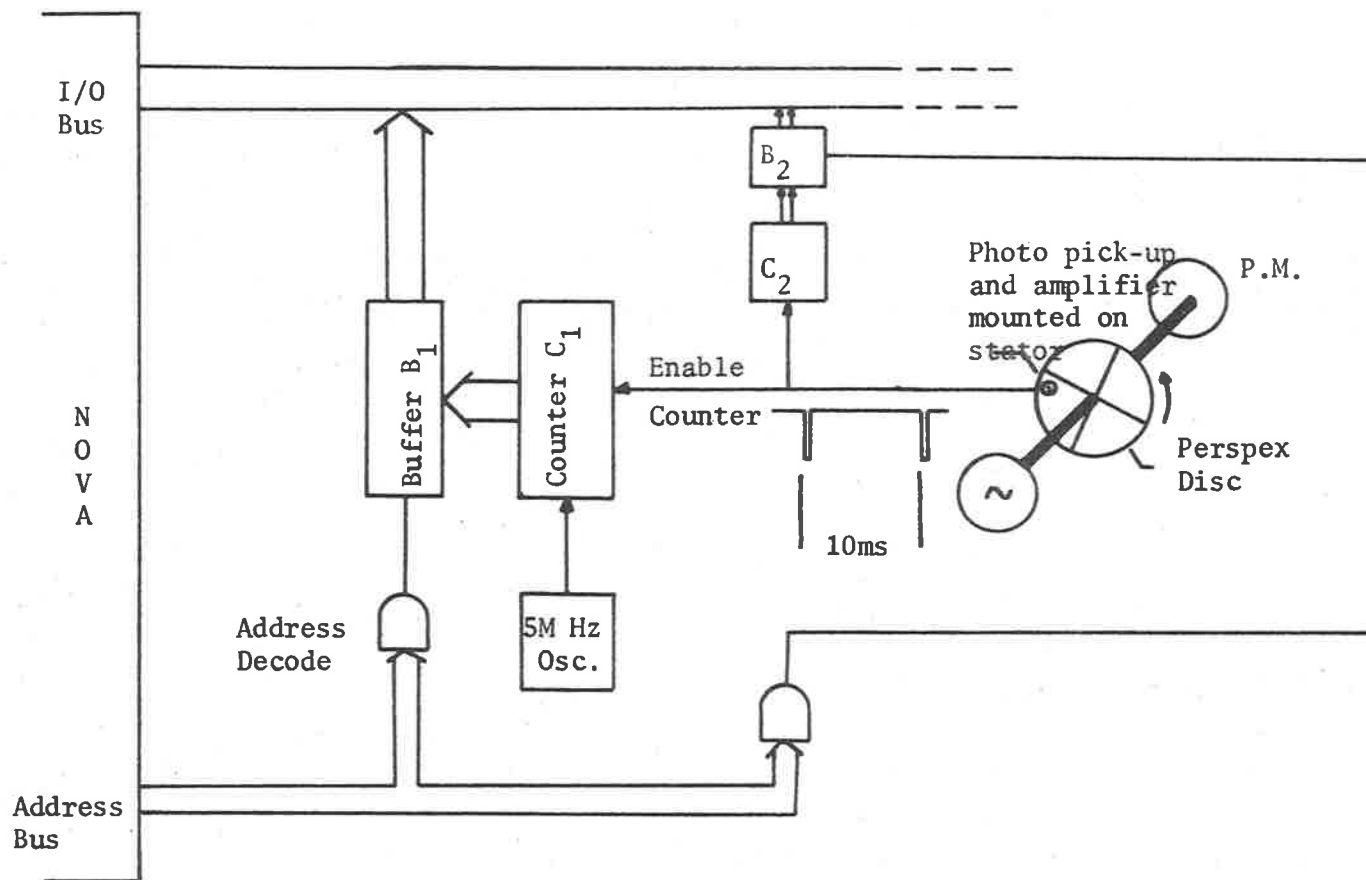


Figure 5.4 The machine speed transducer

A transducer producing an output pulse every one quarter of a revolution was mounted on the alternator shaft, and a binary counter gated between pulses to count the output of a 5MHz crystal oscillator, as shown in figure 5.4

At one per unit speed therefore, the counter would contain 50,000 after one quarter of a revolution (i.e. a 10ms gating period). A value of 49,913 would indicate a machine speed of 0.99826 p.u. The count is shifted into a buffer stage and the counter reset to zero after each quarter revolution. From the buffer it may be read into the NOVA via the I/O bus when required by the program.

The required resolution is achieved simply by having a high enough oscillator frequency. For the device to have an accuracy of 1 part in 50,000 however, the pulses from the shaft transducer must divide the rotor into quarters with precision. The shaft transducer used consists of a perspex disc, blacked out except for four radial lines, mounted between a light source and a photo sensor. The photo sensor pulses every quarter revolution. With a 6 inch diameter disc, an accuracy of 1 part in 50,000 could only be obtained if the clear lines on the disc were positioned to the nearest 0.0001 of an inch. This is assuming that the boss supporting the disc was precision machined and the alternator shaft is true running. However as deviation from synchronous speed and not the synchronous speed is used as the system state, the following technique is employed whereby the deviation is measured with a high degree of accuracy.

Another counter, marked C2 in figure 5.4, is used. This is a 2 bit ring counter. Each time a line on the disc passes the sensor this counter is incremented. Thus each quarter of the disc is effectively labelled by the value held in the counter during the time that quarter is passing the pickup. Whenever the program requires a value for the deviation of the machine

speed from one per unit, the contents of both buffers b_1 and b_2 are read into the computer. Thus the program obtains the oscillator count in a quarter revolution plus an indication of the section of the disc which enabled that count. This information is used as follows.

At some initial time, under steady state conditions at synchronous speed, a measurement is made for each quarter of the disc. At four labelled locations in store the computer records these oscillator counts applicable to each particular sector of the disc. The four counts in store (Q_1, Q_2, Q_3, Q_4) will be approximately 50,000. How near to 50,000 depends on the accuracy with which the lines on the disc were drafted (and their width) and the precision of the supporting boss. During a disturbance then, a count of W may be obtained from sector 3. The speed deviation is calculated by the program as $(W_3 - Q_3)/Q_3$ per unit. The accuracy of this measurement is limited by the vibration of the photo sensor and light source. Attention was given to the mounting of the devices to ensure that the accuracy was comparable to the resolution.

The specifications of the speed transducer are:

- (1) Resolution 0.00002 p.u.
- (2) Repeatable accuracy ± 0.00004 p.u.
- (3) Aperture time 10ms
- (4) Direct binary output of 16 bits.

5.7 MEASUREMENT OF THE INFINITE BUS FREQUENCY

From records of the Electricity Trust of S.A., it was seen that short term deviations in the mains frequency as high as $\pm 0.1\%$ can occur. This figure is comparable with the small signal deviations predicted for the alternator speed response (see Chapter 7). Therefore to determine in practise the effect of bus frequency deviation on the behaviour of the SGTL system, this frequency is measured with the same resolution as the alternator speed.

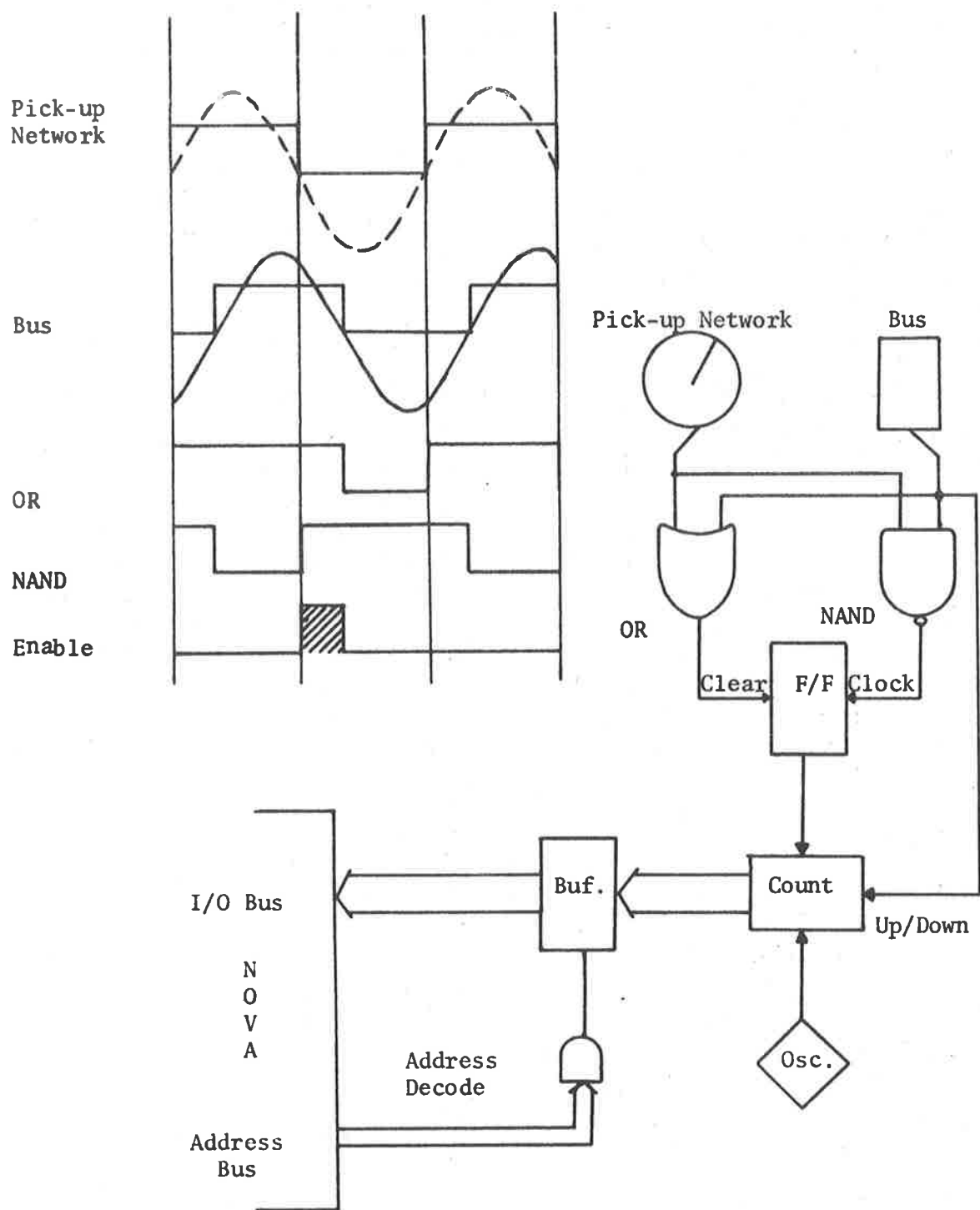


Figure 5.5 The power angle transducer.

A system similar to that of the speed transducer is used. Pulses from the same 5MHz oscillator are counted between zero crossings of one of the phases of the infinite bus. Thus 50,000 pulses, each $2\mu\text{s}$ long, are counted between zero crossings of the 50Hz signal. The count is shifted into a buffer at each zero crossing. It is then available to the program as required.

5.8 THE POWER ANGLE TRANSDUCER

Although the power angle δ is not used by the optimal controller it is required for the experimental determination of the K parameters as detailed in section 4.4.

δ is the angle between the rotor of the synchronous machine and an equivalent rotor of the infinite bus. The alignment of the alternator rotor with a reference point on the stator is marked by a pulse from a photo sensing arrangement similar to that used in the speed transducer. This pulse is used to control a digital circuit which produces a square wave of the same period as the pulses and locked to them. With the alternator open circuited, the photo sensor is rotated relative to the alternator shaft until the square wave output from the pick-up network is in phase with the open circuit terminal voltage. Then, under all loading conditions of the SCTL system, this square wave may be regarded as being in phase with the theoretical quadrature axis voltage of the equivalent two axis machine.

As the zero crossing of the open circuit machine voltage is used as the reference for the alternator rotor, the zero crossing of the same phase of the bus voltage waveform becomes the reference for the equivalent infinite bus rotor. This bus waveform is squared and reduced to TTL compatible logic levels. The power angle δ is then the phase angle between the two square waves in figure 5.5.

These signals are ORed and NANDed to produce the waveforms shown. These in turn control the setting of a flip flop which is set to 1 for a time which is proportional to the phase angle. This 1 level enables a binary counter.

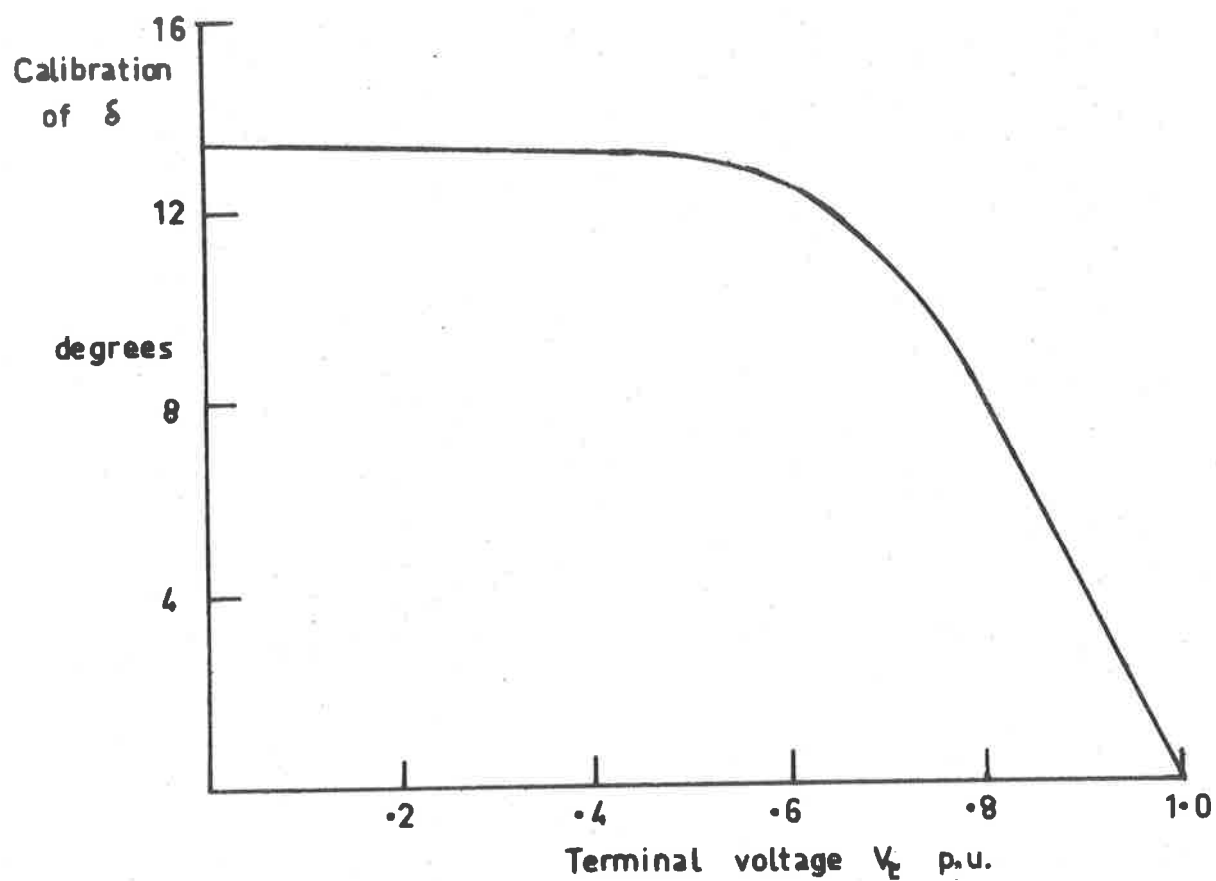


Figure 5.6 The calibration curve for the δ transducer, showing the effect of saturation on the alignment of the quadrature axis with the rotor axis.

When the enable signal returns to zero the contents of the counter, containing a representation of δ , are clocked into a buffer from where they may be read into the NOVA as required. The oscillator frequency was chosen as 18kHz (360 pulses every 20ms) so that the digital readout is in electrical degrees.

Positive or negative readings of δ (i.e. generating or motoring) are obtained by using an UP/DOWN counter and controlling the direction of the count with the bus waveform. If the bus voltage is positive when enable is 1 then the count is positive (generating), as shown in the waveforms of figure 5.5.

The quadrature axis of the equivalent 2 axis machine may be considered fixed with respect to the rotor of the synchronous machine if the effect of saturation is ignored. Due to saturation however, the alignment of the quadrature axis on the rotor varies with air gap voltage. To allow for this variation a calibration curve of the measured value of δ against open circuit terminal voltage was produced (figure 5.6). Knowing the terminal voltage at any particular operating, the correction to be subtracted from the measured value of δ is obtained from this curve.

The specifications of the transducer are:

- (1) Resolution 1 degree.
- (2) Accuracy (using calibration) ± 1 degree.
- (3) Range ± 180 degrees.
- (4) Sampling period 20ms.
- (5) 9 bit 2's complement binary output

5.9 THE TERMINAL VOLTAGE TRANSDUCER

The alternator terminal voltage, an analogue quantity, is sampled by the digital controller via the A/D interface. The controller uses only the magnitude of this voltage. The simplest way of making this available to the controller is to convert the terminal voltage to a proportional D.C. value before it is sampled by the A/D converter. A 3 phase rectifier followed by a fifth

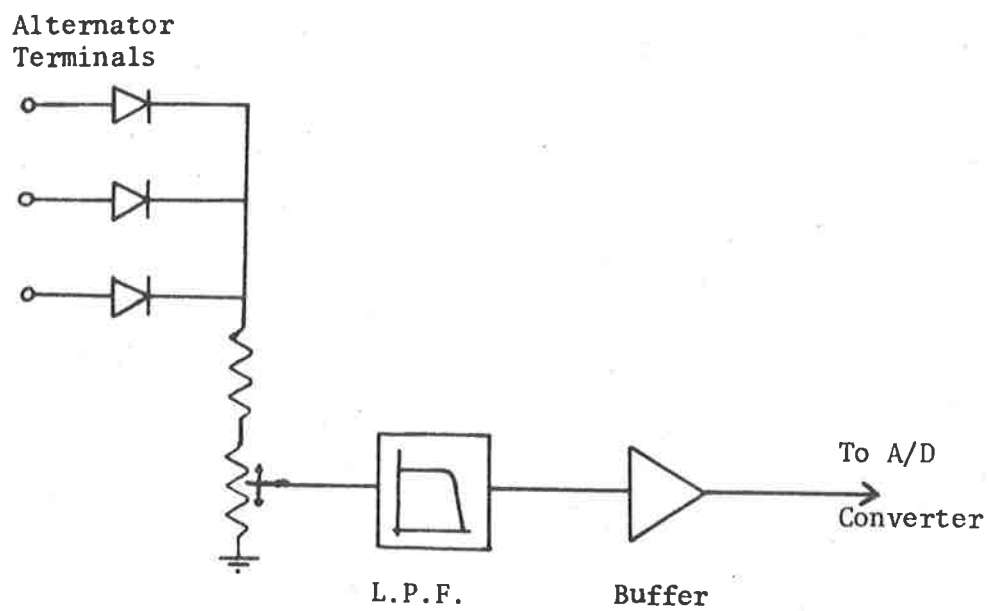


Figure 5.7 The terminal voltage transducer

order Butterworth low pass filter, as shown in figure 5.7, performs this function. The rectifier voltage is resistively divided to a suitable level before being applied to the active filter. The high order filter reduces the rectifier ripple to 1mV (0.1% of F.S.D.), while having an effective step response time constant of 15ms.

5.10 THE REAL AND REACTIVE POWER TRANSDUCERS

The real and reactive powers transferred between the alternator and the infinite bus are measured at the bus. These measurements are used by the controller to determine the gains it is to use (section 6.2), and to estimate the electrical torque (section 6.3).

Analogue techniques were used throughout to determine the powers rather than computing them from v_b , v_b and δ because of the limited storage and computing capabilities of the small computer. The power transducers produce proportional D.C. voltages which are sampled by the controller via the A/D interface. A schematic of the system used for measuring reactive power is shown in figure 5.8. Real power is measured by a similar system without the integrators.

The devices, C.T., transform the current in each line to a voltage which is proportional to and in phase with it. For reactive power measurement these signals are phase shifted ninety degrees by the integrators and then multiplied by a voltage proportional to the infinite bus voltage. The multiplier outputs are proportional to the instantaneous reactive power delivered by each phase. As shown in appendix A.8 the sum of these outputs is a D.C. voltage proportional to the total reactive power. Some ripple on the output of the summing amplifier, resulting from system imbalance, is reduced by an active low pass filter. The filter output passes to an A/D converter.

An accurate transformation of line current to a proportional voltage is achieved using the circuit in figure 5.9. Simple resistive loading of the secondary of a current transformer is unsatisfactory as a load high enough to produce a useful voltage (say 10 volts at a maximum line current of 20 amps)

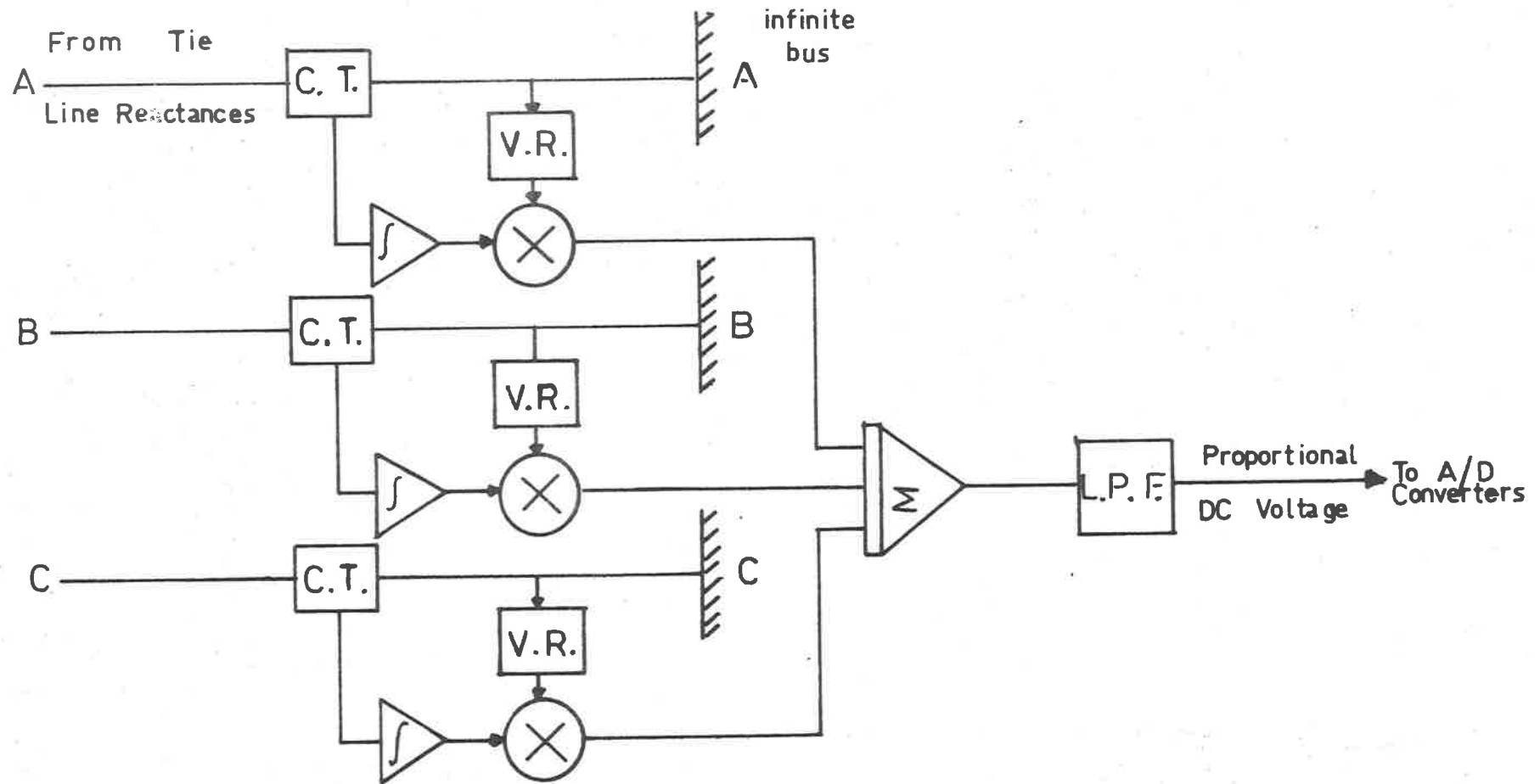


Figure 5.8 The system for measuring the reactive power delivered to the infinite bus.

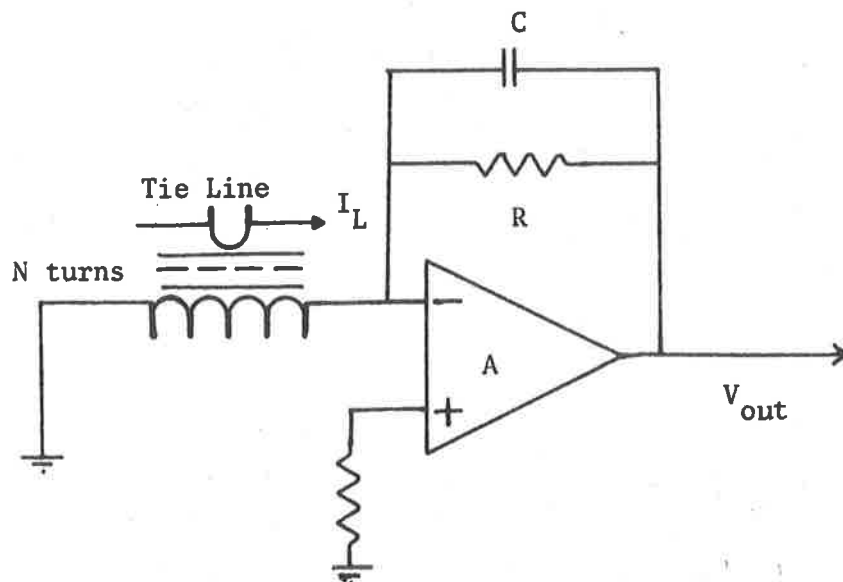


Figure 5.9 The circuit used to transform line current to a proportional voltage.

produces a non linear relationship between primary current and output voltage, as well as a variable phase shift between the waveforms. The effective resistance inserted in the secondary of the current transformer in figure 5.9 is R/A ohms where A is the open loop gain of the amplifier. The output voltage is $I_L R/N$. With the values chosen for R and N , and with the amplifier used (see appendix A8.1), these figures become typically:

- (a) effective secondary resistance = 0.044 ohms
- (b) output voltage 7V for $I_L = 20A$
- (c) burden = 0.7 μ VA

With this extremely low burden a suitable current transformer is easily wound. The capacitor, C , removes some slight phase shift. Overall the device is linear to better than $\pm 0.1\%$ in magnitude and has a phase shift of less than ± 0.5 degrees over its full range.

Simple dividers are used to reduce the infinite bus voltage to the required level for the multipliers. The integrators are active devices using capacitive feedback around an operational amplifier. Integrated circuit voltage multipliers were chosen as the analogue multipliers. The summing amplifiers are of the standard form.

The specifications of the power transducers are:

- (1) Line current 3.5 p.u. max.
- (2) Bus voltage 1.25 p.u. max.
- (3) Real power ± 1.0 p.u. max.
Reactive power ± 2.0 p.u. max.
- (4) Output ± 1 volt D.C. max.
- (5) Accuracy $\pm 0.5\%$ of full scale
- (6) Response time 15ms
- (7) Output impedance 0.01 Ω

5.11 THE FIELD EXCITER

So that the characteristics of the exciter could be ignored in the analysis of the system, an electronic amplifier was used. A suitable commercially produced type was chosen. This device is a high power operational amplifier with 3dB frequency of 300kHz and output ratings of ± 36 volts at 5 amps. The external feedback components of the amplifier were adjusted to give the required gain from the D/A output to the alternator field.

CHAPTER 6

IMPLEMENTATION OF THE CONTROLLER

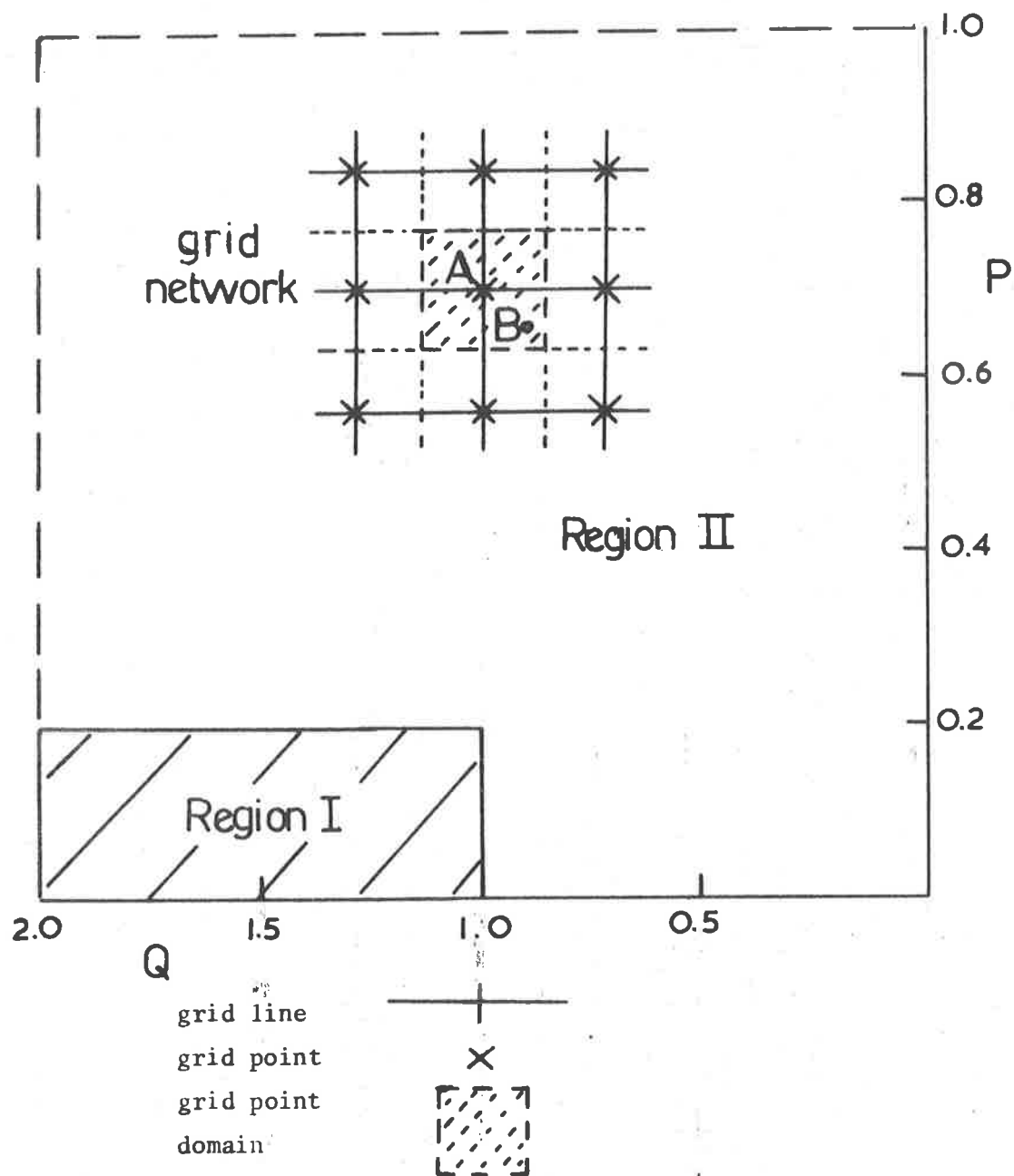


Figure 6.1 The formation of the grid points and surrounding domains in the PQ plane. The plane is divided into two regions of uniform grid spacings.

6.1 INTRODUCTION

The control strategy to be applied to the SGTL system has been detailed in Chapter 3, working from a mathematical model of the experimental system. The parameters of the model were evaluated in Chapter 4 and its accuracy verified under open loop conditions. Consideration is now given to the implementation of the controller.

The concept, mentioned in section 3.4 of dividing the PQ plane by lines of constant power is expanded and consideration is given to the choice of a suitable spacing for these grid lines.

The NOVA control program is discussed to indicate the manner in which a small digital computer can be used on line to control and monitor this complex non linear system.

6.2 THE GRID OF OPTIMAL GAINS

The elements of the matrices in the state equation are shown in section 3.4 to be constant at any particular steady state operating point, resulting in a linear state equation and controller. Since the matrix elements vary with operating point, the controller gains vary non linearly with P and Q, as shown in figures 4.19 to 4.23. To implement a controller which is optimal throughout the operating plane, continuous variation of gains with operating point is required. However this is not possible due to the limited storage, computing power and speed of the small digital computer. Instead the gains are computed off-line at a series of grid points in the PQ plane and held constant over a small rectangular domain surrounding each grid point (figure 6.1). The control is then truly optimal only at the grid point, and deviates from optimality in the surrounding domain. The degree of this deviation is dependent on the spacing of the grid points.

The gains used by the controller are those computed for the grid point of the domain in which the operating point lies.

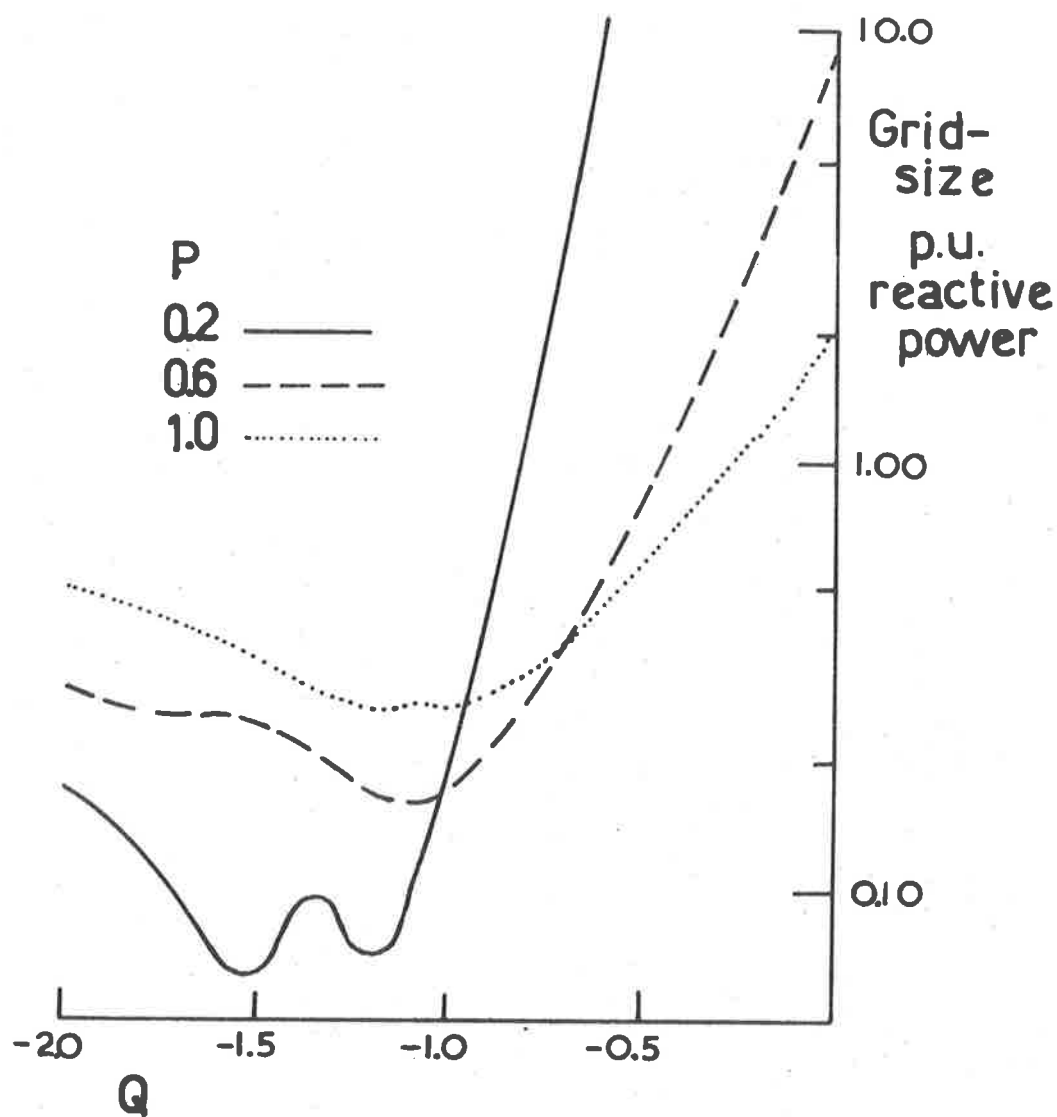


Figure 6.2 The variation of grid spacing for a 10% RMS deviation from optimal response. Operating point varying along lines of constant real power.

6.2.1 THE GRID SPACING

As a guide in choosing a suitable grid spacing, a comparison is made between the true optimal response at an operating point B, and the response obtained when the gains which are optimal for the nearest grid point A, are applied at B (see figure 6.1).

This comparison is made as follows. The optimal gains are computed at a series of steady state operating points over the PQ plane. At each of these points the response of the terminal voltage to a step change in reference voltage is predicted when the controller uses the gains computed for optimality at that point (point B). The response is also predicted when the controller uses the gains computed for optimality for the grid points in the immediate vicinity of that operating point (point A). The RMS deviations of the predicted non-optimal responses from the true optimal response are computed over the 2% settling time of the optimal voltage response, and expressed as a percentage of the input step. The spacing between grid points which would result in a 10% RMS deviation is then estimated as:

$$S_{p10} = \frac{S_p \times 10\%}{\% \text{ RMS observed}}$$

where S_{p10} is the spacing for a 10% RMS deviation, S_p is the actual spacing used, and the denominator is the percentage RMS deviation observed with this spacing. This is only an approximation as it assumes a linear relationship between RMS deviation and grid spacing, but since the result is used only as a guide it is considered satisfactory.

Figures 6.2 and 6.3 show how the grid spacing varies over the operating plane if a maximum RMS deviation of 10% from the true optimal response is to be maintained. Figure 6.2 shows the operating point varying along lines of constant real power and in 6.3 the point

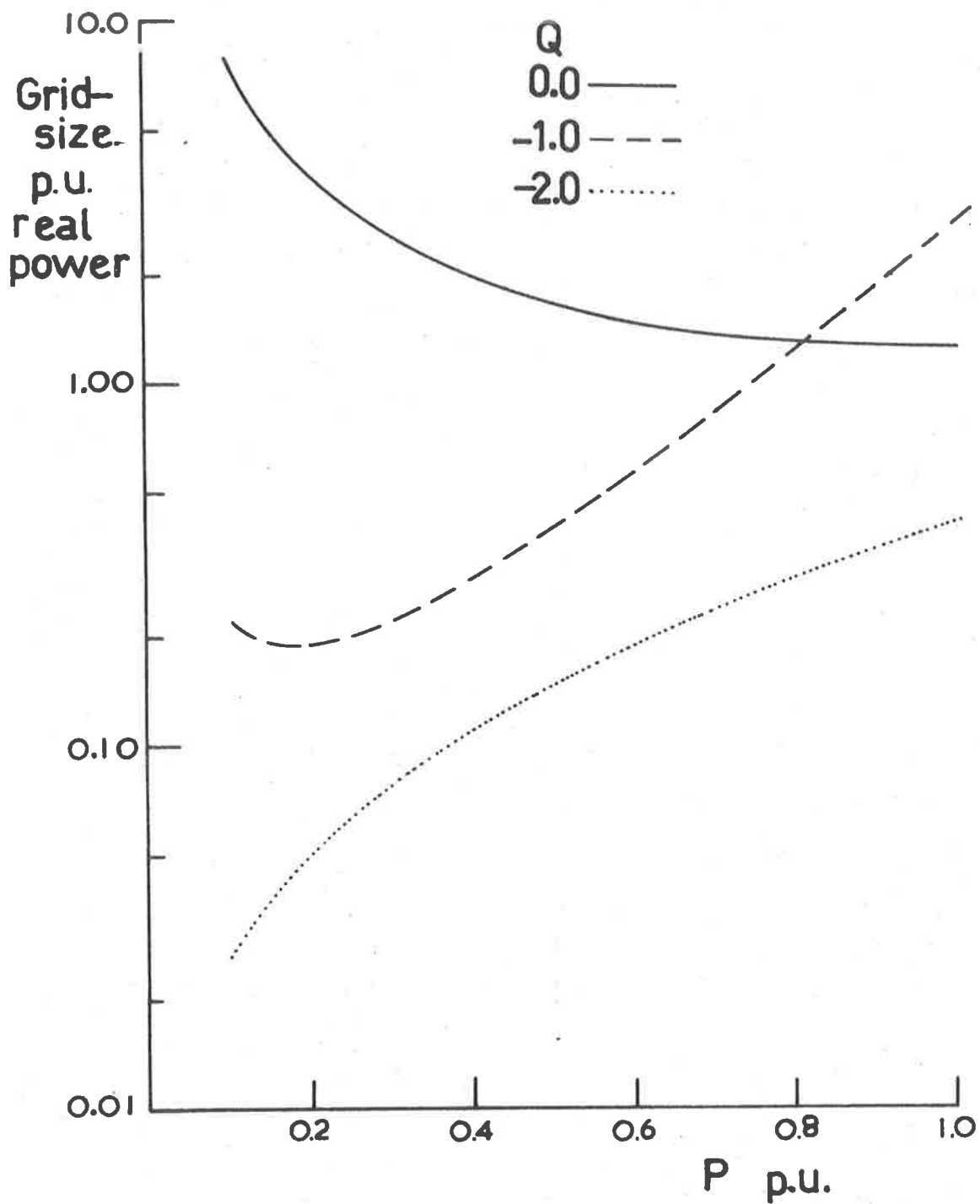


Figure 6.3 The variation of grid spacing for a 10% RMS deviation from optimal response. Operating point varying along lines of constant reactive power.

varies along lines of constant leading reactive power. The grid spacing is measured in per unit real or reactive power.

To facilitate the determination of the gain value the controller should use during operation (see section 6.33) even spacing between grid lines is desired. Considering figures 4.19 to 4.23 which show the variation of the optimal gains with operating point, and comparing these with figures 6.2 and 6.3, it is seen that the required grid is very fine in regions where the rate of change of optimal gains with operating point is large. This occurs for high leading reactive power and low real power. The limited storage of the controller does not allow the fine grid necessary in this region of the PQ plane to be applied over the whole plane. Therefore the plane is divided into regions I and II of figure 6.1. Region I is bounded by a line at 0.2 p.u. real power and lines at -1.0 and -2.0 p.u. reactive power. Grid spacings of 0.05 p.u. along the Q axis (from figure 6.2) and 0.025 p.u. along the P axis (from figure 6.3) were chosen in this region. For the remainder of the plane a grid spacing of 0.1 p.u. along both axis is used.

Having chosen a grid spacing which satisfies the 10% RMS error criterion, a check is made on the margin of safety which exists in selecting the set of gains to be used at any particular operating point. At a series of steady state operating points lying beyond the open loop stability limit, feedback gains computed for all grid points in the plane are used for the controller, and the stability of the system with these gains applied at each operating point in turn is predicted. Figure 6.4 shows a series of operating points (1) to (4) at a real power of 0.5 p.u. The system remains stable if the feedback gains used by the controller at these points are obtained from grid points lying on or within the boundaries (1) to (4).

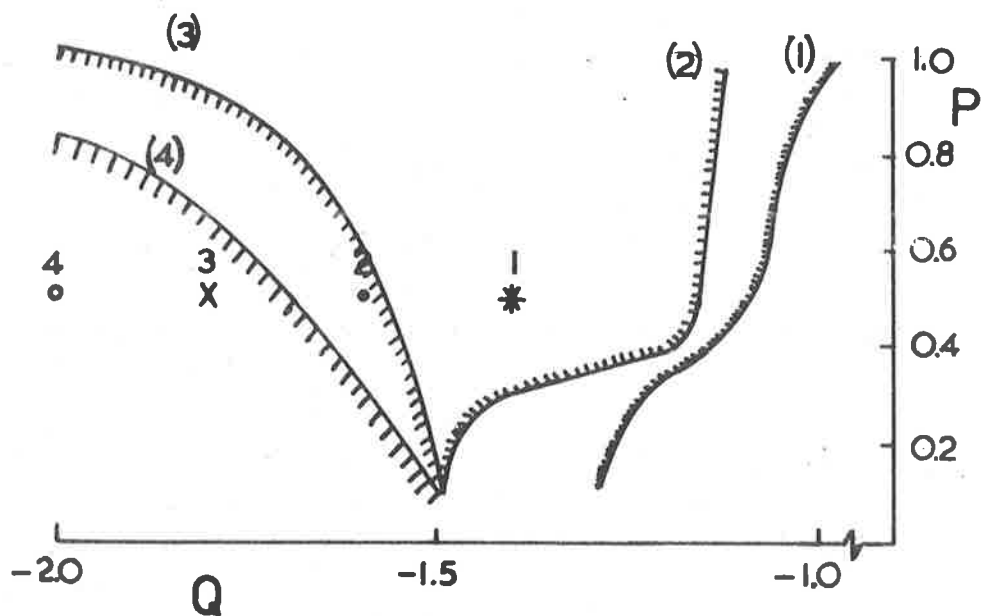


Figure 6.4 Controller gains from grid points to the left of lines (1) to (4) result in stable operation if applied to operating points (1) to (4). Operating points are at $P = 0.5$ p.u. $Q = -1.4, -1.6, -1.8$ and 2.0 p.u.

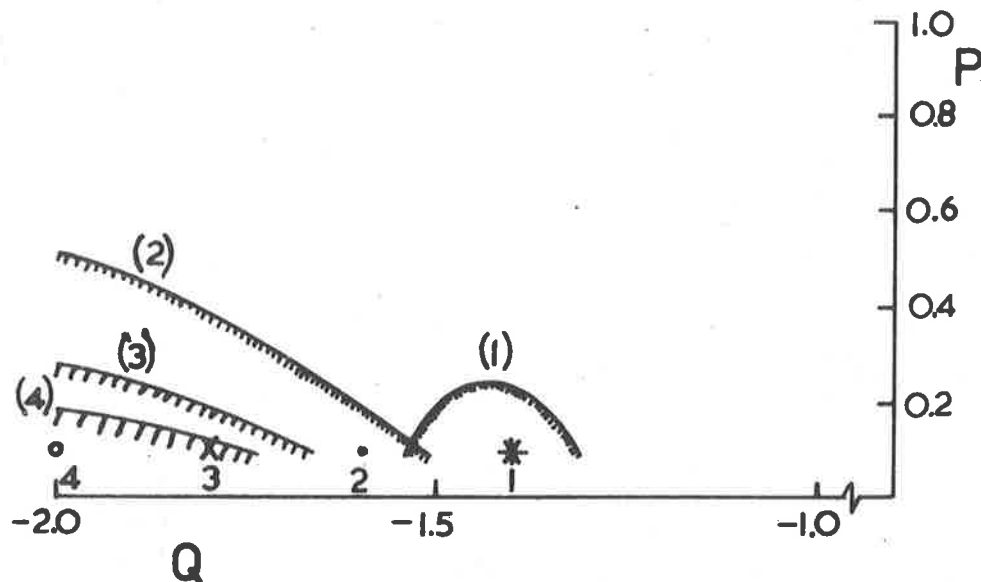


Figure 6.5 Controller gains from grid points to the left of lines (2) to (4) or bounded by (1) result in stable operation if applied to operating points (1) to (4). Operating points are at $P = 0.1$ p.u. $Q = -1.4, -1.6, -1.8$ and 2.0 p.u.

Thus for a real power of 0.5 p.u. there is a large safety margin in the proposed system. For low real power this margin is reduced as shown in figure 6.5. This analysis was performed to determine how accurately the operating point of the system must be determined if stability is to be maintained and to indicate how critical any inaccuracies in the mathematical model may be with respect to the stability margin of the system. Thus for low real power good accuracy is required in determining the operating point, and an accurate mathematical model is necessary if the 10% RMS error criterion is to be satisfied in practice, or even if stable operation is to be obtained.

6.3 THE CONTROL PROGRAM

The NOVA computer is programmed to function as the system controller, incorporating the control philosophies discussed in the previous section. The controller senses deviations in terminal voltage, speed and electrical torque from their steady state values, these deviations having been chosen as the states of the system. The computed control force is a perturbation superimposed on the steady state value of the applied field voltage.

The program was developed from the point of view of system assessment and experimentation rather than as a general purpose operational system. The following is a brief outline of its operation and may be read in conjunction with the flow diagram.

A voltage is applied to the field of the alternator from the computer via a D/A converter and the electronic field amplifier - exciter. This permits the machine set to be run up to the correct speed and the system synchronised (see section 4.2).

The program is then restarted to provide a simple proportional voltage control which allows a smooth transition into the optimal control section.

Output a voltage to the
alternator field via a D/A
converter and field amplifier

↓
Run up the machine set

↓
Synchronize the system

↓ manual transfer of
program control

Read the terminal voltage V_t

↓
Assume $V_{ref} = 1 \text{ p.u.}$

↓
Error = $V_{ref} - V_t$

↓
Field voltage $V_f = 4 \times \text{Error}$

↓
Output V_f via D/A converter

↓ manual transfer of
program control

Start the real time clock

↓
Read V_t

↓
Initial $V_t = V_{to} = V_t$

↓
Read P

↓
Assume $C_{eo} = P$

↓
 $\Delta V_{ref} = 0$

↓
 $\Delta \omega = 0$

S
Y
N
C
H
R
O
N
I
Z
E

P
R
O
P
O
R
T
I
O
N
A
L
C
O
N
T
R
O
L

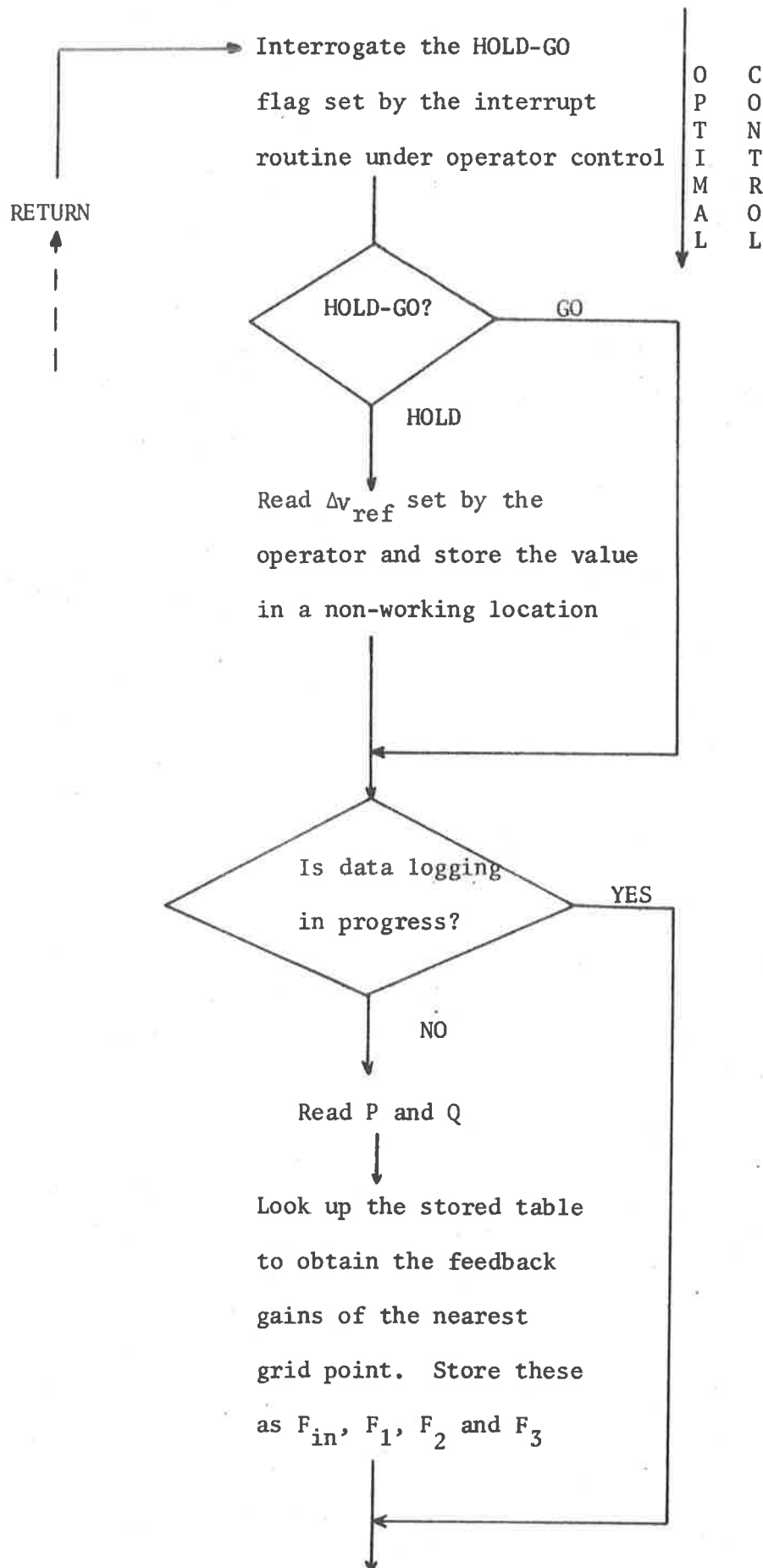
T
R
A
N
S
I
T
I
O
N

This simple controller reads the terminal voltage and applies a voltage to the field of the alternator proportional to the deviation of the terminal voltage from one per unit. Once the system has reached a steady state condition the computer is stopped then restarted at another entry point in the program to allow the transition to the optimal controller to be made. When the computer is stopped, or if no signal is sent out, the D/A converter acts as a zero order hold, maintaining the field voltage at its present value.

The real time clock on the NOVA is initiated with a 10ms period. The clock operates by interrupting the main program every period allowing a supplementary program, or interrupt routine, to perform its specific function at set time intervals. Once this function is completed, control is returned to the main program at the point at which the interrupt occurred. The interrupt routine is outlined in the following section (6.3.1).

The terminal voltage is read and assumed to be the steady state or initial value. The initial electromagnetic torque is approximated by the measured real power, assuming that the small perturbations in speed have negligible effect on this estimation. The machine speed is assumed to be one per unit, i.e. $\Delta\omega = 0$, and the perturbation in the reference voltage is zero. The present applied field voltage becomes the initial field voltage on entry to the optimal control section. Having set up the initial conditions of the system at the time of the transition, the main control program is entered automatically.

Operator control over the program is provided by a switch whose status is read by the computer which then sets a flag designated HOLD-GO. The switch is actually interrogated by the interrupt routine which in turn sets a HOLD-GO flag in the main program. This allows the operator to set the required perturbation in reference voltage while the flag is set to HOLD, and



also indicates to the interrupt routine that a new set of initial conditions for the system is required.

The operation of the controller is such that the perturbations in reference voltage applied to the system are set by the operator to any one of a programmed range of discrete steps. There are eleven steps in the range and for small perturbation studies these were programmed to be -0.05 p.u. to +0.05 p.u. in 0.01 p.u. steps. This range can be changed simply by altering program constants. If the required perturbation in reference voltage is set to zero and the operator HOLD-GO switch activated, then new initial conditions are established for the system and the controller acts to maintain the states of the system at zero. That is, the initial conditions of the system are maintained since the system states are chosen as perturbations from these initial conditions. When operating in this way the controller becomes an output regulator rather than a tracking regulator, and as explained in section 2.6 the control force which returns the disturbed states to their required value of zero is optimal.

Thus, if the operator control switch is in the HOLD position a new set of initial system conditions are obtained and the required perturbation in reference voltage is read by the computer and stored in a temporary location. If the switch is set to GO this section is skipped over.

The program checks whether data is being recorded for later analysis (see section 6.3.5). If data logging is not in progress then the PQ operating point of the system is determined and the control gains applicable to the nearest grid point 'looked-up' from the stored table of gains (see section 6.3.3). If data logging is in progress then the gains used by the controller are not changed from the current values.

6-7a

↓
 Δv_{ref} is picked up
from its working
location

↓
$$I_n = \Delta v_{\text{ref}} * F_{\text{in}}$$

↓
Read current V_t

↓
$$\Delta v_t = V_{t0} - V_t$$

↓
$$v_{\text{tf.b.}} = \Delta v_t * F_1$$

↓
Read current P

↓
$$\Delta C_e = P_o - P$$

↓
$$C_{\text{ef.b.}} = \Delta C_e * F_3$$

↓
$$\omega_{\text{f.b.}} = \Delta \omega * F_2$$

↓
$$\Delta v_f = I_n + v_{\text{tf.b.}} + \omega_{\text{f.b.}} + C_{\text{ef.b.}}$$

↓
$$V_f = V_{f0} + \Delta v_f$$

↓
Output V_f via D/A converter
and field amplifier to the
alternator field.

↑
|
|
|
RETURN

The contribution of the input and the states of the system to the control force are now determined. The input term is the product of the perturbation in reference voltage and the input multiplier F_{in} .

The machine terminal voltage is read and its deviation from the initial terminal voltage determined. (Once the transfer to the optimal control program has been made, all initial conditions are set up in the interrupt routine). The voltage feedback signal is then computed.

The real power delivered to the bus is measured and the deviation from the initial power found. The deviation in electrical torque is assumed to be equal to the deviation in power. Considering the following equations this is seen to be a close approximation when losses are negligible.

$$\begin{aligned}
 P &= \omega C_e \\
 P + \Delta P &= (\omega_s + \Delta\omega)(C_e + \Delta C_e) \\
 \Delta P &= \Delta\omega C_e + \Delta C_e \omega_s \quad \omega_s = 1 \text{ p.u.} \\
 \Delta C_e &= \Delta P - \Delta\omega C_e \\
 &= \Delta P - \Delta\omega P
 \end{aligned}$$

The ratio of ΔC_e to $\Delta\omega$ predicted by the small signal linear model is approximately 40:1 (see chapter 7) so that with P less than 1 p.u., $\Delta C_e = \Delta P$ with little error. The torque feedback signal is computed as $\Delta C_e * F_3$.

The deviation of machine speed from the initial value is measured within the interrupt routine and stored. This is recalled from its storage location by the main program and the feedback signal computed as $\Delta\omega * F_3$.

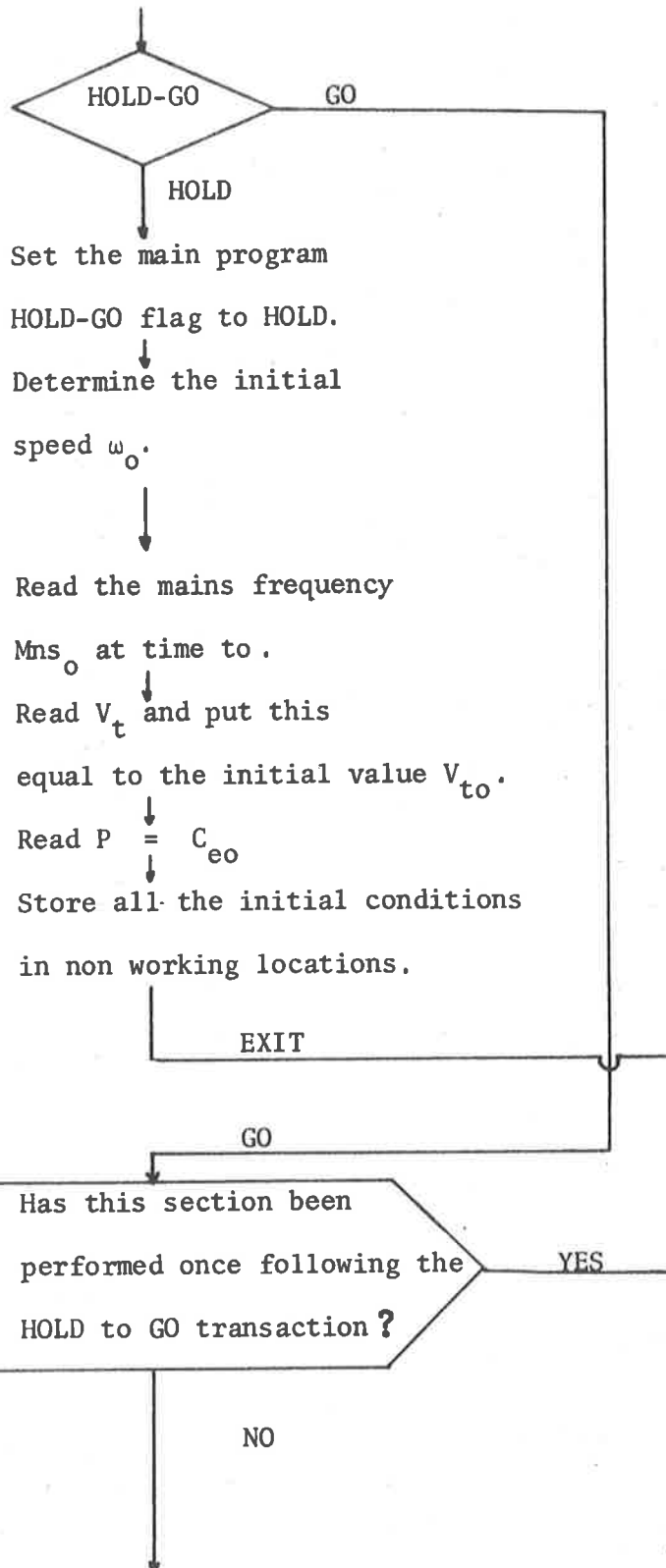
The perturbation to be applied to the field voltage is the sum of these feedback signals and the input signal. This is added to the initial value of field voltage and the total applied to the field of the alternator via a D/A converter and the field amplifier. The program sequence

THE INTERRUPT ROUTINE

Restart the clock.

↓
Interrogate the clock operator

HOLD-GO control switch.



counter then returns to the start of the optimal control section and the above cycle is repeated.

6.3.1 THE INTERRUPT ROUTINE

As mentioned previously the NOVA real time clock operates by interrupting the main program and transferring program control to the interrupt routine. This routine is now outlined.

The clock is restarted so that it will interrupt again in a further 10ms.

The operator set HOLD-GO switch is interrogated. If the switch is set to HOLD the sequence is as follows. The HOLD-GO flag in the main program is set to HOLD. The initial steady state machine speed is measured. (Small deviations in the infinite bus frequency are followed by the synchronized alternator (see section 8.4) so that although the system is in a steady state condition the initial speed may not necessarily be the nominal synchronous speed of 1500 rpm. Therefore each time initial conditions are established a value is determined for the initial speed). The mains frequency is also recorded at this time.

The present terminal voltage becomes the initial terminal voltage. The real power is recorded as the initial torque and the present field voltage as an initial value.

All of the initial conditions are stored in temporary (non-working) locations. With the HOLD flag set, the program follows the above sequence then skips the next section.

If the GO flag had been set originally the program is transferred to this point. The HOLD-GO flag in the main program is set to GO. The perturbation in reference voltage set by the operator and stored by the main program in a non

6-9a



Set main program HOLD-GO
flag to GO.



Transfer Δv_{ref} from its
non working location to a
working location.



Transfer initial conditions
to working locations.



Read present ω .



$$\Delta\omega = \omega_0 - \omega$$



Read present mains frequency.



$$\Delta mns = mns_0 - mns.$$



Data log if required.



Return to the main
program at the point
at which the interrupt
occurred.

working location is transferred to a working location. The initial conditions are also transferred to working locations. Thus with the flag set to HOLD the controller operates with the original set of initial conditions and reference input. As soon as GO is set the new conditions apply. Another program flag is used to ensure that the above transfer of initial conditions from temporary to working locations occurs only on the first program cycle after the HOLD to GO transition.

The program branches meet at this point to determine the deviations of the machine speed and the mains frequency from their initial values.

The relevant operating data and the deviations of the states are recorded in the computer store if the operator requires them (see section 6.3.5).

Program control is now returned to the point in the main program at which the interrupt occurred.

Data logging is performed within the clock interrupt routine so that the states may be recorded at 10ms intervals. Speed measurements are made at 10ms intervals as this is the aperture time of the transducer (see appendix A4). The initial conditions are set up in this routine so that they may be applied at the initial time t_0 of a data logging sequence.

6.3.2 FIXED POINT ARITHMETIC

All arithmetic operations are performed by the program in a fixed point mode. While the NOVA has as part of its software library a floating point interpreter, floating point operations are 2 to 3 orders slower than fixed point. This would make the control program cycle time so long that the

controller would have to be regarded as a discrete sampling system, and the simplicity afforded by the continuous system would be lost. The present cycle time is approximately 2ms and the control is assumed continuous.

When operating in fixed point mode the computer regards all numerical quantities as 2's complement integers, and when writing the program the inferred position of the binary (c.f. decimal) point must be noted by the programmer. For example a terminal voltage of 0.625 p.u. is read into a 16 bit NOVA register as (see Appendices A3, A7);

Data bit No	15	14	13	12	11	10	9	8	7	6	5	4	3	2	1	0
Register Contents	*0	1	0	1	0	0	0	0	0	0	0	0	0	0	0	0

*Sign bit

The inferred position of the binary point is between bits 15 and 14 so that the inferred value is $2^{-1} + 2^{-3} = 0.625$. The number is taken by the computer to be $2^{12} + 2^{14}$. In writing the program it is necessary to keep track of the inferred position of the binary point throughout all arithmetic operations, so that the correct field voltage is computed.

Double precision arithmetic (that is two registers combined to form a double length (32 bit) word) is used where necessary to obtain sufficient resolution and to ensure register over flow (saturation) does not occur in the intermediate stages of computation.

6.3.3 GAIN TABLE LOOK-UP

The optimal gains are held in the NOVA store in locations whose addresses are determined by the PQ position of the grid point for which the gains are computed. Thus during operation

		+4K _g										Q					P					Gain						
Bit No. of register		15	14	13	12	11	10	9	8	7	6						5	4	3	2						1	0	
		0	0	0	0																							
The address of the storage location for gain F _i at grid point P,Q, _i region II						1	0	0	0	0	0	-0.0					0	0	0	1	0.1					0	0	F _{in}
						1	0	0	0	0	1	-0.1					0	0	1	0	0.2					0	1	F ₁
						1	0	0	0	1	0	-0.2					0	0	1	1	0.3					1	0	F ₂
						⋮						⋮									⋮					1	1	F ₃
						⋮						⋮									⋮							
						1	1	0	0	1	1	-1.9					0	1	0	1	0.9							
						1	1	0	1	0	0	-2.0					0	1	1	0	1.0							
4 K _g		-Q x 10											Px 10						F									

Figure 6.6a The technique for determining storage locations of the feedbacks; region II

		Q										P				Gain				
Bit No. of register	15	14	13	12	11	10	9	8	7	6		5	4	3	2		1	0		
	0	0	0	0	0															
The address of the storage location for gain F at grid point P,Q; region I						0	0	0	0	0		0	1	0	0	0.100	0	0	F_{in}	
						0	0	0	0	1	1.05	0	1	0	1	0.125	0	1	F_1	
						0	0	0	1	0	1.10	0	1	1	0	0.150	1	0	F_2	
																0	1	1	F_3	
																1	0	0		
																0	200			
						1	0	1	0	0	2.00									
7K ₈		-(Q+1)x20											Px40					F		

Figure 6.6b The technique for determining storage locations of feedback gains; region I

the storage locations of the gains to be used by the controller can be identified after determining the present PQ operating point of the system. The following procedure was adopted for storage and look up.

The address of the storage location for a particular gain, at a particular grid point, is made up as a composite 16 bit NOVA word. Referring to figure 6.6, the two least significant bits (0, 1) indicate whether the location holds F_{in} , F_1 , F_2 or F_3 . The next four bits (2, 3, 4, 5) indicate the real power of the grid point, and the next five (6, 7, 8, 9, 10) the reactive power.

If the grid point lies in Region II (figure 6.1) then bits 2 to 5 are calculated as $P*10$ and bits 6 to 10 as $-Q*10$ where P and Q are the coordinates of the point. (Region II has an 0.1 p.u. grid spacing). Four thousand octal is added to this composite word (figure 6.6a) to shift the locations to a higher part of store. Thus for Region II the gains are stored in locations with addresses 04040_8 for the gain F_{in} at PQ point (0.1, 0.0) to 06433_8 for the gain F_3 at (1.0, -2.0).

If the grid point lies in Region I then bits 2 to 5 are $P*40$ (0.025 p.u. grid spacing on P) and bits 6 to 10 are $-(Q + 1)*20$ (0.05 p.u. grid spacing on Q). Seven thousand octal is added to this composite word. The addresses of the storage locations for Region I range from 07020_8 for F_{in} at (0.1, -1.0) to 11443_8 for F_3 at (0.2, -2.0).

The gains F_{in} , F_1 and F_3 are stored in the appropriate locations with the inferred binary point lying between bits 8 and 9; that is a gain of one is stored in the NOVA as the

```

.LOC 07020      ; Q=-1.00,P=0.1 TO .2,ETA,F1,F2,F3
2905.
-1116.
27.
45.
2865.
-1142.
36.
53.
2820.
-1172.
44.
61.
2773.
-1208.
52.
68.
2725.
-1247.
61.
75.
.LOC 07120      ; Q=-1.05,P=0.1 TO .2,ETA,F1,F2,F3
2881.
-1132.
32.
57.
2828.
-1167.
41.
68.
2770.
-1210.
51.
78.
2711.
-1259.
61.
87.
2653.
-1312.
70.
95.

```

Figure 6.7 A listing of section of the 6400 punched tape output of controller gains.

integer 2^9 . For F_2 , one is stored as $2^{20}/50,000$. These multiplicands are determined by the magnitudes of the gains to be stored to give the best resolution without overflow, and for the speed feedback gain F_2 , by the fact that the transducer records 50,000 as one per unit.

After computing the feedback gains on the large computer as per section 4.5, F_{in} , F_1 and F_3 are multiplied by 2^9 and F_2 by $2^{20}/50,000$ before being punched onto paper tape. In addition the address of the storage location for each set of gains is computed and punched. The data on the paper tape from the 6400 computer is then compatible with the NOVA assembler and may be used directly as program constants at assembly time. A listed segment of this output appears in figure 6.7.

To look up a controller gain during operation the PQ point of the system is read by the computer. The program then determines whether this operating point lies in Region I or II of the plane. If it is in Region I then P is rounded to the nearest 0.025 p.u., and multiplied by 40. Q is rounded to the nearest 0.05 p.u., increased by one, inverted and multiplied by 20. The composite word in figure 6.6b is formed to give the address of the storage location of the gains applicable to the grid point nearest the operating point. Knowing the address of the storage location, the gains are available as required. A similar procedure is followed if the operating point lies in Region II.

6.3.4 ALLOWING FOR LOSSES

Losses in the alternator and tie lines are neglected in the mathematical model of Chapter 3. However they affect the measurement of real power delivered to infinite bus, and for

reasons detailed in section 8.3 it was found necessary to allow for these losses in determining the electromagnetic torque of the alternator and also the system operating point.

The effect of these losses on the performance of the controller is most marked in the region of high leading reactive power where the torque feedback gain F_3 is comparatively high (figure 4.22). In this region the line current is about 1.4 to 2.0 p.u. and the terminal voltage 0.6 to 0.4 p.u. Measurements of the losses were made under these conditions.

The armature losses consist of armature copper losses, stray losses, and iron losses. The copper and stray losses are measured under short circuit conditions and an effective armature resistance determined. For a particular short circuit armature current I_{sc} , the total short circuit loss is equal to $3 R_{eff} I_{sc}^2$. Since the stray loss is not proportional to I_{sc}^2 , R_{eff} is dependent on the armature current. R_{eff} is determined over a range of I_{sc} of 1.0 to 2.0 p.u.

For terminal voltages about 0.5 p.u. the iron losses were found to be negligible compared to the effective copper losses, and all losses are assumed to be expressible as $I^2 R_{eff}$.

From a plot of the variation of the effective series resistance of the armature and tie line against line current, a constant value of 0.023 p.u. is chosen for use in an on line estimation of the power loss. The controller estimates the line current from the measured values of P and Q and calculates the $I^2 R$ losses. This is added to the measured value of real power to provide an estimate of generated power. This corrected value is used by the controller in determining the operating

point of the system, and also in estimating the electrical torque perturbation.

Operation without the correction factor is considered in section 8.3.

6.3.5 DATA LOGGING

A data logging option is included in the interrupt routine. The values of the states may be recorded in the computer store at 10ms intervals for up to 8 seconds following a step change in reference voltage and the associated establishment of a new set of initial conditions. Also recorded is the relevant operating data: the initial values of P and Q; the controller gains and the grid point from which they were obtained; the magnitude of the applied step; the final steady state value of Q; and the deviation of bus frequency from its initial value throughout the period.

The operator instigates the data logging process by setting a switch while the program control HOLD-GO switch is in the HOLD position. Data logging starts when the control switch is returned to the GO position and the new initial conditions have been transferred to their working locations. Once the buffers carrying the recorded states have been filled the final steady state value of Q is automatically recorded, and the logging sequence terminated.

Data logging will only occur if the data switch is on when the HOLD to GO transition is made. Therefore when a logging sequence has been completed the data switch may be set off and the system stepped to a point in the operating plane where it is open loop stable. The computer is then stopped

(opening the control loop) and using a high speed punch, the stored information punched onto paper tape for future analysis.

When studying system behaviour under small signal perturbations, the controller gains are held constant during data logging. This means that if the operating point does penetrate the domain of another grid point in response to an input step, the gains applying at the start of the logging sequence are retained. Consequently the recorded response cannot be influenced by a change in controller gains.

6.4 SUMMARY

The program outlined in this chapter is primarily intended to control the experimental system. The use of the HOLD-GO switch to initiate the establishment of a new set of initial conditions could be automated in an operational system. Since the control must operate on small deviations of the states from their initial values, a maximum allowable limit could be set on the magnitude of these deviations. Beyond this limit a new set of initial conditions would be established as soon as the system reached its new steady state operating point. The only feature of the implemented controller which must be retained is the use of the small signal deviations of the system variables, as these deviations and not the absolute quantities constitute the state variables in the mathematical model.

Only a simplified outline of the program has been given. A full listing incorporating operating instructions and functional comments appears in Appendix B. The program has not been specifically optimized with respect to cycle time or storage requirements, having been written in a way which makes its method of operation easier to follow. The program is written entirely in NOVA assembly language as the higher level BASIC available on NOVA is not only unsuitable as a real time language, but uses most of the limited store of the computer.

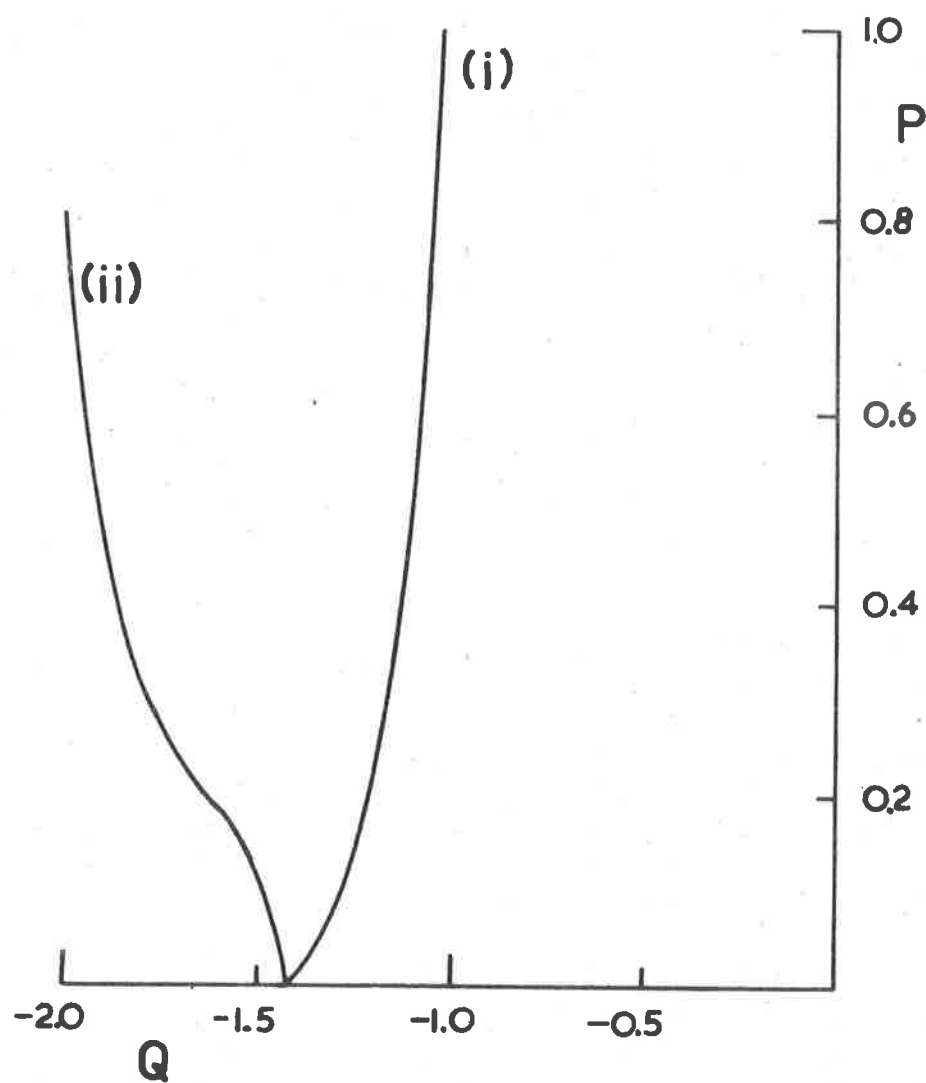


Figure 7.1 Operation of the S.G.T.L. system in the leading power factor region

- (i) The stability limit of the unregulated machine.
- (ii) The stability limit with a proportional voltage regulator whose gain is continuously adjusted for maximum extension of the limit.

7.1 INTRODUCTION

With the infinite bus voltage maintained constant at 1.0 p.u., the small signal behaviour of the experimental system was investigated at leading power factors throughout Regions I and II of the PQ plane (as defined in figure 6.1). The reactive power is limited to -2.0 p.u. by the 15 amp rating of the 3 phase supply mains used as the infinite bus. The real power available has a maximum of 1.0 p.u. determined by the rating of the D.C. motor prime mover, and a minimum of 0.12 p.u. determined by the maximum resistance available in series with the armature of the D.C. motor. In addition, operation in the region of -2.0 p.u. reactive power is time limited by the thermal rating of the alternator, as currents in this region are more than 2.0 p.u.

The closed loop system was stable throughout the defined Regions. It should be noted that the stability limits shown in figure 7.1 for both the open loop system and the system under the control of a proportional voltage regulator, whose gain is adjusted for maximum extension of the limit, lie within these Regions.

This Chapter deals with the small signal behaviour of the experimental system for operation in Region II. Operation in Region I, which is characterised by low real power and high leading reactive power, is considered in sections 8.3 and 8.4.

7.2 SOME THOUGHTS ON THE PRESENTATION OF THE RESULTS

Before displaying the behaviour of the closed loop experimental system the reasons for establishing this system are recalled. Basically these were to permit the assumptions made in deriving the mathematical model to be validated and to test the control strategy formulated, and so assess the practical feasibility of applying linear optimal control theory to the non linear system. Thus the purposes in presenting the experimental results are

to indicate the nature of the system's behaviour, and to demonstrate the validity of the theory by showing that theoretical predictions are correct, or in the event of discrepancies existing, to indicate their form and postulate reasons for their existence.

The theoretical (mathematically modelled) system can be characterized in a number of ways. From the recorded operating point of the system during a test sequence (section 6.3.5), the K parameters of the model can be computed. (These have already been shown accurate under steady state open loop conditions in section 4.4). The gains used by the controller during a data logging sequence are also recorded for convenience (to save recalculating them), and from these and the calculated K parameters, the theoretical transfer functions of the closed loop system for each of the states can be computed. Thus the closed loop system can be characterized by its poles and zeros. Also from the transfer functions the response of each state variable to a change in reference voltage can be predicted and presented graphically for comparison with measured data, or some other criterion used for comparing predicted and actual responses.

Before proceeding further, consideration must be given to the nature of the data available from the experimental system. The states of the perturbed system are recorded in the NOVA store using the data logging facility of section 6.3.5. Thus the information available is of a discrete (digital) form limited by the accuracy and resolution of the transducers. In appendix A4 the resolution in measuring the states of the system is seen to be approximately 1 part in 80 when a sufficiently small input step (0.04 p.u.) is applied that the non linear system can be regarded as approximately linear. Besides the resolution limit, the environment of the transducers and recording system introduce some noise into the data.

With the foregoing facts in mind the problem to be considered is that of analysing the system performance to show that the closed loop system behaves in the manner predicted by its mathematical model, or to explain why it deviates.

Much research has been done in recent years on the problem of identification, whereby the system's dynamic characteristics are evaluated in either a stochastic or deterministic manner. These techniques could possibly be applied to the closed loop system and a comparison made between the model and the experimental system characteristics in the closed loop. For example, the cross correlation of input and output, when a pseudo random sequence of appropriate nature is applied, produces the impulse response of the system. However in a noisy environment, the impulse response so obtained can be significantly in error.²⁸ Even after the impulse response has been obtained, there are difficulties associated with determining the transfer function of the system. It has been shown that even when extremely accurate data is available for analysis, two transfer functions of widely different pole-zero configurations will produce responses fitting the available data equally well^{29,30}. During the course of determining the optimal controller, and predicting its behaviour, an example such as this arose, where the mathematical model had the following pole-zero configurations for two different operating and feedback conditions:

$$\begin{aligned} \text{T.F. (1)} &= \frac{\Delta v_t}{\Delta v_{\text{ref}}} = A \frac{(S + 4.2)(S - 3.6)}{(S + 0.67)(S + 2.3 + 0.9j)(S + 2.3 - 0.9j)} \\ \text{T.F. (2)} &= \frac{\Delta v_t}{\Delta v_{\text{ref}}} = B \frac{(S + 3.3 + 3.2j)(S + 3.3 - 3.2j)}{(S + 4.9)(S + 1.4)(S + 0.6)} \end{aligned}$$

Although of widely different pole-zero configuration, the difference in the step responses, up to the 2% settling time, of the two transfer functions was within 3% RMS of the steady state value. Thus after consideration of the

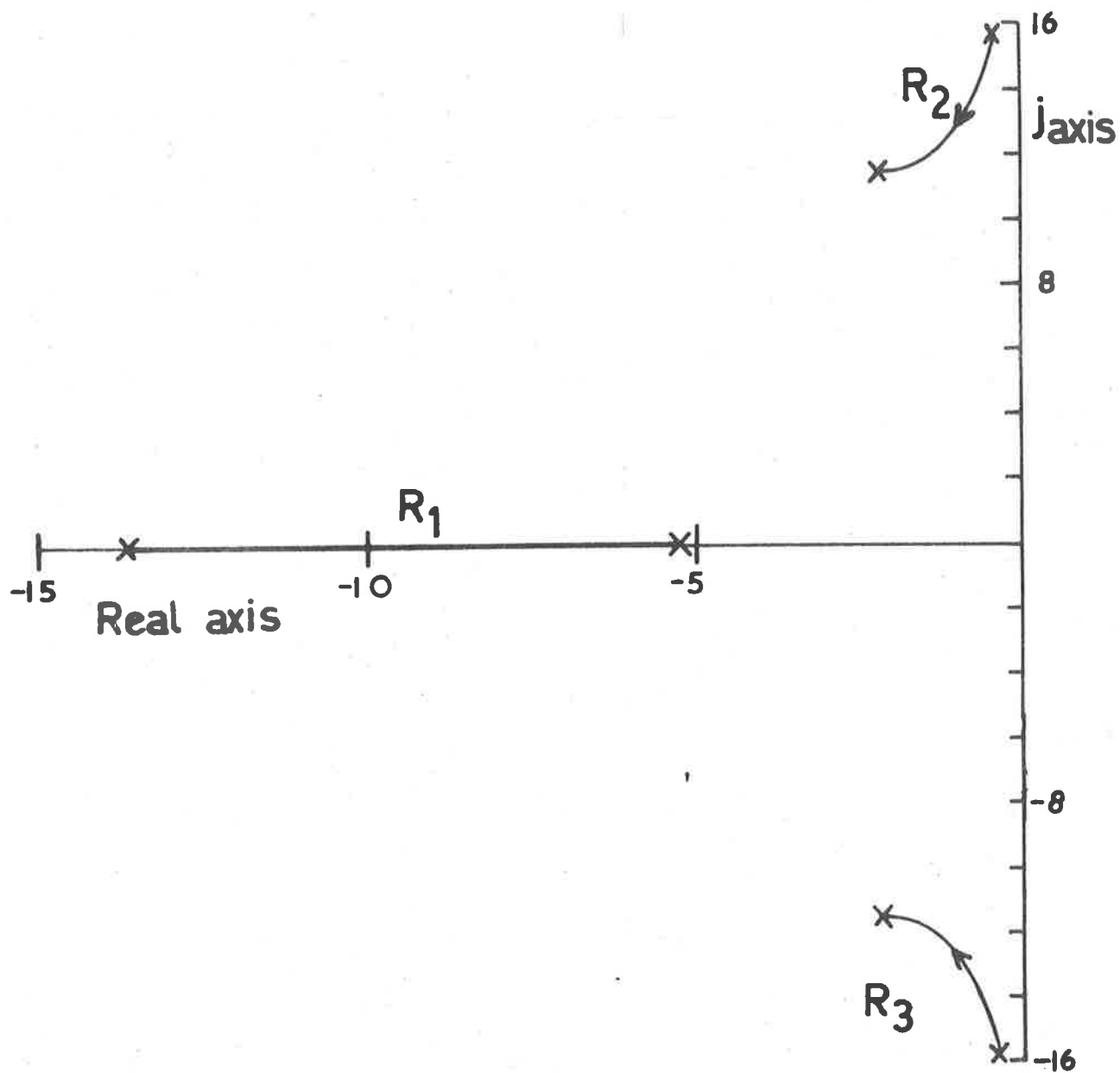


Figure 7.2 Predicted root locus plot of the optimal closed loop system for an operating point variation of $(P,Q) = (0.6, 0.0)$ to $(0.6, -1.0)$

available data, and the lengthy procedure required for an analysis of this type, it appears that system identification in this form is not particularly suited to performance appreciation and analysis.

The controller has been designed to provide optimization of the terminal voltage response to an input step perturbation. Thus a comparison of predicted and actual voltage responses would be one way of evaluating system performance. One method used for comparing the responses is to determine the RMS error between them, expressed as a percentage of the steady state value. This criterion is used for comparing predicted and actual impulse responses^{28, 29, 31} (expressed as a percentage of peak deviation) when attempting a pole-zero identification.

Graphing the response of each of the states to an input step allows immediate appreciation of the small signal behaviour of the system at various operating points in the PQ plane, and gives a quick visual comparison between the predicted and actual performance. In addition, an immediate qualitative assessment of the nature of any discrepancies can be made. The response waveforms of each of the states are graphed in section 7.3.

One further technique is used to compare the predicted and actual behaviour of the system. Figure 7.2 shows how the locations of the three poles of the closed loop mathematical model vary with the PQ operating point. The transfer function of input step to machine speed deviation takes the form

$$\frac{\Delta\omega}{\Delta v_{ref}} = \frac{N_1 S}{(S + R_1)(S + R_2)(S + R_3)}$$

where R_1 , R_2 and R_3 are the roots of the characteristic polynomial (real or complex). Comparing this with figure 7.2 the contribution of the real root, R_1 , to the time response of the system is quickly damped ($Ae^{R_1 t} \rightarrow 0$ rapidly), and, as can be seen in figure 7.5, the response has the characteristic form

of a second order, under damped, oscillatory system, at least for leading reactive powers up to -1.0 per unit. Ignoring the first swing of the oscillatory response (during which time $Ae^{-R_1 t}$ becomes negligible) the characteristic polynomial is regarded as being $p^2 + 2\zeta \omega_n p + \omega_n^2$, and the damping factor ζ and the natural frequency ω_n are determined from the plotted system responses³². The predicted values are obtained from the complex conjugate pole pairs R_2 and R_3 ($p^2 + 2\zeta \omega_n p + \omega_n^2 \equiv p^2 + (R_2 + R_3)p + R_2 R_3$).

While a comparison of damping ratios and natural frequencies is equivalent to the previously rejected comparison of pole locations, the distinctive 2nd order nature of the speed response in this region of the PQ plane is suited to a quantitative comparison of this type. This comparison is tabulated in section 7.4 together with an evaluation of the RMS error between predicted and actual voltage response.

7.3 RESPONSE WAVEFORMS

The observed responses presented in this section were obtained by applying a small step change in reference voltage to the experimental system. Initially the system was operating in the steady state in the vicinity of one of the gain grid points, and then perturbed across the domain of that point. Following the step change the controller gains were held constant even when the instantaneous operating point moved into an adjacent domain. The values of the states and the relevant operating data were recorded using the data logging facility described in section 6.3.5.

Predicted responses were calculated on the 6400 computer using the values for the model parameters which are applicable to an operating point lying midway between the initial and final steady state points, and using the same gains selected by the regulator during the test sequence.

Initially 0.01 p.u. steps in reference voltage were used to perturb the system. This particularly small step was used to minimize the

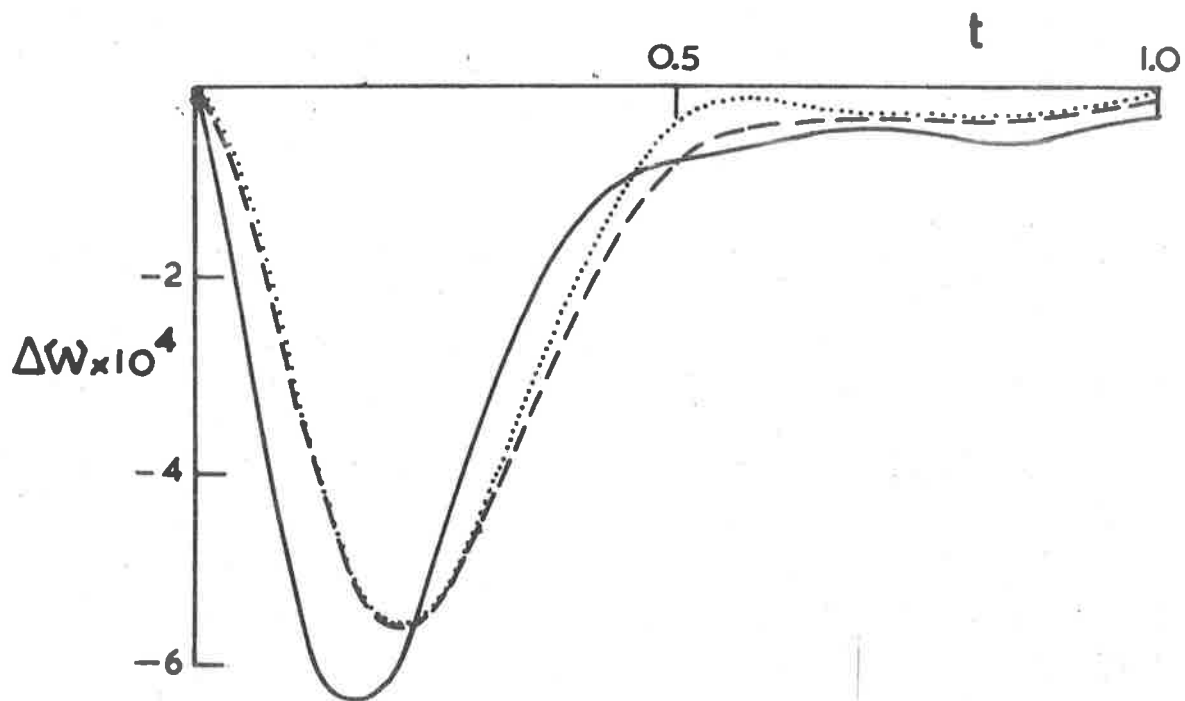


Figure 7.3(a) Operating point $P = 0.6$ $Q = -1.6$

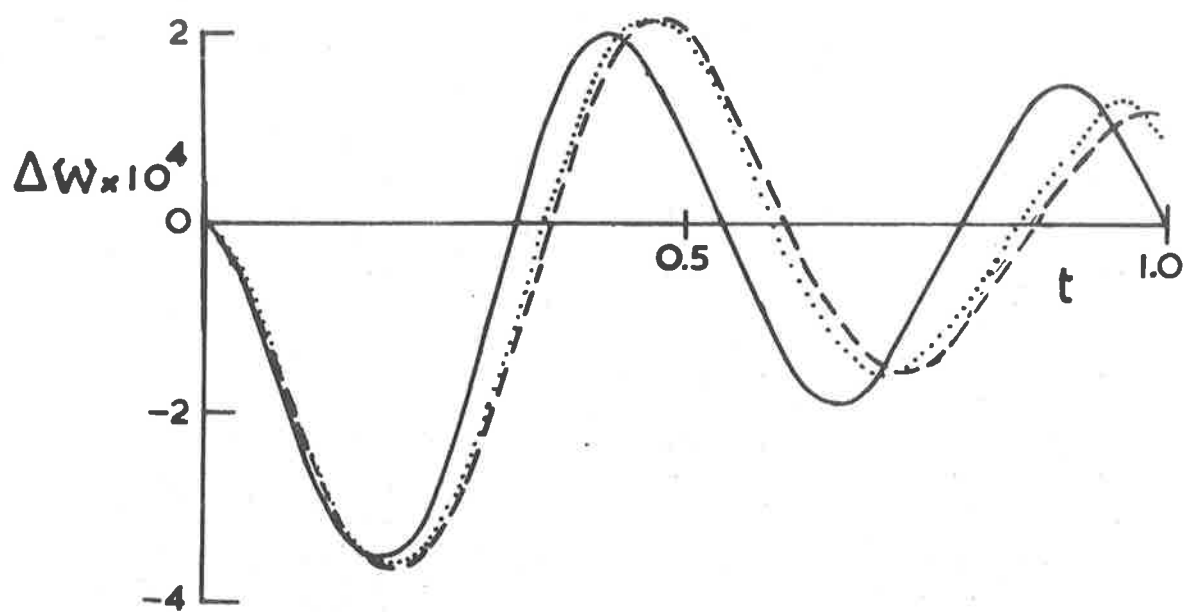


Figure 7.3(b) Operating point $P = 0.6$ $Q = -0.8$

Figure 7.3 The effect of the magnitude of the step change in reference voltage on speed response.

- Predicted response
- Observed response 0.01 p.u. step
- Observed response 0.04 p.u. step
scaled by $\frac{1}{4}$ for purposes of comparison.

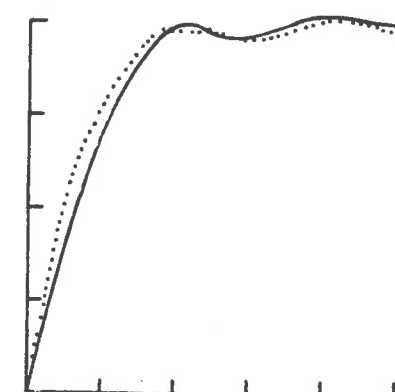
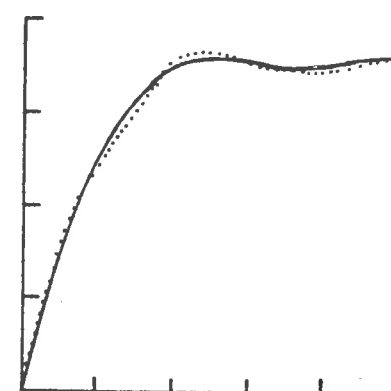
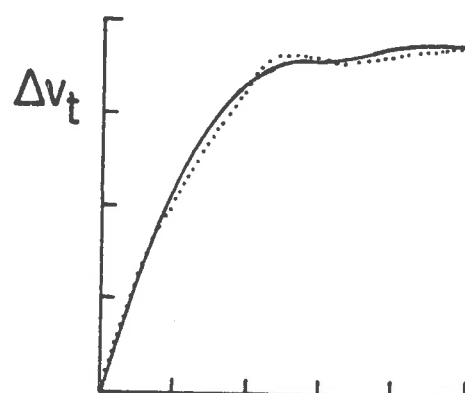
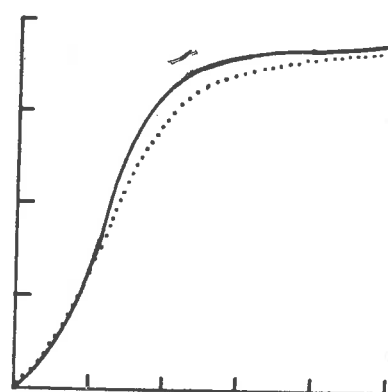
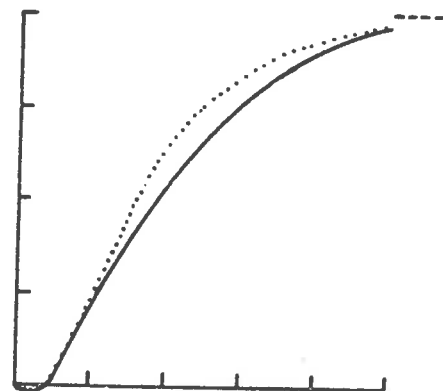
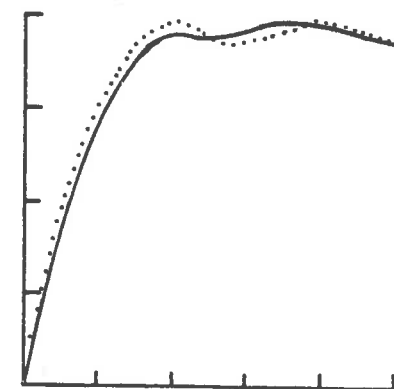
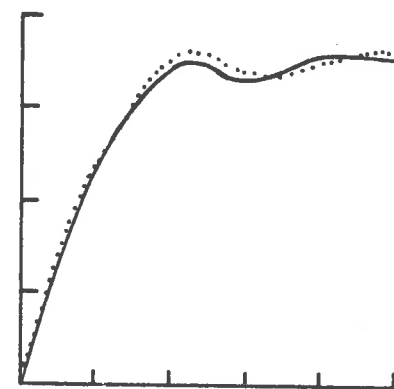
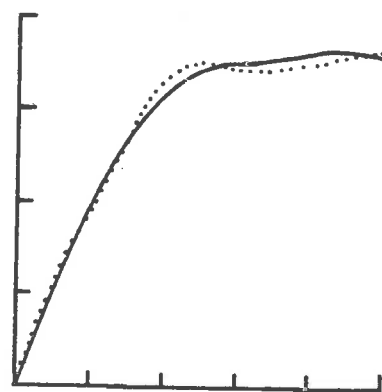
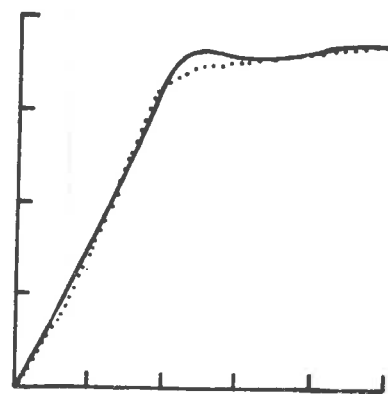
effects of the system non linearity on the observed responses. Close agreement between the predicted and observed responses was obtained in a number of cases where errors due to noise were small. Because the signal levels associated with the states are low, the effect of transducer resolution and input noise, particularly that due to mains frequency variations (see section 8.4), in general precluded accurate analysis of experimental responses. Subsequent tests were conducted using an input step of 0.04 p.u. The increase in step size produce an improvement in the signal to noise ratio and caused no significant deterioration in the agreement between predicted and actual terminal voltage response. However some change was evident in the torque and speed responses. Examples of the effect of the increased step size on the speed response are shown in figure 7.3.

In response to a voltage reference change of 0.04 p.u., the system operating point shifts from its initial steady state value by approximately 0.13 p.u. parallel to the Q axis to reach its end point steady state position. With constant prime mover power (and assuming constant losses) the real power remains unchanged. To ensure that the deterioration between the predicted and actual speed responses with the increased step size, as seen in figure 7.3, did not originate from an incorrect choice of the effective operating point for the linear mathematical model, performance was predicted for the system operating at both the initial and end point steady state conditions. Little change in the predicted performances was observed. From this and subsequent large signal tests (see section 8.2), it was concluded that the non linear nature of the experimental system was influencing the observed responses.

The form of the voltage response varies little from point to point in the operating plane and in general is that of a third order system with a dominant real pole. Figure 7.4 shows the predicted and recorded terminal voltage responses in the vicinity of selected grid points in the PQ plane.

A wide variation in the behaviour of both machine speed and electrical torque was observed over the operating plane. Figures 7.5 and 7.6 show predicted and measured responses for each of these system states.

CURRENT
RATING
EXCEEDED



Δv_t

time

09

P

06

03

-20

Q

-1.6

-1.2

-0.8

-0.4

Figure 7.4 Terminal voltage responses to an 0.04 p.u. step change in reference voltage. The grid point about which the system is perturbed is given by the coordinates of the origin of the response curves.

Response curve scales:

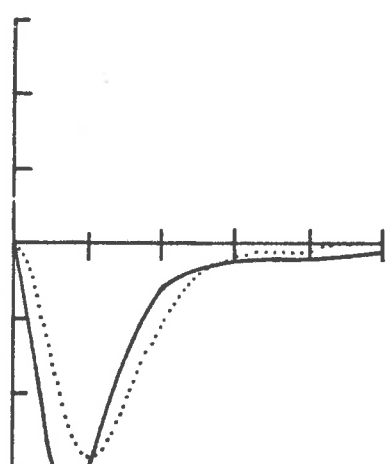
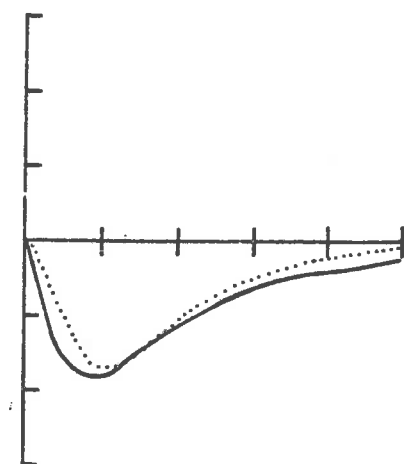
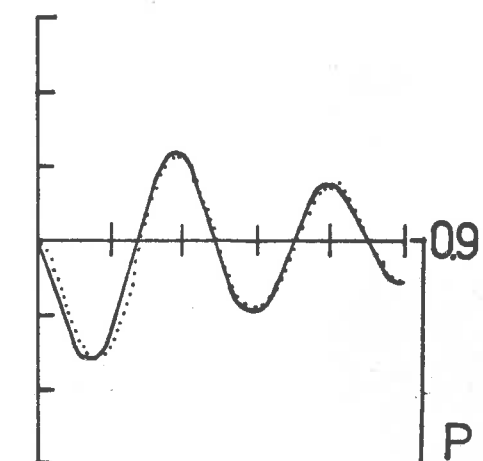
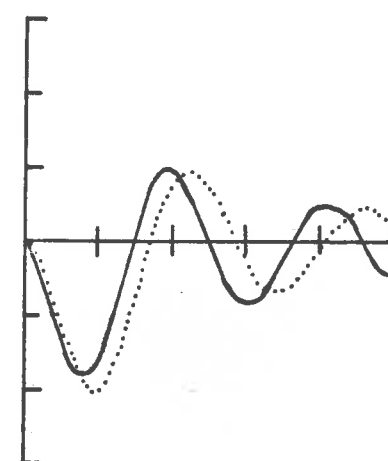
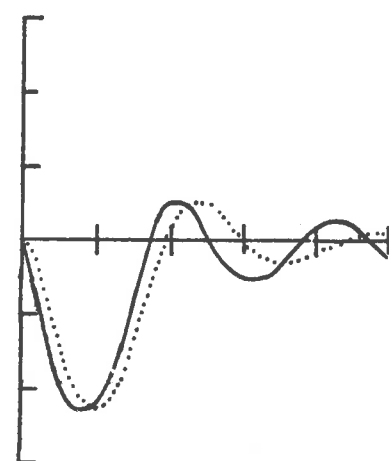
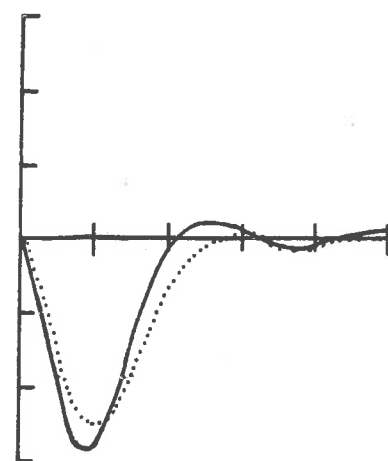
Vertical Δv_t .01 p.u. / div.

Horizontal time 0.2 s / div.

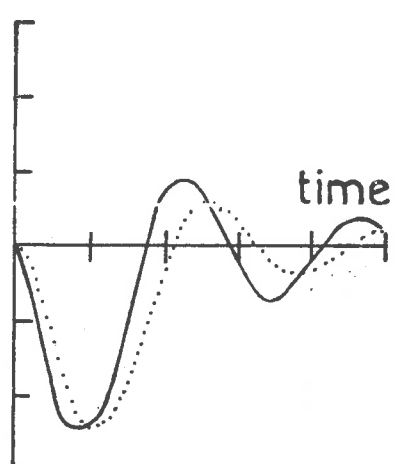
Predicted responses

Recorded responses —————

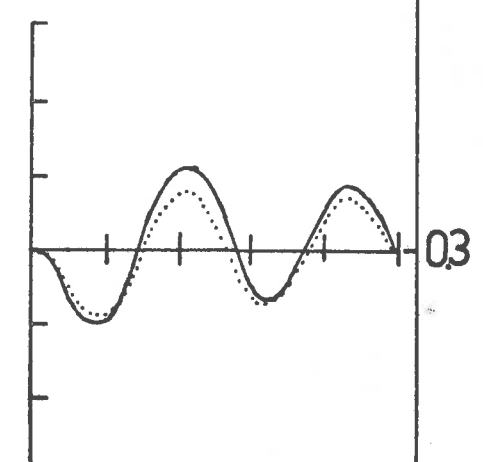
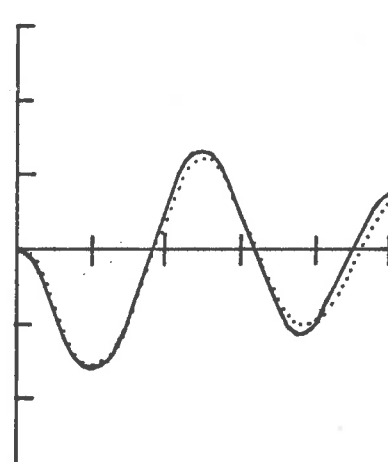
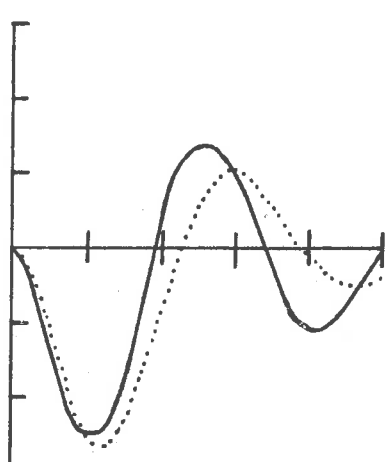
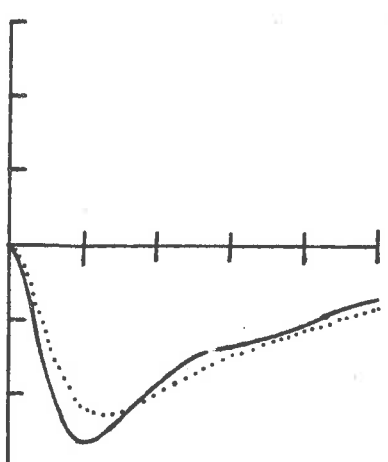
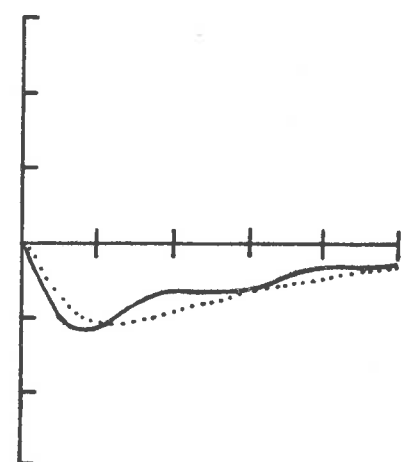
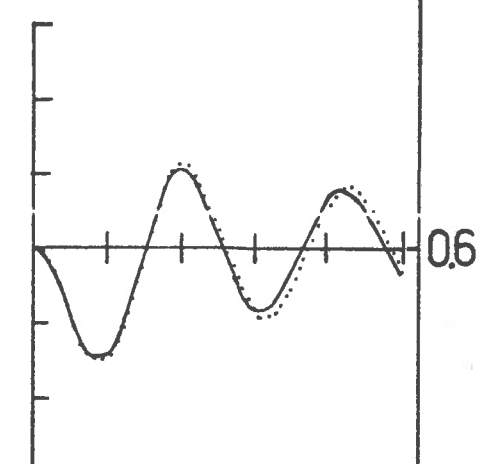
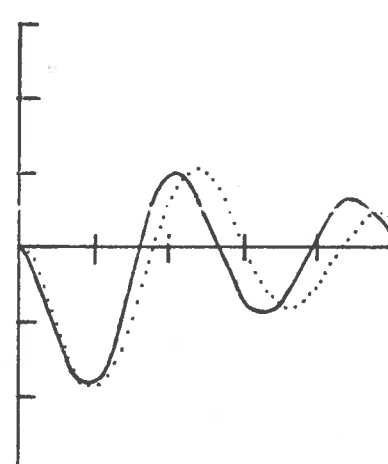
CURRENT
RATING
EXCEEDED



Δw



time



2.0

Q

1.6

1.2

0.8

0.4

0.9

P

0.6

0.3

Figure 7.5 Machine speed responses to an 0.04 p.u.
step change in reference voltage. The grid
point about which the system is perturbed
is given by the coordinates of the origin
of the response curves.

Response curves scales:

Vertical $\Delta\omega$ 0.0008 p.u. / div.

Horizontal time 0.2 s / div.

Predicted responses

Recorded responses —————

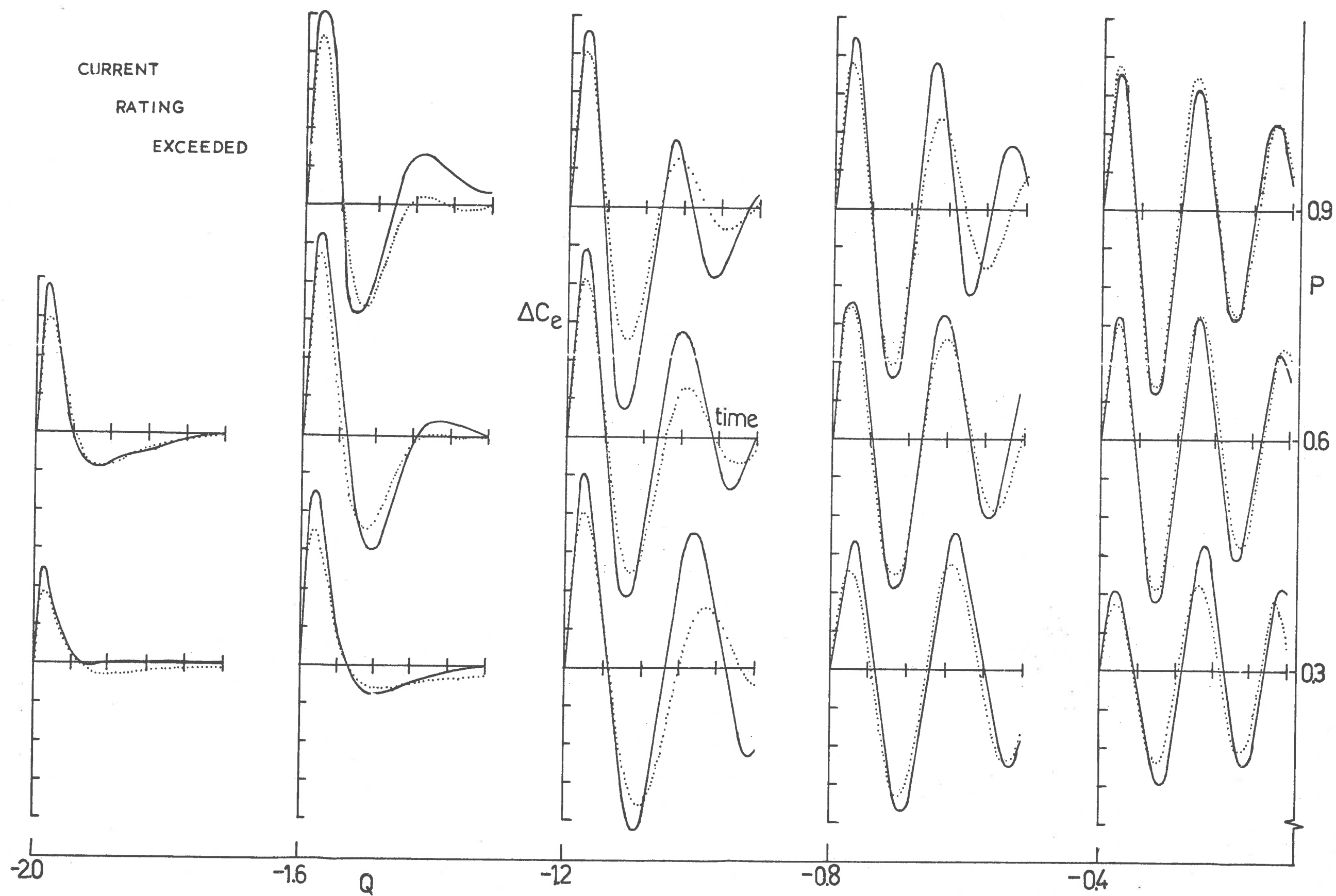


Figure 7.6 Electrical torque responses to an 0.04 p.u.
step change in reference voltage. The grid
point about which the system is perturbed
is given by the coordinates of the origin
of the response curves

Response curve scales:

Vertical ΔC_e 0.008 p.u. / div.

Horizontal time 0.2 s / div.

Predicted responses

Recorded responses _____

7.4 TABULATION OF THE SYSTEM BEHAVIOUR

As explained in section 7.2 the speed response curves are regarded as being those of a second order system up to leading reactive powers of 1 p.u., and the natural frequency of oscillation and the damping ratio are determined from plotted response data.

The same test conditions apply as for the previous section. That is, the system is perturbed from an initial steady state operating point by an 0.04 p.u. step change in reference voltage, and the states of the system and the relevant operating information recorded. The data recorded in the NOVA store is punched onto paper tape and read into the 6400 computer. The information describing the response of the system states is digitally filtered using a fourth order Bessel filter (minimum group delay) with a cut off frequency of 5Hz. This reduces the quantization noise (due to the 1 part in 80 transducer resolution) and other wide band noise picked up by the transducers. The filtered responses together with the filtered version of the predicted responses are then plotted by the computer. In addition the RMS deviation of the terminal voltage from its predicted response is computed over the 2% settling time and expressed as a percentage of the input step (of 0.04 p.u.).

The frequencies and damping ratios of the experimental system shown in the following tables were derived from the plotted speed responses. As mentioned previously the predicted values are obtained from the complex pole pair R_2 and R_3 .

Five test runs were performed in the domains of each of the grid points to produce the following data.

Q_{bus}	Frequency ω_n (Hz)		Damping Ratio		% RMS voltage error
	Predicted	Deviation in observed	Predicted	Deviation in observed	
-0.2	2.67	0.00	0.008	+0.003	9.5
		+0.01		+0.003	8.1
		0.00		+0.006	9.1
		-0.02		+0.002	8.9
		-0.01		+0.001	8.5
-0.4	2.22	-0.02	0.018	+0.002	9.8
		0.00		-0.001	9.3
		+0.03		+0.001	9.5
		0.00		-0.001	10.2
		-0.01		+0.004	9.7
-0.6	2.01	-0.01	0.024	+0.005	8.3
		+0.01		+0.010	8.4
		-0.01		-0.006	9.9
		+0.02		+0.003	9.2
		0.00		-0.005	8.9
-0.8	1.84	+0.02	0.045	-0.005	8.6
		+0.06		-0.006	7.4
		+0.02		-0.001	6.0
		+0.03		-0.001	6.9
		+0.01		-0.005	7.9
-1.0	1.64	+0.21	0.078	-0.018	4.3
		+0.08		-0.004	6.6
		+0.19		-0.008	7.4
		+0.10		-0.005	6.3
		+0.13		-0.025	5.0

Table 7.1 System operating about a grid point of real power 0.3 p.u., reactive power given by Q_{bus} .

Natural frequencies and damping ratios of speed response curves. Observed values are presented as deviations from those predicted.

% RMS deviation of terminal voltage from its predicted response.

Q_{bus}	Frequency ω_n (Hz)		Damping Ratio		% RMS voltage error
	Predicted	Deviation in observed	Predicted	Deviation in observed	
-0.2	2.42	+0.01 0.00 +0.03 -0.04 0.00	0.034	-0.001 -0.009 +0.004 -0.003 0.000	8.1 9.8 8.9 10.5 9.6
-0.4	2.25	+0.13 +0.06 +0.10 +0.06 0.00	0.044	+0.004 +0.001 +0.001 -0.001 -0.001	7.9 7.6 7.5 8.9 8.4
-0.6	2.16	+0.14 +0.08 +0.12 +0.08 +0.11	0.058	-0.08 +0.06 -0.05 0.00 -0.07	8.6 5.9 7.1 6.1 6.4
-0.8	2.00	+0.06 +0.15 +0.10 +0.15 +0.17	0.092	-0.014 -0.018 -0.020 -0.018 -0.020	5.3 7.1 8.5 3.4 4.2
-1.0	1.90	+0.10 +0.10 +0.16 +0.16 +0.10	0.144	-0.042 -0.047 -0.049 -0.047 -0.042	4.1 4.2 4.5 6.0 6.8

Table 7.2 System operating about a grid point of real power 0.6 p.u., reactive power given by Q_{bus}

Natural frequencies and damping ratios of speed response curves. Observed values are presented as deviations from those predicted.

% RMS deviation of terminal voltage from its predicted response.

Q_{bus}	Frequency ω_n (Hz)		Damping Ratio		% RMS voltage error
	Predicted	Deviation in observed	Predicted	Deviation in observed	
-0.2	No tests performed as terminal voltage exceeds the 1.0 p.u. maximum of the transducer.				
-0.4	2.36	+0.13 +0.14 +0.14 +0.14 +0.16	0.069	-0.021 -0.019 -0.014 -0.003 -0.008	9.5 10.5 8.8 8.0 9.1
-0.6	2.24	+0.20 +0.12 +0.18 +0.14 +0.14	0.093	-0.008 -0.013 -0.011 -0.012 -0.009	6.9 9.3 9.9 6.6 10.3
-0.8	2.14	+0.17 +0.16 +0.16 +0.16 +0.18	0.120	-0.011 -0.018 -0.010 -0.018 -0.015	6.1 3.9 5.9 5.6 6.3
-1.0	2.08	+0.30 +0.21 +0.20 +0.25 +0.28	0.150	-0.030 -0.025 -0.012 -0.033 -0.033	6.4 2.1 5.4 5.1 2.6

Table 7-3 System operating about a grid point of real power 0.9 p.u., reactive power given by Q_{bus} .

Natural frequencies and damping ratios of speed response curves. Observed values are presented as deviations from those predicted.

% RMS deviation of terminal voltage from its predicted response.

7.5 DISCUSSION

7.5.1 ON THE NATURE OF THE SYSTEM BEHAVIOUR

As detailed in Chapter 2 the physical objectives of the controlled system are defined mathematically to produce a cost function. Mainly from considerations of the ease of implementation, the cost function for the experimental system was chosen in section 3.3 as:

$$J = \frac{1}{2} \int_{t_0}^{t_f} (\Delta v_t - \Delta v_{ref})^2 + \Delta v_f^2 g dt$$

With $g = 0.05$, the resulting controller operates mainly to minimize voltage error. This produces the type of voltage response shown in figure 7.4, but, with no explicit expression for torque and speed in the cost function, provides only incidental control on the damping of these two states.

Figures 7.4, 7.5 and 7.6 show that the amplitude of the perturbations remain relatively constant across the PQ plane for each state. Figures 4.21 and 4.22 show that for operation near the unity power factor region the feedback gains F_2 and F_3 for speed and torque are relatively small, so that the contribution of the feedback signals from these states to the control force are negligible in comparison with the voltage feedback term. Thus in the unity power factor region the effective damping is small (as is the case with the unregulated machine).

As the leading reactive power increases the speed and torque feedback gains increase. Thus, with little change in the magnitude of the perturbations, the contributions of these states to the control force increase, resulting in the higher degree of damping obtained.

This effect highlights the need for careful consideration of the cost function chosen for any specific system. It is apparent that to gain satisfactory damping on speed and torque these states must appear explicitly in the cost function, and be weighted by an appropriate penalty factor to obtain the desired behaviour. The work of Tyler and Tuteur¹⁴, and other similar investigations, could be used to select weighting matrix elements which position the roots of the characteristic polynomial at a satisfactory distance from the imaginary axis of the s plane. However all the other factors detailed in section 3.3 which govern the choice of the cost function, must still be observed, and once again the difficulties associated with choosing a suitable cost function are encountered.

7.5.2 ON THE ACCURACY OF THE PREDICTED PERFORMANCE

The K parameters of the open loop model have been shown in section 4.4 to be accurate when measured under steady state conditions. For an 0.01 p.u. step perturbation, figure 7.3 shows the close agreement between the predicted and measured speed response. These two observations tend to confirm the accuracy of the mathematical model for the small signal system. The deterioration of the agreement between the predicted and actual speed responses for an 0.04 p.u. step is attributed to the non linear nature of the system, although it is noted that no significant change occurs in the voltage response with the larger step. This is because the chosen cost function ensures insensitivity of the heavily weighted state (or deviation of this state from a required value) to variations in the system parameters¹².

The non linear nature of the system is reflected by the variation of the computed optimal gains with the steady state operating point. Referring to figures 4.20 to 4.23 the regions in the PQ plane where the gains vary rapidly correspond to the regions where the predicted and measured frequency and damping ratios of the speed response, given in tables 7-1 to 7-3, most obviously disagree. That is, as the leading power factor increases so does the margin between the two sets of figures, at least up to the value of Q (-1.0 p.u.) for which these measurements were made.

While the comparisons for speed and torque are deteriorating with increasing leading Q, (up to -1.0 p.u.) the agreement between predicted and actual voltage response is improving. From the values in tables 7.1 and 7.3 the average RMS error varies from 8.8% at $(P, Q) = (0.3, -0.2)$ to 5.9% at $(0.3, -1.0)$ and from 9.2% at $(0.9, -0.4)$ to 4.3% at $(0.9, -1.0)$. Thus the observation that the terminal voltage response is unaffected by the non linear nature of the system is apparently confirmed, as the best agreement between predicted and actual voltage responses are obtained in the region of the PQ plane where the rate of change of feedback gains, and the comparison of speed responses, indicate that the system is highly non linear.

From figure 7-3, the disagreement between the voltage curves for low reactive powers arises from the fact that the observed response is slower than that predicted. In this region the dominant real pole of the voltage transfer function produces an effective time constant of about 80 ms and it would appear that the time constants of the transducers (about 15 ms ; see Appendix

A) are influencing the behaviour of the closed loop system. In general, however, agreement between predicted and observed behaviour is very close, particularly when it is remembered that an error of 8% RMS is expressed as a percentage of the input step perturbation of 0.04 p.u.; that is the voltage errors are of the order of 0.003 p.u.

Thus linear optimal control theory is successfully used to control the non linear system under conditions where the small signal mathematical model is a satisfactory representation of the system, and allowing for some evidence of non linearity, predictions of its behaviour can be accurately made.

CHAPTER 8

EFFECTS OF DEPARTURE FROM
THE IDEAL MODEL

8.1 INTRODUCTION

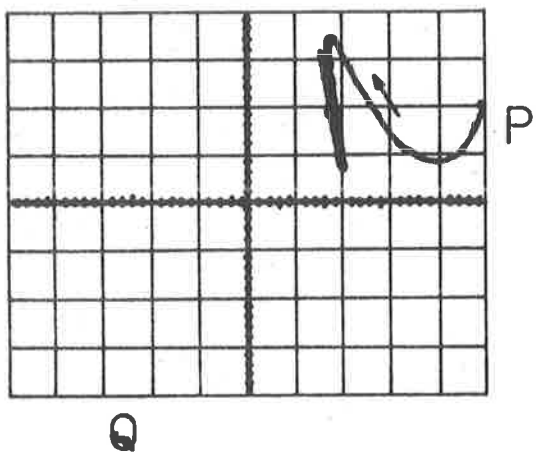
Optimal (or near optimal) control of the generator excitation is achieved when the operating conditions are such that the assumptions made in the derivation of the mathematical model are valid (section 3.6). System operation was investigated in three cases where these assumptions are clearly invalid, and the effect of varying the tie line reactance from its assumed value of 0.29 p.u. is also assessed.

8.2 LARGE SIGNAL BEHAVIOUR

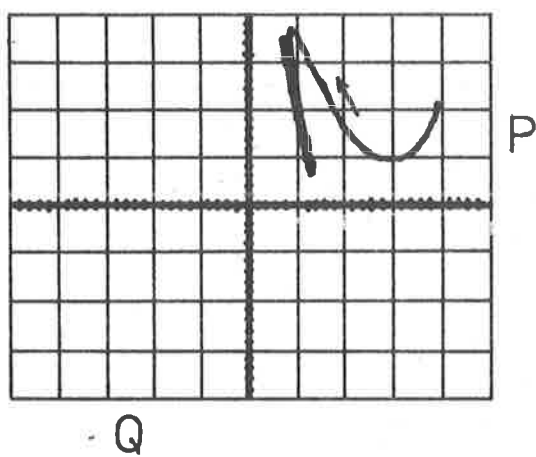
The mathematical model of the experimental system derived in Chapter 3 was obtained basically from the small signal linearisation of Park's equations. Therefore the model can be strictly accurate only at a given steady state operating point. In section 7.3 it was shown that system behaviour is accurately predicted by the model when it is perturbed by a step of 0.01 p.u. A larger step of 0.04 p.u. caused noticeable discrepancies in the speed and torque responses. To assess the effect of a further increase in the input step size, the system was perturbed by an 0.20 p.u. change in reference voltage, and it was found that system performance could no longer be accurately predicted.

Figure 7.5 shows the machine speed response to an 0.04 p.u. step in reference voltage and from these curves (or tables 7.1, 7.2, 7.3) it is seen that the observed speed response has a higher frequency and is more lightly damped than the predicted response, particularly in the region of the open loop stability limit of the system (that is around $Q = -1.0$ p.u.). As the step size is increased, this trend continues to the point where an 0.20 p.u. step results in a limit cycle of machine speed and torque in this region of the operating plane. Elsewhere in the operating region investigated, the observed performance corresponds more closely to that predicted for the small signal model. For higher values of leading reactive power, the speed and torque

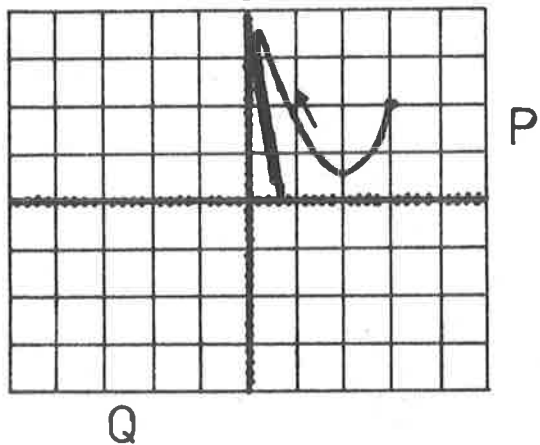
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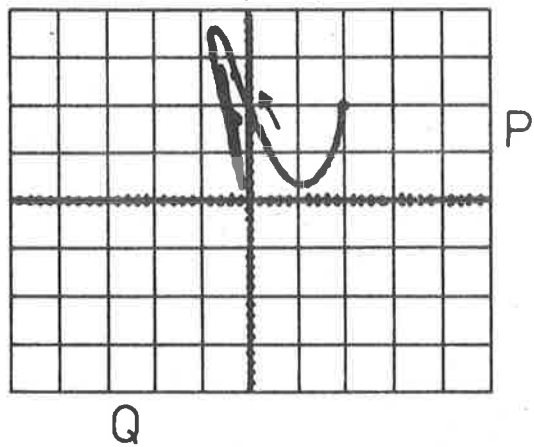
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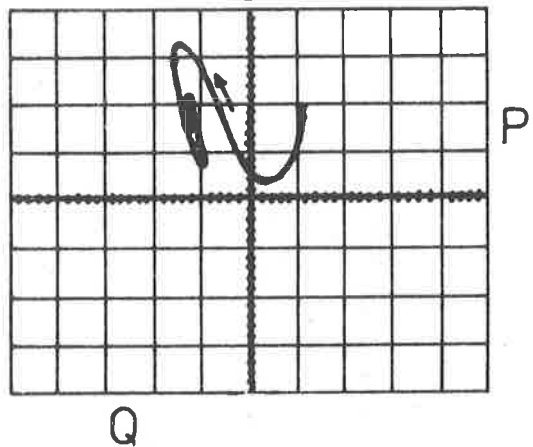
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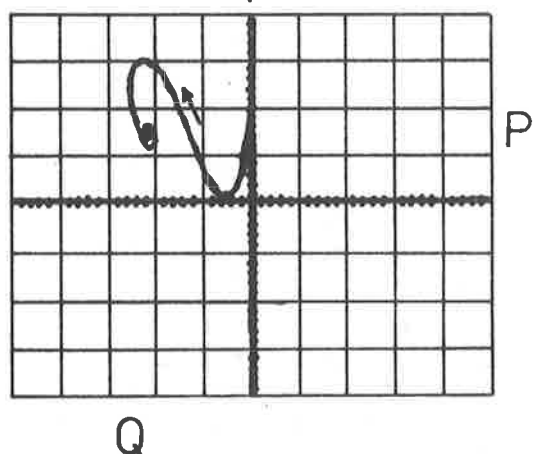
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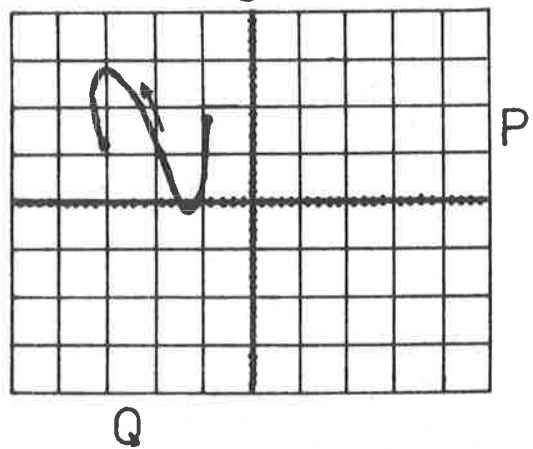
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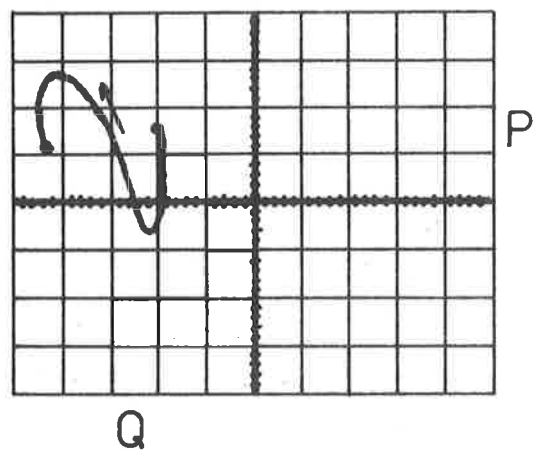


Figure 8.1 Trajectories of the operating point
in the PQ plane for the closed
loop system subjected to a step
change in reference voltage of -0.20 p.u.

The vertical axis is the real power measured
by the transducer at the infinite bus, scaled to
 0.1 p.u./div. (i.e. 0.8 p.u. max.)

The horizontal axis is the reactive power delivered
to the bus, scaled to 0.2 p.u./div. (i.e. -2.0 p.u. max.)

Movement along the trajectory is in the direction
indicated by the arrows.

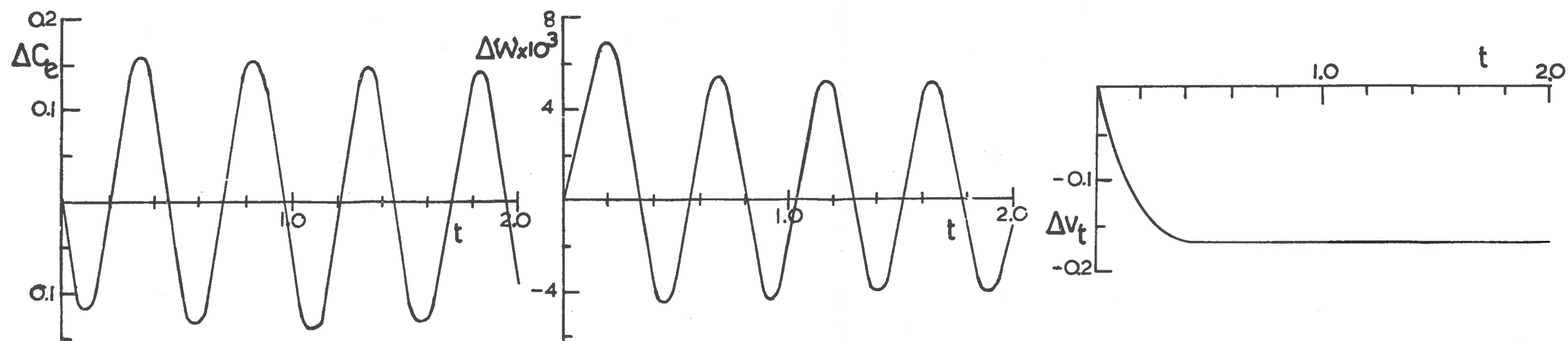


Figure 8.2 The recorded deviations in torque, speed and terminal voltage resulting from a step change in reference voltage of -0.20 p.u. Initial operating point $P'=0.6$, $Q=-0.4$ p.u.
Final operating point - oscillatory about $P'=0.57$, $Q=-0.92$ p.u. (P' recorded at the bus, without power correction)

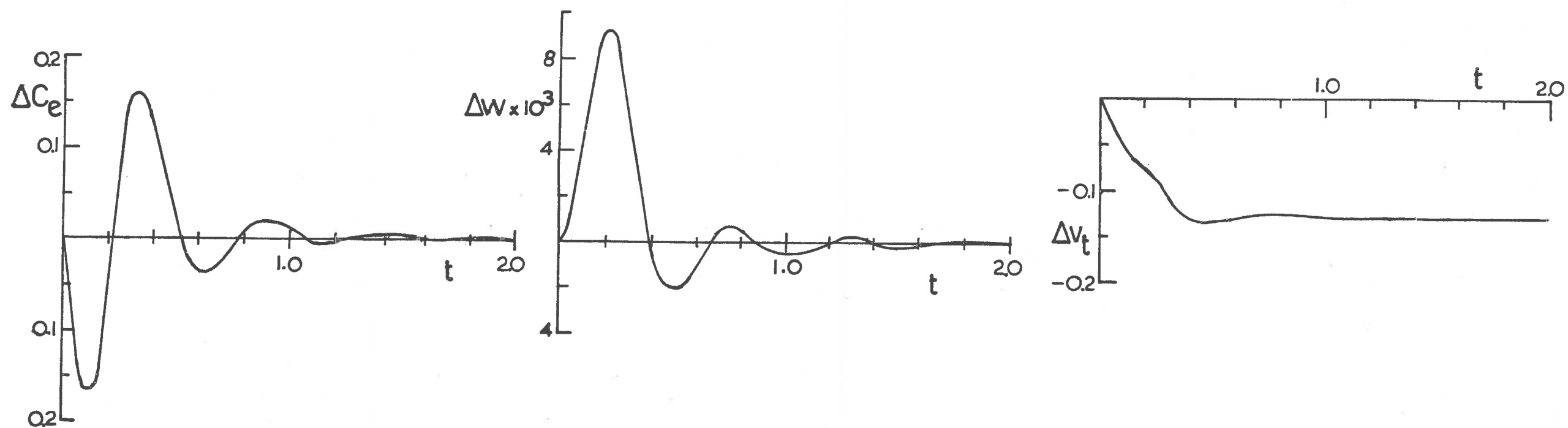


Figure 8.3 The recorded deviations in torque speed and terminal voltage resulting from a step change in reference voltage of -0.20 p.u. Initial operating point $P'=0.58$, $Q=-1.0$ p.u.
Final operating point $P'=0.54$, $Q=-1.42$ p.u. (P' recorded at the bus, without power correction)

responses become heavily damped and the system quickly settles to its final steady state operating point.

Figure 8.1 shows a series of observed trajectories of the operating point in the PQ plane resulting from an input step of 0.20 p.u. in reference voltage.

Two examples of the limit cycling are seen in figures 8.1b and 8.1c where the operating point oscillates about the end point of the trajectory, along a line of constant machine terminal voltage. In figure 8.2 the recorded deviations of terminal voltage, speed and torque are shown corresponding to the trajectory of figure 8.1c. The terminal voltage is seen to reach a steady state value with an effective time constant of approximately 0.14 seconds. Comparing the torque response in figure 8.2 with the operating point trajectory in figure 8.1c, it is seen that the system enters the limit cycle about the end point shortly after the first positive swing in electrical torque ($\Delta S \Delta P \equiv \Delta C_e$, this corresponds to the first positive peak in the real power of the operating point trajectory). From the torque response curve this is seen to take approximately 0.4 seconds which is the time taken for the terminal voltage to settle to its steady state value. Subsequently the terminal voltage is held constant by the regulator while the torque and speed continue in an undamped oscillatory mode.

Figure 8.3 shows the recorded states of the system when operating in a region of higher leading reactive power. This corresponds to the operating point trajectory shown in figure 8.1f. In this case speed and torque are well damped, and the terminal voltage settles to its steady state value almost coincidentally with the torque returning to zero deviation.

It should be noted that the torque responses were recorded in the NOVA store with the loss correction factor of section 6.3.4 operating, whereas the operating point trajectories were obtained by applying the outputs

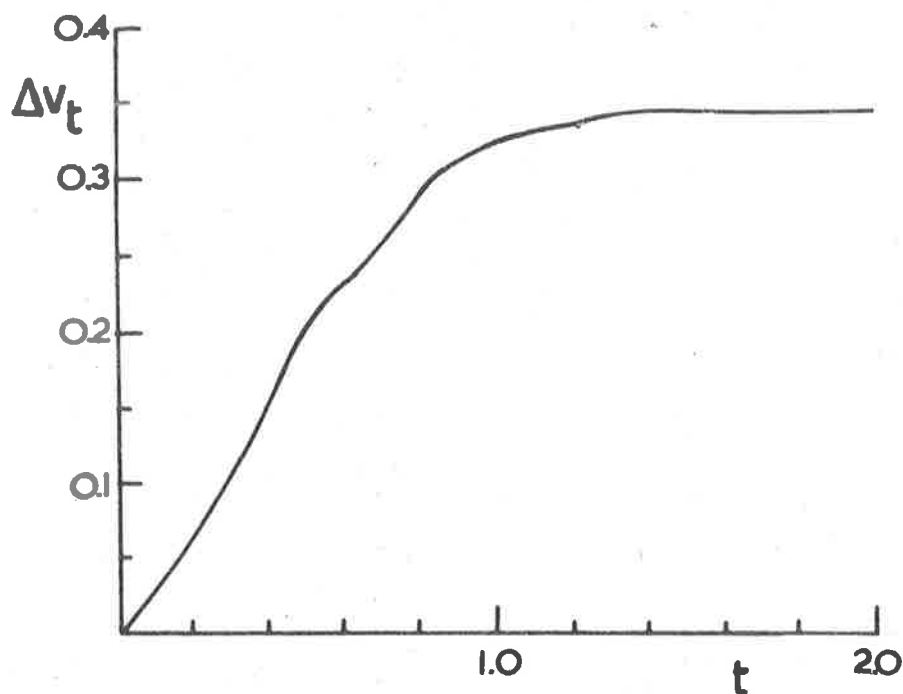


Figure 8.4a

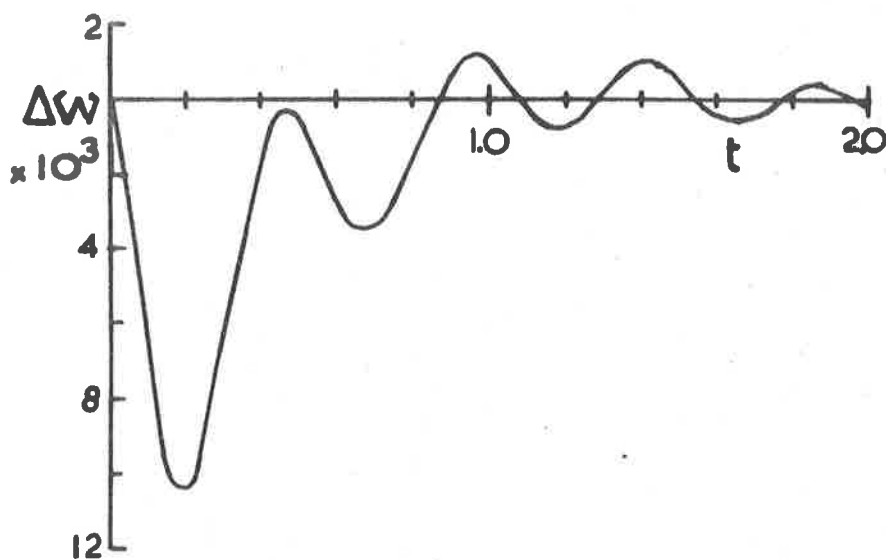


Figure 8.4b

Figure 8.4 The recorded response of the system states to an input step of +0.10 p.u. Initial operating point $P = 0.55$, $Q = -1.44$ p.u. Final operating point $P = 0.60$, $Q = -0.56$ p.u. (P recorded at the bus).

of the power transducers (measuring power at the infinite bus) directly to the X and Y inputs of a storage CRO. Thus the trajectories are shown without corrections for power losses.

So far in this section large signal behaviour has been demonstrated for a negative step in reference voltage; that is the operating point traverses from low to high leading reactive power. Figures 8.4 and 8.5 provide an example of the observed performance when the system is perturbed by a step of +0.20 p.u. Some interesting points arise.

When a step of -0.20 p.u. is applied the operating point is seen to shift by approximately -0.40 to -0.60 p.u. parallel to the Q axis, and the terminal voltage changes by -0.13 to -0.20 p.u. (figures 8.1, 8.2, 8.3). In the given example (which is typical of the observed behaviour) the operating point changes from $Q = -1.44$ to $Q = -0.56$ p.u., and the terminal voltage increases by 0.35 p.u. Thus a much larger swing results from a positive than from a negative input step. Unlike the test conditions of Chapter 7, for these large perturbations regulator gains are allowed to vary as the operating point traverses successive domains. In Chapter 7 operation was confined within (or almost within) one grid point domain by virtue of the applied step size, and other than inverting the responses, no appreciable variation resulted from changing the sign of the input step. The magnitudes of the torque and speed transients are not noticeably affected by changing the sign of the input step.

A detailed investigation of this phenomena was not performed and this characteristic of the non linear system remains unexplained.

One other point which bears comment is the change in nature of the system response. By perturbing the operating point from a region of high leading Q, the large speed and torque feedback gains applied initially quickly damp these two states, so that by the time the operating point has

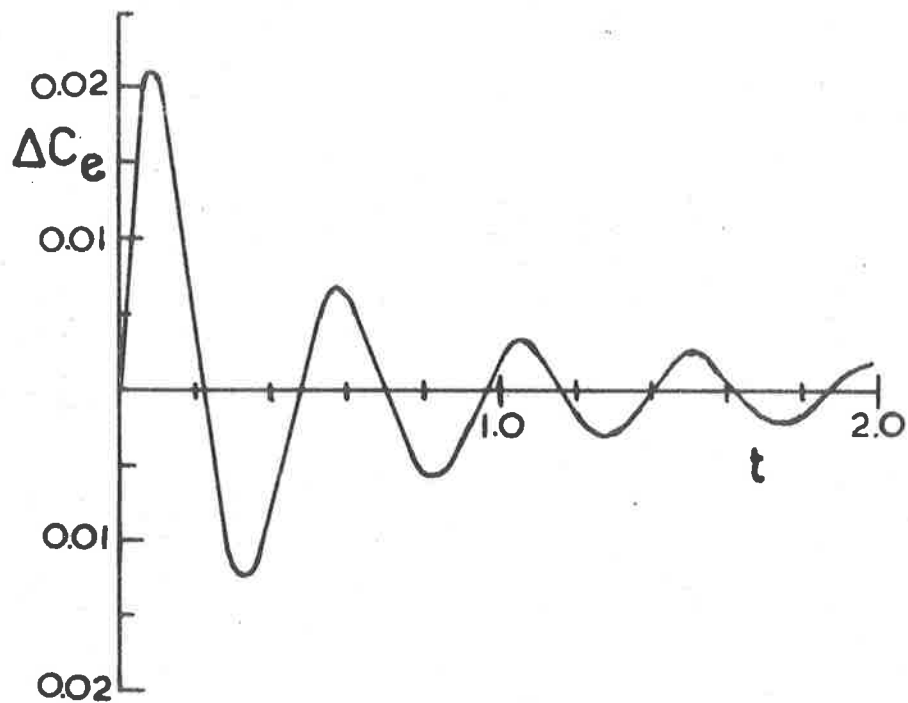


Figure 8.4c

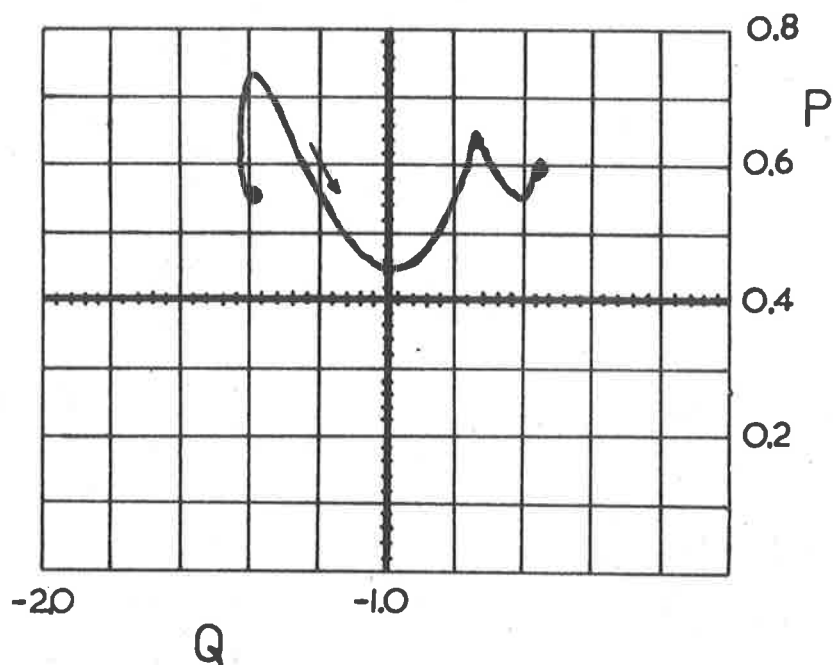


Figure 8.5 Trajectory of the operating point in the PQ plane for a + 0.20 p.u. step change in reference voltage.
Initial operating point $P = 0.55$, $Q = -1.44$ p.u.
Final operating point $P = 0.60$, $Q = -0.56$ p.u.

traversed to the region of the plane where speed and torque feedback gains are negligible, the magnitude of these states have been greatly reduced. Lightly damped oscillations continue for a short period until steady state operating conditions are achieved. In perturbing the operating point from near unity power factor to a low value of leading Q (i.e. in the other direction to reach the same operating region) the initial large swings are not damped and the observed limit cycling results.

Checks were made throughout these large signal tests to ensure that limiting was not occurring anywhere within the control loop, and the observed behaviour cannot be attributed to this possibility. In general, the nature of the large signal response of the system states are as predicted by the mathematical model. The terminal voltage response has the appearance of a system with a dominant real pole, and the torque and speed responses are underdamped oscillator for low leading reactive power (although prone to the demonstrated limit cycling) changing to an overdamped response at high leading Q . Probably the most interesting feature of the large signal performance is the ability of the controller to maintain constant terminal voltage even when the system enters a limit cycle. Coupled with the fact that the controller was optimized for voltage regulation, this suggests that a cost function could be chosen which would result in an improved overall system performance under large signal disturbances. Thus a cost function producing a controller which provides adequate speed and torque damping in the small signal model over the whole operating plane, instead of just the high leading Q region as at present, could be expected to provide this damping under large signal conditions as now occurs in the high leading Q region only.

Consideration could also be given as to whether there is any merit in attempting to establish a controller which would produce a voltage

response whose magnitude is less dependent on the sign of a large signal disturbance.

8.3 EFFECT OF MACHINE AND TIE LINE LOSSES

To simplify the mathematical model of the SCTL system, the resistances of the armature and tie line are neglected. In the experimental system, for operation with low currents or high real power, the resistive losses are in fact a negligible fraction of the real power delivered to the bus. However for conditions of low real power and high leading reactive power the resistive losses become more significant. The use of real power measured at the infinite bus affects the behaviour of the system and its analysis in three ways.

The first is concerned with the prediction of system performance. The real power, recorded on line to determine the operating point of the system, is in error by an amount equal to the resistive losses. Therefore the K parameters computed to predict the system behaviour are inaccurate, as they are based on incorrect information regarding the operating point of the system. Consequently observed and predicted performance differ.

The second point concerns the estimation of the perturbation in torque used as a state variable by the controller. As the steady state operating point of the system shifts in the PQ plane in response to a disturbance, the resistive losses vary due to the change in line current. Thus the real power measured at the infinite bus changes, and the initial estimate of electrical torque is different to that at the final steady state operating point. For a lossless system, with constant prime mover power, these values should be the same.

The third effect arises in the selection of the feedback gains appropriate to the nearest grid point. If an inaccurate value of power is used to look-up the gain table, the resulting controller does not produce

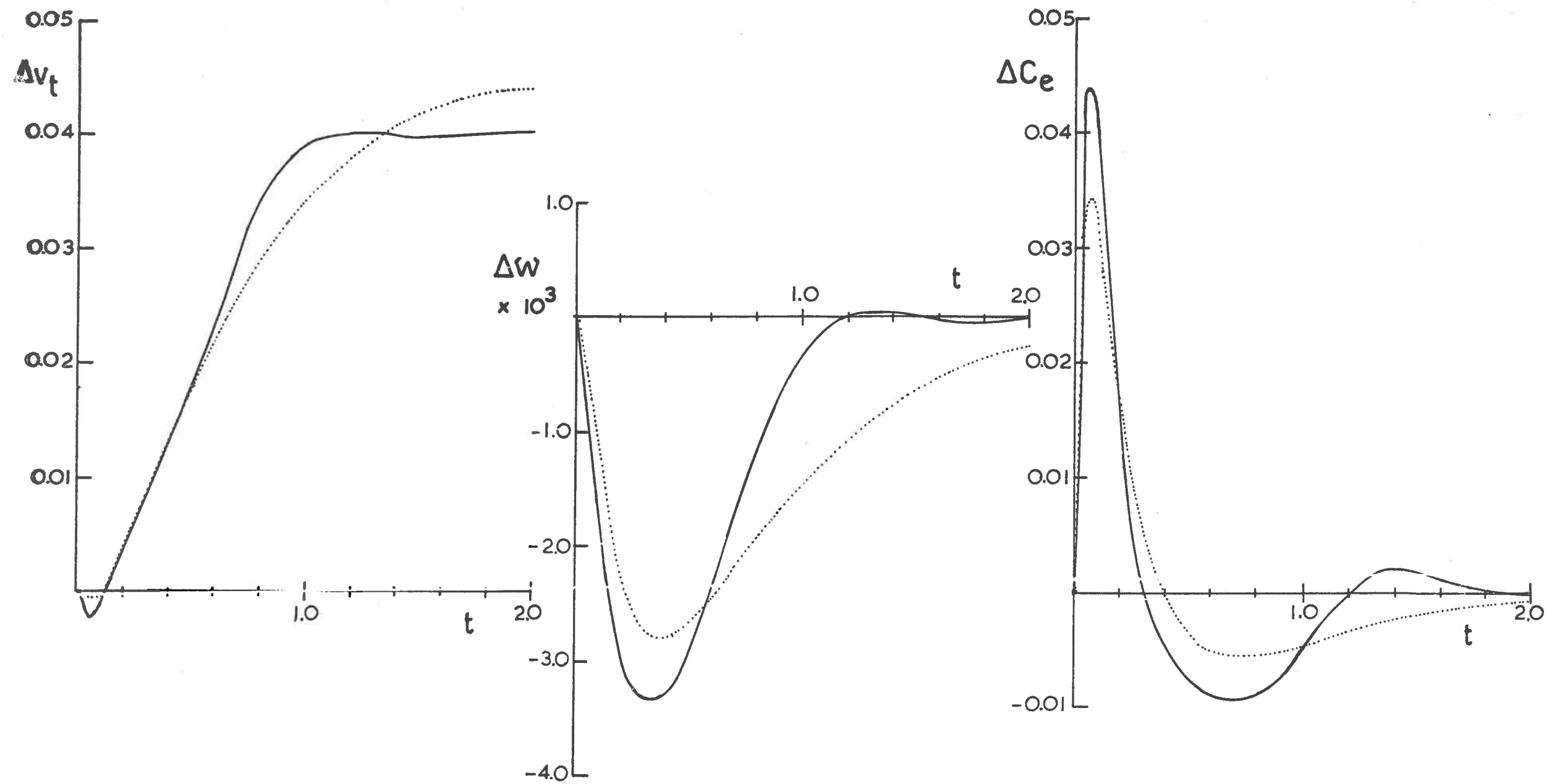


FIGURE 8.6 Response of the system states to an 0.04 p.u. step change in reference voltage.
 Initial operating point $P=0.15$ $Q=-1.52$
 Final operating poing $P=0.15$ $Q=-1.40$
 Observed response — Predicted Response

optimal system performance (or near optimal according to the 10% RMS error criterion of section 6.2.1).

Without the power correction factor outlined in section 6.3.4, system behaviour in the sensitive region of low real power, high leading reactive power is markedly affected. System operation was attempted at a generated real power of about 0.2 p.u. (measured power at bus about 0.1 p.u.) and high leading reactive power.

Figure 6.5 shows that the selection of the correct grid point for gain look-up is critical in this region of the PQ plane if stability is to be maintained. The selection of incorrect regulator gains coupled with an incorrect estimate of the torque perturbation resulted in system instability. Stable operation is possible when the power correction factor is utilized.

Figure 8.6 shows the response of the system states to an 0.04 p.u. input step when the correction factor is incorporated. The system is operating about $Q = -1.5$ p.u. and with a corrected real power of 0.15 p.u. The real power delivered to the bus is 0.07 p.u. As can be seen the system operates in a stable and predictable manner. The discrepancies between predicted and actual performance are attributed to the non linear nature of the system, recalling that the 10% RMS error criterion demands an 0.05 p.u. grid along the Q axis in this region of the PQ plane (Region I), and observing that an 0.04 p.u. input step perturbs the system by 0.12 p.u. in Q during this test.

Stable operation in the absence of the power correction factor is shown in figure 8.7. To obtain these response curves the system was operated in the region of $Q = -1.6$ p.u. with a measured real power of 0.22 p.u. The correct value of real power for this point is 0.30 p.u. Figures 7.4, 7.5 and 7.6 show that close agreement exists between the predicted and actual response curves in this region of the plane when the power correction factor is used.

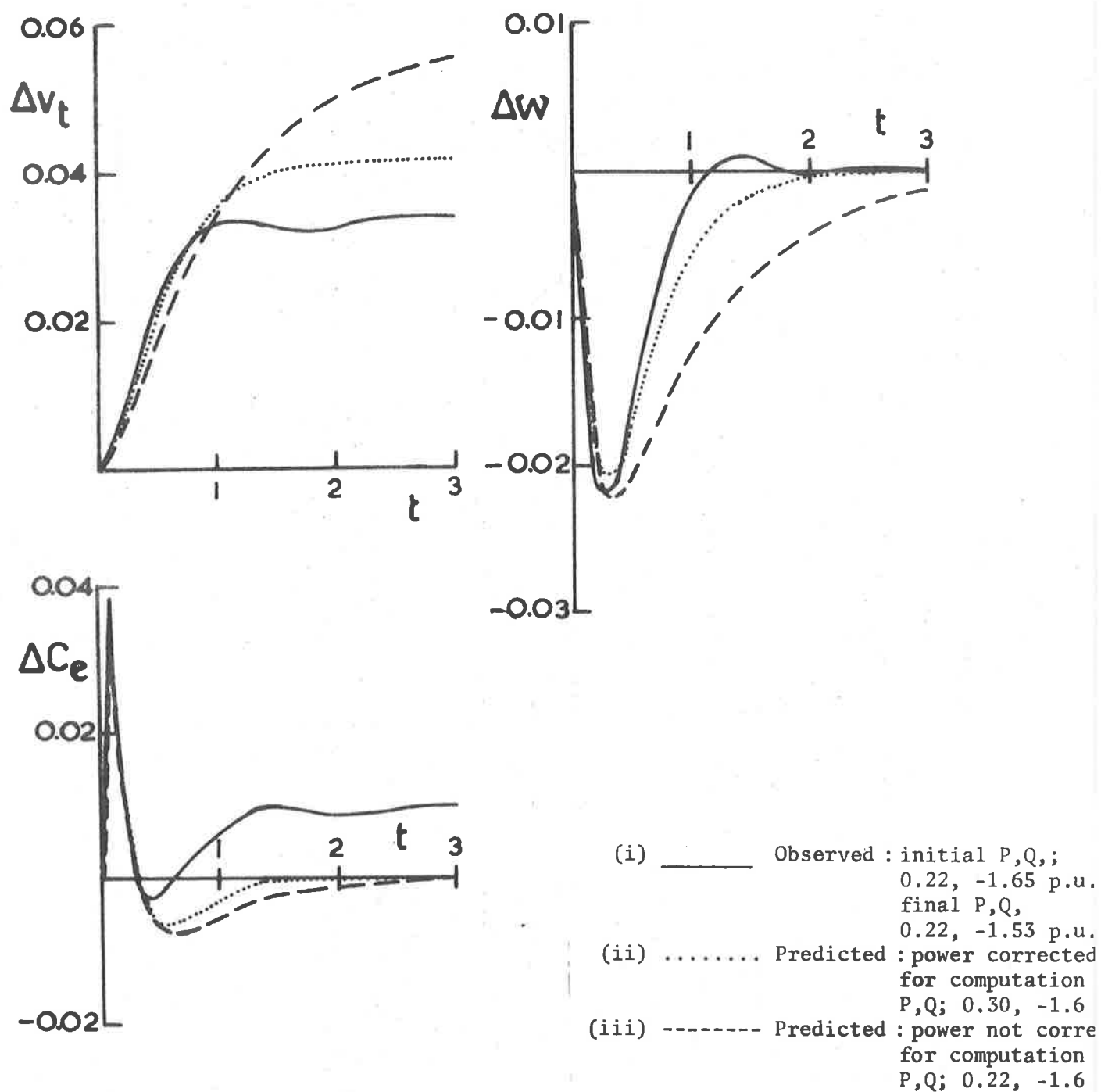


Figure 8.7 Response of the states to an 0.04 p.u. step.
No power correction used.

In figure 8.7, the predicted performances for the system operating at the measured real power, and also at the corrected real power, are plotted to demonstrate two of the effects of omitting this factor.

Comparing the observed performance with that predicted by calculations based on the corrected operating point, an offset is observed in the steady state value of the on line estimated torque deviation, being equal to the difference between the resistive losses at the initial and steady state operating points. This produces an offset in the steady state value of the terminal voltage deviation due to the presence of a non zero torque feedback signal in the steady state. Thus with the voltage and torque feedback gains used by the controller having the same sign and about the same magnitude, a positive error in the steady state torque produces a negative error in the steady voltage as seen in figure 8.7.

When the predicted performance is based on the measured operating point, the comparison is far from satisfactory as the effective operating point of the experimental system is quite different from this.

Thus the three major effects of neglecting the power correction factor have been demonstrated: the use of incorrect regulator gains may result in system instability; for stable operation, offsets in torque and terminal voltage deviations exist; and predictions are based on incorrect estimates of the operating point.

This simple technique of neglecting resistive losses in the derivation of the mathematical model, but allowing for them in the measurement of real power appears to be a satisfactory solution to the problem of system losses. The armature and tie line resistance of 0.03 p.u., applicable to the experimental system, is comparable to that of operational systems. Thus the preceding compensation technique could well be applied to such systems with similar satisfactory results.

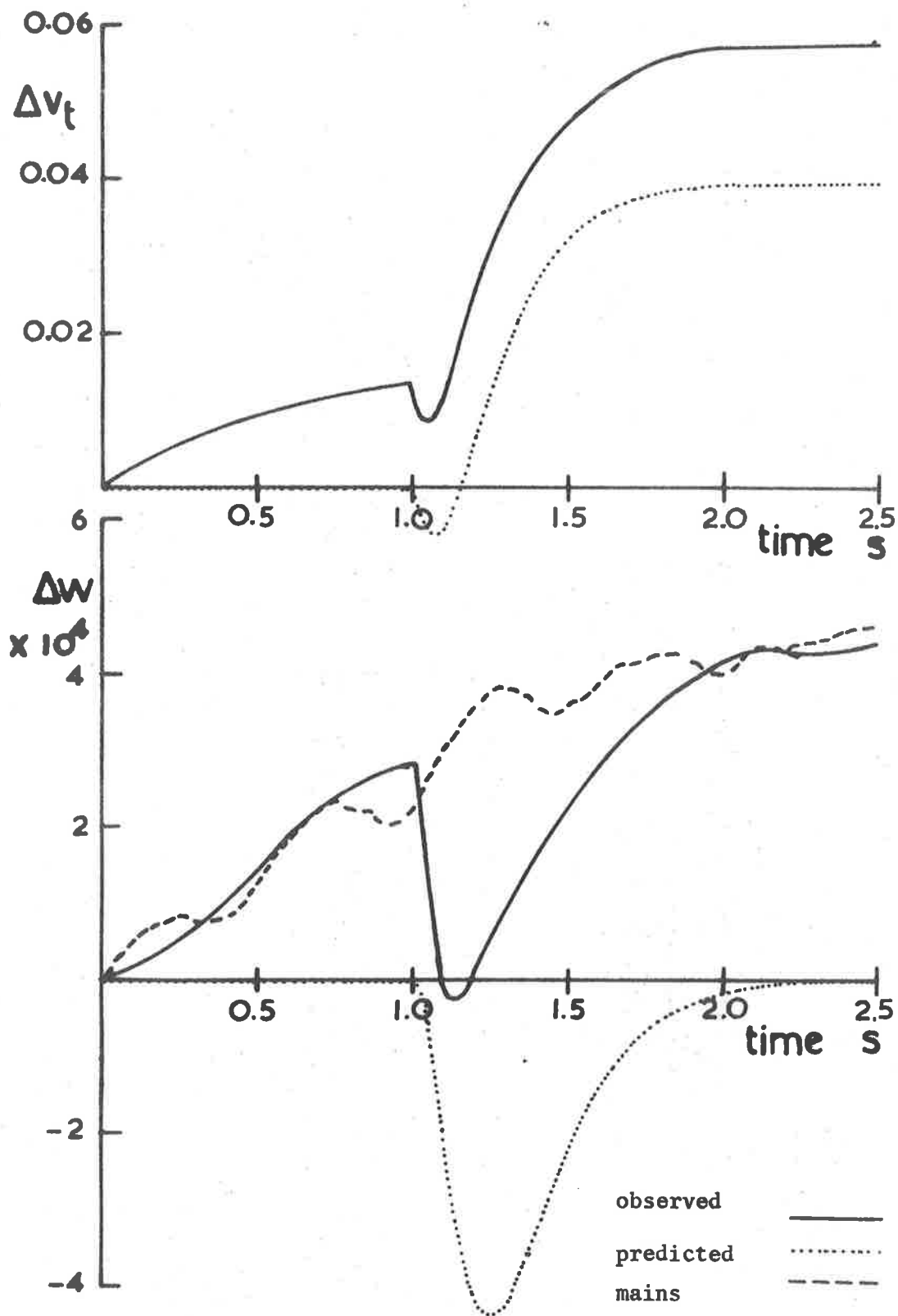


Figure 8.8 Terminal voltage and machine speed responses to a step change in reference voltage of 0.04 p.u. Operating points : initial; 0.16, -2.05; final; 0.16, -1.95. Power measured at bus 0.03 p.u.

8.4 BUS FREQUENCY DEVIATION

Constant frequency of the infinite bus has been assumed throughout the analysis although deviations as high as $\pm 0.1\%$ p.u. were observed.

When performing the perturbation tests on the system, initial values of machine speed and infinite bus frequency are recorded. As the infinite bus frequency drifts during the test period, the observed machine speed deviation has components due to both this drift and the step change in reference voltage. In the unity power factor region of the PQ plane, where the speed feedback gain is negligible, the principle of superposition may be applied to the small signal linear system. That is, the observed machine speed response is equal to the deviation resulting from a step change in reference voltage alone, added to the deviation which would result from a drift in bus frequency alone.

For operation at low real power and high leading reactive power the speed feedback gain is high and a significant component of the control force may be due to the deviation in mains frequency. This deviation therefore influences the observed responses of all of the states via the control force. Figure 8.8 shows this effect when the system is operating in the region of 0.16 p.u. real power and -2.0 p.u. reactive power. For the first second of the test the perturbation in reference voltage is zero. During this period an increase in mains frequency and a corresponding increase in the machine speed is observed. This gives rise to a change in terminal voltage due to the feedback signal from speed. At $t = 1$ s a step change in reference voltage of 0.04 p.u. is applied. The states respond to this change as shown, still being influenced by the increasing drift in mains frequency.

Fluctuations in mains frequency about the initial value produce the change in terminal voltage shown, and this causes a change in the reactive power of the operating point. Such a drift of Q and V_t with bus frequency

is characteristic of this form of controller if the reference speed differs from the bus frequency. Slow variations in Q as high as 0.15 p.u. due solely to bus frequency drifts were observed in the region of 0.15 p.u. real power and -2.0 p.u. reactive power. The system remained stable under these conditions.

To minimize errors in the recorded responses of the system, mains frequency deviations were measured throughout all test sequences and the observed speed corrected by subtracting these deviation. Also as the effect can be more than simple superposition, particularly when the speed feedback gain is appreciable, test results were rejected in cases where the p.u. change in bus frequency exceeded 10% of the corrected p.u. peak swing. Despite this precaution some of the voltage response curves shown earlier are slightly influenced by mains fluctuations, particularly in the region of low P and high leading Q .

8.5 TIE-LINE REACTANCE VARIATION

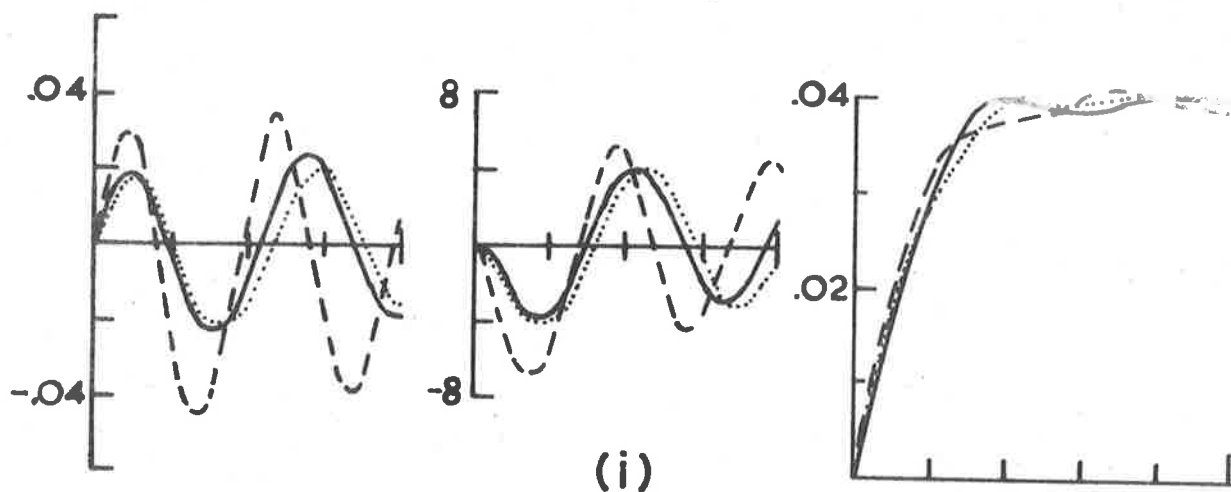
In practical power system operation, the tie-line reactance, representing the Thevenin equivalent reactance of the external system, varies according to the system loading and configuration. However, in determining the optimal controller, and for the tests described previously, a constant value of tie-line reactance has been assumed.

To assess the effect of line variations in the experimental system, the line reactors were varied over a range of 50-125% of the constant value for which the controller gains were calculated.

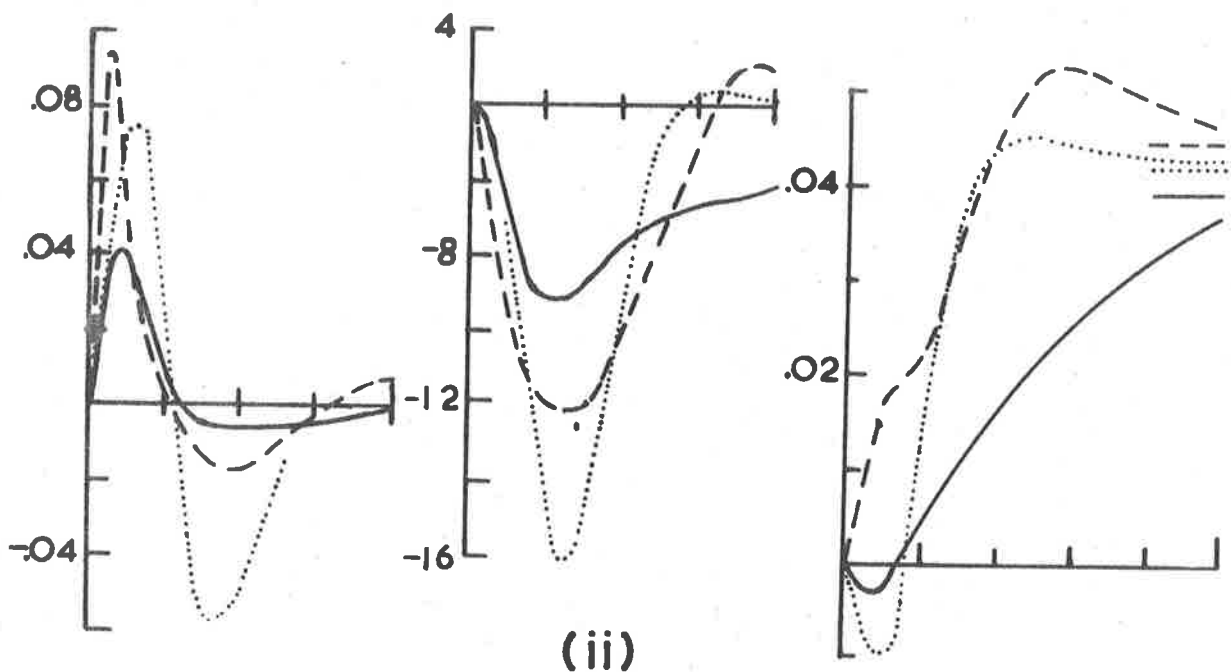
Tests similar to those described in Chapter 7 were carried out. The system remained stable for this range of reactances (the maximum available from the reactors) throughout the operating region of the system. Examples of the system behaviour are shown in figure 7.9. For power factors near unity little change occurs in the step voltage response, although speed and torque deviations are noticeably effected. For high leading reactive power however,

changing the line reactances appreciably effects the behaviour of all three states.

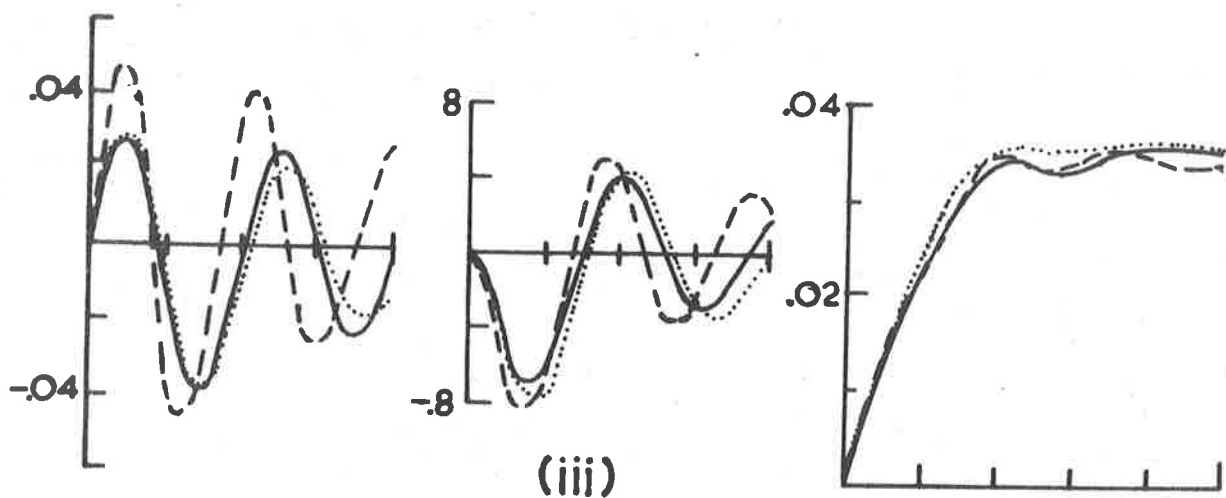
It can be seen in figure 8.9 that the peak deviations of speed and torque are at a minimum when the tie line reactance is 0.29 p.u., the value used to compute the optimal controller.



(i)



(ii)



(iii)

Electrical Torque
 ΔC_e p.u.

Machine Speed
 $\Delta \omega \times 10^4$ p.u.

Terminal voltage
 Δv_t p.u.

Figure 8.9 Observed state variable responses to a
step change in reference voltage of
0.04 p.u.

Grid point of the domain in which the
system is operating:

- (i) $P = 0.3$ p.u., $Q = -0.4$ p.u.
- (ii) $P = 0.3$ p.u., $Q = -1.6$ p.u.
- (iii) $P = 0.9$ p.u., $Q = -0.8$ p.u.

External reactance x_e used:

0.145 p.u. (5 ohms) -----

0.29 p.u. (10 ohms) _____

0.36 p.u. (12.5 ohms)-

Horizontal scale (time base) 0.2 s/div.

In choosing a suitable cost function consideration could be given to the effect of tie line variation on the system behaviour. That is, if possible, choose a cost function such that the controlled system is minimally affected by tie line variation. If the nature of such a cost function conflicts with other design requirements, the variation of the equivalent tie line impedance could be monitored and a table look-up system, similar to that used at present for the operating point, used to allow the gains of the controller to vary with line reactance.

8.6 SUMMARY

The established experimental SGTL system proved to be particularly useful in determining the system behaviour under non ideal conditions. It allows a simple, economical and accurate investigation into those aspects of practical power system operation which are extremely difficult to model, or which, in a theoretical study, may be overlooked.

The effect of tie line variation was investigated using small perturbations in reference voltage, and, having shown the mathematical model to be an accurate small signal representation in Chapter 7, could have been carried out theoretically. However, the large signal behaviour with the observed limit cycling, the effect of resistive losses and a technique to allow for them, and the effect of infinite bus frequency deviations on the controlled system are certainly most readily and accurately evaluated on an experimental system.

The question which arises from these investigations is what can be done to improve performance under these conditions? It was suggested earlier that by using a cost function which contains speed and torque explicitly, improved large signal behaviour would be obtained. This improvement would be achieved by larger feedback gains for these states, as exists

at present for operation in the high leading reactive power region of the PQ plane. However any increase in the speed feedback gain would accentuate the present problem existing in this region, where deviations in the infinite bus frequency cause disturbances to the states of the system and its operating point. Thus a compromise may be necessary or, alternatively, different weighting matrices used for different regions of the operating plane.

It is considered that the technique used to allow for resistive losses in the system has been proven satisfactory. In a practical operating system the real power transfer would be measured at the terminals of the machine and tie line resistance would not contribute to the effect of the armature resistance as with the present system.

Consideration has already been given to possible procedures to allow for tie line impedance variations. In view of the current conflicting requirements of the cost function, it is felt that the further complications which would arise in trying to minimize system susceptibility to tie line variation, would produce an unwieldy and unworkable problem. Thus the use of a look-up table, with gains dependent on the equivalent tie line reactance, appears to be the most suitable solution.

CHAPTER 9

SUMMARY, COMMENTS AND CONCLUSIONS

9.1 SUMMARY

The object of many previous studies into the small signal behaviour of the SGTL system has been to determine theoretically the stability limits which exist when generating leading reactive power. Various combinations of feedback control signals have been investigated, and in a few cases the theoretical predictions substantiated by experimentation. Having established these limits, consideration is given to designing a regulator with fixed gains which will yield adequate system performance over a range of possible loading conditions, although, due to the non linear nature of the system, a compromise is necessary between the range of operation and the performance obtained within this range.

While at first sight modern optimal control theory appears to provide all the answers, the application of this theory to a complex non linear system is fraught with difficulties. The objective of this study has been to develop and implement a strategy and design and produce the required hardware to prove that it is possible to successfully apply the elements of this theory to such a system. As a result the system designer is allowed the benefits of this approach, with the assurance of a proven solution.

9.2 THE OPTIMIZATION PROBLEM

Optimal control theory requires the designer to have an intimate knowledge of the plant and the goals of the controlled system to satisfactorily formulate the problem in mathematical terms. However, it has the distinct advantage over other design procedures in that an ideal solution to the idealized problem can be found. The greatest difficulty in applying this theory to practical systems is that few such systems fall into a category for which direct analytical procedures exist for the mathematical solution of the problem. However, by liberally interpreting the rules governing the use

of the theory, making approximations both in the derivation of a mathematical model and, if necessary, in obtaining the solution to the formulated problem, it is possible to achieve useful guides to the solution of a wide range of problems. This was the philosophy adopted in developing a controller for the SGTL system.

The mathematical model derived for the system is the widely used small signal linear model, based on Park's non-linear equations for the synchronous machine. Besides the approximation of representing a non linear system with small signal linear equations, several simplifying assumptions were made in deriving the model. These reduce the order of the equations describing the system and allow the relatively simple and well defined linear optimization procedures to be applied. The performance of the system under conditions where these approximations are obviously invalid was investigated to indicate whether the operation of the controller is critically affected, and in order that techniques to improve the performance under these conditions might suggest themselves.

9.2.1 THE COST FUNCTION

With the plant described by a linear set of equations a quadratic cost function could be, and was in fact, used to describe the required behaviour of the controlled system. While this form of cost function allows an analytical solution to be obtained which produces a stable closed loop form of control, the problem of choosing suitable elements within the quadratic frame work still exists.

In this study a simple cost function was chosen which would produce a physically realizable controller acting to minimize voltage errors, with some penalty on the control force required to achieve this defined goal. In light of the experience

gained in operating the experimental system several points arose which would influence the choice of a more suitable cost function. These have been dealt with as they occurred in other Chapters but are summarized collectively here to indicate their multiple and sometimes conflicting demands.

The small signal terminal voltage response is satisfactory throughout the PQ operating region studied, although increased damping on the speed and torque responses appears warranted for operation around unity power factor. This would be achieved, and, it is reasoned, the large disturbance limit cycles eliminated, by explicitly penalizing deviations of these variables in the cost function, resulting in a controller with large feedback gains for speed. However a general increase in this gain over the whole PQ plane would accentuate one problem already existing, and possibly initiate another. Firstly in the high leading reactive power region where the speed feedback gain is already high, the system is particularly susceptible to deviations in the infinite bus frequency. Secondly, in this region at present, particularly at low real power, this feedback gain, and that for torque, increase very rapidly with leading reactive power, so that speed and torque deviations are already kept small. Any further increase in these feedbacks would present the dual problem of accurately measuring very small perturbations in these quantities and adequately accommodating the very large gains so that the $F_2 \times \Delta\omega$ and $F_3 \times \Delta C_e$ products could be computed with sufficient accuracy.

A uniform type of voltage response is obtained across the PQ plane with the present controller, and overdamped speed and

torque responses are obtained at high leading reactive power, although the control force is required to minimize voltage errors only. That is, in this region of the plane, minimum voltage errors and heavy speed and torque damping are complementary goals. It is therefore considered that by easing the penalty on voltage deviation, and introducing an explicit penalty on speed and torque, a controller would result producing acceptable and relatively uniform responses from all of the states throughout the operating plane. If it proved difficult to find such a controller using a single criterion, the elements of the cost function could be changed for different operating regions of the plane to achieve the desired result.

While it might be possible to minimize the dependence of the controlled system on tie line reactance variation and the effects of machine losses by a judicious choice of cost function, both these points can be adequately covered when implementing the controller, and the added complications and possible compromises with other requirements imply that these points need not be considered.

It would be convenient, but not essential, that the optimal gains vary slowly with the steady state operating point to keep the required grid spacing large, so minimizing the number of computations required in establishing the controller, and reducing the storage demands on the controlling computer.

Finally, whatever cost function is used, the resulting controller must be technically and economically feasible.

9.3 THE EXPERIMENTAL SYSTEM

A simple SGTL model was established in the laboratory using a 5kVA alternator with a D.C. motor as prime mover. The regulator is a small digital computer (NOVA), monitoring the states of the system and computing the control signal applied to an electronic amplifier exciting the alternator field.

No complex or expensive transducers are used, the most stringent specification being the 1 in 50,000 resolution required for the speed transducer. This is achieved by using simple hardware backed by a few logical and arithmetic operations in the controlling computer.

The ability of the regulator to achieve the desired goals (i.e. to perform as predicted) is dependent on the accuracy of the mathematical model derived for the SGTL system. With the procedure used in the derivation, it is possible to check and confirm the accuracy of the model parameters under open loop conditions before attempting to implement the regulator. Although neglected in the derivation, saturation effects are allowed for from graphs of alternator reactances and the field time constant against terminal voltage.

9.4 OPERATION OF THE CLOSED LOOP SYSTEM

A sub-optimal form of adaptive control is achieved by automatically switching the regulator gains to the values most appropriate to a particular operating condition.

The small signal control strategy is applied by sensing deviations of voltage, speed and torque from their initial values. Using these deviations as the system states, the required perturbation in field voltage is computed and added to its initial value before being applied to the alternator field. In the experimental system the initial conditions are established under operator control. In an operational system this function could be automated so that new initial conditions are established whenever the steady state

voltage or operating point deviations exceed a predetermined level. In the event of large disturbances producing unacceptable responses, or forcing the system to instability as a result of system nonlinearities, new initial conditions could be established under transient conditions. In the experimental system it was found that all modes of the large disturbance responses (including the sustained limit cycling) could be immediately modified to approximately the small signal responses by setting up new initial conditions during the transient period.

The small signal dynamic response of the closed loop system is accurately predicted, confirming the accuracy of the small signal (steady state) mathematical model. Stable operation is achieved over an extended region of the complex power plane to regions well beyond the stability limit of a conventional voltage regulator. The range of operation of the system studied is limited only by the thermal ratings of the equipment and not by instability. As mentioned previously, practical limits (other than equipment ratings) may eventually prevent continued extension of possible operating regions as the required regulator gains or the necessary accuracy of the state measurements may be unrealizable.

Operating the experimental system over the extended power plane highlights features of the system which may be overlooked or about which misleading information may be gathered in a limited theoretical investigation. For example, from a theoretical study in a confined operating region of the PQ plane, Anderson⁵ concludes that the neglected resistance has no effect on the controlled system, and observed no difference between the transient responses of closed loop linear and non linear mathematical models. While this is undoubtedly true for the PQ region investigated (and while using that

particular non linear mathematical model), the assumption could readily be drawn that these features are generally applicable to the SGTL system. It has been shown that this is not so.

It appears that full appreciation of the SGTL system can best be gained from experimentation with a laboratory model of the system.

9.5 FINAL COMMENTS

Besides the obvious fact that a small digital computer is ideally suited to the role of an adaptive controller, there are many significant advantages to a regulator of this type. The system is simple, of modest cost, fast, accurate, reliable and requires little specialized hardware. As well as this, the control strategy is easily modified if required, and additional functions (VAR limiting, monitoring, protection, alarms etc.) can be readily incorporated.

When modern optimization theory is used as a guide to determine the gains of such a regulator, stable system operation is possible under previously intolerable loading conditions. The theory can be extended to include the dynamics of the field exciter if necessary, and also the prime mover and its governing system. The prime mover power may then be regarded as an input to the system. This would completely define an operational power plant and allow an improved regulator, based on the foregoing study, to be implemented.

APPENDIX A

DETAILS OF THE TRANSDUCERS

AND ANALOGUE INTERFACE

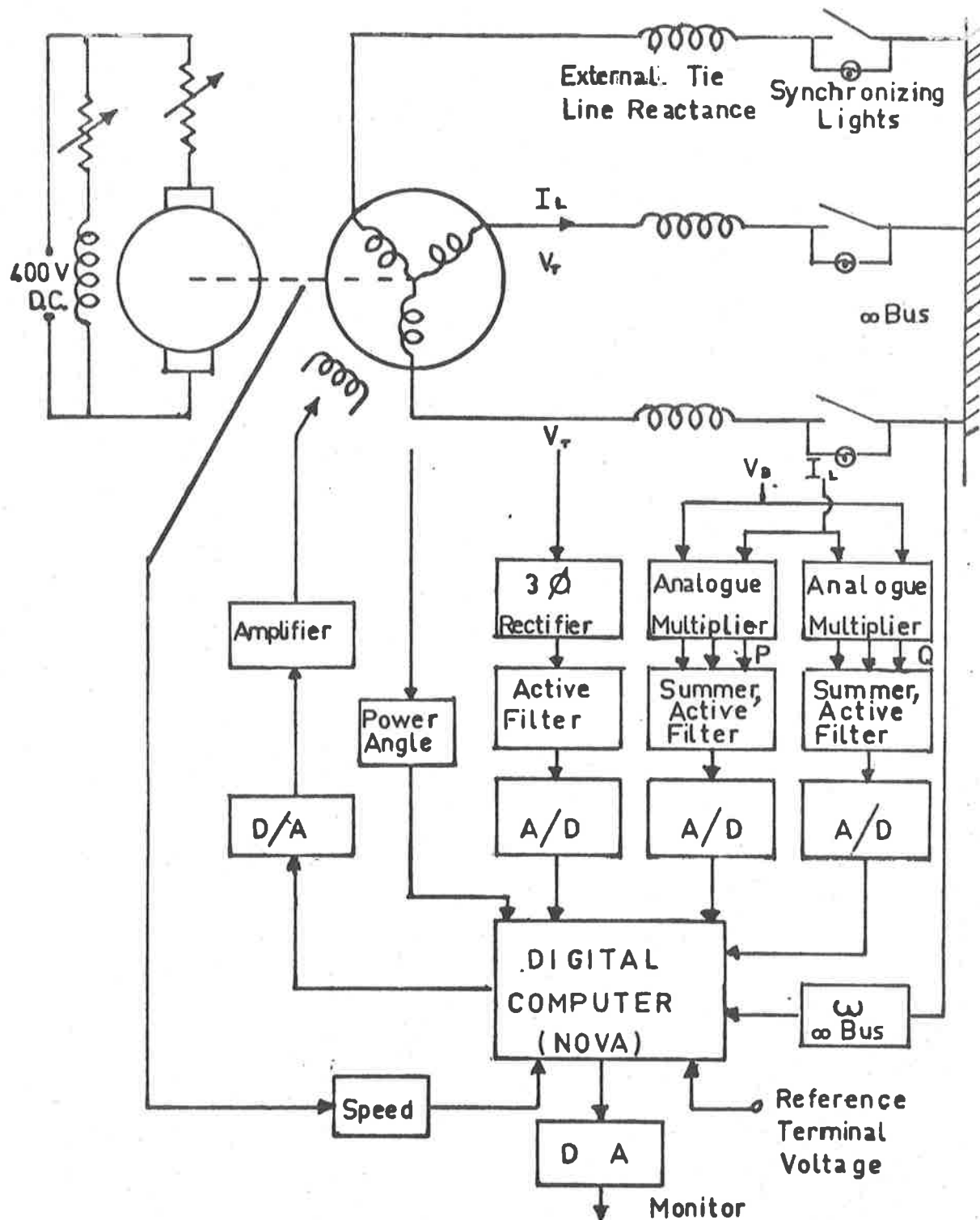


Figure A1 The S.G.T.L. system with the digital controller and transducers.

A INTRODUCTION

This appendix contains a detailed description of the A/D interface and transducers which were developed for this project. An outline of the equipment is given in Chapter 5. Some of the material presented in that chapter (e.g. the specifications) is repeated here to make this appendix a complete and self contained hardware description.

Figure A1 is a schematic of the experimental model, showing the analogue SGTL system and the digital controller, with the A/D interface and the transducers forming the link between the two parts.

In the following sections frequency compensating networks and D.C. zero biasing circuits for the operational amplifiers are not shown where the manufacturer's recommended procedures have been adopted.

A.1 ANALOGUE TO DIGITAL CONVERSION

The successive approximation type of converter was built as their method of operation results in a more versatile converter with a wider range of applications.

In this type of converter the analogue signal is compared with a set of binary weighted voltages or currents. Starting with the largest binary bit, these are successively tested against the analogue signal and are either left in or switched out, depending whether, when summed with previously tested and accepted bits, they are larger or smaller than the analogue signal.

Consider the schematic of figure A2. $G_0, G_1, \dots, G_{10}$ are a set of binary weighted conductances; i.e. $G_0 = 2 \times G_1$, $G_1 = 2 \times G_2$ etc. The section shown as switching control initially closes switch S_0 , and in so doing produces a voltage $V_D = (G_0/G) \times V_{ref}$. V_D is compared with the analogue input V_{in} . If V_D is larger than V_{in} , then the control line signals the switching control, which opens the switch

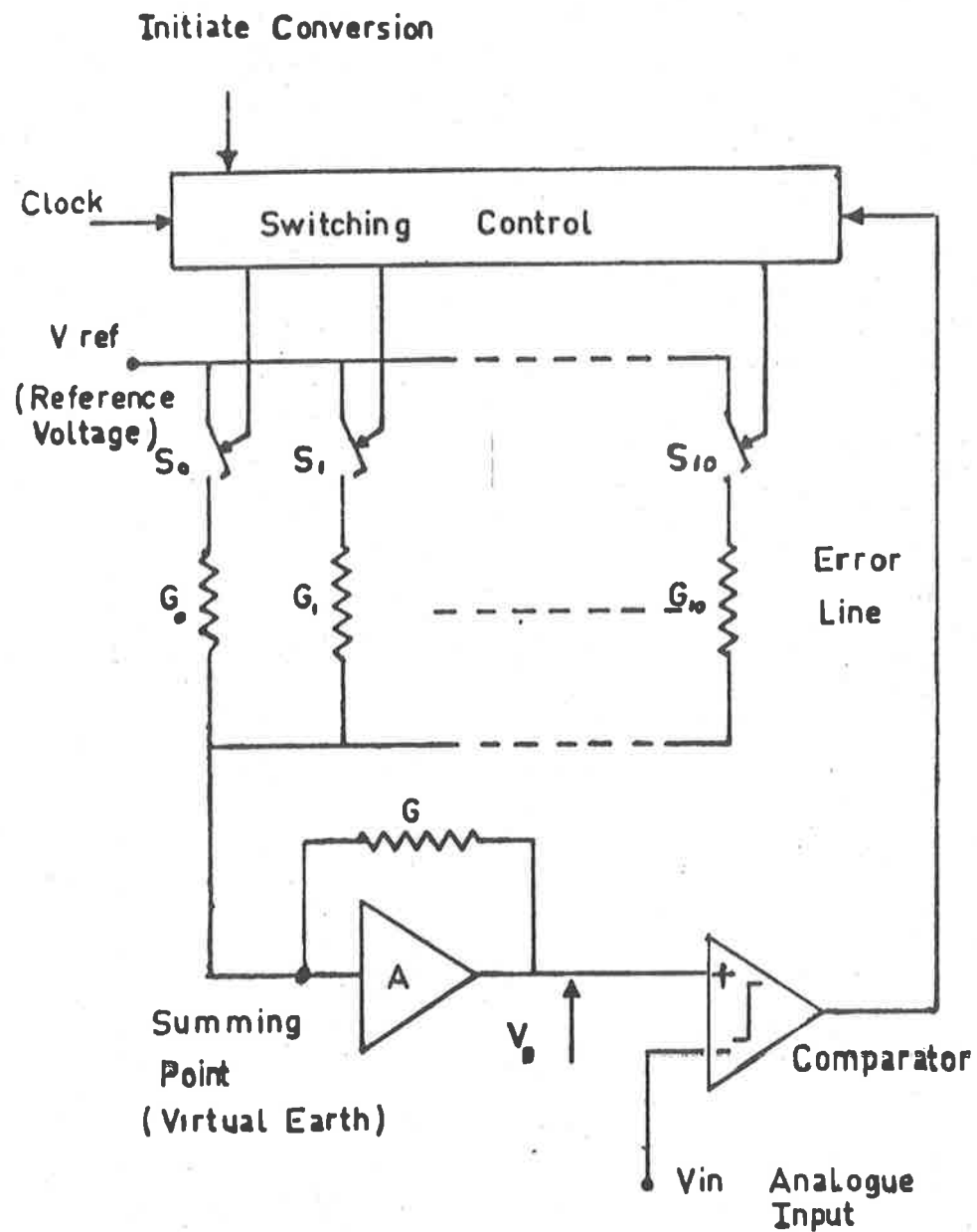


Figure A2 Schematic of a successive approximation A/D converter.

S_0 again. If V_D is less than V_{in} , then S_0 is left closed. Assume that V_D is greater than V_{in} in this instance, and therefore S_0 is reopened by the switching control under command from the comparator. The switching control then closes switch S_1 producing $V_D = (G_1/G) \times V_{ref}$. Now let $V_D < V_{in}$, and so after comparison switch S_1 is left closed. Switch S_2 is now closed and $V_D = (G_1 + G_2) \times V_{ref}/G$, i.e. V_D is proportional to the sum of the binary weighted conductances which have their switches closed.

This sum is now compared with the analogue input signal undergoing conversion. If the sum is larger than V_{in} then switch S_2 (only) is opened, and if V_D is less than V_{in} , S_2 is left closed. The switching control thus proceeds along the line of switches, testing the magnitude of the equivalent analogue signal, produced by a specific set of switch conditions, against the signal at the input. At the completion of the conversion the setting of the switches represents in binary form (to some scale) the analogue input signal, and V_D is equal to V_{in} to within the magnitude of the least significant binary bit.

As mentioned previously the analogue to digital interface was developed as a general purpose unit. Specifications were chosen with this in mind. Ten binary bits plus a sign bit, giving a resolution of $\pm 0.05\%$ of full scale were considered adequate, and the maximum conversion rate which would permit operation with the same accuracy was determined experimentally. Advances in electronic technology have superseded some design considerations used in the development of the units, although the overall specifications and cost per converter still compare favourably with current commercial productions.

For the purpose of the following discussion, the digital control and switches have been separated from the analogue section.

A 1.1 THE SWITCHING CONTROL, SWITCHES AND CONDUCTANCES

The switching control register of figure A2 must sequentially control the switches, and obey commands from the comparator. On initiation, the register is required to propagate a "1" switching level down its line synchronous with the clock input, but be capable of having any particular bit left set, or being reset, depending on the level of the comparator output when that bit has just been set and tested. The required pattern of settings is not known in advance, and may only be found by switching in and testing each conductance in turn. A digital register to operate in this way was developed and named the rollup register because of its action³³. This unit reduced the number of components required and yet operated more reliably and effectively than other switching systems. The required control signal gating for the register, shown in figure A5, was performed using diodes and transistors, after consideration of the cost and performance possible, using other methods. Having completed a conversion sequence the roll up register holds the binary representation the analogue input voltage allowing it to be read into the computer, or some other register, before proceeding with a new switching sequence. This it does on receipt of an external command.

The digital register is used to control the switching of the binary weighted conductances, designated G_0 to G_{10} in figure A2, via S_0 to S_{10} . The conductances are brought to a summing point which becomes a virtual earth by the action of the operational amplifier A.

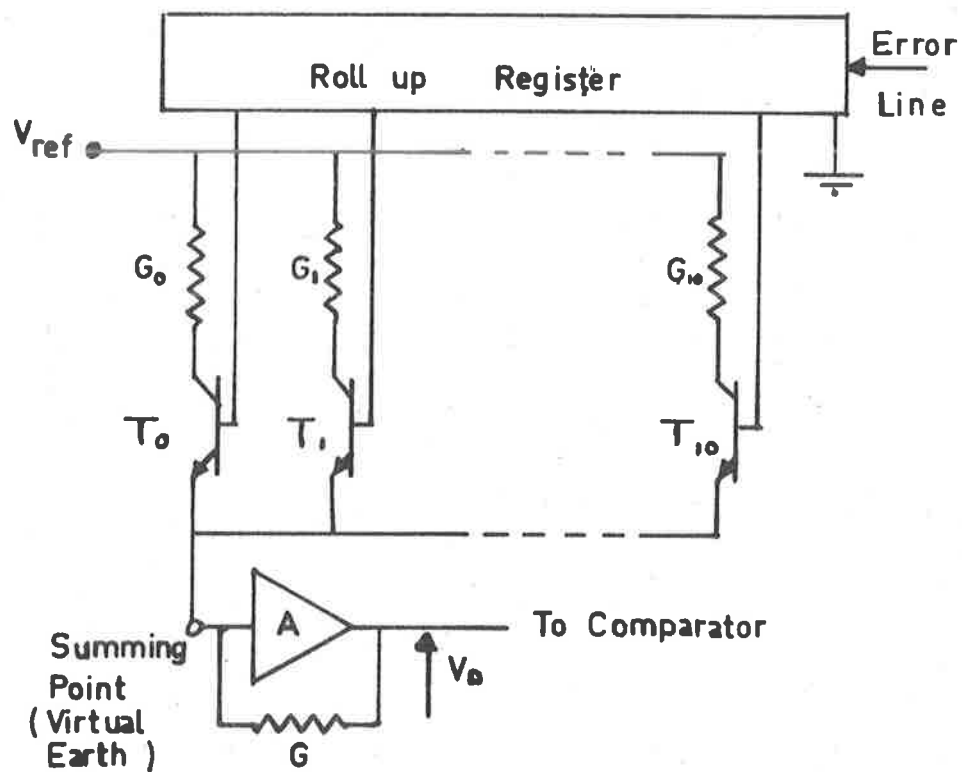


Figure A3 A/D converter with T_0 to T_{10} replacing S_0 to S_{10}

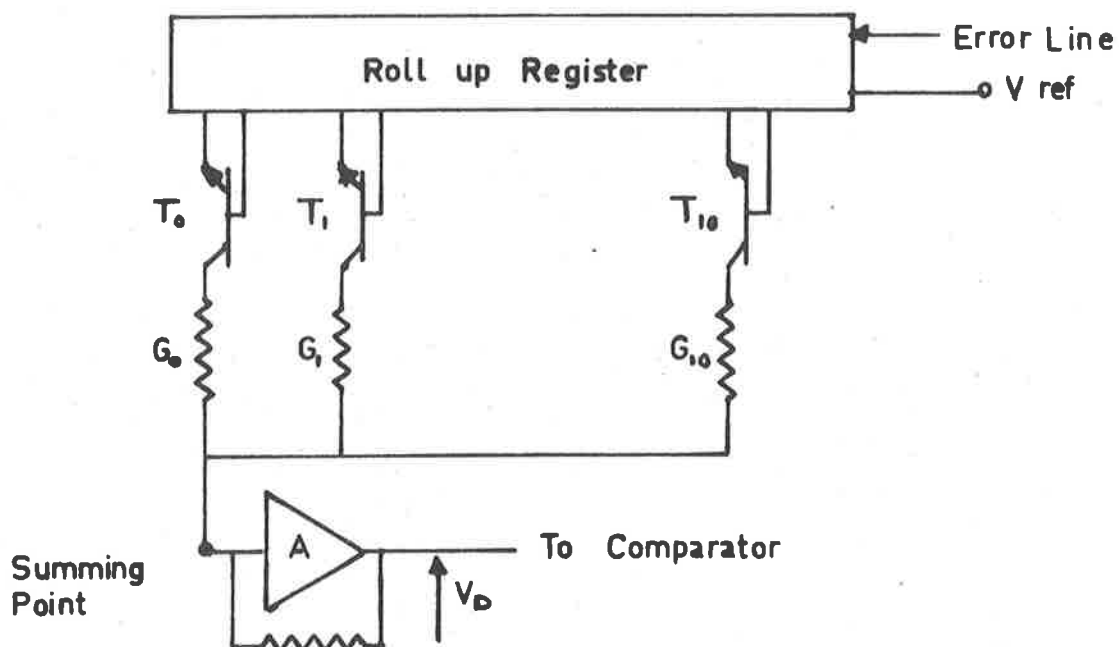
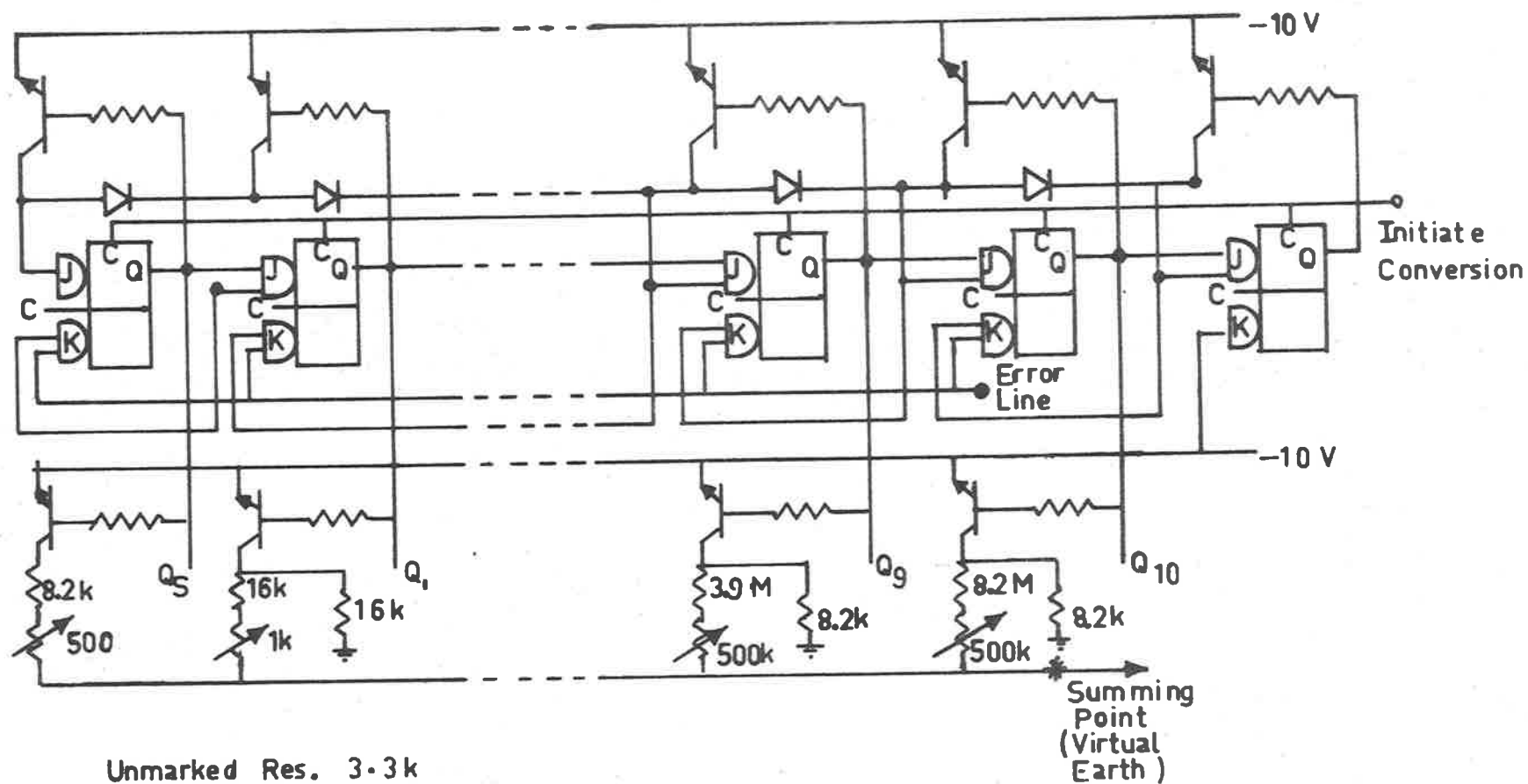


Figure A4 Switching configuration used in the A/D converters.

At the time of development there were three recognized semiconductor switches used to perform the function of S_0 to S_{10} , these being the field effect transistor, the bilateral transistor and transistor controlled diode switches. From considerations of switching speed and junction temperature coefficients, detailed by Pucknell³⁴, the bilateral transistor was considered the most suitable. If the switches S_0 to S_{10} of figure A2 are replaced directly by bilateral transistors, the configuration of figure A3 results. The transistors T_0 to T_{10} are used to switch the conductances under control of the register. However in this circuit, the base current which flows to turn the transistor on passes through the base-emitter junction into the summing point, and so contributes to the value of V_D . Since the pattern of switches for a series of conversions is an unpredictable quantity this base-emitter current cannot be compensated for. By floating the power supply for the rollup register and its clock on the reference supply rail, the system in figure A4 was obtained. Here the base-emitter currents no longer pass to the summing point. The voltages of the 'high' and 'low' logic levels of the rollup register are shifted by this technique, although when originally developed for operation with the CIRRUUS computer, built within the University of Adelaide, this was not a disadvantage. When the interface was later modified for operation with the Data General NOVA computer, simple circuits were required to shift the logic levels so that they became compatible with the NOVA.

Various types of resistor were considered for use as the conductances. Although the ratio of the conductances is more



Unmarked Res. 3.3k
 Diodes 0A47
 Transistor 2N3646
 F/Fs FCJ101

Figure A5 Realization of the Rollup Register,
 switches and conductances

important than their absolute values, most common resistors have a temperature coefficient lying within a large range. Worst case design considerations favoured the use of the highly stable metal oxide resistor. Figure A5 shows the final circuit configuration. The resistors $B_0 - B_{10}$ are bleed resistors chosen to provide approximately equal collector currents in each of the transistors T_0 to T_{10} , so as to produce constant switching times. This current was selected to optimize switching times for the transistor type used.

A 1.2 THE SUMMING AMPLIFIER AND COMPARATOR

With the circuit configuration shown in figure A4, $V_D = G_s \times V_{ref}/G$, where G_s is the sum of the switched in conductances. Applying feedback principles to the circuit the accurate value of V_D is $(G_s/G) \times (A/A + 1) \times V_{ref}$. Therefore, with an amplifier of gain A greater than 2000, $A/(A + 1)$ may be assumed equal to one with an error less than 0.05%. The open loop gain of most integrated circuit operational amplifiers is greater than 2000 at D.C., and therefore there is no difficulty in choosing an amplifier with sufficient open loop gain. However all operational amplifiers produced at the time of investigation had as part of their offset cancelling network a pair of bilateral PNP transistors, which have a non linear phase characteristic with frequency. To overcome this it is necessary to compensate the amplifier with frequency dependent feedback components (either internal or external), which provide a predictable phase and gain characteristic. The effect of these components is to limit the open loop small signal bandwidth and the slew rate (power bandwidth) of the amplifier.

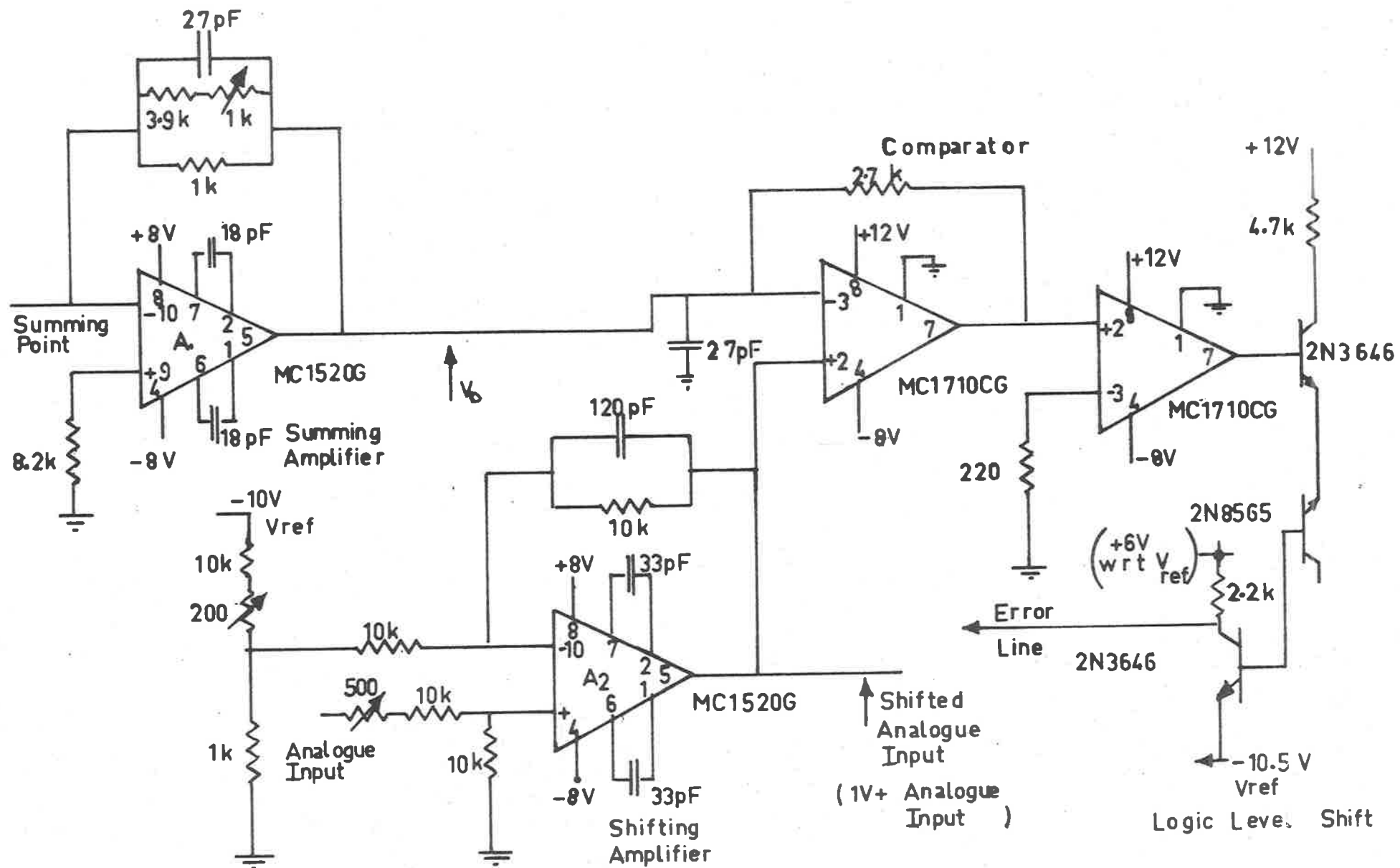


Figure A6 Realization of the summing amplifier, shifting amplifier and comparator.

As the digital register switches in the conductances a staircase type of waveform appears at V_D , the largest step occurring when the largest conductance G_0 is switched in. V_{ref} and G were chosen to produce a step of 1 volt where G_0 was switched. Larger voltages than this are limited by slew rate considerations, and a smaller voltage would have produced problems with amplifier noise, as voltages less than 1mV would have been associated with the smallest conductance G_{10} . For the system to operate accurately V_D must settle to within the value of the least significant bit (1mV) before a comparison with the analogue signal is made (this is the period of one clock pulse on the register). Therefore an operational amplifier with a high slew rate must be chosen, and any overshoot on the step response must be quickly damped. For a clock rate of 1MHz and a maximum step input of 1 volt a slew rate of greater than 5 volts/ μ sec must be achieved so that the signal can reach its peak value and settle to within 1mV within 1 μ sec.

Various types of integrated hybrid and discrete component operational amplifiers were investigated. After balancing cost against performance an MC1520G integrated circuit was chosen. Using this amplifier with minimum compensation and some phase lead feedback it was possible to achieve a clocking rate of 1MHz with the desired accuracy.

The circuit configuration for the summing amplifier appears in figure A6. The reference supply was chosen as -10.5 volts as relatively simple but highly stable reference supplies can be produced to operate at these voltage levels.

Conductances were chosen so that the output range of V_D is 0 to 2 volts in steps of approximately 1mV ($1V/2^{11}$). A reference of -10.5 volts was chosen rather than -10.0V to enable simple trimming of the conductances to be achieved while using resistances in the 2% preferred range.

So that bipolar input signals could be accepted another amplifier, A2, was used to offset by one volt an input of ± 1 volts to the range of 0 to 2 volts for comparison with V_D . Two MC1720C integrated circuits were used as comparators, two being used to minimize the indecisive state which occurs for near equal inputs.

The final section of the circuit in figure A6 shifts the binary output of the comparator to the logic levels used in the rollup register, where it becomes the error line shown in previous sections.

A 1.3 THE A/D CONVERTER SPECIFICATIONS

To provide a multichannel system, six analogue to digital converters were built on printed circuit cards. As the cost of the converter had been kept to a minimum, it was decided that the greater flexibility afforded by six converters, rather than a six channel analogue multiplexer and single converter warranted their production. Also, at the time, accurate high speed analogue multiplexing was neither cheap nor easy. With six such converters it is possible to perform simultaneous sampling and immediate conversion of any required analogue signals, which may then be held in the individual registers until required by the computer.

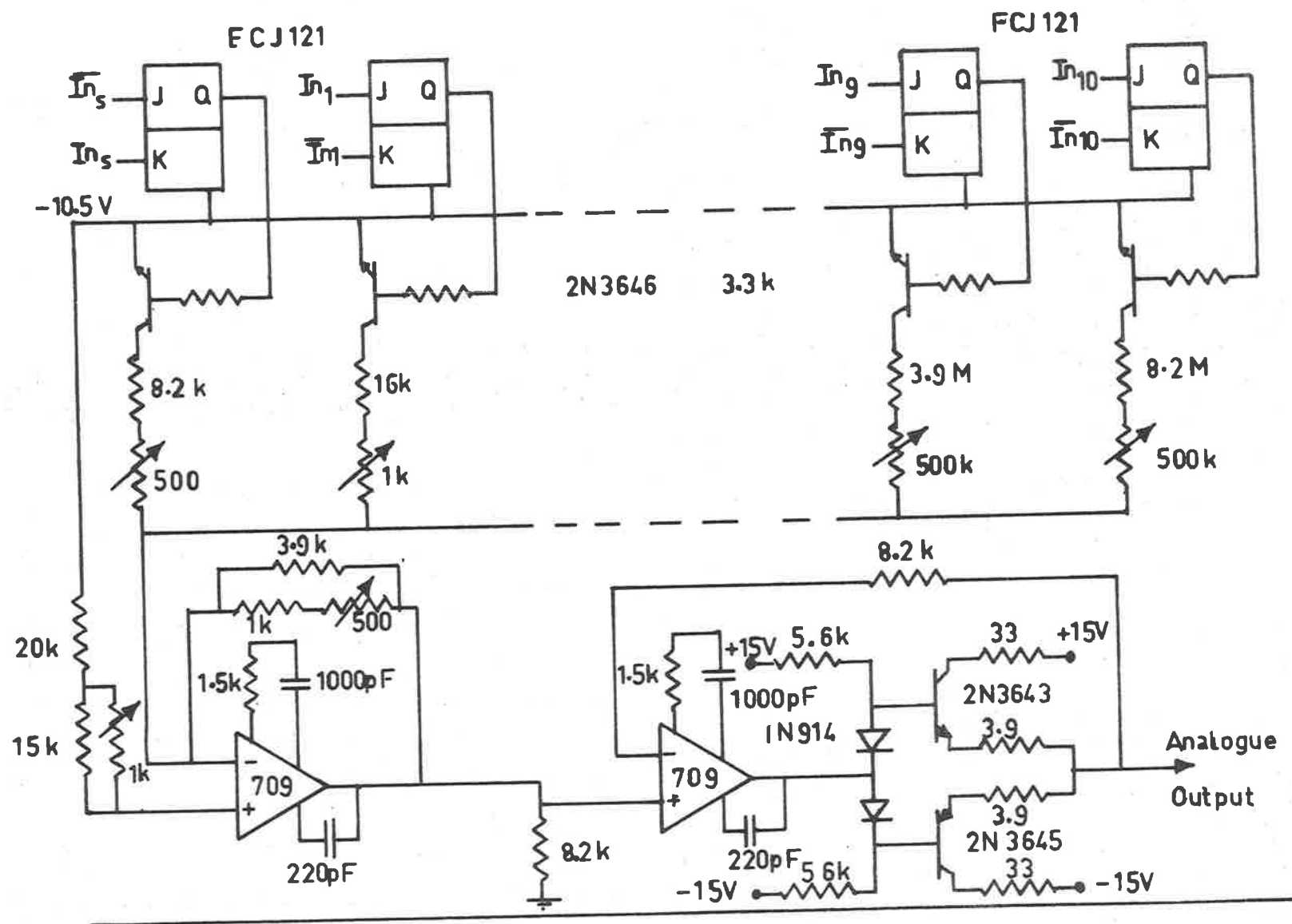


Figure A7 The circuit for the D/A converters

The following specifications are met by each of the six converters:

- (1) Bipolar input range ± 1 volt.
- (2) Digital output 10 bits + sign in two's complement.
- (3) Conversion time 11 μ secs (clocking frequency 1MHz).
- (4) Resolution ± 1 mV ($\pm 0.05\%$ F.S.D.)
- (5) Repeatable conversion accuracy under operating conditions ± 2 mV ($\pm 0.1\%$ F.S.D.)
- (6) Automatic hold of converted input.
- (7) Conversion restart from single 'low' level pulse.
- (8) Input impedance 10k Ω .

The converter circuit can be obtained by combining figures A5 and A6.

A.2 DIGITAL TO ANALOGUE CONVERSION

The D/A converters are similar in configuration to the A/D converters. The digital input is clocked into a holding register which operates the conductance switches. These are brought together at a summing point and the operational amplifier output is the analogue representation of the digital input.

Figure A7 shows the circuit used for the D/A converters. No bleed resistors are provided across the conductance and an MC1709C is used as the summing amplifier for economy. These features limit the speed of response (settling time) of the converter, but the time constants of many analogue systems are sufficiently long that this would not affect overall system performance. While data can be clocked into the digital holding register at any rate up to 15MHz, the slew rate of the analogue stage limits the useful rate to approximately 50kHz.

A unity gain buffer stage with a push-pull emitter follower output was incorporated to provide a low impedance output stage with a current capability of 300mA, as compared with the 5mA maximum current output of the operational amplifier alone.

The specifications of the D/A converters are:

- (1) Digital input 10 bits + sign in two's complement.
- (2) Bipolar output range ± 1 volt.
- (3) Resolution $\pm 1\text{mV}$ ($\pm 0.05\%$ of F.S.D.)
- (4) Repeatable conversion accuracy $\pm 1\text{mV}$ ($\pm 0.05\%$ F.S.D.)
- (5) Zero order digital hold.
- (6) Dynamic output impedance 0.5Ω
- (7) Output current 300mA maximum.

Four D/A converters were built with the above specifications

A.3 COMPUTER CONTROL OF THE ANALOGUE CONVERTERS

The transfer of information to and from the NOVA central processor and its peripherals occurs via a 16 bit I/O bus. The particular peripheral involved in the transfer at any one time is selected by a code applied to the six bit NOVA address bus. The only device which reacts to this address is the one whose prewired selection decode network is activated by the execution of a programmed instruction requesting that device.

In addition to the six bit address the device code may be extended by the use of 3 input or 3 output registers A, B and C. For example, a programmed input instruction such as DIB, Acc, 72 would transmit device code 72B on the address bus and B register line, allowing this particular device to set up on the I/O bus the contents of its register. This information is then clocked by the

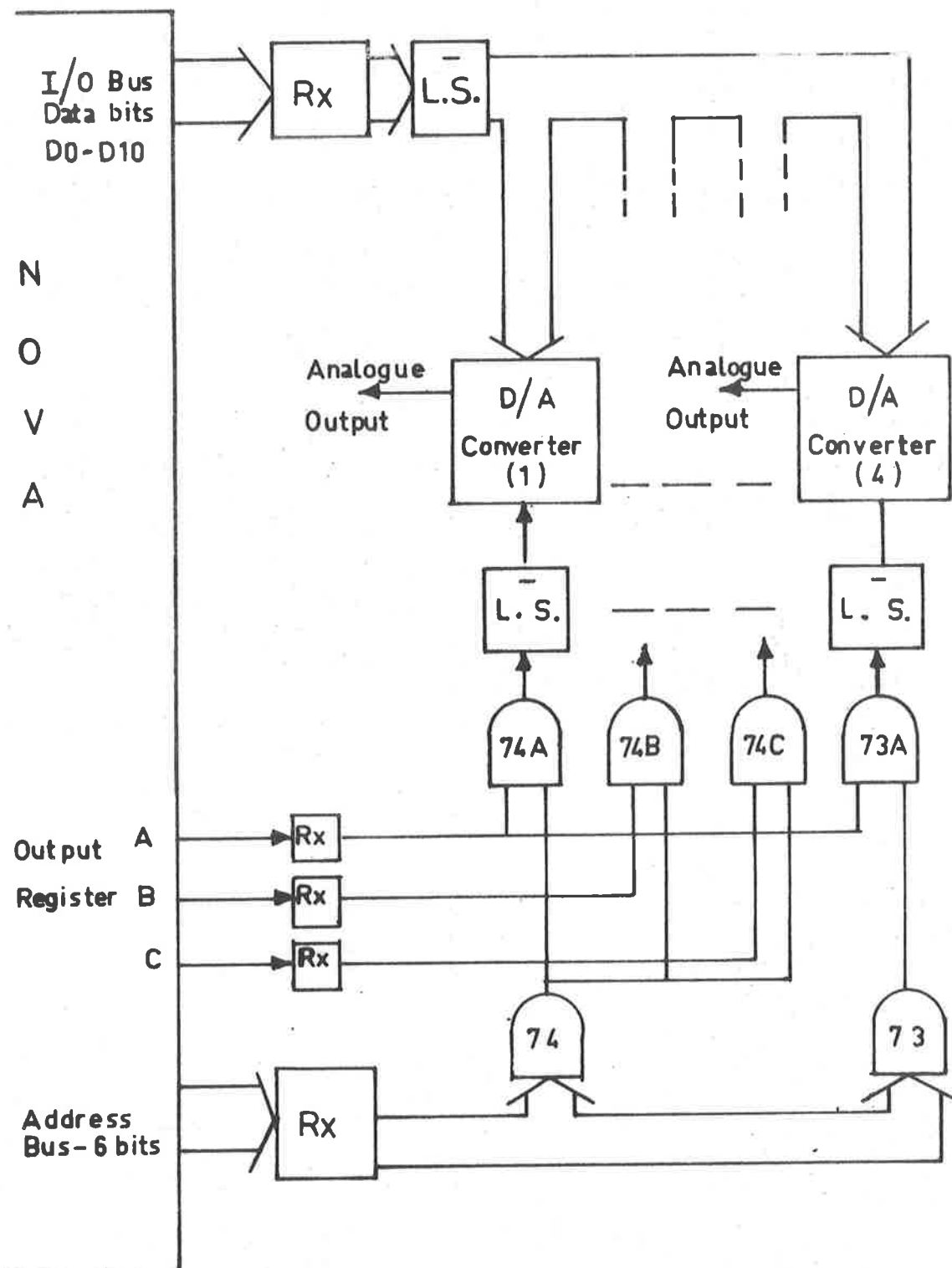


Figure A8 NOVA control of the D/A converters.

central processor into the accumulator whose number is designated by Acc (0 to 3).

Further control lines are available on the NOVA, but the only other one used in this interface is the DONE line. This allows the central processor to ascertain whether a particular device has completed its required function.

As it is the simpler of the two, the control of the D/A converters is considered first. Throughout this and the following sections details of high and low logic levels are not considered, so that inverters, although required in parts, are not shown unless essential to the explanation of system operation. Logic functions are implemented with standard 7400 series TTL. The detailed requirements of a NOVA interface are set out in the NOVA Users Handbook.²⁴

A 3.1 NOVA CONTROL OF THE D/A CONVERTERS

Referring to figure A8, device codes 74A, 74B, 74C and 73A were used for the four D/A converters. These are therefore controlled by the program using the following instructions:

DOA Acc, 74

DOB Acc, 74

DOC Acc, 74

DOA Acc, 73

For example, executing instruction DOA Acc, 74 would cause the central processor to apply to the I/O bus, the contents of the accumulator designated by Acc. This is a binary number in two's complement, the sign bit being data bit number 0. As the D/A converters are of 10 bits plus sign, only data bits 0 to 10 are

actually wired to the converter. This output data is shifted in its logic voltage level down to levels compatible with the D/A converters. While the output data is on the I/O bus, the device code 74 is applied to the address bus. This code, plus output register A, enable gate 74A, which, after a logic voltage shift, clocks the I/O data into the holding register of the D/A converter. An analogue voltage representative of the binary data on the I/O bus (originating from accumulator Acc) then appears as the output of the D/A converter).

A 3.2 NOVA CONTROL OF THE A/D CONVERTERS

Control of the A/D converters is more involved than that of the D/A's, as the conversion process is time separated from the transfer of information to the computer, i.e. the unit must be initiated and complete its conversion before the binary data can be read into the central processor. Separate program instructions are used to initiate the conversion and to read the converted data.

The instruction initiating the conversion is an output instruction, with control signals originating from the central processor on program command. A mask is set up in one of the accumulators and outputted as device code 73C with the instruction

DOC Acc, 73

The mask bit in Acc required for each individual unit to initiate a conversion is:

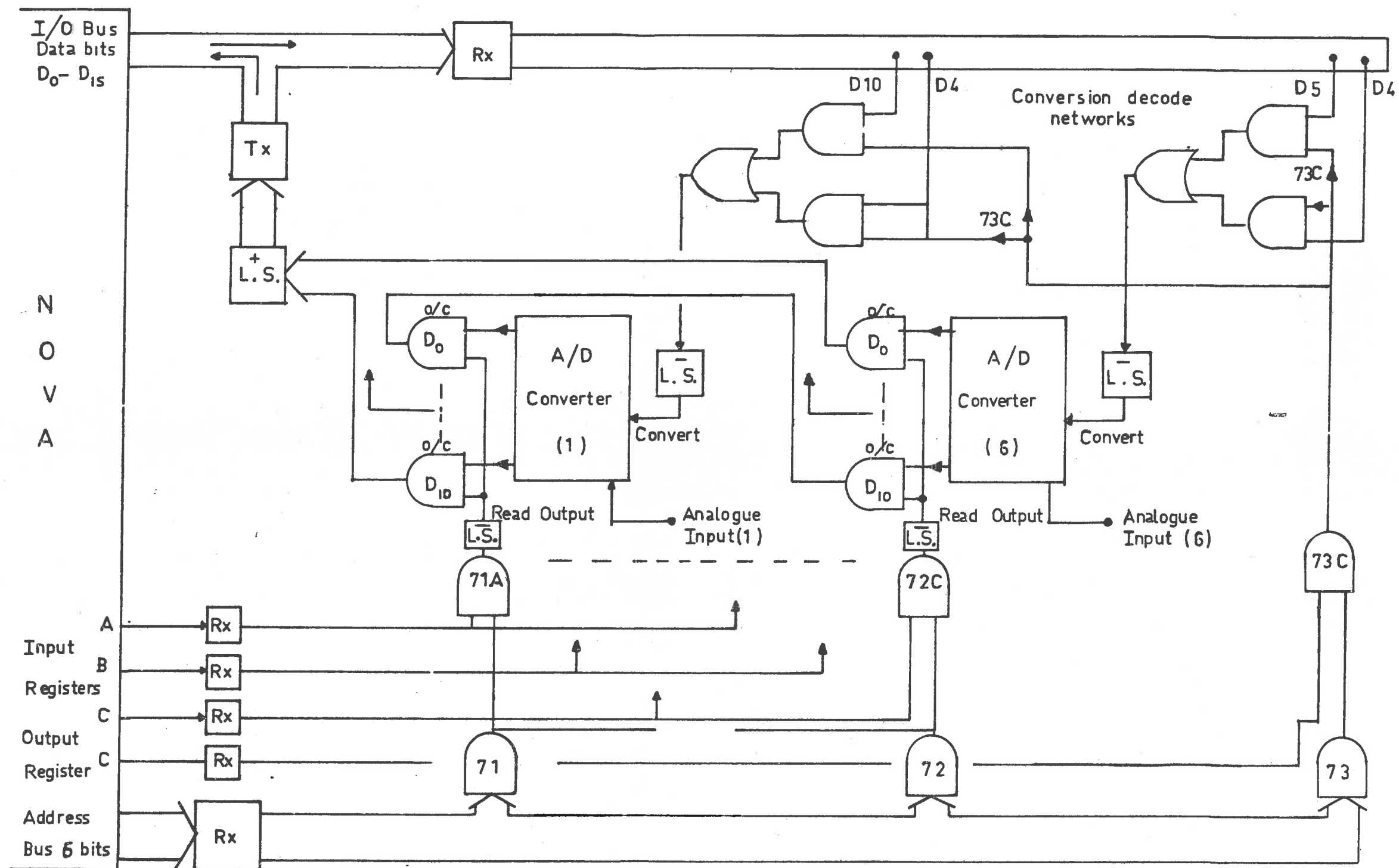


FIGURE A9. NOVA control of the A/D converters.

A/D Number	Mask Bit	"1" i.e. Octal Number in Acc.
(1)	10	40
(2)	9	100
(3)	8	200
(4)	7	400
(5)	6	1000
(6)	5	2000
All units convert	4	4000

Thus if an A/D conversion on unit 6 only is required a mask of 2000 octal is set up in accumulator number 1 say. The program instruction

DOC 1,73

is then given. Referring to figure A9, this instruction sets up a logic "1" on data line D_5 of the I/O bus, device code 73 on the address bus, and output register number C on the register line. This enables AND gate 73C. The output of this gate together with D_5 , enable the conversion decode network which pulses the conversion line of A/D converter number 6. This unit then converts its analogue input to the equivalent two's complement binary form. Simultaneous conversion of all units is achieved by setting data bit 4 to a logical "1", i.e. using a mask of 4000 octal in the accumulator. Any combination of the converters can be initiated by combining the appropriate bits into the mask.

Once the converted data has been set up in the holding registers of the units, an input instruction transfers this information into the central processor.

Device codes 71 and 72 with input registers A, B and C are used for the six converters. For example, data is read from converter number 1 with the instruction,

DIA Acc, 71

This enables gate 71A which then enables the gates D_0 to D_{10} of figure A9. The data appears on the I/O bus through the wired OR function of these open collector gates and is clocked into the accumulator by the central processor. Data bit D_0 is the sign bit and D_1 to D_{10} the magnitude bits of the two's complement binary number.

After a conversion has been initiated, at least 11 μ secs must be allowed for its completion before the data can be read into the computer. The program may either continue with instructions not associated with the A/D converters, or wait for the conversion to be completed by interrogating the DONE flag. Once the conversion is completed the DONE flag is set by the converter. (DONE flags are not shown in figure A9).

A complete sequence which could be used to convert the analogue signal at the input of A/D number 1 and read the binary equivalent into accumulator number 2 of the central processor is;

```

LDA 1, MSK 1      ; set up a mask of
                   40 octal in acc. 1
DOC 1, 73          ; output this mask to
                   initiate a conversion
SKPDN 71           ; interrogate the DONE
                   flag. If set then

```

skip the next instruction.

JMP -1 ; DONE flag is not set
 therefore jump back
 and interrogate it again

DIA 2,71 ; The DONE flag has been
 set, therefore read the
 (MSK 1 = 40_8) holding register of the
 converter into accumulator
 number 2.

A.4 THE MACHINE SPEED TRANSDUCER

Figure A1 shows how the alternator terminal voltage and the real power delivered to the infinite bus are sampled by the computer via the A/D converters. The resolution of the converters of 1 in 2000 of full scale is therefore the resolution on the measurement of the terminal voltage and electrical torque states. The predicted deviation of these two states from their steady state for an input step of 0.04 p.u. is approximately 0.04 p.u., so that the deviations may be resolved to 1 in 80 (0.04×2000) of their peak response. The peak deviation of speed from synchronous speed for an 0.04 p.u. input step however is of the order of 0.0016 p.u. Thus to obtain the same resolution on speed deviation as for the other states, a resolution of 1 in 50,000 is required in the measurement of synchronous speed. This is summarised in table A1.

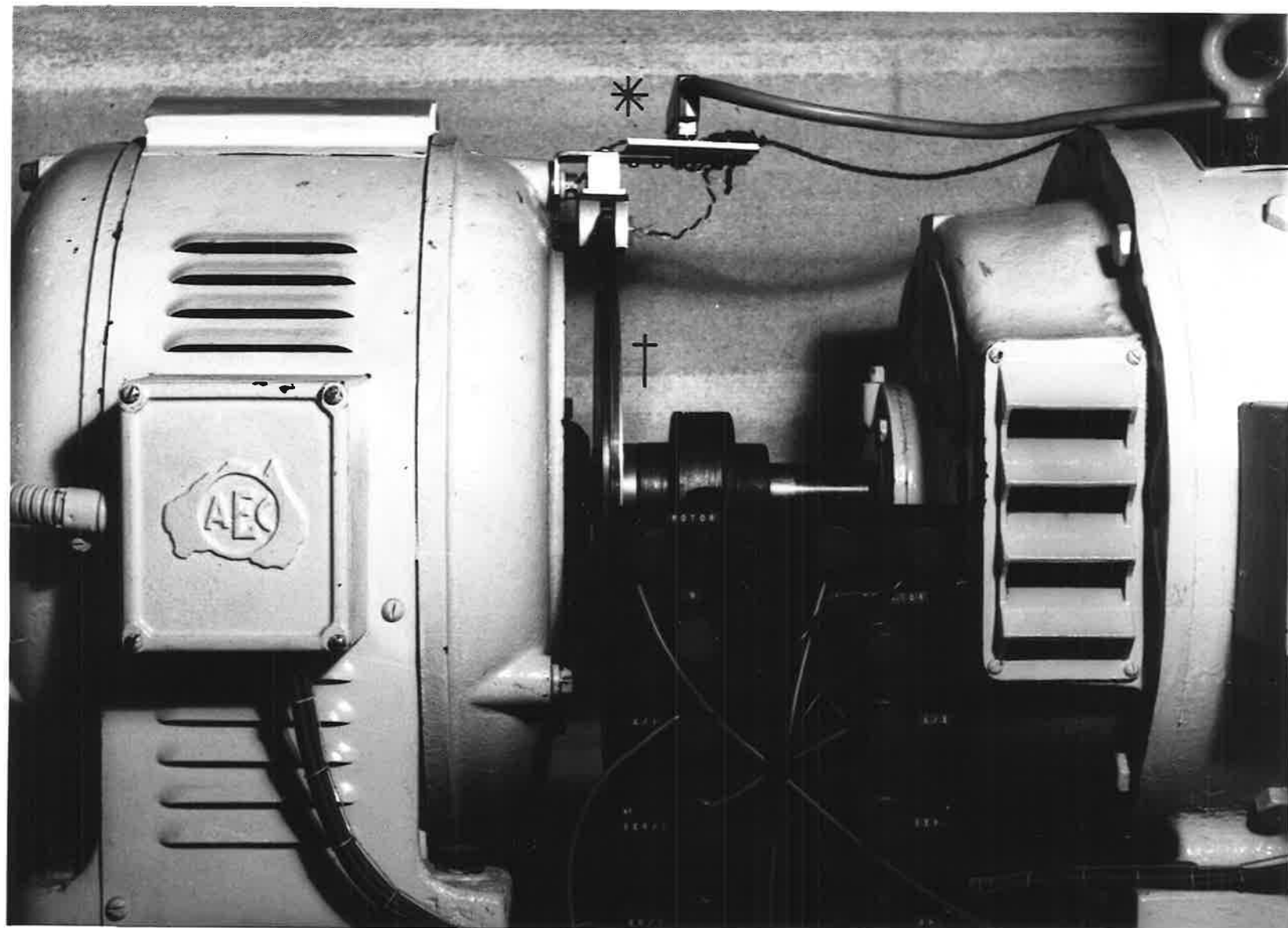


Figure A10 The photodisc⁺ and pickup*
for the speed transducer.

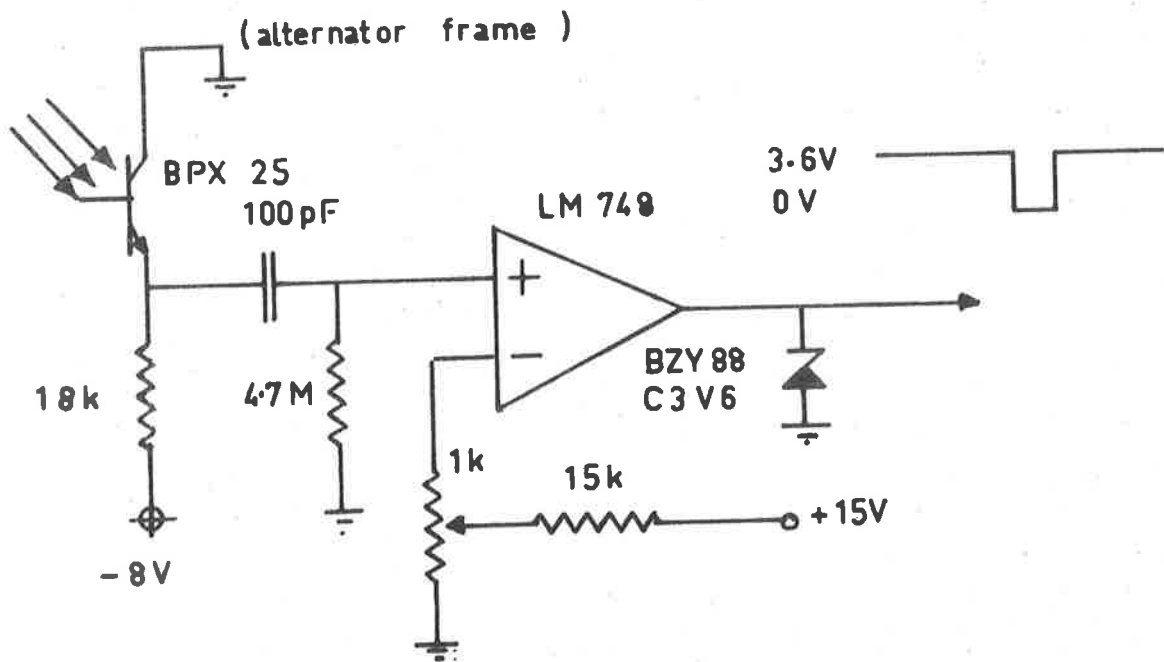


Figure A11 The photo pick up, amplifier and squaring circuit.

State	Resolution in measurement of 1 p.u.	Magnitude of deviation for an 0.04 p.u. step input.	Resolution on deviation
v_t	1 in 2000	0.04	1 in 80
C_e	1 in 2000	0.04	1 in 80
ω	1 in 50000	0.0016	1 in 80

Table A1 Resolution in the measurement of the states

When operating near unity power factor, the system time constants are of the order of 200 mseconds. Thus for transducer time constants to be neglected in system analysis, a sampling time of no more than 10 mseconds was considered necessary. Synchronous speed for the 50Hz 4 pole alternator is 1500 rpm, or 1 mechanical revolution per 40 mseconds. The speed of the rotor must therefore be resolved to 1 part in 50000 during one quarter of a revolution. Also as the A/D converters do not have this resolution, a direct digital output is required. A system to meet these specifications was developed.

A transducer producing an output pulse every quarter revolution (every 10ms at synchronous speed) was mounted on the alternator shaft, and a binary counter gated between pulses to count the output of a 50MHz crystal oscillator (counting 50,000 pulses in the 10ms).

The shaft transducer consists of a perspex disc blacked out except for four orthogonal radial lines. A photo-transistor is activated every time one of these lines passes between it and a light source. The transistor output is amplified and squared to give TTL compatible logic levels. The light source, photo pick-up and squaring circuit are mounted on the frame of the alternator as shown in figure A10. The circuit itself is shown in figure A11.

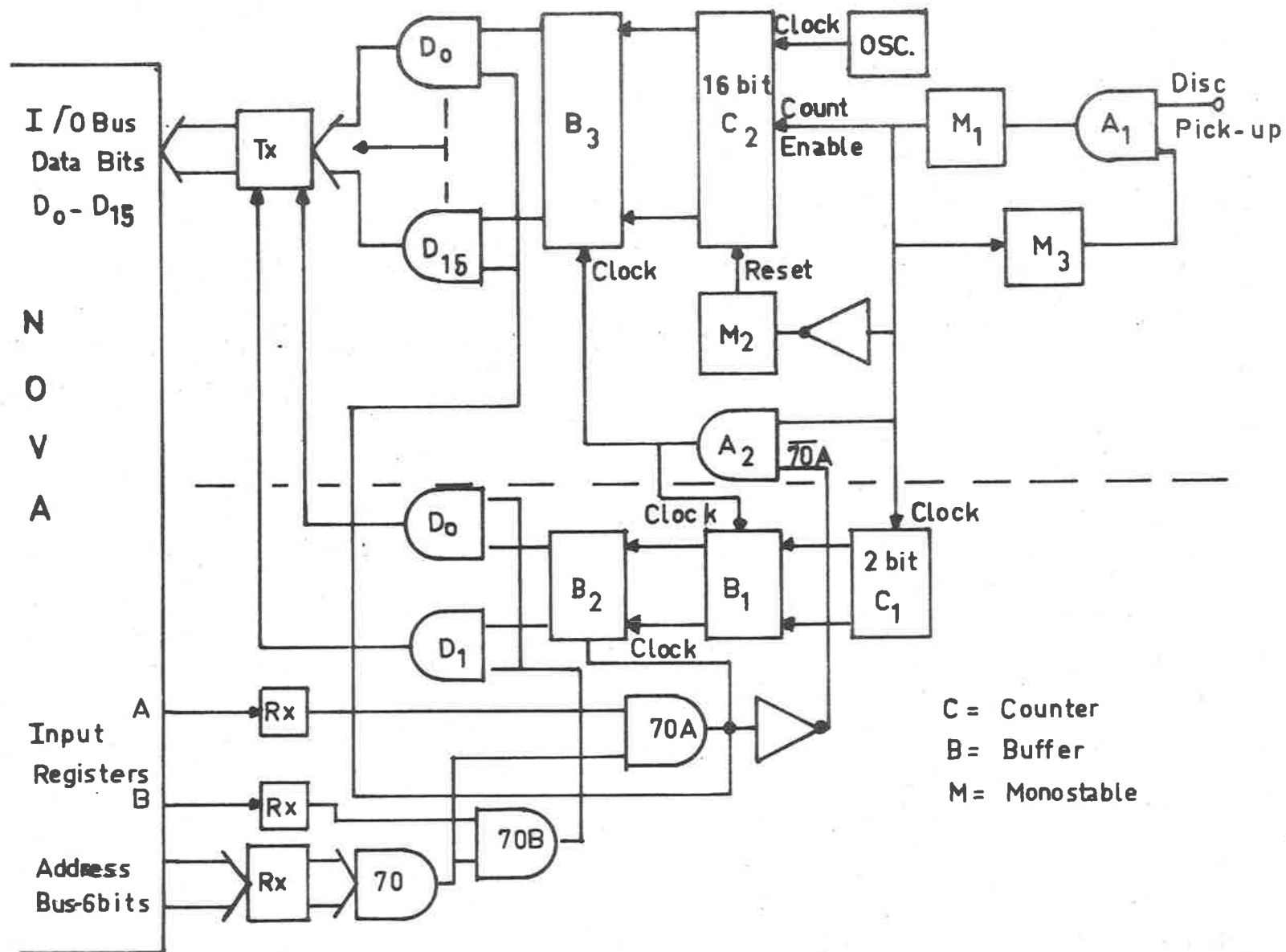


Figure A12 The binary counters and their control networks for the speed transducer.

Referring to figure A12, the output of the disc pick-up becomes the input of the logic ring A1, M1, M3, which removes any noise or ringing appearing on the edges of the pick-up output pulse. Waveforms of this circuit are shown in figure A13. Assume initially that the disc pick-up is at a logical "1", with gate A1 and monostalbes M1 and M3 also at "1" levels. As a line on the disc passes the pick-up its output drops to "0", monostalbe M1 is triggered, and its output becomes "0" triggering M3 to "0". When the line passes from the pick-up after approximately 15 μ seconds, its output returns to a "1" level, but gate A1 is still disabled by the 40 μ seconds pulse from M3. Thus as soon as M1 is triggered once, noise on the trailing or leading edges of the pick-up pulse do not affect the operation of the logic circuits.

During the 2 μ seconds that M1 is low the counter C2 is disabled. As M1 returns to a logic "1" level the information held in the counter is clocked into the buffer B3, and monostalbe M2 is triggered for a 2 μ second pulse which resets the counter C2, after which C2 restarts counting the output pulses of the 5MHz crystal oscillator for a further 10 msec. The above cycle is then repeated.

Input device code 70A was allotted for the speed transducer. The programmed instruction;

DIA Acc, 70

is therefore used to read the contents of the buffer B3 into the accumulator Acc, via the open collector AND gates D_0 to D_{15} . While the NOVA central processor is reading the buffer B3, the buffer clock is inhibited by AND gate A2 to prevent indeterminate data being read in.

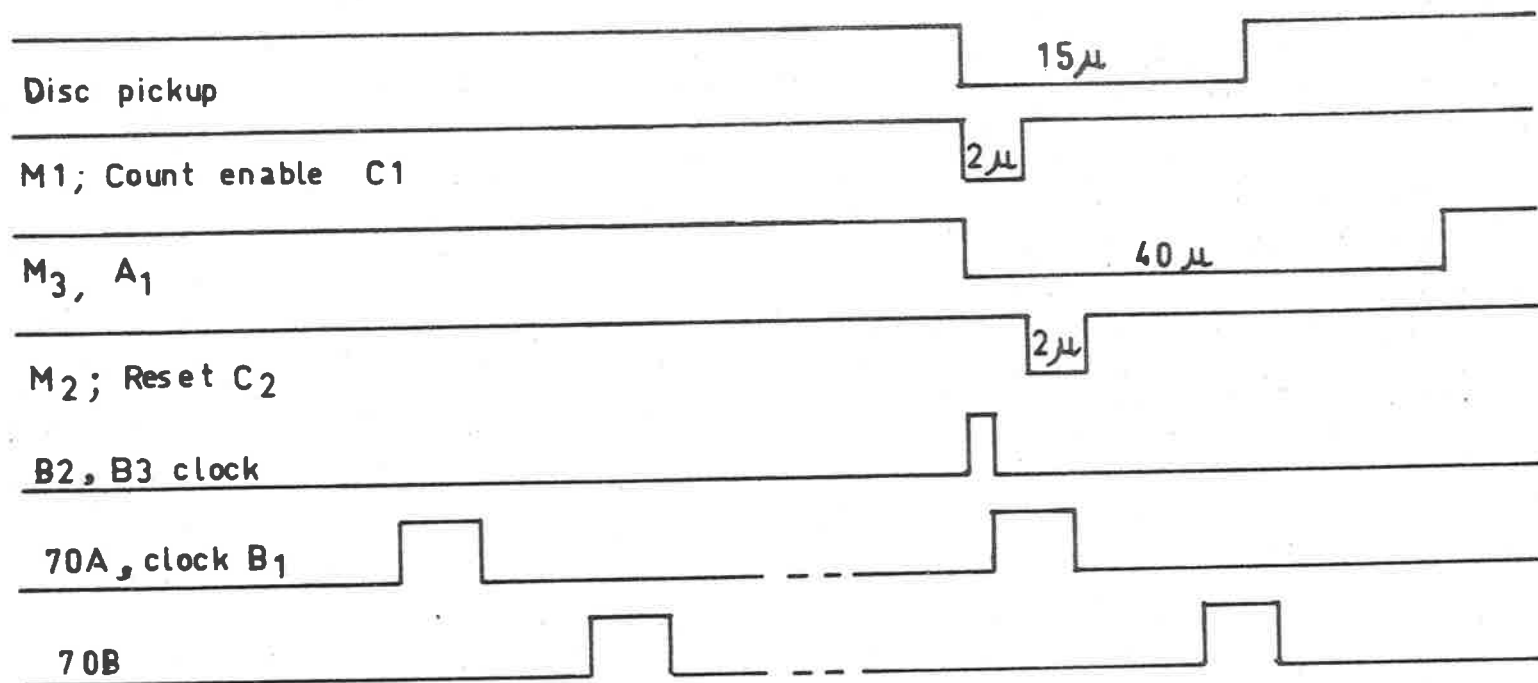


Figure A13 Waveforms associated with the circuit of figure A12.

The resolution of the transducer is a function only of the clock frequency and the number of bits in the counter. The accuracy however is dependent on the positioning of the lines on the photo-disc. To obtain an accuracy of 1 in 50,000 on a six inch radius disc (approximately 36" circumference), the lines would have to be drafted with an accuracy of 0.001 inches. Precision machining of the supporting boss and true running of the alternator shaft would also be required. However as deviation from the steady state synchronous speed and not the synchronous speed is used as the system state, the following technique was employed whereby the deviation is measured with the same percentage accuracy as synchronous speed.

A system which effectively labels the four sectors of the disc was developed. Each time a line passes the disc pick up and monostable M1 is triggered, a two bit counter C1 (in figure A12) is incremented. This is clocked into a buffer B1 by the same pulse which clocks the new oscillator count in C2 into buffer B3. When the oscillator count is read from B3 into the computer with the DIA Acc, 70 instruction, the data in B1 is clocked into buffer B2. The programmed instruction;

DIB Acc, 70

then reads the number of the sector from B2 into the accumulator Acc. Thus when the alternator is synchronized with the infinite bus, the two bit ring counter C1 indicates which sector is controlling the oscillator count in C2. Some of the gating and buffers in figure A12 which may appear superfluous are necessary as the programmed reading of the counters and the pulses from the photodisc are not synchronized.

The two programmed instructions;

DIA 1, 70

DIB 2, 70

therefore read the number of pulses from the 5MHz oscillator in approximately a quarter revolution of the shaft, and also the sector of the disc to which this reading applies. An initial set of readings (taken at 10 msecond intervals) may be;

Sector No.	Initial Oscillator count for $\frac{1}{4}$ of a rev.
0	49876
1	50137
2	50103
3	49938

At a later stage following a disturbance the readings may be;

Sector No.	Subsequent Oscillator Count	Deviation
0	49926	+50
1	50188	+51
2	50155	+52
3	49990	+52

Per unit speed deviation is then computed as deviation divided by initial count. If synchronous speed (= 1 p.u.) is assumed to be a count of $50,000 \times 0.2 \mu\text{second}$ pulses for a quarter of a mechanical revolution, the error generated by this assumption is equal to the deviation of the initial counts from 50,000 due to mechanical mis-alignment. Comparison between the per unit deviations with 50,000 as 1 per unit and the initial readings for the various sectors as 1 per unit, shows that the error introduced is much less than the figure of +0.00004 p.u. resulting from machine vibration and noise in the photo pick-up.

The logic functions were implemented using standard 7400 series T.T.L.

The following specifications apply to the transducer:

- (1) Resolution 0.00002 p.u.
- (2) Repeatable accuracy +0.00004 p.u.
- (3) Aperture time 10ms
- (4) Direct binary output of 16 bits.

A.5 MEASUREMENT OF THE INFINITE BUS FREQUENCY

The infinite bus frequency is measured with the same resolution and accuracy as the machine speed using a similar method. A binary counter is again used to count pulses from a 5MHz oscillator, with the gating of the counter controlled in this case by the zero crossing of one phase of the infinite bus voltage waveform. Thus 50,000 pulses, each 2 μ s long, are counted between zero crossings of a 50Hz signal.

Referring to figure A14, the waveform of the A phase of the infinite bus is squared using a series 100K Ω resistor and two 3.6 volt zener diodes in parallel. Further squaring is provided by an LM748 operational amplifier with its output clamped to TTL compatible logic levels. With this low impedance, noise free source the logic ring used at the input of the speed transducer is not necessary, and the squared bus waveform is applied directly to the input of the monostable M1 and the count enable input of the 16 bit binary counter C.

As the mains voltage changes from positive to negative, the negative going edge of the squared waveform W triggers monostable M1 to "0", and the "0" level of W inhibits the counter C. Monostable M1 clocks the count into the buffer B, and as it returns to "1" level triggers monostable M2. The 2 μ second pulse from M2 resets the counter C.

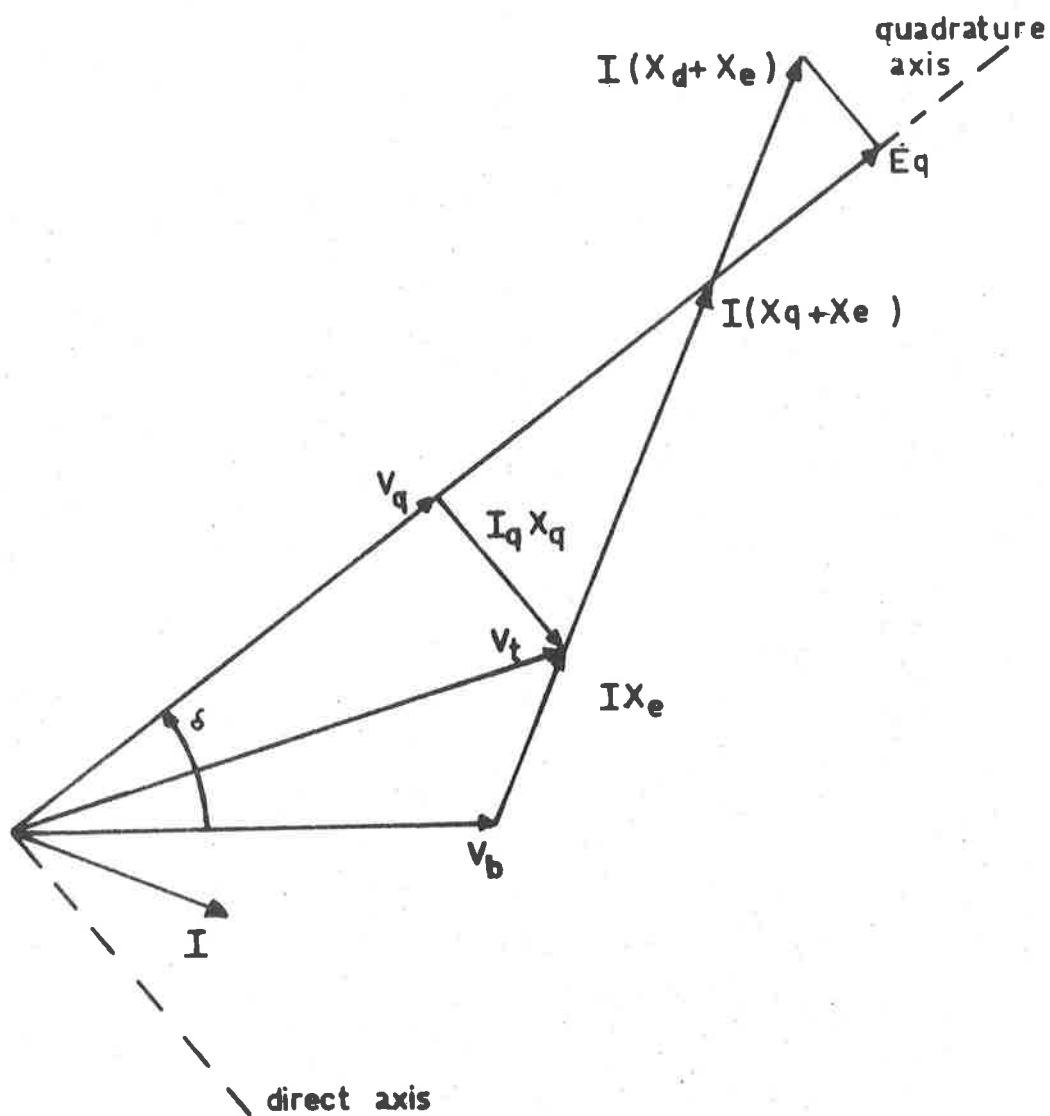


Figure A15 Voltage vector diagram of the S.G.T.L. system.

Input device code 70C was used for the transducer, so that the programmed instruction

DIC Acc, 70

reads the contents of the buffer B into the accumulator Acc. As for the speed transducer the buffer clock is inhibited when data is being read by the computer.

While the mains voltage is negative the counter C is disabled. As it goes positive, the change to a "1" level in W allows the count to continue. The count proceeds until the mains voltage crosses zero again and the above cycle is repeated. As the count is inhibited for every half cycle, a new reading is only taken every 20 mseconds. Also as the device operates on zero crossings of the mains voltage, any waveform distortion due to external disturbances results in noise in the reading of frequency. However the only first order effect of mains frequency deviation is on the speed deviations of the alternator, and the inertia of the machine set is sufficient to effectively filter any very short term disturbances. Therefore the limitation of the 20 msecond sampling period and the noise due to waveform distortion can be removed by digital filtering during subsequent analysis of the recorded mains frequency deviations, without losing any information necessary to assess the effect of these deviations on the system.

A.6 MEASUREMENT OF THE POWER ANGLE

The power angle δ , used in the equations describing the SCTL system, is the angle between the quadrature axis of the synchronous machine and the vector representing the infinite bus voltage, as shown in figure A15. The quadrature axis is aligned with the rotor of the machine, and the busbar voltage vector may be regarded as

a synchronously rotating reference.

If the time t_1 at which the rotor passes a reference point on the stator, and the time t_2 at which the synchronous reference passes an equivalent point are recorded, the angle between the axis can be determined with a knowledge of the synchronous speed i.e. $\delta = \omega_s (t_2 - t_1)$. The positive to negative zero crossing of the phase to neutral busbar voltage of phase A is used to mark the time t_2 . An equivalent mark to t_1 is achieved as follows.

A blacked out disc with a clear diametric line is mounted on the alternator shaft, and a photo activated pick up is mounted on the frame to form a system similar to that used for the speed transducer (figure A10). This device produces an output pulse every electrical revolution. With the armature of the alternator open circuited, or with zero real power in the SGTL system, the vector representing the terminal voltage in figure A15 is aligned with the quadrature axis. Under these conditions δ is zero, and the positive to negative zero crossing of the A phase machine terminal voltage is coincident with the zero crossing of the infinite bus A phase voltage. The photodisc is positioned on the alternator shaft to produce its pulse as the open circuit terminal voltage crosses zero. When the alternator load is varied and the operating point shifts in the PQ plane, this pulse marks the time t_1 at which the rotating quadrature axis passes the pick up reference point. δ is measured by counting the pulses from an oscillator for the time t_1 to t_2 , i.e. from when the disc pick up pulses until the positive to negative zero crossing of the A phase busbar voltage occurs. For an oscillator frequency of 18kHz (i.e. 360 pulse every 20ms) the counter records δ in electrical degrees.

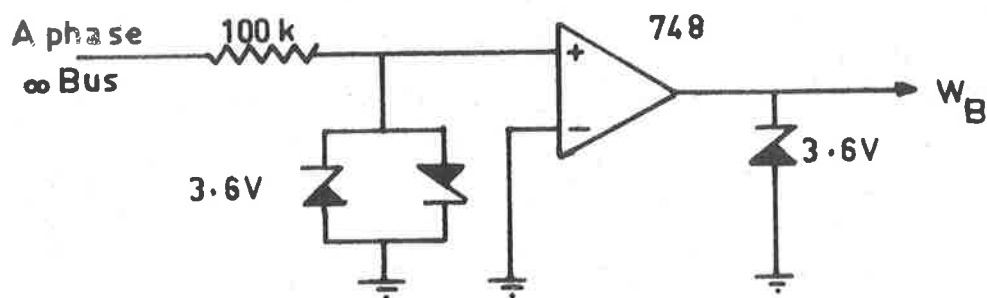


Figure A16 Circuit for squaring infinite bus waveform to TTL logic levels.

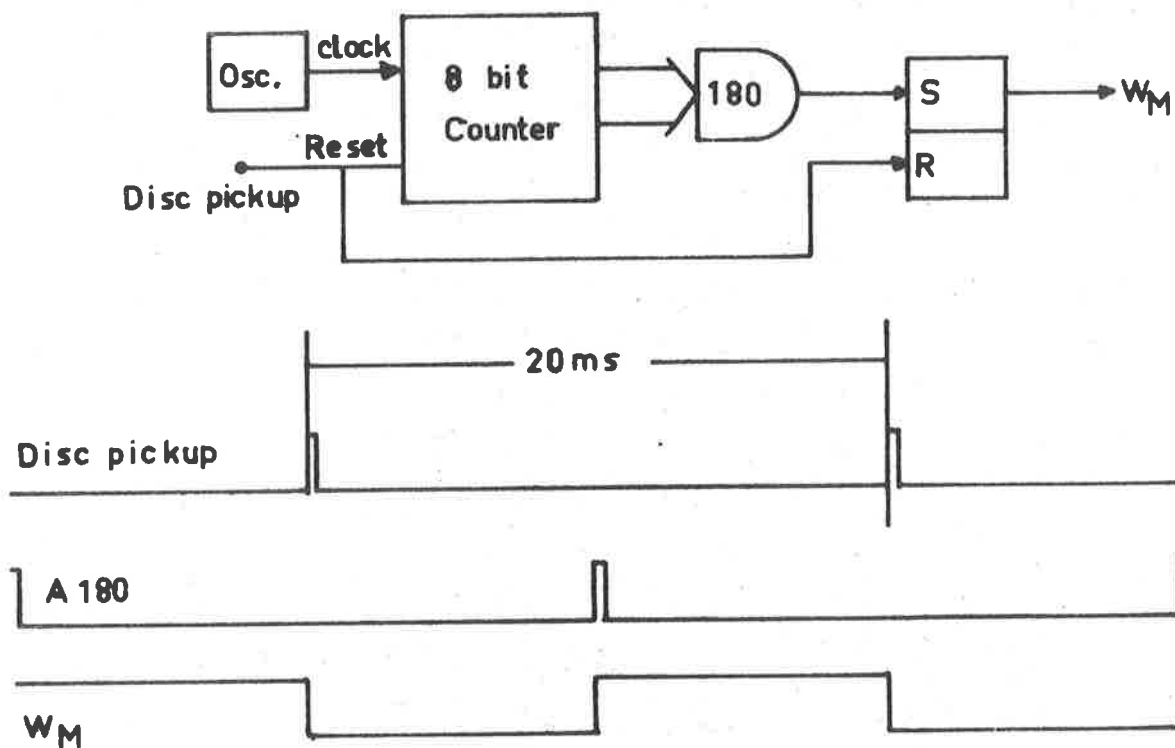


Figure A17 Construction of the W_M waveform.

The mounting of the photodisc and pick up was similar to that shown in figure A10, and the same pick up circuit as shown for the speed transducer in figure A11 was used. In this case the disc has only one diametrical line instead of the two used in the speed transducer.

Figure A16 shows the circuit used to square the infinite bus voltage waveform to TTL compatible logic levels. (This is the same as the squaring circuit used in the measurement of bus frequency, figure A14). A square wave whose trailing edge is coincident with the pulses from the disc pick-up was generated using the circuit of figure A17. An 18kHz (360 in 20ms) oscillator clocks an 8 bits binary counter. The output of the counter is wired to an AND gate which produces a pulse when the count reaches 180 degrees (264 octal). This pulse sets an R.S. flip flop to a logical "1" level. Both the flip flop and the counter are reset to "0" by the disc pick up (every 20 msecs at synchronous speed). The square wave generated is labelled W_M .

A disc with four radial lines and the pick up directly clocking a flip flop would not have been satisfactory as there would be no way of marking which pulses were coincident with which zero crossing (i.e. positive to negative or vice versa) on the open circuit terminal voltage, and a waveform 180 degrees out of phase with that required could be produced.

The squared infinite bus voltage waveform W_B and the waveform W_M become the inputs to the transducer in figure A18. These inputs are NANGED to produce the waveform N in figure A19, and ORED to form waveform O. The positive going edge of N clocks a D type flip flop to a "1" level, and when O drops to "0" the D flip flop

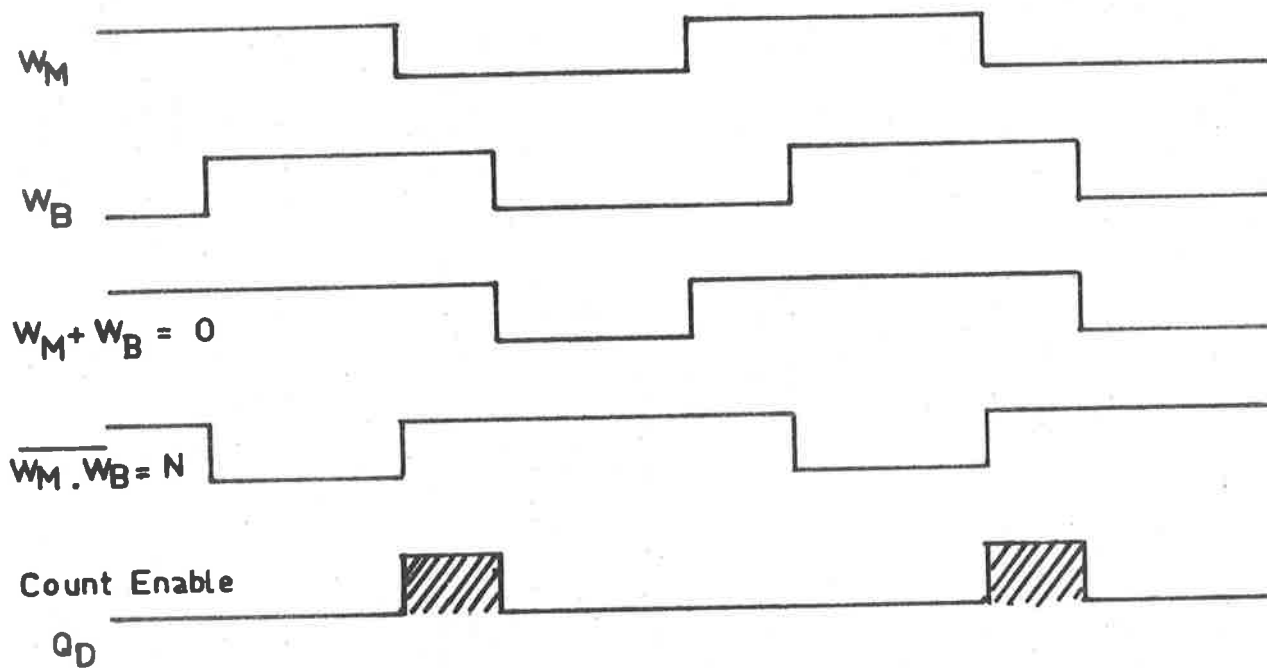


Figure A19 The waveforms associated with the δ transducer of figure A17.

is reset to "0". Thus the time for which D is at a "1" level is from when the quadrature axis passes its reference point (and W_M drops to a "0" level) until the infinite bus voltage crosses zero (and W_B drops to "0"). During this time the 9 bit binary counter, C, is enabled and counts the pulses from the 18kHz oscillator to determine δ in degrees. The count is disabled by OR gate D dropping to "0", simultaneously triggering monostable M for 2 μ seconds to clock the value of the count into the buffer B.

As this configuration is symmetrical, (i.e. waveforms W_M and W_B could at this stage be interchanged with no change in operation) it is possible to obtain positive (generating) or negative (motoring) readings for δ by using an UP/DOWN counter. One of the signals W_M or W_B determines the direction of the count. As shown in figure A18, W_B is used and a positive count occurs if the counter is enabled while W_B is positive. The waveform of figure A19 correspond to W_B lagging W_M with the alternator in a generating mode.

Input code 70A was used for this device. This is the same code as used by the speed transducer, but the angle transducer was removed when no longer required to be replaced by the speed transducer with a minimum of rewiring of the interface. The NOVA address decoding network operates as described for previous devices, and also inhibits the buffer clock via AND gate A while a read cycle is in progress. The counter output is 9 bits in 2's compliment binary form. The least significant 8 bits are the magnitude bits, connected to the data lines D_8 to D_{15} of the I/O bus on NOVA. The 9th bit of the counter is the sign bit. By connecting this one bit to the data lines D_0 to D_7 of the I/O bus,

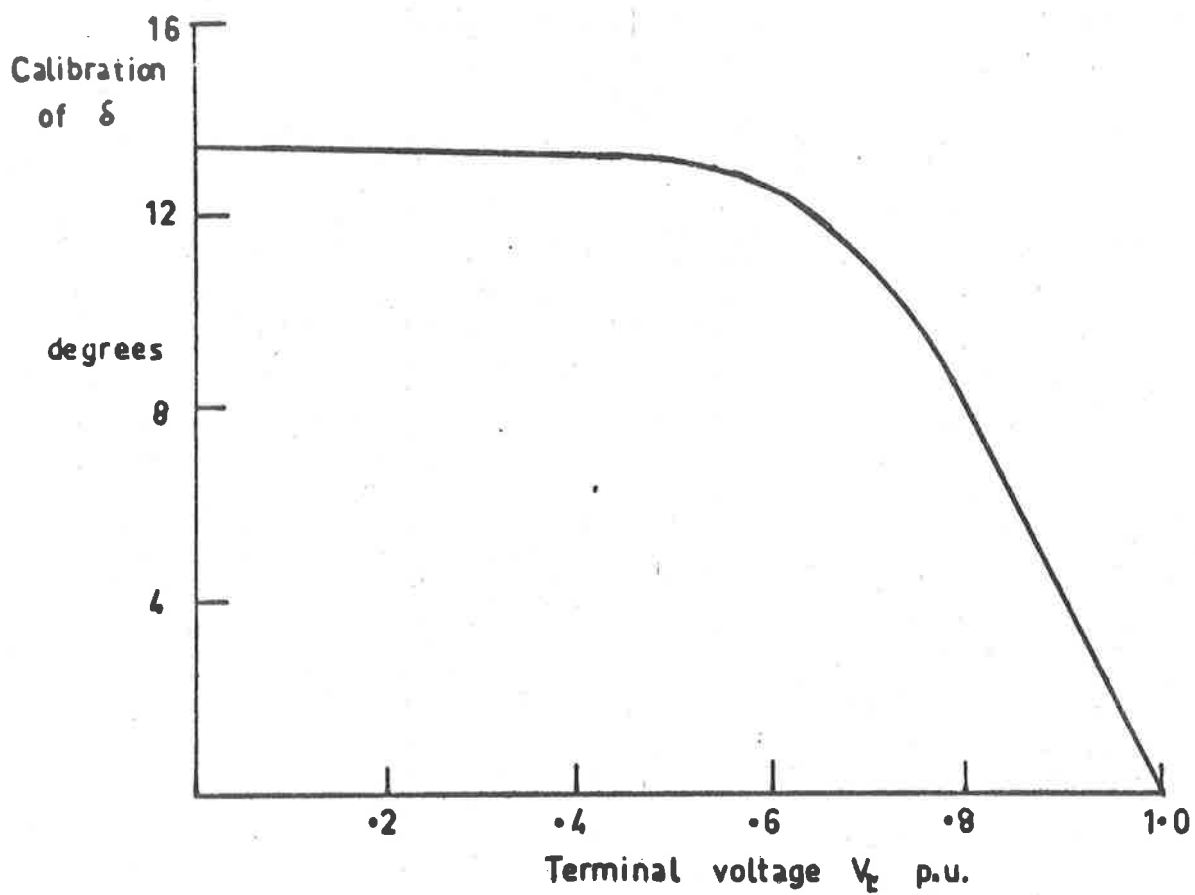


Figure A20 The calibration curve for the δ transducer, showing the effect of saturation on the alignment of the quadrature axis with the rotor axis.

the 9 bit two's complement counter output is transformed to a 16 bit two's complement number in the NOVA accumulators, the least significant bit representing one degree.

The quadrature axis of the equivalent 2 axis machine may be considered fixed with respect to the rotor of the synchronous machine, if the effect of saturation is ignored. Due to saturation however, the alignment of the quadrature axis on the rotor varies with air gap voltage. The point in time at which the quadrature axis passes a reference point on the rotor has been taken as the time at which the positive to negative zero crossing of the open circuit terminal voltage occurs. The time at which the rotor passes its reference point is given by the pulse from the disc pick up circuit. These two references were aligned when the open circuit terminal voltage was 1 per unit, so that the rotor axis and the quadrature axis may be considered coincident when the terminal voltage is one per unit. As the field current of the open circuited machine is reduced the two reference points separate. A plot of this separation in degrees against open circuit terminal voltage is shown in figure A20.

For this particular machine, the Potier reactance is negligible and the air gap voltage may be approximated by the terminal voltage. Therefore the effect of saturation on the measurement of δ can be allowed for by using figure A20 as a calibration curve. The terminal voltage at any particular operating point is used to determine the correction to be subtracted from the measured value of δ .

The curve of figure A20 was checked by operating the SGTL system at zero real power ($\delta = 0$) while varying the reactive power

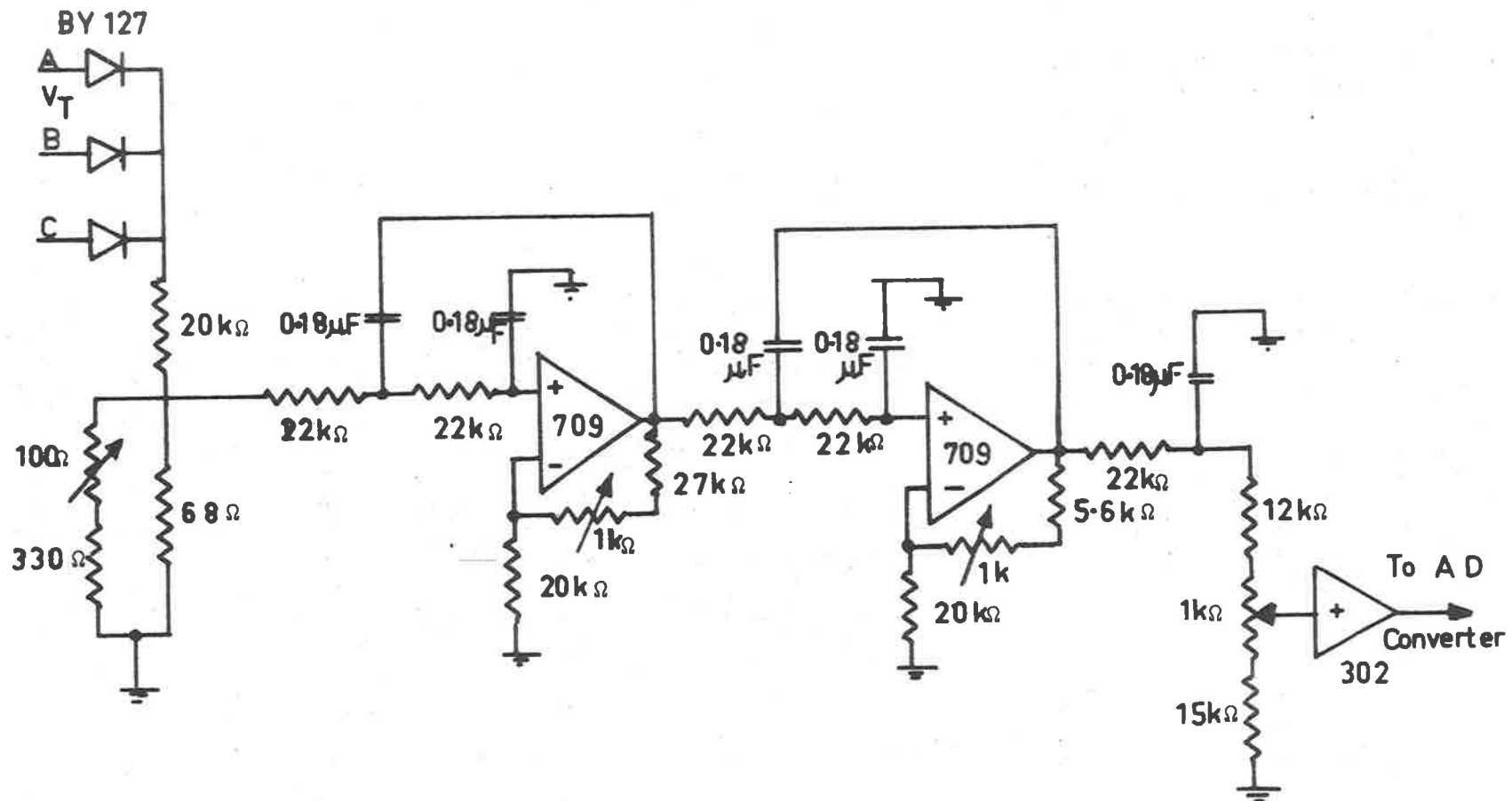


Figure A21 The terminal voltage A/C to D/C converter.

from 0 to -1.0 p.u. This corresponds to a terminal voltage range of 1.0 to 0.7 p.u. The accuracy of the corrected reading of δ was confirmed in the open loop stable region of the PQ plane by recording the operating point from which the angle was computed.

As δ is used only to plot the static curves from which the K parameters are evaluated, the 20 msec sampling period and the relatively low resolution of the transducer are not limiting.

The specifications for the unit are

- (1) Resolution 1 degree (electrical)
- (2) Accuracy (using calibrated correction) ± 1 degree
- (3) Range ± 180 degrees (generating or motoring)
- (4) Sampling period 20 ms
- (5) 9 bit binary output in 2's complement.

A.7 THE TERMINAL VOLTAGE TRANSDUCER

The voltage at the terminals of the machine is converted into an equivalent, small D.C. value which the controller samples via the A/D converter interface.

(The Electricity Trust of S.A. operates with a grounded neutral system, so that earth and neutral are interchangeable in this transducer and also in the power transducers in section A8).

The circuit in figure A21 rectifies the terminal voltage and a unity D.C. gain, fifth order active Butterworth filter reduces the rectifier ripple to 1mV (0.1% of F.S.D.). The filter output, which is equal to the peak of the input voltage waveform, is applied to an A/D converter with an input range of ± 1 volt peak. Thus the resistive divider following the three phase rectifier reduces 1 p.u. terminal voltage to 1 volt peak (a ratio of $\sqrt{2} \times 240$)

The filter consists of two second order active networks cascaded with one passive first order network. The transfer function of a 5th order Butterworth filter, frequency scaled to 1 radian per second is:

$$\frac{K}{(s + 1)(s^2 + 0.62s + 1)(s^2 + 1.62s + 1)}$$

For the actual R.C. filter this corresponds to:

$$\frac{A_1}{R^2 C^2 s^2 + RC(3-A_1)s + 1} \times \frac{A_2}{R^2 C^2 s^2 + RC(3 - A_2)s + 1} \times \frac{1}{RCs + 1}$$

assuming all resistors and capacitors are of equal value throughout the filter. A_1 and A_2 are the closed loop gains of the amplifiers

With $RC = 1$ ($= 1/\omega_{3dB}$), comparing coefficients of the two functions gives $A_1 = 2.38$ and $A_2 = 1.38$. The feedback resistors about the amplifiers in figure A20 set their gains to these values. The output ripple is reduced to 1mV by choosing a 3dB cut off frequency of 40Hz i.e. the RC product is frequency scaled by $40 \times 2\pi$ so that $RC = 1/80\pi$. A value of 22 Kohms for R provides suitable impedance levels for the operational amplifiers and forces C to 0.18 μ F. The D.C. gain of the filter is $A_1 \times A_2 = 3.28$. This is reduced to unity by the resistive divider following the filter. The final stage is a unity gain, low output impedance buffer amplifiers.

By using a high order filter with a 40Hz cut off frequency the ripple is reduced to the required level without a long time constant being introduced. The effective time constant of this circuit for an input step is 15 ms.

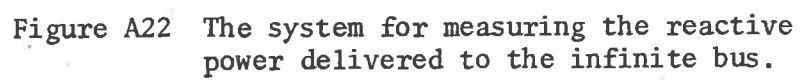


Figure A22 The system for measuring the reactive power delivered to the infinite bus.

A.8 THE REAL AND REACTIVE POWER TRANSDUCERS

Figure A22 shows the system used for measuring the reactive power. A similar system, without the integrators, measures real power.

The devices labelled C.T. transform the line current of each phase to a voltage which is proportional to and in phase with the current waveform. For reactive power measurement these signals are phase shifted 90 degrees by the integrators and then multiplied by a voltage proportional to the infinite bus voltage. The output of the multipliers is proportional to the instantaneous reactive power delivered by each phase. Assume each of the devices has unity gain. If a current A p.u. is flowing with a power angle ϕ , and the phase to neutral infinite bus voltage is V p.u. then the output of the A phase multiplier may be derived as follows:

$$\begin{aligned}
 M_A &= V \text{ p.u.} \times A \text{ p.u.} \\
 &= 2 V \sin \omega_s t \times 2 A \sin \left(\omega_s t - \frac{\pi}{2} - \phi \right) \\
 &= 2 V \sin \omega_s t + 2 A \cos \left(\omega_s t - \phi \right) \\
 &= 2 VA \sin \omega_s t \left(\cos \omega_s t \cos \phi + \sin \omega_s t \sin \phi \right) \\
 &= 2 VA \left(\frac{1}{2} \sin 2 \omega_s t \cos \phi + \frac{1}{2} \cos 2 \omega_s t \sin \phi + \frac{1}{2} \sin \phi \right) \\
 &= VA \left(\sin 2 \omega_s t \cos \phi + \sin \phi (1 + \cos 2 \omega_s t) \right) \\
 &= VA \sin (2 \omega_s t + \phi) + VA \sin \phi
 \end{aligned}$$

For a balanced system, all phase currents are A amps in magnitude with the phases of V and A shifted by ± 120 degrees. The outputs of the multipliers on these phases is therefore,

$$\begin{aligned}
 M_B &= VA \sin \left(2 \omega_s t + \phi + \frac{2\pi}{3} \right) + VA \sin \phi \\
 M_C &= VA \sin \left(2 \omega_s t + \phi - \frac{2\pi}{3} \right) + VA \sin \phi
 \end{aligned}$$

These signals are added to produce an output at the summing amplifier of;

$$\begin{aligned}
\Sigma &= M_A + M_B + M_C \\
&= VA (\sin (2\omega_s t + \phi) + \sin (2\omega_s t + \phi + \frac{2\pi}{3}) \\
&\quad + \sin (2\omega_s t + \phi - \frac{2\pi}{3}) + 3 \sin \phi) \\
&= VA (\sin (2\omega_s t + \phi) + \sin (2\omega_s t + \phi) \cos \frac{2\pi}{3} \\
&\quad + \sin (2\omega_s t + \phi) \cos \frac{2\pi}{3} + 3 \sin \phi) \\
&= 3VA \sin \phi
\end{aligned}$$

i.e. for a balanced system the output of the reactive power transducer is a D.C. voltage proportional to the VA product and the sin of the power angle. Similarly the output of the real power transducer can be shown to be $3VA \cos \phi$. Due to inbalance in the alternator some ripple at a frequency of $2\omega_s$ is present on the output of the summing amplifiers. This is reduced to less than 0.1% of full scale by an active fifth order Butterworth filter.

A 8.1 THE CURRENT AND VOLTAGE TRANSFORMERS

The maximum allowable voltage input to the analogue multipliers is 10 volts. It was assumed that any surge on the infinite bus voltage would not exceed 1.25 p.u. (300 volts) so that the phase to neutral voltage was reduced in the ratio of $300\sqrt{2}$ to 10. A simple resistive divider labelled V.R. in figure A22 performs this function.

To obtain maximum accuracy from the transducers, the current transformers must be capable of producing a peak output voltage of 10 volts to be used as the input of the analogue multipliers. In addition their outputs must be in phase with the line currents, and they must be linear to within 0.1% of full scale. By loading a simple current transformer with virtually zero burden, the problems associated

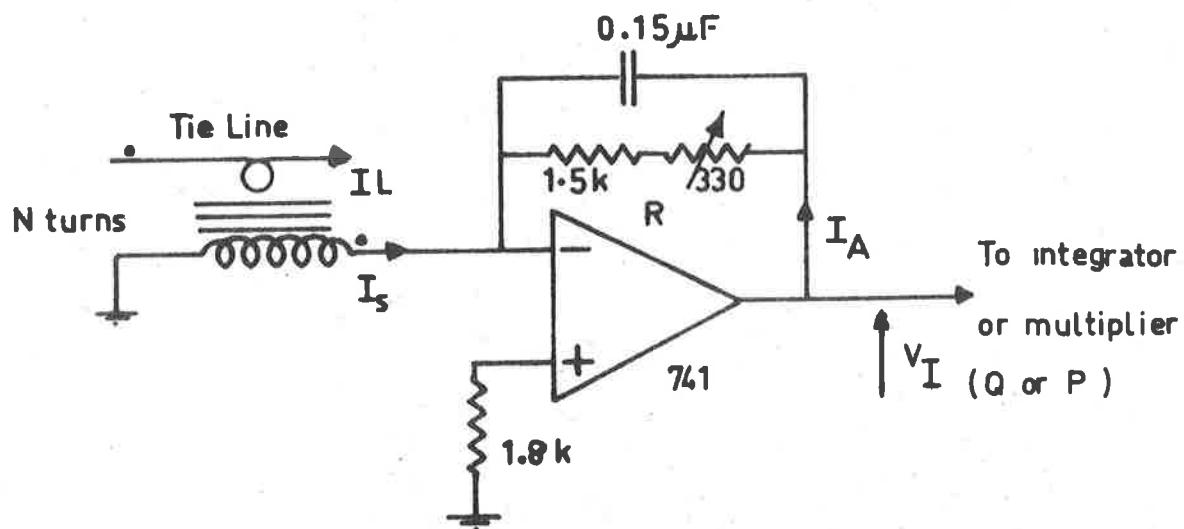


Figure A23 The circuit used to transform line current to an equivalent voltage.

with producing a suitable device are greatly reduced. The circuit shown in figure A23 achieves this.

The current flow in the secondary (I_S) is matched by the feedback current (I_A) of the operational amplifier to achieve a virtual earth at the input of the amplifier, and virtually zero burden on the current transformer. The output voltage from the amplifier is:

$$V_I = I_A R = I_S R = I_L R/N$$

Originally the line current of the SGTL system was expected to have a maximum of 3.5 p.u. The peak output must be 10 volts (= maximum allowable input to the multipliers), so that a line current of 3.5 x 6.95 amps in the primary should produce only $10/\sqrt{2}$ (7.07) volts. The maximum output current (I_A) of the amplifier used is 5mA so that R cannot be less than 1.414 kohms ($V_I = I_A R = 7.07 = 5 \times 1.414k$), and the turns ratio N must be at least 4.84k ($V_I = I_L R/N = 7.07 = 6.95 \times 3.5 \times 1.414k/4.84k$). With a single loop primary, these requirements are satisfied by a six thousand turn secondary and $R = 1.75$ kohms.

The effective resistance inserted in the secondary of the current transformer is equal to R/A where A is the open loop gain of the amplifier. This is typically 40,000 at 50Hz so that the resistance is $1.75k/40k = 0.044$ ohms. The voltage at the virtual earth input to the amplifier is $I_S R/A = 0.044 \times 6.95 \times 3.5/6000 = 0.176$ mV maximum, and the maximum burden on the current transformer is 0.176 mV x $4mA = 0.7\mu$ VA. With this extremely low burden, and low secondary currents, suitable cores and winding wire for the transformer

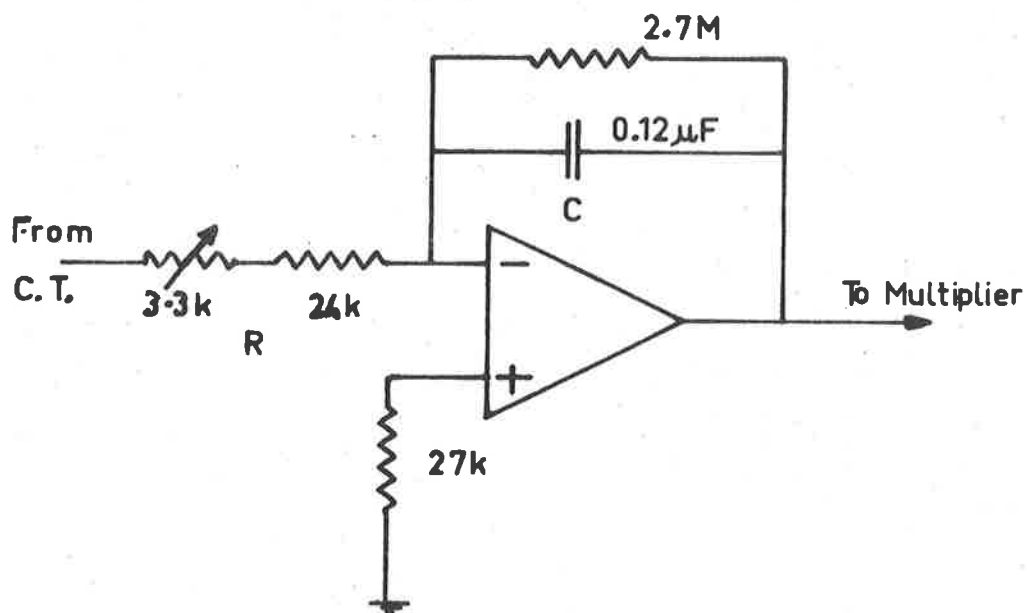


Figure 24 The integrator circuit for unity gain and $\pi/2$ phase shift at 50Hz.

are easily chosen.

The capacitor shown across the feedback resistor in figure A23 is used to remove the 3 degrees of phase shift originating from the transformer.

Under test, these transducers proved to be accurate in magnitude to within the limits of the test equipment (0.1% of full scale) and to have a phase shift between the current waveform and voltage output of less than ± 0.5 degrees over their full range.

The specifications are:

- (1) input currents 0-24 amps (3.5 p.u.)
- (2) output voltage 0-7.07 ($\frac{10}{2}$) volts RMS.

(These input and output ratings are due to the input limit of the multipliers).

- (3) magnitude accuracy $\pm 0.1\%$ of full scale
- (4) linearity over range $\pm 0.1\%$
- (5) phase shift between I_{in} and V_{out} ± 0.5 degrees over full range.
- (6) output impedance 0.01 ohms
- (7) maximum output current (at full scale) 1mA
maximum output current (at zero) 5mA

A 8.2 THE INTEGRATOR

A 90 degree phase shift of the output waveform of the current transducer is required when measuring reactive power with an analogue multiplier. This is achieved with the simple active integrator shown in figure A24. Unity gain and 90 degree phase shift is achieved when the frequency $\omega = 1/RC$. At 100π rads per second this is achieved with $R = 26.5k\Omega$ and

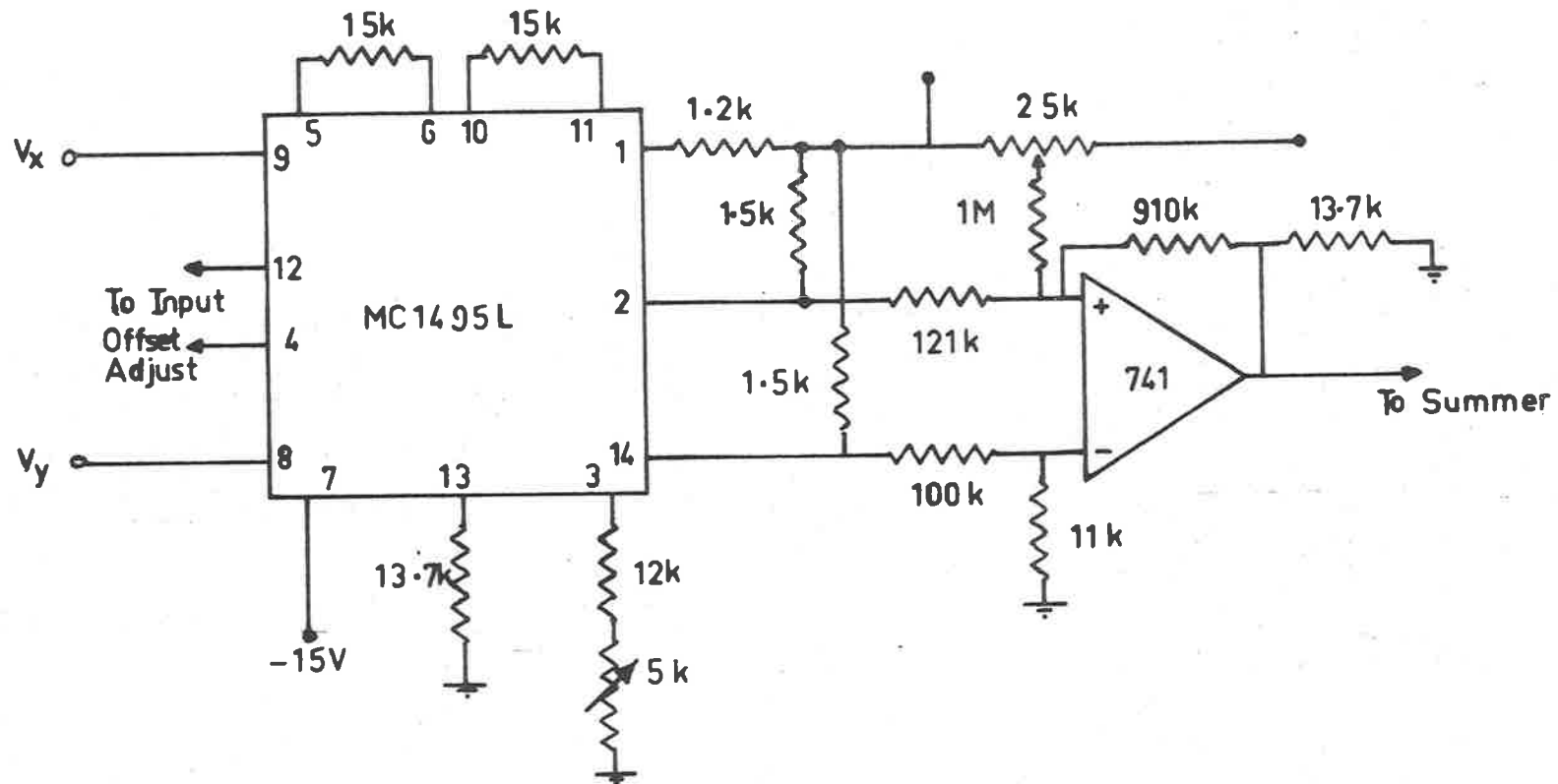


Figure A25 The four quadrant linear multiplier with an operational amplifier common mode level shift.

$C = 0.12\mu\text{F}$. The $2.7\text{M}\Omega$ feedback resistor is to prevent a D.C. offset from developing at the amplifier input, although its value is high enough to have negligible effect on the integration.

A 8.3 THE MULTIPLIERS

After consideration of performance, price and ease of application, integrated circuit voltage multipliers were chosen as the most suitable form of analogue multipliers. The type used was the MC1495L and the circuit in which these are incorporated is shown in figure A25. This forms a linear four quadrant multiplier with maximum X and Y inputs of ± 10 volts, and an overall gain factor of 0.1 i.e. 10 volts applied to each input produces 10 volts at the output.

A 8.4 THE SUMMING AMPLIFIER

The outputs of the multipliers on each phase are summed to produce a D.C. voltage proportional to the total real or reactive power transferred to the infinite bus. The required gain for the summing amplifier for real power is derived as follows. The maximum real power at which the system operates is 1 per unit. The active Butterworth filter following the summer has unity gain, and the maximum input to the A/D converters is 1 volt. Thus 1 per unit real power must be represented by 1 volt at the output of the summer. The maximum product of voltage and current (as used in deriving the multiplier input levels) is $1.25 \times 3.5 \text{ p.u.} = 4.375 \text{ p.u.}$, assuming unity power factor. This would be represented by the maximum output of the multipliers of $5 + 5 \sin \omega_s t$ (on phase A). The sum of each of the phases would

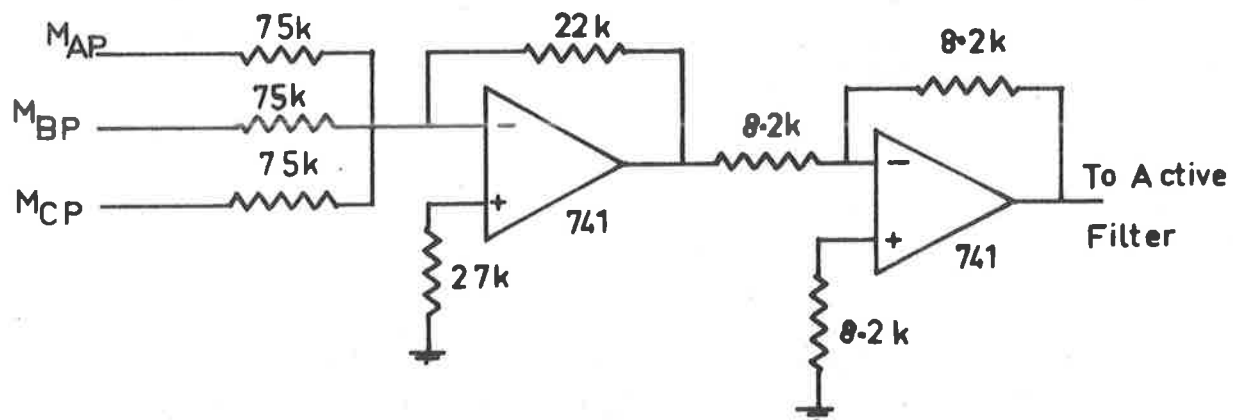


Figure A26 The summing amplifier for real power measurement

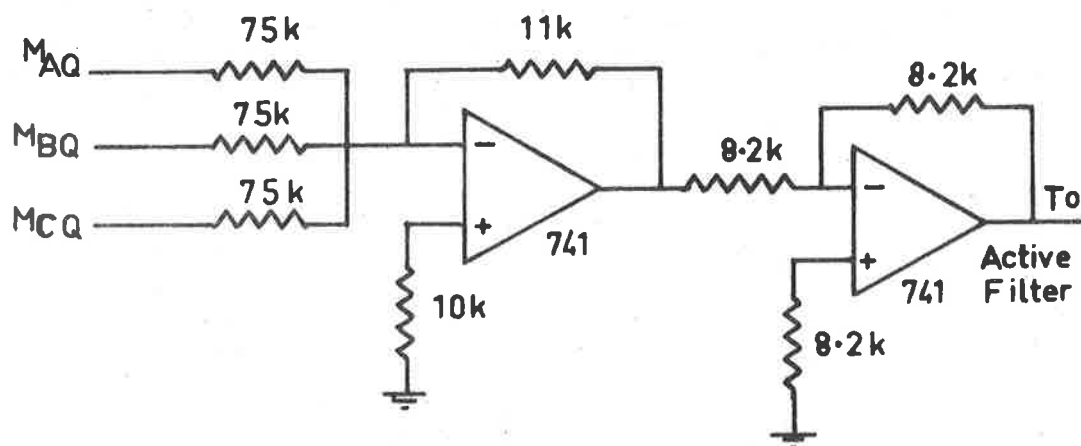


Figure A27 The summing amplifier for reactive power measurement

produce 15 volts i.e. for a unity gain summing amplifier 4.375 p.u. would be represented by 15 volts. Thus for 1 p.u. to be represented by 1 volt an amplifier gain of $4.37/15$ on each input is required. This is achieved by the summer in figure A26. This is an inverting amplifier so that the unity gain inverting buffer shown in figure A25 is required to give the output the correct sign.

The maximum leading reactive power at which the system can operate is 2.0 p.u. (This is limited by the supply current ratings). Thus the required gain of the reactive power summing amplifier is found by a similar process to be 0.1408. This amplifier is shown in figure A27.

A 8.5 THE ACTIVE BUTTERWORTH FILTERS

The same filter design as shown in figure A20 (the terminal voltage transducer) was used for each of the power transducers.

A 8.6 THE SPECIFICATIONS

Calibration tests were performed on the transducers. These showed an accuracy of $\pm 0.5\%$ of F.S.D. over their operating range. The specifications are:

- (1) Line current 3.5 p.u. max.
- (2) Bus voltage 1.25 p.u. max.
- (3) Real power ± 1.0 p.u. max.
Reactive power ± 2.0 p.u. max.
- (4) Output ± 1 volt D.C. max.
- (5) Accuracy $\pm 0.5\%$ of full scale.
- (6) Response time 15 ms.
- (7) Output impedance 0.01 ohms.

APPENDIX B

THE NOVA CONTROL PROGRAM

```
;*****OPTIMAL CONTROL***CORRECTED POWER*****
;*****DATA LOGGING*****
;INITIALLY USE SIMPLE VOLTAGE FEEDBACK TO
;POSITION THE SYSTEM WITH Q APPROXIMATELY ZERO FOR
;THE PARTICULAR VALUE OF P,BEFORE ENTERING
;THE OPTIMAL CONTROL.AS WILL BE SEEN THE OPTIMAL
;CONTROL IS APPLIED IN AN INCREMENTAL FORM
```

.LOC 376

```
DOA 1,73      ;OUTPUT A FIELD VOLTAGE TO
JMP .-1        ;ALLOW THE MACHINES TO
                ;BE RUN UP AND THE SYSTEM
                ;SYNCHRONISED.
```

```
;FOR P=0.4 A SUITABLE VALUE IN ACC1 IS 14400
;=-0.437VOLTS FROM D/A 73A,PRODUCING +18 VOLTS
;FROM THE FIELD POWER AMPLIFIER.
```

```
;
;
;                HAVE HOLD-GO IN GO
;POSITION.THEN RESTART AT 400 FOR VOLTAGE FEED
;-BACK CONTROL.
```

.LOC 400

```
START: LDA 0,ALLCON ;CAUSES ALL A/D CHANNELS
DOC 0,73 ;TO INITIATE A CONVERSION
SKPDN 71 ;WAIT UNTIL CONVERSION
JMP .-1 ;COMPLETED.
DIA 1,71 ;READ HOLD-GO
MOVL 1,1,SNC ;SKIP IF HOLD-GO NEG.(HOLD).
LDA 2,N1V ;PUT VREF=1 P.U.
DIA 0,72 ;READ VT(FROM A/D NO.72A).
SUB 2,0 ;ERROR=VT-VREF IN ACC0
MOVZL 0,0 ;MULTIPLY ERROR BY 2
                ;I.E. GAIN OF LOOP IS 4
                ;SINCE 1 P.U. ERROR=1VOLT
                ;AND 1 P.U. FIELD=0.5VOLT
                ;WHEN CONSIDERED AS
                ;ANALOGUE QUANTITIES.
DOA 0,73 ;OUTPUT THE FIELD VOLTAGE
                ;TO THE POWER AMPLIFIER.
                ;N.B. ERROR=VT-VREF
                ;SINCE POWER AMP. INVERTS
STA 0,VFIELD ;STORE THE FIELD VOLTAGE
STA 0,VFLDO ;FOR USE ON INITIAL
                ;ENTRY OF THE OPTIMAL
                ;CONTROL SECTION.
JMP START ;RETURN TO START.
```

```
;*****THE OPTIMAL CONTROLLER*****
;AFTER SYNCHRONISATION THE ABOVE SECTION OF THE
;PROGRAM IS USED TO PUT THE SYSTEM NEAR THE Q=0
;AXIS FOR THE PARTICULAR VALUE OF P USED.
;THE FOLLOWING SECTION OF THE MAIN PROGRAM
;INITIALISES THE CLOCK ,WHICH MAY INTERRUPT
;ANYWHERE WITHIN ONE PERIOD WHEN FIRST STARTED.
;TO ENTER THE OPTIMAL CONTROL RESTART AT
;LOCATION 500
```

```
.LOC 500
```

LDA 0,CLOCK	;START THE CLOCK AT
DOAS 0,RTC	;100HZ,I.E.10M.SEC PERIOD.
LDA 0,AD1INT	;SET UP THE FIRST INTERRUPT
STA 0,1	;SERVICE LOCATION.
INTEN	;ENABLE THE INTERRUPT.
JMP .+1	;WAIT FOR THE
JMP .-1	;FIRST INTERRUPT.
INT1: NIOS RTC	;RESTART THE CLOCK.
LDA 0,ZERO	;PUT DSPED=0 INITIALLY.
STA 0,DSPED	
LDA 0,SETGO	;PUT MAIN PROGRAM
STA 0,HLDGO	;HOLD-GO TO GO.
INTEN	;REENABLE THE INTERRUPT.
LDA 0,AD2INT	;MODIFY THE INTERRUPT SERVICE
STA 0,1	;LOC FOR NORMAL RUNNING.

;THE FOLLOWING SECTION ALLOWS A SMOOTH TRANSITION
 ;FROM THE INITIAL VOLTAGE FEEDBACK CONTROL
 ;TO THE OPTIMAL CONTROL.
 ;(INCREMENTAL IN NATURE).

```

      LDA 1,EIGHT      ;PUT 8 IN A TEST LOCATION
      STA 1,TEST8      ;TO ALLOW AN 8 TIMES LOOP.
      LDA 2,ZERO       ;CLEAR ACC2 TO CARRY THE
                      ;AVERAGE OF VT.
      STA 2,DVREF      ;INITIALLY DVTREF=0.
RDVT1: LDA 0,VTCON      ;CAUSES VT A/D TO
      DDC 0,73         ;CONVERT.
      SKPDN 72         ;WAIT UNTIL CONVERSION
      JMP .-1          ;COMPLETED.
      DIA 1,72         ;READ VT.
                      ;FORM AN AVERAGE OF THE 8
      MOVZR 1,1        ;VALUES OF VT READ IN.
      MOVZR 1,1        ;DIVIDE BY 16 AND ADD TO THE
      MOVZR 1,1        ;SUM OF VT(VT POSITIVE)
      MOVZR 1,1        ;TO HOLD IN THE CORRECT
                      ;FORMAT.
      ADD 1,2          ;VT AVERAGE IN ACC2.
      DSZ TEST8        ;READ 8 TIMES YET.
      JMP RDVT1        ;NO,READ AGAIN.
                      ;YES,THEN STORE THE PRESENT
      STA 2,VTO        ;VALUE OF VT AS VT BASE.
;HAVING FOUND THE PRESENT VALUE OF VT SET UP BY
;THE PREVIOUS CONTROL (WHICH INITIALLY BECOMES
;THE REFERENCE VOLTAGE),AND MADE THE ASSUMPTION
;THAT THE INITIAL PERTURBATION IN SPEED IS
;ZERO ,IT IS NECESSARY TO FIND THE INITIAL
;VALUE OF ELECTRICAL TORQUE.SINCE IN THE
;STEADY STATE SPEED IS ONE P.U.,TORQUE=POWER,
;SO THAT THE INITIAL ELECTRICAL TORQUE IS THE
;PRESENT VALUE OF REAL POWER.THE INCREMENTAL
;CONTROL IS THEN APPLIED USING THE ABOVE
;QUANTITIES AS THE INITIAL BASE VALUES.
;READ THE CORRECTED VALUE OF PWR USING THE
;SUBROUTINE.P IS PASSED WITH THE BINARY
;POINT BETWEEN BITS 15 AND 14.

```

```

      JSR [CRCTP
      MOVZR 1,1        ;STORE THE CORRECTED
      STA 1,TORQO      ;VALUE OF P AS THE
                      ;BASE TORQUE IN THE
                      ;CORRECT FORMAT.

```

;HAVING INITIATED THE CONSTANTS FOR A CONTINUOUS
 ;TRANSFER ,THE MAIN PROGRAM IS ENTERED.THE HOLD-GO
 ;DEVICE ALLOWS STEPS OF REFERENCE VOLTAGE TO BE
 ;EASILY APPLIED AND STEADY STATE OPERATING CONSTANTS
 ;TO BE ACCURATELY EVALUATED.
 ;INITIALLY GO IS SET.
 ;THE HOLD-GO FLAG IS SET VIA THE INTERRUPT ROUTINE
 ;(WHICH SETS UP INITIAL SPEED READINGS ON EACH OF
 ;THE 4 SECTORS OF THE DISC,IMEDIATELY AFTER EVERY
 ;GO FLAG).FOR GO HOLD-GO IS +VE ,-VE FOR HOLD.
 ;WHILE IN THE HOLD POSITION DVREF IS
 ;STEPPED TO THE REQUIRED VALUE.

RESTT:	LDA 0,HLDGO	;CHECK MAIN HOLD-GO.
	MOVL 0,0,SNC	;IF HOLD CONTINUE.
	JMP GO	;IF GO,JUMP OVER THE NEXT
		;SECTION,WHICH SETS FLAGS.
	LDA 1,ONE	;SET AN AFTER-HOLD FLAG.
	STA 1,AFTHG	
	JMP GO,ON	;GO ON WITH GENERAL RUNNING.
GO:	DSZ AFTHG	;AFTER HOLD FLAG SET.
	JMP NOFLG	;NO,CONTINUE GENERAL RUNNING.
		;YES THEN DO NOT SET FLAG
		;AFTER DSZ HAS RESET IT.

;AFTER A HOLD FLAG HAS BEEN SET,AND GO IS SET,THE
 ;STATES OF THE SYSTEM BECOME BASE VALUES TO WHICH
 ;SMALL SIGNAL DEVIATIONS ARE ADDED.THESE ARE SET UP IN
 ;THIS SECTION.(AFTER BEING COMPUTED IN THE INTERRUPT
 ;ROUTINE.)THE INCREMENT IN VREF IS READ IN.
 ;THIS HAS A RANGE OF +,-0.05 P.U. IN 0.01 STEPS.

DIB 0,71	;READ STEP POSITION.
LDA 1,N25	;CHECK IF POSITION IS
SUB 0,1,SNR	;FOR +0.05
LDA 2,NP.05	;YES,PUT VREF=0.05.
LDA 1,N26	;NO.TRY FOR +0.04.
SUB 0,1,SNR	;E.T.C.
LDA 2,NP.04	
LDA 1,N27	
SUB 0,1,SNR	
LDA 2,NP.03	

```

LDA 1,N28
SUB 0,1,SNR
LDA 2,NP.02
LDA 1,N29
SUB 0,1,SNR
LDA 2,NP.01
LDA 1,N210
SUB 0,1,SNR
LDA 2,ZERO
LDA 1,N211
SUB 0,1,SNR
LDA 2,NN.01
LDA 1,N212
SUB 0,1,SNR
LDA 2,NN.02
LDA 1,N213
SUB 0,1,SNR
LDA 2,NN.03
LDA 1,N214
SUB 0,1,SNR
LDA 2,NN.04
LDA 1,N215
SUB 0,1,SNR
LDA 2,NN.05
STA 2,TEMPR
LDA 1,ONE
STA 1,HGSET
NOFLG: LDA 0,ZERO
STA 0,AFTHG
GO.ON: LDA 0,ALLCON
DOC 0,73
SKPDN 72
JMP .-1
JSR [CRCTP
;DVREF SET UP.
;SET A FLAG FOR USE IN DATA
;LOGGING ROUTINE.
;CLEAR AFTER HOLD FLAG.
;ALL A/D CHANNELS
;CONVERT.
;WAIT UNTIL CONVERSION
;COMPLETED.
;READ P AND PUT IT IN A FORM
;SUITABLE TO LOOK-UP THE
;FEEDBACKS GAINS.

```

STA 1,PSTOR	;STORE P
LDA 0,NO.1	;CHECK $P > OR = 0.1$ P.U.
SUBL] 0,1,SZC	;YES,THEN JUMP OVER
LDA 1,NO.1	;NO THEN PUT $P = 0.1$ P.U.
LDA 0,NO.21	;CHECK $P < 0.21$ P.U.
SUBL] 0,1,SNC	;YES THEN JUMP OVER
JMP LRGRD	;NO THEN USE LARGER GRID.
LDA 0,QONE	; $P < 0.21$ P.U. THEREFORE
	;CHECK $Q < -1.0$ P.U.
DIC 2,72	;READ Q
ADDL 0,2,SNC	; $Q + 1$ P.U. $Q > OR = 0$ USE
JMP LRGRD	;LARGER GRID.
STA 2,QPLS1	; $Q < -1$ P.U. USE SMALLER
	;GRID
SMLGD: LDA 0,ZERO	;CLEAR ACCO,P IN ACC1.
DIC 2,71	;CHECK IF RECORDING.
MOVL 2,2,SNC	;+VE IF RECORDING THUS
JMP COMPT	;DO NOT CHANGE GAINS.IF
	;NOT RECORDING THEN
	;CONTINUE
LDA 2,N160	;MULTIPLY $P * 160$,TAKE
JSR .MPYU	;UPPER HALF = $P * 160 * 2^{15}$
	; $2^{16} - 16 = P * 80$
MOVZR 0,0,SZC	;DIVIDE $P * 80$ BY 2
INC 0,0	;AND ROUND IF NECESSARY.
STA 0,PMOD1	;I.E. $PMOD1 = 40 * P$ ROUNDED.
	;NOW MODIFY Q.
LDA 2,QPLS1	;ACC2 CONTAINS $Q + 1$ L.S.
MOVOR 2,1	;REPOSITION BY R.S.IN 1
NEG 1,1	;INVERT SO THAT $Q + 1$ POST.
LDA 0,ZERO	;CLEAR ACCO
LDA 2,N160	;MULTIPLY $(Q + 1) * 160$,TAKE
JSR .MPYU	;UPPER HALF = $(Q + 1) * 160 * 2^{14}$
	; $2^{14} * 2^{16} - 16 = (Q + 1) * 40$
MOVZR 0,0,SZC	;DIVIDE BY 2 AND
INC 0,0	;ROUND UP IF NECESSARY.
MOVZL 0,0	;ACCO CONTAINS MODIFIED Q
MOVZL 0,0	;WHICH IS LEFT S. 6
MOVZL 0,0	;TIMES TO FORM A
MOVZL 0,0	;COMPOSITE ADDRESS FOR
MOVZL 0,0	;THE TABLE LOOKUP OF
MOVZL 0,0	;STORED FEEDBACK GAINS

LDA 3,PMOD1	; WHEN ADDED
MOVZL 3,3	; TO THE 2 TIMES LEFT
MOVZL 3,3	; SHIFTED MODIFIED
ADD 0,3	; VALUE OF P.
LDA 0,N7K	; SHIFT THE ADDRESS TO
ADD 0,3	; THE HIGHER PART OF
	; STORE WHERE THE SMALL
	; GRID SIZE F.B. GAINS
INTDS	; ARE STORED.
STA 3,ADPQS	; THE ADDRESS OF ETA
STA 3,ADPQ	; FOR THIS P,Q POINT.
LDA 1,0,3	; PICK UP ETA AND STORE
STA 1,TETA	; IT IN A TEMPORY LOC.
LDA 1,1,3	; SIMILARY FOR F1
STA 1,TF1	
LDA 1,2,3	; F2
STA 1,TF2	
LDA 1,3,3	; AND F3.
STA 1,TF3	
INTEN	
JMP COMPT	; NOW COMPUTE THE CONTROL.
LRGRD: LDA 0,ZERO	; CLEAR ACCO,P IN ACC1
DIC 2,71	; CHECK: IF RECORDING
MOVL 2,2,SNC	; DO NOT CHANGE GAINS.
JMP COMPT	
LDA 2,N40	; MULTIPLY P*40, TAKE UPPER
JSR .MPYU	; HALF=P*40*2 ¹⁵ *2 ¹ -16
MOVZR 0,0,SZC	; =P*20.DIVIDE BY 2
INC 0,0	; AND ROUND IF NEEDED.
STA 0,PMOD2	; I.E. PMOD2=P*10 ROUNDED.
DIC 1,72	; NOW MODIFY Q.
LDA 2,N2	; ENSURE THAT 2 ¹⁵ DOES
ADD 2,1	; NOT OCCUR IN ACC,
	; SINCE NEG 2 ¹⁵ =2 ¹⁵ .
MOVL] 1,1,SNC	; CHECK THAT Q IS
LDA 1,QMOD0	; NOT POST.IF YES PUT
	; Q=-0.001 P.U.
NEG 1,1	; INVERT TO MAKE Q +VE.
LDA 0,ZERO	; CLEAR ACCO
LDA 2,N80	; Q*80, TAKE UPPER HALF
JSR .MPYU	; =Q*80*2 ¹⁴ *2 ¹ -16=Q*20.
MOVZR 0,0,SZC	; DIVIDE Q*20 BY 2 AND
INC 0,0	; ROUND UP IF NECESSARY.

B-8

MOVZL 0,0	;FORM THE COMPOSITE P,Q
MOVZL 0,0	;ADDRESS AS BEFORE.
MOVZL 0,0	
MOVZL 0,0	
MOVZL 0,0	
MOVZL 0,0	
LDA 3,PMOD2	
MOVZL 3,3	
MOVZL 3,3	
ADD 0,3	
LDA 0,N4K	;FOR LARGE GRID F.B.
INTDS	;GAINS ARE STORED
ADD 0,3	;FROM LOC 4000 ONWARDS.
STA 3,ADPQL	;THE ADDRESS OF ETA FOR
STA 3,ADPQ	;THIS P,Q POINT.
LDA 1,0,3	;PICK UP ETA AND STORE
STA 1,TETA	;IT IN A TEMP. LOCATION.
LDA 1,1,3	;SIMILARLY FOR F1
STA 1,TF1	
LDA 1,2,3	;F2
STA 1,TF2	
LDA 1,3,3	;AND F3.
STA 1,TF3	
INTEN	

;THE STATES OF THE SYSTEM ARE STORED IN P.U.
 ;IN A FIXED POINT FORMAT WITH THE DECIMAL POINT
 ;BETWEEN BITS 14 AND 13 OF THE REGISTER I.E.
 ;1 P.U. APPEARS AS 2^{14} IN THE REGISTERS.THE
 ;FEEDBACK GAINS F1,F3 AND THE FORWARD
 ;MULTIPLIER ETA ARE STORED IN P.U. WITH THE
 ;DECIMAL POINT BETWEEN BITS 9 AND 8 I.E.1 P.U.
 ;APPEARS AS 2^9 IN THE REGISTERS.(HOWEVER F2
 ;HAS 1 P.U. AS $2^{20}/51669$ IN THE REGISTER-DSPEED
 ;IS STORED EFFECTIVELY AS $51669=1$ P.U.(SET BY
 ;OSCILLATOR FREQUENCY)).THUS THE
 ;PRODUCT OF STATE AND FEEDBACK HAS 1 P.U. AS 2^{23}
 ;IN A DOUBLE LENGTH WORD.IN ORDER TO FORM THIS
 ;INTO A SINGLE LENGTH WORD LEFT SHIFT THE
 ;DOUBLE LENGTH WORD 4 TIMES TO GIVE 1 P.U.=
 ; 2^{27} .TAKE THE UPPER HALF OF THE WORD(EQUIVILENT
 ;TO 2^{16}),SO THAT THE SINGLE LENGTH WORD HAS
 ;1 P.U. $=2^{27} \cdot 2^{16} = 2^{11}$, I.E. THE DECIMAL
 ;POINT LIES BETWEEN 10 AND 11,GIVING A RANGE OF
 ;VALUES FOR THE PRODUCT OT +,-16 IN STEPS
 ;OF 0.0005,ONE HALF OF THE LEAST SIGNIFICANT
 ;BIT OF THE ANALOGUE CONVERTERS.

COMPT: LDA 0,ZERO	;CLEAR ACC0.
LDA 1,TETA	;PUT ETA IN ACC1.
LDA 2,DVREF	;PUT DVREF IN ACC2.
JSR [MPY	;DINPUT=ETA*DVREF
MOVZL 1,1	;LEFT SHIFT THE DOUBLE
MOVL 0,0	;LENGTH WORD 4 TIMES
MOVZL 1,1	
MOVL 0,0	
MOVZL 1,1	
MOVL 0,0	
STA 0,VREFB	;AND TAKE THE UPPER
LDA 3,DVREF	;HALF OF DINPUT.
MOVZL 3,3	;MULTIPLY DVREF BY 32
MOVZL 3,3	;SO THAT 8 VOLTS AT D/A
MOVZL 3,3	;ON OUTPUT=1 P.U.
MOVZL 3,3	
MOVZL 3,3	
DOB 3,74	;AND OUTPUT TO MONITOR.

LDA 1,ZERO	;CLEAR ACC1.
LDA 2,EIGHT	;SET UP AN 8 TIMES
STA 2,ROOT8	;LOOP TEST.
LDA 0,VTCON	;CAUSE VT A/D TO
READV: DOC 0,73	;CONVERT AND WAIT
SKPDN 72	;UNTIL CONVERSION
JMP .-1	;COMPLETED.READ VT
DIA 2,72	;WITH BINARY POINT
MOVZR 2,2	;BETWEEN BITS 14.15
MOVZR 2,2	;FORM AN AVERAGE OF
MOVZR 2,2	;THE 8 VALUES READ
MOVZR 2,2	;READ IN BY DIVIDING
ADD 2,1	;BY 16 TO PUT THE BINARY
	;POINT IN THE CORRECT
	;POSITION.
DSZ ROOT8	;READ 8 TIMES YET.
JMP READV	;NO READ AGAIN.
LDA 0,VTO	
SUB 0,1	;DVT=VT-VTO IN ACC1
STA 1,DVT	;STORE IT.
MOVZL 1,0	;MULTIPLY DVT BY 32
MOVZL 0,0	;AS FOR VREF.
MOVZL 0,0	
MOVZL 0,0	
MOVZL 0,0	
DOA 0,74	;OUTPUT DVT TO MONITOR.
LDA 0,ZERO	;CLEAR ACC0
LDA 2,TF1	;PICK UP F1
JSR [MPY	;FORM VTFB=DVT*F1
MOVZL 1,1	;LEFT SHIFT THE DOUBLE
MOVL 0,0	;LENGTH WORD 4 TIMES
MOVZL 1,1	;AND USE THE UPPER HALF.
MOVL 0,0	
MOVZL 1,1	
MOVL 0,0	
MOVZL 1,1	
MOVL 0,0	
STA 0,VTFB	;STORE VOLTAGE CONTRIB-
	;UTION TO THE CONTROL.

;NOW FIND THE ELECTRICAL TORQUE AND COMPUTE ITS
 ;CONTRIBUTION TO THE CONTROL FORCE.
 ;SINCE THE DEVIATION OF SPEED FROM
 ;SYNCHRONOUS SPEED SHOULD NOT EXCEED 0.1 PERCENT
 ;THE ELECTROMAGNETIC POWER FROM THE MACHINE
 ;CAN BE EQUATED TO THE TORQUE WITH LITTLE ERROR
 ;0.1 PERCENT BEING COMPARABLE TO THE ACCURACY
 ;OF THE READING OF POWER.
 ;1 P.U. POWER= $P \times 2^{15}$ THUS RIGHT SHIFT ONCE
 ;TO HAVE TORQUE IN THE CORRECT FORMAT
 ;I.E. 1P.U.= 2^{14} .

LDA 1,PSTOR	;PICK UP P
MOVZR 1,1	;PUT 1 P.U. $P=P \times 2^{14}$
LDA 0,TORQO	
SUB 0,1	;DTORQ=TORQ-TORQO.
STA 1,DTORQ	;STORE IT.
LDA 0,ZERO	;CLEAR ACCO
LDA 2,TF3	;PICK UP F3
JSR [MPY	;TQFB=DTQ*F3
MOVZL 1,1	;L.S. THE DOUBLE LENGTH
MOVL 0,0	;4 TIMES AND USE
MOVZL 1,1	;UPPER HALF.
MOVL 0,0	
MOVZL 1,1	
MOVL 0,0	
STA 0,TQFB	;STORE TORQUE CONTRIB
	;UTION TO THE CONTROL.

LDA 0,ZERO	;CLEAR ACCO
LDA 1,DSPED	;PICK UP THE CURRENT
	;VALUE OF DSPED,
MOVZL 1,1	;AND POSITION SUCH
MOVZL 1,1	;THAT F2*DSPED HAS
MOVZL 1,1	;1 P.U.=2'23.
LDA 2,TF2	;GET F2
JSR [MPY	;SPEED F.B.=DSPED*F2
	;RETURNED IN ACCO.
MOVZL 1,1	
MOVL 0,0	
MOVZL 1,1	
MOVL 0,0	
MOVZL 1,1	
MOVL 0,0	
MOVZL 1,1	
MOVL 0,0	
LDA 1,TQFB	;FORM THE SUM OF ALL
ADD 1,0	;THE ELEMENTS CONSTITUT-
LDA 1,VTFB	;ING THE CONTROL FORCE.
ADD 1,0	
LDA 1,VFFFB	
ADD 1,0	;ACCO CONTAINS THE
	;INCREMENTAL CONTROL.
MOVZL 0,0	;L.S. 3 TIMES TO PUT THE
MOVZL 0,0	;BINARY POINT BETWEEN
MOVZL 0,0	;BITS 13 AND 14 FOR
	;CORRECT OUTPUT FORMAT,
	;SINCE 1P.U. FIELD IS
	;STORED AS 14.13
NEG 0,0	;INVERT SINCE THE
	;POWER AMP. INVERTS.
LDA 2,VFLDO	;ADD THE INCREMENT IN
ADD 0,2	;FIELD VOLTAGE TO THE
	;CURRENT BASE FIELD.
STA 2,VFIELD	;STORE THE PRESENT TOTAL
	;FIELD VOLTAGE AS BASE
	;WHEN REQUIRED.
DOA 2,73	;OUTPUT FIELD VOLTAGE.
JMP [RSTOPT	;RETURN TO THE START
	;OF THE OPTIMAL CONTROL.

;THE FOLLWING IS THE MAIN INTERUPT SERVICE PROGRAM,
 ;WHICH IS USED TO FIND THE SPEED DEVIATION, AND
 ;PERFORM THE DATA LOGGING FUNCTION.

```

SNDIN:  NIOS RTC           ;RESTART THE CLOCK.
        INTEN             ;REENABLE THE INTERUPT.
        STA 0,SAVO        ;SAVE THE ACCUMULATORS
        STA 1,SAV1        ;AND CARRY
        STA 2,SAV2
        STA 3,SAV3
        MOVL 0,0
        STA 0,SAVC
        LDA 1,HGCON       ;CAUSE HOLD-GO A/D
        DOC 1,73          ;CHANNEL TO CONVERT.
        SKPDN 71          ;WAIT UNTIL CONVERSION
        JMP .-1           ;COMPLETED.
        DIA 0,71          ;READ HOLD-GO
        MOVL 0,0,SNC      ;IF HOLD JUMP OVER.
        JMP INTGO         ;IF GO JUMP TO NEXT PART.
INTHD:  LDA 1,N16          ;SET UP A 16 TIMES LOOP
        STA 1,TST16       ;TEST.
        LDA 1,SETHD       ;SET MAIN PROGRAM HOLD
        STA 1,HLDGO       ;WHICH WILL ALSO BE USED
                          ;TO CHECK WHETHER THE 16
                          ;LOOP IS FINISHED.
        STA 1,FN100       ;100 LOOP FLAG.
        LDA 0,N100        ;SET UP 100 TIMES
        STA 0,TS100       ;LOOP TEST.
        LDA 0,ZERO        ;CLEAR THE STORAGE
        STA 0,2           ;LOCATIONS FOR CARRYING
        STA 0,3           ;THE SUM OF THE INITIAL
        STA 0,4           ;SPEED READINGS.
        STA 0,5
        STA 0,MNSUM
        STA 0,SUMVT
        LDA 0,N65         ;SET UP A 64
        STA 0,TST65       ;TEST LOOP.
        JMP INTON         ;CONTINUE GENERAL
                          ;OPERATION.

```

;THIS SECTION IS USED TO DETERMINE INITIAL
 ;SPEED CONDITIONS.

```

INTGO:  LDA 1,HLDGO       ;CHECK IF 16 TIMES LOOP
        MOVL 1,1,SNC      ;COMPLETED.
        JMP INTON         ;THE LOOP HAS BEEN
                          ;COMPLETED THEN JUMP ON
                          ;NOT COMPLETED CONTINUE.

```

```

DSZ TST65
JMP VTRED
JMP SPDRD
VTRED: LDA 0,SUMVT      ;SET UP A VALUE FOR VTO.
      LDA 1,VTCON
      DOC 1,73
      SKPDN 72
      JMP .-1
      DIA 1,72
      MOVZR 1,1
      MOVZR 1,1
      MOVZR 1,1
      MOVZR 1,1
      MOVZR 1,1
      MOVZR 1,1
      MOVZR 1,1
      ADD 0,1
      STA 1,SUMVT
SPDRD: DIA 1,70          ;READ SPEED COUNTER
      DIB 2,70          ;AND SECTOR NUMBER.
      LDA 3,2,2         ;PICK UP THE SUM OF
      ADD 3,1           ;COUNTER,ADD
      STA 1,2,2         ;NEW READING AND RESTORE.
      LDA 3,ONE         ;CHECK IF ALL SECTORS
      LDA 0,TST0        ;HAVE BEEN READ ONCE.
      SUB 2,0,SNR
      STA 3,RD0
      LDA 0,TST1
      SUB 2,0,SNR
      STA 3,RD1
      LDA 0,TST2
      SUB 2,0,SNR
      STA 3,RD2
      LDA 0,TST3
      SUB 2,0,SNR
      STA 3,RD3
      LDA 1,ONE
      DSZ RD0
      JMP RDON
      DSZ RD1
      JMP RD1N
      DSZ RD2
      JMP RD2N
      DSZ RD3
      JMP RD3N
      JMP ALRD1        ;ALL READ ONCE.

```

RD3N:	STA 1,RD2	;SECTOR 3 NOT READ.
RD2N:	STA 1,RD1	;2 NOT READ.
RD1N:	STA 1,RD0	;1 NOT READ.
RDON:	JMP INTON	;0 NOT READ GO ON WITH
		;GENERAL RUNNING.
ALRD1:	DIC 1,70	;READ THE INITIAL
	LDA 2,MNSUM	;MAINS FREQUENCY AND
	ADD 2,1	;FORM A SUM.
	STA 1,MNSUM	
	DSZ TST16	;ALL READ 16 TIMES YET.
	JMP INTON	;NO GO ON WITH G.R.
	LDA 1,2	;YES,THEN FORM AN
	MOVZR 1,1	;AVERAGE OF THE INITIAL
	MOVZR 1,1	;SPEED READINGS AND STORE
	MOVOR 1,1	;THEM IN POSITION FOR USE.
	MOVOR 1,1	
	STA 1,6	
	LDA 1,3	
	MOVZR 1,1	
	MOVZR 1,1	
	MOVOR 1,1	
	MOVOR 1,1	
	STA 1,7	
	LDA 1,4	
	MOVZR 1,1	
	MOVZR 1,1	
	MOVOR 1,1	
	MOVOR 1,1	
	STA 1,10	
	LDA 1,5	
	MOVZR 1,1	
	MOVZR 1,1	
	MOVOR 1,1	
	MOVOR 1,1	
	STA 1,11	
	LDA 1,MNSUM	
	MOVZR 1,1	
	MOVZR 1,1	
	MOVOR 1,1	
	MOVOR 1,1	
	STA 1,MNS	

	LDA 1,PSTOR	;FIND PWR AND PUT IT
	MOVZR 1,1	;IN THE CORRECT FORMAT.
	STA 1,TMPTQ	;=BASE TORQUE IN STORE.
	LDA 1,VFIELD	;STORE PRESENT VFIELD
	STA 1,TMPVF	;IN A TEMPORARY LOCATION.
	LDA 1,SETGO	;SET HLDGO +VE TO
	STA 1,HLDGO	;SHOW LOOP COMPLETED.
INTON:	LDA 1,HLDGO	;CHECK HOLD.
	MOVL 1,1,SZC	;IF GO (+VE) SKIP.
	JMP NODEC	;HOLD-DO NOT DECREMENT.
	LDA 1,FN100	;CHECK 100 LOOP.
	MOVL 1,1,SNC	;FINISHED IF +VE.
	JMP NODEC	;JUMP ON IF FINISHED.
	DSZ TS100	;NOT FINISHED-DEC.
	JMP REFZR	;NOT ZERO-DVREF=0.
	LDA 1,SETGO	;ZERO-SET FLAG.
	STA 1,FN100	;SHOWS FINISHED.
	JMP OKRF	
REFZR:	LDA 1,ZERO	;PUT DVREF=0
	STA 1,DVREF	
	LDA 1,SUMVT	;NEW VALUE OF VTO,
	STA 1,VTO	
	LDA 1,TMPTQ	;TORQO,
	STA 1,TORQO	
	LDA 1,TMPVF	;AND VFEILD0
	STA 1,VFLDO	
	JMP NODEC	
OKRF:	LDA 1,TEMPR	;LOAD REQUIRED
	STA 1,DVREF	;VALUE OF DVREF.
NODEC:	DIA 1,70	;READ SPEED COUNTER
	DIB 2,70	;AND SECTOR NUMBER.
	LDA 0,6,2	;PICK UP INITIAL COUNT.
	SUB 1,0	;DSPEED=SPEED-SPEEDO
	STA 0,DSPED	;SPEED PERTURBATION.
	DIC 1,70	;READ MAINS COUNTER
	LDA 0,MNS	;AND FIND THE
	SUB 1,0	;DEVIATION.
	STA 0,DMNS	

;THE FOLLOWING SECTION DETERMINES WHETHER STATE DATA
 ;LOGGING IS REQUIRED FOR RESPONSE CORRELATION,AND
 ;CONDITIONAL ON THIS STORES THE DATA IN THE CORRECT
 ;LOCATIONS.THE FIRST SECTION DETERMINES THAT LOGGING
 ;IS TO COMMENCE,INITIALIZES THE STORAGE LOCATIONS
 ;AND STORES THE FIRST BITS OF DATA.

DIC 0,71	;CHECK IF RECORD WANTED.
MOVL 0,0,SZC	;IS STOREIT +VE.
JMP EXIT	;NO THEN EXIT.
DSZ HGSET	;YES.IS HGSET=1.IF=1 THEN
JMP HGNOT	;STORE INITIAL DATA.HGSET=0,
	;GO AND SEE IF SUBSEQUENT
	;RESPONSE WANTED.

LDA 1,DVTST	;INITIALIZE ALL STORAGE
LDA 2,DWST	;LOCATIONS

LDA 3,DCEST

LDA 0,DFQ

STA 0,24

STA 1,21

STA 2,22

STA 3,23

LDA 1,DVT

STA 1,[21

LDA 2,DSPEED

STA 2,[22

LDA 3,DTORQ

STA 3,[23

LDA 0,DMNS

STA 0,[24

LDA 1,ADPQ

STA 1,[ADPQH

LDA 1,PSTOR

MOVZR 1,1

STA 1,[PSTOS

LDA 1,QTEMP

STA 1,[QSTOR

LDA 1,TETA

STA 1,[TETAS

LDA 1,TF1

STA 1,[TF1S

LDA 1,TF2

STA 1,[TF2S

LDA 1,TF3

STA 1,[TF3S

LDA 1,ONE

STA 1,FSTIN

JMP EXIT

;STORE THE INITIAL
 ;VALUES OF DVT,DSPEED
 ;AND DTORQUE.

;GRID POINT FOR F.B.GAINS,

;ACTUAL PQ OPERATING
 ;POINT,
 ;(N.B. P,Q ARE STORED 14.13
 ;WITH Q INVERTED)
 ;AND THE F.B. GAINS.

;SET FSTIN FLAG=1.

;EXIT FROM RECORDING LOOP.

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HGNOT:	LDA 1,ZERO	
	STA 1,HGSET	; RESET HGSET TO ZERO
	DSZ FSTIN	; IS DATA WANTED(FSTIN=1)
	JMP EXITO	; NO, THEN EXIT
	LDA 1,21	; YES, CHECK IF BUFFER
	LDA 0,VTFUL	; STORING DVT IS FULL.
	SUB 1,0,SNR	
	JMP EXIT1	; BUFFER FULL THUS EXIT.
	LDA 1,DVT	; NO THEN STORE DVT,
	STA 1,[21	; DSPEED, AND DTORQUE.
	LDA 2,DSPED	
	STA 2,[22	
	LDA 3,DTORQ	
	STA 3,[23	
	LDA 0,DMNS	; STORE MAINS
	STA 0,[24	; DEVIATION.
	LDA 1,ONE	; SET FSTIN=1
	STA 1,FSTIN	
	JMP EXIT	; EXIT RECORDING LOOP.
EXIT1:	LDA 1,QTEMP	
	STA 1,[QST2	
	LDA 1,DVREF	
	STA 1,[DVRFS	
EXITO:	LDA 1,ZERO	; RESET FSTIN=0.
	STA 1,FSTIN	
EXIT:	LDA 0,SAVC	; RESTORE ACC'S AND CARRY.
	MOVR 0,0	
	LDA 0,SAVO	
	LDA 1,SAV1	
	LDA 2,SAV2	
	LDA 3,SAV3	
	JMP [0	; RETURN TO MAIN PROGRAM.

```

.LOC 0
0      ;CLEAR LOCATION ZERO TO SAVE
      ;P.C. DURING INTERRUPT
0      ;CLEAR LOCATION 1 TO SAVE INT.
      ;SERVICE ROUTINE ADDRESS.

.LOC 6
143305 ;INITIAL APPROXIMATIONS
144704 ;OF COUNTER READINGS
144576 ;FOR FIRST RUN UP.
144240

.LOC 42
ALLCON: 4000
CLOCK: 2. ;SET CLOCK TO 100 HZ.
AD1INT: INT1 ;ADDRES OF 1ST SERVICE ROUTINE
AD2INT: SNDIN ;ADDRES OF MAIN INTERRUPT ROUTINE
EIGHT: 8. ;USED FOR EIGHT TIMES
TEST8: 0 ;LOOPING TESTS
VTCON: 400 ;A/D 72A (VT) TO CONVERT WHEN OUTPUTED
ZERO: 0 ;USED AS ZERO FOR REGISTER CLEARING
VREF: 0 ;REFERENCE VOLTAGE
QMOD0: 177740 ;PUT Q= -0.001 WHEN Q>0 (AS READ)
HGCON: 40
N160: 160. ;160 FOR ROUNDING AND LOOK UP
AFTHG: 0
VREF0: 0
ONE: 1
RSTOPT: RESTT
VTO: 0
VFIELD: 0
VFLD0: 0
DVREF: 0
NO.1: 6340
N40: 40.
PMD01: 0
N4K: 4000
VRFFB: 0
VTFB: 0
SAV0: 0
SAV1: 0
SAV2: 0
SAV3: 0
SAVC: 0

```

TQFB: 0
DTORQ: 0
DVT: 0
ADPQL: 0
N80: 80.
PMOD2: 0
PCON: 0
TORQ0: 0
PSTOR: 0
NO.21: 15000
QONE: 40000
QPLS1: 0
N7K: 7000
ADPQS: 0
TETA: 0
TF1: 0
TF2: 0
TF3: 0
MPY: .MPY
DSPED: 0
CRCTP: PCRCT
RETRN: 0
PQCON: 3000
QMOD2: 177774
PTEMP: 0
QTEMP: 0
PSQD: 0
REST: 27172
ROOT8: 0
N2: 0
HLDGO: 1
SETGO: 1
SETHD: -1
N16: 20
TST16: 0
TST0: 0
RD0: 0
TST1: 1
RD1: 0
TST2: 2
RD2: 0
TST3: 3
RD3: 0
MNSUM: 0
DMNS: 0
MNS: 0

```

.MPYU: DOA 0,1      ;UNSIGNED MULTIPLY ROUTINE.
        DOB 1,1
        INTDS
        DOCP 2,1
        NIO 0
        JMP .+1
        DIA 0,1
        DIB 1,1
        INTEN
        JMP 0,3

```

```

DVTST: 11777      ;1ST WORD OF DVT IS IN LOC12000.
DWST: 13477      ;STORE 1500(8) VALUES 832(10).
DCEST: 15177
DFQ: 16677
QST2: 11767
DVRFS: 11770      ;STORAGE LOCATIONS FOR
ADPQH: 11771      ;REQUIRED DATA.ALL STATES
PSTOS: 11772      ;ARE STORED 14.13.
QSTOR: 11773
TETAS: 11774
TF1S: 11775
TF2S: 11776
TF3S: 11777
FSTIN: 0
VTFUL: 13477
ADPQ: 0
HGSET: 0
FN100: 0
TS100: 0
N100: 100.
TEMPR: 0

```

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N1V: 32760.
N25: 40
N26: 100
N27: 200
N28: 400
N29: 1000
N210: 2000
N211: 4000
N212: 10000
N213: 20000
N214: 40000
N215: 100000
NP.05: 819.
NP.04: 655.
NP.03: 491.
NP.02: 328.
NP.01: 164.
NN.01: -164.
NN.02: -328.
NN.03: -491.
NN.04: -655.
NN.05: -819.
SUMVT: 0
N65: 65.
TST65: 0
TMPTQ: 0
TMPVF: 0

;CORE PUNCHING ROUTINE TO OUTPUT ON PTP
 ;THE DATA LOGGED DURING A TEST SEQUENCE.
 ;THE S.A. IS .LOC. 260.AFTER COMPLETION
 ;OF ATEST RUN THE CONTROL PROGRAM IS
 ;RESTARTED AT LOC 400 AS DURING INITIAL
 ;RUN-UP.

.LOC 41

ADOUT: OUTPT

.LOC 260

IN: LDA 1,[LOC1
 JSR .BIND
 LDA 0,SP
 JSR OUTPT
 JSR OUTPT
 LDA 0,CR
 JSR OUTPT
 LDA 0,LF
 JSR OUTPT
 ISZ LOC1
 LDA 2,LOCEN
 LDA 1,LOC1
 SUBL 1,2,SNC
 JMP IN
 HALT

LOC1: 11770

LOCEN: 16677

CR: 15

LF: 12

SP: 40

OUTPT: DOAS 0,PTP

SKPDN PTP

JMP .-1

JMP 0,3

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```
; BINARY TO DECIMAL ASCII CONVERT
; CONVERTS A SINGLE PRECISION, TWOS COMPLEMENT
;     NUMBER TO AN ASCII CHARACTER STRING

; INPUT:      N IN AC1

; OUTPUT:      ASCII CHARACTER STRING, TERMINATED BY A
;              NULL WORD
;              CHARACTERS ARE RIGHT ADJUSTED IN ACO
;              PASSED TO THE ROUTINE WHOSE ADDRESS
;              MUST BE IN LOCATION 41 OF PAGE 0
;              STRING OF FORM:
;                  +DDDDD(NULL)
;              OR
;                  -DDDDD(NULL)

; CALLING SEQUENCE:
;     JSR     .BIND
;     RETURN

; DESTROYED:   AC1, AC3, CARRY
; UNCHANGED:   ACO, AC2
```

```
.BIND:  STA 3,.ED03      ; SAVE RETURN
        STA 2,.ED02      ; SAVE AC2
        STA 0,.ED00      ; SAVE ACO
        LDA 3,.ED30      ; ADDRESS OF POWER OF TEN TABLE
        STA 3,.ED10      ; INITIALIZE POINTER
        LDA 0,.ED20      ; ASSUME NEGATIVE
        MOVL] 1,1,SZC
        NEG 1,1,SKP
        LDA 0,.ED21      ; NO, IT IS POSITIVE; GET PLUS
.ED97:  STA 1,.ED11      ; SAVE N
        JSR [.ED40      ; PUT OUT SIGN OR DIGIT
        LDA 1,.ED11      ; GET CURRENT VALUE OF N
        LDA 3,[.ED10     ; GET CURRENT POWER OF TEN
        ISZ .ED10        ; BUMP POINTER
        MOV 3,0,SNR
        JMP .ED98        ; PUT OUT NULL
        LDA 0,.ED22      ; GET ASCII '0'
.ED99:  SUBZ 3,1,SZC     ; DOES POWER OF TEN GO IN
        INC 0,0,SKP      ; YES, BUMP RESULT DIGIT
        ADD 3,1,SKP      ; NO, RESTORE PREVIOUS VALUE
```

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```

        JMP .ED99          ; CONTINUE SUBTRACTING
        JMP .ED97          ; PUT OUT DIGIT

.ED98:   JSR [ .ED40        ; PUT OUT NULL WORD
        LDA 0, .ED00        ; RESTORE ACO
        LDA 2, .ED02        ; RESTORE AC2
        JMP [ .ED03        ; RETURN

.ED00:   0                  ; SAVE ACO
.ED02:   0                  ; SAVE AC2
.ED03:   0                  ; SAVE AC3

        .RDX 10
.ED05:   10000              ; POWER OF TEN TABLE
        1000               ; 10**3
        100                ; 10**2
        10                 ; 10**1
        1                  ; 10**0
        0                  ; END OF TABLE INDICATION
        .RDX 8

.ED10:   0                  ; ADDRESS OF CURRENT POWER OF
        ; TEN ENTRY
.ED11:   0                  ; RUNNING SUM WORD

.ED20:   '-'               ; ASCII '-'
.ED21:   '+'               ; ASCII '+'
.ED22:   60                ; ASCII '0'

.ED30:   .ED05              ; ADDRESS OF POWER OF TEN TABLE

.ED40=41 ; PAGE ZERO PUT CHARACTER
        ; ROUTINE ADDRESS

```

.LOC 3000

;SUBROUTINE FOR FINDING THE ELECTROMAGNETIC
 ;POWER WITHIN THE MACHINE ALLOWING FOR R,FE,
 ;LEAKAGE LOSSES ,USING AN EFFECTIVE VALUE OF R
 ;TO CORRECT THE VALUE OF P READ AT THE
 ;INFINITE BUS. $P(\text{CORRECT})=P(\text{READ})+1'2*R(\text{EFFECTIVE})$.
 ;A CONSTANT VALUE OF R IS ASSUMED.

PCRCT:	STA 3,RETRN	;SAVE THE RETURN ADDRESS.
	LDA 1,EIGHT	;SET UP AN 8 TIMES
	STA 1,ROOT8	;LOOP TEST
	LDA 2,ZERO	;CLEAR ACCS 2,3 TO CARRY
	LDA 3,ZERO	;THE AVERAGES OF P,Q.
READ:	LDA 0,PQCON	;CAUSE P,Q A/D'S
	DOC 0,73	;TO CONVERT.
	SKPDN 72	;WAIT UNTIL
	JMP .-1	;CONVERSION COMPLETED.
	DIB 1,72	;READ P WITH BINARY POINT
	MOVZR 1,1	;BETWEEN BITS 15 AND 14
	MOVZR 1,1	;FORM AN AVERAGE OF THE 8
	MOVZR 1,1	;VALUES READ IN BY DIVID
	MOVZR 1,1	;ING BY 16 TO PUT THE
	ADD 1,2	;BINARY POINT BETWEEN BITS
		;14 AND 13.
	DIC 1,72	;READ Q,CHECK IT IS NEG.
	MOVZL] 1,1,SNC	;IF NOT PUT Q=SMALL NEG.
	LDA 1,QMOD2	
	NEG 1,1	;INVERT Q TO MAKE IT +VE.
	MOVZR 1,1	;Q READ WITH 14.13 THUS
	MOVZR 1,1	;ONLY DIVIDE BY 8 TO FIND
	MOVZR 1,1	;THE AVERAGE IN THE
	ADD 1,3	;SAME FORMAT AS P.
	DSZ ROOT8	;READ 8 TIMES YET.
	JMP READ	;NO--READ AGAIN
	STA 2,PTEMP	;YES--THEN STORE THE
	STA 3,QTEMP	;AVERAGES OF PAND Q
	LDA 0,ZERO	;CLEAR ACCO.
	MOV 2,1	;P IS IN ACC'S 2 AND 1.
	JSR .MPYU	;P*2'14*P*2'14=P*2'28.
		;TAKE UPPER HALF(ACCO).
	STA 0,PSQD	;GIVES P'2*2'12

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LDA 1,QTEMP	
LDA 2,QTEMP	
LDA 0,ZERO	;SIMILARLY FIND $Q'2 \cdot 2'12$
JSR .MPYU	;IN ACC0.
LDA 1,PSQD	$P'2 + Q'2 = (V \cdot 1)'2$
ADD 0,1	;ASSUME VBUS=1 P.U. THEN
	$P'2 + Q'2 = 1'2$ IN ACC1.
LDA 2,REST	$1'2 \cdot 2'12 \cdot R \cdot 2'19$, TAKE
LDA 0,ZERO	;UPPER HALF,
JSR .MPYU	$= 1'2 \cdot R \cdot 2'15$.
LDA 1,PTEMP	;PICK UP VALUE OF P READ
MOVZL 1,1	;L.S. TO CORRECT FORMAT.
ADD 0,1	$P + 1'2 \cdot R \cdot 2'15 = P$ CORRECT
	;IN THE RIGHT FORMAT AS
	;IF READ FROM THE A/D.
LDA 3,RETRN	;RETURN TO MAIN
JMP 0,3	;PROGRAM.

```

; MULTIPLY
; MULTIPLIES TWO FIXED POINT, SINGLE PRECISION,
;     TWOS COMPLEMENT NUMBERS

; INPUT:      N1 IN AC1, N2 IN AC2

; OUTPUT:     N1*N2, HIGH ORDER IN ACO, LOW ORDER IN
;             AC1

; CALLING SEQUENCE:
;     JSR     .MPY
;     RETURN

; UNCHANGED:  AC2
; DESTROYED:  ACO, AC1, AC3, CARRY

; REQUIRES:   .MPYU (UNSIGNED MULTIPLY)

```

```

.MPY:  STA 3,.AB03      ; SAVE RETURN
        SUB 3,3          ; AC3 WILL CONTAIN CORRECTION
                        ; FACTOR
        MOVL] 1,1,SZC   ; TEST SIGN OF N1
        ADD 2,3          ; ADD N2 TO CORRECTION
        MOVL] 2,2,SZC   ; TEST SIGN OF N2
        ADD 1,3          ; ADD N1 TO CORRECTION
        STA 3,.AB10     ; SAVE CORRECTION FACTOR
        JSR [.AB30      ; CALL UNSIGNED MULTIPLY
        LDA 3,.AB10     ; GET CORRECTION
        SUB 3,0         ; TRUE RESULT IN 2S COMPLEMENT
                        ; FORM
        JMP [.AB03      ; RETURN

.AB03:  0                ; SAVE AC3

.AB10:  0                ; STORAGE FOR CORRECTION FACTOR

.AB30:  .MPYU           ; UNSIGNED MULTIPLY ADDRESS

```

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